Group-Wise V-BLAST Detection in Multiuser Space-Time Dual-Signaling Wireless Systems

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***Abstract*— This paper studies the V-BLAST detection in a general multiuser space-time wireless system, in which each user's data stream is either (orthogonal) space-time block coded (OSTBC) for transmit diversity or spatially multiplexed (SM) for high spectral efficiency. The motivation behind this work is that each user adopting a signaling scheme better matched to his own channel condition proves to improve the individual link performance but the resultant co-channel interference mitigation problem is scarcely addressed thus far. By exploiting the algebraic structure of orthogonal code, it is shown that the V-BLAST detector in the considered dual-signaling environment allows for an attractive group-wise implementation: at each iteration a group of symbols, transmitted either from an OSTBC station or from an antenna of an SM terminal, are jointly detected. The group detection property, resulting uniquely from the use of orthogonal codes, potentially improves the dual-mode signal separation efficiency, especially when the OSTBC terminals are dense in the cell. The imbedded structure of the channel matrix is also exploited for deriving a computationally efficient detector implementation. Flop count evaluations and numerical examples are used for illustrating the performance of the proposed V-BLAST based solution.**

***Index Terms*— Multiuser detection, spatial multiplexing, space-time block codes, vertical Bell Labs layered space-time (V-BLAST), multiple-input multiple-output (MIMO).**

I. Introduction

A. Motivations

MULTIUSER space-time block coded (MU-STBC) systems [13]-[15], [23] can provide multiple fading-resistant links but the co-channel user interference then becomes a crucial factor for system performance. In stead of resorting to the computationally-intensive joint maximum-likelihood (ML) decoding for signal recovery, there are various proposals exploiting the algebraic structures of the orthogonal ST block codes for facilitating interference mitigation, e.g., the Naguib's parallel interference cancellation (PIC) scheme [16], and the Stamoulis's decoupled-based method [19]. Recently it is suggested in [11] and [24] to tackle the problem alternatively via the ordered successive interference cancellation (OSIC) approach, which is also known as the V-BLAST algorithm [7]. This is because the OSIC mechanism shows a different approach as compared with [16], [19], and it is believed to maintain a reasonable trade-off between performance and complexity with respect to the joint ML decoding [6], [17]. In [24], the usage of the orthogonal codes is shown to have an attractive impact on the V-BLAST detector: it allows a user-wise group detection property. Such a V-BLAST based detection is also considered in [3], [20], [22] to resolve multi-antenna ST coded streams; the presented frameworks therein, however, are mainly for ST trellis codes, and do not explicitly exploit the codeword's algebraic structures whenever block codes are considered.

In this paper we consider a more general class of MU ST wireless systems, in which each user's data stream is either orthogonal ST block coded (OSTBC) for transmit diversity or spatially multiplexed (SM) for high spectral efficiency. Such a system configuration has been suggested for future MIMO uplink transmission: users nearby the base station could usually send high-rate data due to relatively reliable channel conditions, whereas the far-end users might sacrifice data rate for transmit diversity in order to guard against the channel impairments like large path-loss or the near-far effect [5]. As reported in [5], [10], different signaling types can also be adopted for improving individual user's link reliability against channel spatial correlations: spatial multiplexing in general fits with independently fading channels, but transmit diversity would be alternatively preferred for correlated low-rank channels. The overall MU link performance in all cases, however, crucially depends on effective interference rejection mechanisms. To the authors' best knowledge, there seems to be yet no related discussions regarding a dual-signaling system; in particular, an investigation of how the orthogonal codes can facilitate the dual-mode signal separation. In this paper, this problem will be addressed based on the V-BLAST detection approach.

B. Design Challenge and Technical Contributions

For the dual-signaling environment, there is a unique receiver design challenge to be addressed. Observe that, since some link users will send data via the OSTBC mode, the base station will have to suffer a certain time latency for data collection/detection so as to exploit the diversity benefit for those OSTBC terminals. For example, if the Alamouti's code [1] is used, two symbol periods are the temporal latent cost for realizing a diversity gain of order two. The inherent time latency produces link robustness for the OSTBC users at the expense of a reduced cell-wide data processing efficiency. This is because, during the processing time required for diversity,

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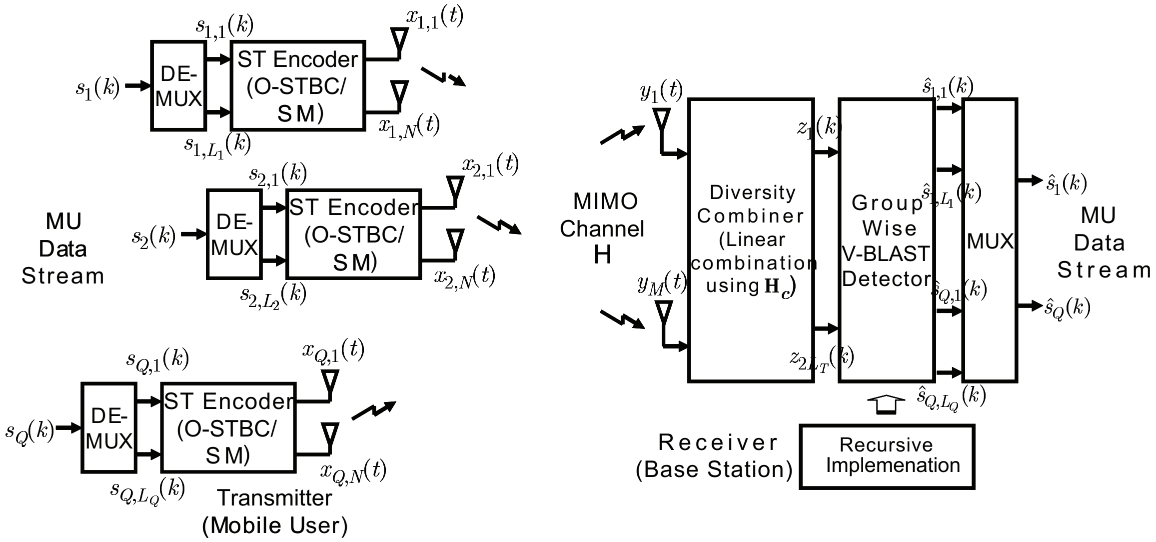

Fig. 1. The schematic diagram of the transceiver.

the base station will receive extra independent source symbols from other high-rate SM terminals: this can enlarge the overall data processing dimension up to a factor equal to the latent cost. As a result, there will be an unavoidable increase in the detector complexity; the computational overload could be significant, especially for the in general one symbol per-layer detection strategy of the V-BLAST algorithm.

This paper proposes a group-wise V-BLAST detection scheme that can effectively tackle such a design challenge. Specifically, it is shown that, even though some users may send the data via the SM mode, the usage of orthogonal codes can induce a distinctive structure on the matched-filtered (MF) channel matrix: it consists of orthogonal design [21] block submatrices. This fact is then exploited for developing a group-wise V-BLAST detector: at each processing step a group of symbols, transmitted either from a particular OSTBC station or from an antenna of an SM terminal during the latent time, can jointly be detected. This result is a generalization of the previous work [24] for MU-STBC systems to the more general dual-signaling scenario. The established group-wise detection property, which reduces the number of iterations, tends to restore the algorithm complexity back, and hence prompts an efficient receiver implementation, despite of the system model expansion for dual-mode data processing. The imbedded structure of the MF channel matrix is moreover exploited for deriving a low-complexity algorithm realization. It can further reduce computations from the following two aspects. First, it is shown that inverting the "big" channel matrix at the initial stage, which would often dominate the overall cost, reduces to solving a set of linear equations of relatively small dimensions. Second, the computation of the initial channel matrix inverse turns out to be the only "direct" inversion operations required; there is an elegant recursive formula for computing the inverse matrix needed at each iteration. Flop cost evaluations and numerical simulations are given, showing the advantages of the proposed solution over existing MU detection schemes applicable to the considered systems. The rest of this paper is organized as follows. Section II describes the system model. Section III specifies the MF channel matrix. The result is then used for developing a group V-BLAST detector in Section IV. Section V proposes a computationally efficient detector implementation and Section VI a simple two-stage processing for dual-mode signals. Section VII shows the simulation results. Finally, Section VIII is the conclusion. Most of the mathematical details required in our discussions are relegated to the appendix.

II. System Model

A. System Description and Basic Assumptions

Consider the MU ST wireless system over the flat fading channels as shown in Fig. 1, in which N transmit antennas are placed at each of the Q user terminals. The data stream of the qth user $s_q(k)$ $(1 \le q \le Q)$ can be either OSTBC [21] for transmit diversity or SM [7] for achieving high data rate. Let $\mathcal{S}_D$ and $\mathcal{S}_M$ be respectively the index sets of the OSTBC and SM users, with $Q_D \triangleq |\mathcal{S}_D|$ and $Q_M \triangleq |\mathcal{S}_M|$ denoting the respective cardinalities. Specifically, Q_D and Q_M are respectively the numbers of the OSTBC users and SM users so that $Q = Q_D + Q_M$. At each OSTBC terminal, consecutive P symbols of the data stream are spatially and temporally encoded according to [21], and are then transmitted across N antenna elements over K time periods. During the same signaling epochs each SM user then sends NK independent symbols; there are thus in total

$$L_T \triangleq PQ_D + NKQ_M \tag{1}$$

data symbols transmitted from the Q users every K symbol periods. The two ST signaling schemes in the considered system can be completely described by the associated $N \times K$ ST codeword matrices. A commonly used codeword description is the linear matrix modulation representation [14, p-97]. Let us divide the data stream of the qth user $s_q(k)$ into groups of substreams as $s_{q,l}(k) \triangleq s_q(L_q k + k - 1)$, $1 \le l \le L_q$, where the number of substreams L_q depends on the signaling mode chosen for the qth user so that $L_q = P$ if $q \in \mathcal{S}_D$ and $L_q = NK$ if $q \in \mathcal{S}_M$. Then the codeword matrix of the qth user can be written as

$$\mathbf{X}_q(k) \triangleq \sum_{l=1}^{2L_q} \mathbf{A}_{q,l}\tilde{s}_{q,l}(k), \quad (2)$$

in which $\mathbf{A}_{q,l} \in \mathcal{C}^{N\times K}$ is the ST modulation matrix, the split real-valued symbols $\tilde{s}_{q,l}(k) \triangleq \mathrm{Re}\{s_{q,l}(k)\}$ for $1 \le l \le L_q$ and $\tilde{s}_{q,l}(k) \triangleq \mathrm{Im}\{s_{q,l-L_q}(k)\}$ for $L_q+1 \le l \le 2L_q$. We note that, for $q \in \mathcal{S}_D$, $\mathbf{X}_q(k)$ is an orthogonal design [21] with[1] $\mathbf{A}_{q,l}\mathbf{A}_{q,l}^H = \mathbf{I}_N$ and $\mathbf{A}_{q,k}\mathbf{A}_{q,l}^H + \mathbf{A}_{q,l}\mathbf{A}_{q,k}^H = \mathbf{O}_N$ for $k \ne l$ [14]. For $q \in \mathcal{S}_M$, there is no imbedded coding structure in $\mathbf{X}_q(k)$, leaving each $\mathbf{A}_{q,l}$ an $N \times K$ matrix with only single nonzero entry equal to 1 or $\sqrt{-1}$. Although the OSTBC and SM schemes are quite different in nature, the linear matrix modulation representation (2) does provide a consistent description of the respective codeword matrices. Moreover, the split of the source symbols into the real and imaginary parts will also unify both the problem formulation and the underlying analysis, regardless of the constellations.

We assume that M antenna elements are located at the receiver. Let $y_m(k)$ be the received discrete-time data, sampled at the symbol-rate, from the mth receive antenna and define $\mathbf{y}(k) \triangleq [y_1(k),\ldots,y_M(k)]^T \in \mathcal{C}^M$. Collecting $\mathbf{y}(k)$ over K successive symbol periods, we have the following ST data model (assuming that the Q users are symbol synchronized[2])

$$\mathbf{Y}(k) \triangleq [\mathbf{y}(k)\cdots\mathbf{y}(k+K-1)] = \sum_{q=1}^{Q}\mathbf{H}_q\mathbf{X}_q(k) + \mathbf{V}(k), \quad (3)$$

where $\mathbf{H}_q \in \mathcal{C}^{M\times N}$ is the channel matrix from the qth user terminal to the receiver, which is assumed to be static during the K signaling periods, and $\mathbf{V}(k) \in \mathcal{C}^{M\times K}$ is the channel noise. The following assumptions are made in the sequel.

- *(a1)* The symbol streams $s_q(k)$, $1 \le q \le Q$, are i.i.d. with zero mean and variance σ_s^2.
- *(a2)* The noise $\mathbf{V}(k)$ is spatially and temporally white, each entry being with zero mean and variance σ_v^2.
- *(a3)* We assume that at least one user signals the data in the STBC mode, hence $Q_D \ge 1$.
- *(a4)* We consider the case $N \le 4$ and hence, according to [21], the symbol block length $P \in \{2,4\}$. The proposed approach is exclusively applicable to this scenario.
- *(a5)* For $3 \le N \le 4$ with complex-valued constellations, the half rate codes [21, p-1464] are used.
- *(a6)* The number of receive antennas is chosen so that $M \ge Q_D + NQ_M$.

B. Vectorized Data Model

We note that the signal part in the matrix data model (3) is a linear mixture of the Q codeword matrices. Toward a compatible MU detection framework, it is common to work with an associated equivalent vectorized linear model that will "restore" each user's symbol block. Specifically, let $\mathbf{s}_q(k) \triangleq [s_{q,1}(k),\ldots,s_{q,L_q}(k)]^T$ be the transmitted symbol block of the qth user. Without loss of generality we assume that, for each $q \in \mathcal{S}_M$, the NK symbols $s_{q,l}(k)$'s are renumbered so that the nth group of K symbols, namely, $s_{q,l}(k)$ for $(n-1)K+1 \le l \le nK$, are precisely those sent via the nth transmit antenna ($1 \le n \le N$). Define $\tilde{\mathbf{s}}_q(k) \triangleq [\mathrm{Re}\{\mathbf{s}_q^T(k)\}\ \mathrm{Im}\{\mathbf{s}_q^T(k)\}]^T \in \mathcal{R}^{2L_q}$ and $\tilde{\mathbf{y}}(k) \triangleq [\mathrm{Re}\{\mathbf{y}^T(k)\}\ \mathrm{Im}\{\mathbf{y}^T(k)\}]^T \in \mathcal{R}^{2M}$ to be the split real-valued symbol block of the qth user and the received vector. Associated with the qth user's channel matrix $\mathbf{H}_q$, we form the following augmented matrix

$$\tilde{\mathbf{H}}_q \triangleq \mathbf{I}_K \otimes \begin{bmatrix} \mathrm{Re}\{\mathbf{H}_q\} & -\mathrm{Im}\{\mathbf{H}_q\} \\ \mathrm{Im}\{\mathbf{H}_q\} & \mathrm{Re}\{\mathbf{H}_q\} \end{bmatrix} \in \mathcal{R}^{2KM\times 2KN}, \quad (4)$$

where the notation $\otimes$ stands for the Kronecker product; also, with $\mathbf{a}_{q,l}^{(j)}$ denoting the jth column of the matrix $\mathbf{A}_{q,l}$ and

$$\tilde{\mathbf{a}}_{q,l}^{(j)} \triangleq \left[\mathrm{Re}\left\{\mathbf{a}_{q,l}^{(j)}\right\}^T\ \mathrm{Im}\left\{\mathbf{a}_{q,l}^{(j)}\right\}^T\right] \in \mathcal{R}^{2N}, \quad (5)$$

we define the $2KN \times 2L_q$ real-valued ST modulation matrix

$$\tilde{\mathbf{A}}_q \triangleq \begin{bmatrix} \tilde{\mathbf{a}}_{q,1}^{(1)} & \cdots & \tilde{\mathbf{a}}_{q,2L_q}^{(1)} \\ \vdots & \ddots & \vdots \\ \tilde{\mathbf{a}}_{q,1}^{(K)} & \cdots & \tilde{\mathbf{a}}_{q,2L_q}^{(K)} \end{bmatrix} \in \mathcal{R}^{2KN\times 2L_q}. \quad (6)$$

Then the matrix data model (3) can be rewritten, after some manipulations, as the following equivalent vectorized linear model

$$\mathbf{y}_c(k) \triangleq \left[\tilde{\mathbf{y}}^T(k)\cdots\tilde{\mathbf{y}}^T(k+K-1)\right]^T = \mathbf{H}_c\mathbf{s}_c(k) + \mathbf{v}_c(k), \quad (7)$$

where $\mathbf{H}_c \triangleq \left[\tilde{\mathbf{H}}_1\tilde{\mathbf{A}}_1\cdots\tilde{\mathbf{H}}_Q\tilde{\mathbf{A}}_Q\right] \in \mathcal{R}^{2KM\times 2L_T}$ is the concatenated equivalent channel matrix, $\mathbf{s}_c(k) \triangleq \left[\tilde{\mathbf{s}}_1^T(k)\cdots\tilde{\mathbf{s}}_Q^T(k)\right]^T \in \mathcal{R}^{2L_T}$ and $\mathbf{v}_c(k) \in \mathcal{R}^{2KM}$ is the corresponding noise component. Through linearly combining the received data $\mathbf{y}_c(k)$ with $\mathbf{H}_c$, we can obtain the MF data vector

$$\mathbf{z}(k) \triangleq \mathbf{H}_c^T\mathbf{y}_c(k) = \mathbf{F}\mathbf{s}_c(k) + \bar{\mathbf{v}}(k), \quad (8)$$

where $\mathbf{F} \triangleq \mathbf{H}_c^T\mathbf{H}_c \in \mathcal{R}^{2L_T\times 2L_T}$ is the MF channel matrix and given by

$$\mathbf{F} = \begin{bmatrix} \tilde{\mathbf{A}}_1^T\tilde{\mathbf{H}}_1^T\tilde{\mathbf{H}}_1\tilde{\mathbf{A}}_1 & \cdots & \tilde{\mathbf{A}}_1^T\tilde{\mathbf{H}}_1^T\tilde{\mathbf{H}}_Q\tilde{\mathbf{A}}_Q \\ \vdots & \ddots & \vdots \\ \tilde{\mathbf{A}}_Q^T\tilde{\mathbf{H}}_Q^T\tilde{\mathbf{H}}_1\tilde{\mathbf{A}}_1 & \cdots & \tilde{\mathbf{A}}_Q^T\tilde{\mathbf{H}}_Q^T\tilde{\mathbf{H}}_Q\tilde{\mathbf{A}}_Q \end{bmatrix}, \quad (9)$$

and $\bar{\mathbf{v}}(k) \triangleq \mathbf{H}_c^T\mathbf{v}_c(k)$. We will hereafter rely on the MF model (8) for detection. To better manifest the core ideas, throughout the context we will focus on the real-valued constellation case, and hence $K = P$ so that $L_T \triangleq PQ_D + PNQ_M$. There are essentially the same results (see Appendices II and IV) for our reports whenever complex-valued constellations are used.

III. Matched-Filtered Channel Matrix

For the particular MU-STBC system, it is shown in [24] that all the $P \times P$ block submatrices of MF channel matrix $\mathbf{F}$ are orthogonal designs [21]. For the considered dual-signaling platform, in which the SM signaling could induce severe

[1] The notations $(\cdot)^T$, $(\cdot)^H$, $\mathbf{I}_m$, and $\mathbf{O}_m$ respectively denote the transpose, the complex conjugate transpose, the $m\times m$ identity matrix, and the $m\times m$ zero matrix.

[2] Symbol synchronization is necessary in TDMA based cellular implementations, e.g., IS-136 and GSM, and 3G TDD CDMA systems such as time division synchronous CDMA (TD-SCDMA). In the literature, this assumption is commonly made when dealing with uplink interference cancellation in MU-STBC systems [16], [18], [19].

coupling effects against the OSTBC signals, the structure of the resultant $\mathbf{F}$ could largely deviate from that of the MU-STBC system. As we will see in what follows, the matrix $\mathbf{F}$ in the dual-signaling environment, however, does preserve the appealing block orthogonal structure. This fact is primarily the guts for developing a group-wise V-BLAST detector.

To characterize $\mathbf{F}$, we should note that its diagonal submatrices (of appropriate dimensions) are basically the effective MF signal components of the Q users' streams, whereas the off diagonal submatrices account for the inter-user signal interference. In view of this observation, one can thus classify the block submatrices of $\mathbf{F}$ based on these signal and interference "signatures". As such, the linear matrix modulation representation of the codeword matrices (2) will allow a systematic way of computing these signature matrices and, based on which, the structure of $\mathbf{F}$ can be readily determined. Since each user sends data via either the SM or the OSTBC mode, there are essentially *two* types of signal signatures, one associated with a signaling scheme. Also, among all the interference signatures there are only *three* distinct canonical building blocks needed to be identified: two of which reflect the interference between each pair of distinct users adopting the same signaling strategy; the other is thus for the different-signaling interference. To further specify these signatures, we shall first determine the respective dimensions. Recall that during consecutive K time slots ($K = P$ for unity code rate), the numbers of symbols sent from an OSTBC and an SM terminal are, respectively, P and NP. As a result, if we denote $\mathbf{F}_{p,q}$ the submatrix of $\mathbf{F}$ representing the interference signature between the pth and the qth users' streams, we then have $\mathbf{F}_{p,q} \in \mathcal{R}^{P\times P}$ for $p,q \in \mathcal{S}_D$, $\mathbf{F}_{p,q} \in \mathcal{R}^{NP\times NP}$ for $p,q \in \mathcal{S}_M$, and $\mathbf{F}_{p,q} \in \mathcal{R}^{P\times NP}$ for $p \in \mathcal{S}_D$ and $q \in \mathcal{S}_M$. All such three $\mathbf{F}_{p,q}$'s, together with the signal signature $\mathbf{F}_{q,q}$ for either $q \in \mathcal{S}_D$ or $q \in \mathcal{S}_M$, are described in the next proposition, whose proof is given in Appendix I.

In the sequel, we denote by $\mathcal{O}(P)$ the set of all $P \times P$ real orthogonal designs with constant diagonal entries as specified in [21, p-1458].

Proposition 3.1: Let $\mathbf{F}_{p,q}$ be the submatrix of $\mathbf{F}$ describing the mutual coupling between the pth and the qth users. Then we have the following results.

(1) If $p,q \in \mathcal{S}_D$, then $\mathbf{F}_{p,q} \in \mathcal{O}(P)$. In particular, we have $\mathbf{F}_{q,q} = \alpha_q \mathbf{I}_P$ for some scalar α_q.
(2) If $p,q \in \mathcal{S}_M$, then each $P \times P$ submatrix of $\mathbf{F}_{p,q} \in \mathcal{R}^{NP\times NP}$ is a scalar multiple of $\mathbf{I}_P$.
(3) If $p \in \mathcal{S}_D$ and $q \in \mathcal{S}_M$, then each $P \times P$ submatrix of $\mathbf{F}_{p,q} \in \mathcal{R}^{P\times NP}$ belongs to $\mathcal{O}(P)$.

Part *(1)* of Prop. 3.1 is known from [21] and is exploited in [24] for developing a user-wise group V-BLAST detector for the MU-STBC systems. The significance of Prop. 3.1 lies in *(2)* and *(3)*, which are more relevant to a dual-signaling scenario. For $p,q \in \mathcal{S}_M$, it is easy to see that the (i,j)th $P \times P$ submatrix of $\mathbf{F}_{p,q}$, $1 \le i,j \le N$, characterizes the coupling effect between the data streams transmitted from the ith antenna of the pth user and from the jth antenna of the qth user; for $p = q$ and $i = j$, this is precisely the single-antenna SM signal signature. Since spatial multiplexing does not impose any spatial and temporal coding structure among the transmitted data, the interference between any pair of SM streams sent from two different antennas, and the respective signal components, would appear to be spatially and temporally decoupled. Hence, as evidenced by *(2)*, the resultant signatures are diagonal matrices; the diagonal entries are all equal since the propagation channels are assumed to be static during K signaling instants. The independently-distributed SM streams, on the other hand, might interfere against the OSTBC signals to induce a coupling matrix that is no longer orthogonal. Part *(3)* of Prop. 3.1 shows that the resultant coupling signature is nonetheless block-wise an orthogonal design. To interpret this result, we first note that a single-antenna SM stream may only temporally interfere with an OSTBC codeword. The temporally-decoupled nature of SM data is likely to render the temporal correlation of OSTBC streams unchanged, thus resulting in an orthogonal type signature. The above discussions show that the symbol stream from a single SM antenna will introduce a set of diagonal blocks (the coupling among the same SM type streams) and a set of orthogonal design blocks (the coupling against the OSTBC signals) in the MF channel matrix $\mathbf{F}$. This result relies solely on the multiplexing nature among all the NQ_M SM signals, *irrespective of the number of transmit antennas placed on each SM terminal*. In light of this observation, the asserted structure of the matrix $\mathbf{F}$ will be preserved particularly when single-antenna cell users are present, as long as each one transmits independent symbols along his own antenna.

Since a matrix being a scalar multiple of $\mathbf{I}_P$ is essentially a $P \times P$ orthogonal design, Prop. 3.1 asserts that in the dual-signaling case, the MF channel matrix $\mathbf{F}$ does consist of orthogonal design based block submatrices. The presence of SM terminals thus substantially preserves the block orthogonal structure of $\mathbf{F}$ as seen in the MU-STBC case [24]. In the next section this fact will be exploited for developing a group-wise V-BLAST detector for the considered MU dual-signaling system.

Remarks:

(a) As we will see in the next section, the proposed group-wise detection property benefits uniquely from the distinctive structures of $\mathbf{F}$ specified as in Prop. 3.1. This attractive property thus also holds whenever single-input terminals exist (since the inherent structure of $\mathbf{F}$ is preserved).
(b) For the complex constellations, there are analogue results as in Prop. 3.1, with possible modifications of the matrix dimensions; these are included in Appendices II and IV.

IV. Group-Wise V-BLAST Detection

For an MU-STBC system, it has been shown in [24] that the V-BLAST detector can per iteration jointly detect a group of P symbols associated with a particular user. By exploiting the distinctive structure of the matrix $\mathbf{F}$ shown in Prop. 3.1, this section proposes a group-wise V-BLAST detection scheme for the considered dual-signaling system. As we will see, the V-BLAST detector can per iteration jointly detect a group of P symbols, transmitted either from an OSTBC terminal or from a single antenna of an SM station. This implies that only $Q_D + NQ_M$ processing layers are required for separating the $P(Q_D + NQ_M)$ transmitted dual-mode symbol streams. As a result, in a dual-signaling case, the "user-wise" group detection property of the OSTBC signals is preserved, and the "antenna-wise" detection for SM streams comes out as a nice byproduct. The proposed group detection capability can prevent significant computational overload due to the data model expansion for processing the dual-mode signals.

A. V-BLAST Ordering

The V-BLAST algorithm resorts to certain ordering strategy, depending on the criterion for signal recovery, for deciding the best reliable symbol to be detected in each processing stage [7]. Based on Prop. 3.1, in what follows we will show that the V-BLAST ordering in each processing layer sorts the symbol streams to be detected into a group-wise basis, which in turn results in a group-wise detector realization.

1) Zero-Forcing (ZF) Criterion: In the ZF V-BLAST detection, the optimal detection order at each step is determined by the index of the symbol decision statistics yielding maximal post-detection SNR [7]. Based on the MF signal model (8), the ZF decision vector in the initial stage is

$$\mathbf{s}_d(k) \triangleq \mathbf{F}^{-1}\mathbf{z}(k) = \mathbf{s}_c(k) + \mathbf{F}^{-1}\bar{\mathbf{v}}(k). \tag{10}$$

Equation (10) shows that, for $1 \leq l \leq L_T$, the lth symbol decision statistics, that is, the lth component of $\mathbf{F}^{-1}\mathbf{z}(k)$, is simply the desired symbol contaminated by the additive noise $\mathbf{e}_l^T\mathbf{F}^{-1}\bar{\mathbf{v}}(k)$, where $\mathbf{e}_l$ is the lth unit standard vector in $\mathcal{R}^{L_T}$. It is straightforward to compute the noise power as

$$\sigma_l^2 \triangleq E\left\{\left|\mathbf{e}_l^T\mathbf{F}^{-1}\bar{\mathbf{v}}(k)\right|^2\right\} = \frac{\sigma_v^2}{2}\mathbf{e}_l^T\mathbf{F}^{-1}\mathbf{e}_l. \tag{11}$$

Since all the transmitted symbols are of equal variance, (10) and (11) imply that the (average) SNR in the lth decision statistics is completely determined by the lth diagonal entry of the noise covariance $\mathbf{F}^{-1}$. In particular, a small $\left[\mathbf{F}^{-1}\right]_{l,l}$ (the lth diagonal entry of $\mathbf{F}^{-1}$) implies a large SNR, and hence better detection accuracy attained by the lth branch. As a result, the optimal detection order at the initial stage, which specifies the symbol with slightest noise corruption, is obtained by searching for the index $1 \leq l \leq L_T$ at which $\left[\mathbf{F}^{-1}\right]_{l,l}$ is minimal. As a result, the optimal index can be found as long as one can explicitly know the diagonal entries of the inverse matrix $\mathbf{F}^{-1}$. In the next proposition we will see that the matrix $\mathbf{F}^{-1}$ "inherits" the key features of $\mathbf{F}$ as established in Prop. 3.1. This result directly alleviates the efforts to search for the optimal index, and allows for a group-wise detection strategy. Also, this will lead to a very efficient procedure for computing the weighting matrices for recovering the symbol groups.

We need the following notation. For a fixed symbol block length P and a positive integer L, let us define $\mathcal{F}(L)$ to be the set of all invertible real symmetric $PL \times PL$ matrices such that, for $\mathbf{X} \in \mathcal{F}(L)$, with $\mathbf{X}_{i,j}$ being the (i,j)th $P \times P$ block submatrix, we have $\mathbf{X}_{i,i} = \gamma_i\mathbf{I}_P$ for some scalar γ_i, and $\mathbf{X}_{i,j} \in \mathcal{O}(P)$, for $i \neq j$, the set of $P \times P$ real orthogonal design with equal diagonal entries.

Proposition 4.1: Let $\mathbf{F}$ be the MF channel matrix as defined in (8). Then the inverse matrix of $\mathbf{F}$ belongs to the set $\mathcal{F}(L)$, where $L \triangleq Q_D + NQ_M$.

Proof: Prop. 3.1 shows that each $P \times P$ block off-diagonal submatrix of $\mathbf{F}$ belongs to $\mathcal{O}(P)$, and hence $\mathbf{F} \in \mathcal{F}(Q_D + NQ_M)$. The following lemma (see Appendix III for outlined proof) characterizes the inverse of the set of matrices in $\mathcal{F}(L)$ and will be used for proving Prop. 4.1.

Lemma 4.1: If $\mathbf{X} \in \mathcal{F}(L)$, then so is $\mathbf{X}^{-1}$.

Since $\mathbf{F} \in \mathcal{F}(Q_D + NQ_M)$, the result then follows immediately from Lemma 4.1. □

Proposition 4.1 asserts that the PL diagonal entries of $\mathbf{F}^{-1}$ assume L distinct levels only. It suffices to search among the L values, one associated with a symbol group transmitted from either an OSTBC user or from a single antenna of an SM terminal, for the optimal detection order. At the initial stage, the ZF V-BLAST ordering thus sorts the symbol to be detected into a group-wise basis; all the P symbols within one group are of equal detection priority. The V-BLAST detector can then jointly detects a group of P symbols, either of a particular OSTBC stream or of a single-antenna SM stream. The optimal group detection index is

$$\bar{l} = \arg\min_{l} \beta_l, \tag{12}$$

where β_l is the $((l-1)P+1)$th diagonal entry of $\mathbf{F}^{-1}$; the ZF weighting matrices are determined from the corresponding indexed columns of $\mathbf{F}^{-1}$.

The detected user's signal is cancelled from the received data (7), yielding a reduced-size data model containing the yet-to-be-detected signals. With such a detect-and-cancel procedure followed by an associated linear combining of the resultant signal as in (8), it can be directly verified that, at the ith iteration, where $1 \leq i \leq L-1$, the noise covariance matrix associated with ZF decision statistics vector is $\mathbf{F}_i^{-1} \triangleq \left(\mathbf{H}_{c,i}^T\mathbf{H}_{c,i}\right)^{-1} \in \mathcal{R}^{(L_T-iP)\times(L_T-iP)}$, where $\mathbf{H}_{c,i}$ is obtained by deleting i block(s) of P columns (corresponding to the previously detected signals) from $\mathbf{H}_c$. Since $\mathbf{F}_i = \mathbf{H}_{c,i}^T\mathbf{H}_{c,i} \in \mathcal{R}^{(L_T-iP)\times(L_T-iP)}$ is simply obtained by deleting the associated i block(s) of P columns and rows from $\mathbf{F}$, the matrix $\mathbf{F}_i$ exhibits the same algebraic structure as $\mathbf{F}$. More precisely, we have $\mathbf{F}_i \in \mathcal{F}(L-i)$, and so is $\mathbf{F}_i^{-1}$ by Lemma 4.1. This shows that the $P(L-i)$ diagonal entries of $\mathbf{F}_i^{-1}$ take on only $(L-i)$ different levels and, by following the previous analysis, group-wise detection can thus be done at each processing step.

2) Minimum Mean Square Error (MMSE) Criterion: The MMSE V-BLAST ordering at each layer picks up the symbol attaining the minimal mean square error for detection. At the initial stage, the joint MMSE weight that minimizes the metric

$$\varepsilon \triangleq E\left\{\left\|\mathbf{s}_c(k) - \mathbf{W}^T\mathbf{z}(k)\right\|_2^2\right\}, \tag{13}$$

is obtained as

$$\mathbf{W} = \left[\mathbf{F} + \frac{\sigma_v^2}{2}\mathbf{I}_{L_T}\right]^{-1}. \tag{14}$$

It is easy to compute the lth symbol mean square error, i.e., $E\left\{\left|\mathbf{e}_l^T\left[\mathbf{s}_c(k) - \mathbf{W}^T\mathbf{z}(k)\right]\right|^2\right\}$, as

$$\varepsilon = \mathbf{e}_l^T\left[\left(\frac{2}{\sigma_v^2}\right)\mathbf{F} + \mathbf{I}_{L_T}\right]^{-1}\mathbf{e}_l. \tag{15}$$

TABLE I
ALGORITHM SUMMARY OF THE PROPOSED GROUP-WISE V-BLAST DETECTOR.

Initialization:
$\mathbf{H}_{c,0}=\mathbf{H}_c$; $\mathbf{F}_0=\mathbf{F}=\mathbf{H}_c^T\mathbf{H}_c$;
$\mathbf{y}_{c,0}(k)=\mathbf{y}_c(k)$; $\mathbf{z}_0(k)=\mathbf{z}(k)=\mathbf{H}_c^T\mathbf{y}_c(k)$;
Recursion: For $0\le i\le Q_D+NQ_M-1$

1. $\mathbf{Q}_i=\begin{cases}\mathbf{F}_i^{-1} & \text{for ZF Criterion}\\ \left[\left(2/\sigma_v^2\right)\mathbf{F}_i+\mathbf{I}_{(L_T-iP)}\right]^{-1} & \text{for MMSE Citerion}\end{cases}$
2. $\bar{l}_i=\arg\min_{1\le l_i\le Q_D+NQ_M-i}\beta_{l_i}$, where β_{l_i} is the $((l_i-1)P+1)$th diagonal entry of $\mathbf{Q}_i$
3. $\mathbf{G}_i=\begin{cases}\mathbf{F}_i^{-1} & \text{for ZF Criterion}\\ \left[\mathbf{F}_i+\left(\sigma_v^2/2\right)\mathbf{I}_{(L_T-iP)}\right]^{-1} & \text{for MMSE Citerion}\end{cases}$
4. $\mathbf{W}_i=\mathbf{G}_i\left[\mathbf{e}_{(\bar{l}_i-1)P+1}\cdots\mathbf{e}_{(\bar{l}_i-1)P+P}\right]$
5. $\hat{\mathbf{s}}_{c,\bar{l}_i}(k)=\mathcal{Q}\left(\mathbf{W}_i^T\mathbf{z}_i(k)\right)$
6. $\mathbf{y}_{c,i+1}(k)=\mathbf{y}_{c,i}(k)-\bar{\mathbf{H}}_{c,\bar{l}_i}\hat{\mathbf{s}}_{c,\bar{l}_i}(k)$
7. $\bar{\mathbf{H}}_{c,\bar{l}_i}$ is obtained from $\mathbf{H}_c$ corresponding to $\mathbf{s}_{c,\bar{l}_i}(k)$
8. $\mathbf{z}_{i+1}(k)=\mathbf{H}_{c,i+1}^T\mathbf{y}_{c,i+1}(k)$
9. $\mathbf{F}_{i+1}=\mathbf{H}_{c,i+1}^T\mathbf{H}_{c,i+1}$
10. $\mathbf{H}_{c,i}$ is obtained by deleting i block(s) of P columns from $\mathbf{H}_c$

Equation (15) shows that the optimal MMSE V-BLAST ordering is determined by the index of the minimal diagonal entry of the matrix $\left[\left(2/\sigma_v^2\right)\mathbf{F}+\mathbf{I}_{L_T}\right]^{-1}$. Since $\mathbf{F}\in\mathcal{F}(L)$ and adding diagonal perturbation $\mathbf{I}_{L_T}$ to a scaled $\mathbf{F}$ essentially preserves the block orthogonal structure, it follows immediately that $\left[\left(2/\sigma_v^2\right)\mathbf{F}+\mathbf{I}_{L_T}\right]\in\mathcal{F}(L)$ and so is $\left[\left(2/\sigma_v^2\right)\mathbf{F}+\mathbf{I}_{L_T}\right]^{-1}$ by Lemma 4.1. The PL symbol mean square errors (diagonal entries of $\left[\left(2/\sigma_v^2\right)\mathbf{F}+\mathbf{I}_{L_T}\right]^{-1}$) can thus be categorized into L groups of constant elements, one associated with a symbol group transmitted from either an OSTBC user or from a single antenna of an SM terminal (as in the ZF case). This thus warrants the group-wise MMSE detection in the initial layer. Through the group detect-and-cancel process, it can be checked that, at the ith iteration ($1\le i\le L-1$), the symbol mean square errors are computed as the diagonal entries of the matrix $\left[\left(2/\sigma_v^2\right)\mathbf{F}_i+\mathbf{I}_{P(L-i)}\right]^{-1}$, where $\mathbf{F}_i=\mathbf{H}_{c,i}^T\mathbf{H}_{c,i}$. Since $\left[\left(2/\sigma_v^2\right)\mathbf{F}_i+\mathbf{I}_{P(L-i)}\right]^{-1}\in\mathcal{F}(L-i)$, so is the inverse matrix (again by Lemma 4.1). This shows that mean square errors in the ith layer are sorted into the same group-wise manner, hence allowing for a group-wise MMSE detection.

B. Algorithm Outline and Related Discussions

The proposed group-wise V-BLAST algorithm is outlined in Table I (in Table I, $\bar{\mathbf{H}}_{c,\bar{l}_i}$ is the channel matrix obtained from $\mathbf{H}_c$ corresponding to the $\bar{l}$th group of data stream in the ith iteration, and $\mathcal{Q}(\cdot)$ is the slicing operation associated with the adopted symbol constellation). Several discussions regarding the proposed method are given as follows.

(a) The group-wise detection property benefits uniquely from the use of the orthogonal codes. When non-orthogonal codes are used, the PL diagonal entries of $\mathbf{F}^{-1}$ in general will take on PL different levels: one has to resort to a symbol-wise based algorithm realization. Therefore, PL stages are needed for separating the PL dual-mode signals; the resultant algorithm complexity, however, could be large.

(b) It is noted that, even if orthogonal codes are used, the appealing group-wise detection property does not hold if the number of the transmit antennas of each user is greater than four (for $N>4$, the matrix $\mathbf{F}$ is observed to lose the special structure as specified in Prop. 3.1). The assumption $N\le 4$ is thus necessary regarding the feasibility of the group-wise detection. This requirement can usually be met in practice since, to reduce the implementation cost and physical size, it is undesirable to place too many antenna elements on the user handsets.

(c) In principle, the ZF metric aims for complete nulling of the interference from other users but is subject to possible noise enhancement, whereas the MMSE criterion focuses on joint suppression of interference and noise. Although the two design metrics resort to different strategies for symbol recovery, in the considered scenario they both lead to the appealing group-wise detection property. This is because the respective optimal detection orders in each processing layer, as shown in the above discussions, are determined by the diagonal entries of certain structured matrices belonging to the family $\mathcal{F}(L)$ (this is also a unique feature pertaining to OSTBC).

(d) Another dual-signaling environment is seen in the mono-link systems recently considered in [9], in which a subset of antenna elements are selected for transmitting the Alamouti coded streams, while the others are left for spatial multiplexing. It is noted that, in this single-user scenario, the (antenna-wise) symbol synchronization assumption is typically satisfied, and the proposed group-wise V-BLAST detector can be used for separating the multi-antenna symbol streams.

C. Conservation of Spatial Resource: CDMA Based Implementation

In the current framework, the assumption on the number of receive antennas $M\ge Q_D+NQ_M$ is needed since only the spatial resource is employed for removing the multi-access interference (MAI). This requirement seems critical in practice because M could be prohibitively large in a user-dense environment. A typical approach to conserving the spatial resource is to alternatively adopt the multi-access techniques, e.g., CDMA, for MAI mitigation [13]. Basically, the MAI can be effectively suppressed through chip despreading, as long as all the cell user are assigned with distinct spreading (signature) codes; in this case, only $M \geq N$ receive antennas are required in order to separate the symbol streams for each SM user. However, since the code resource could usually be limited, especially for the high-speed demand in which small spreading gains are preferred, some users may have to share the same code, such that the antenna resource is still necessary for further MAI rejection. Hence, a conceivable implementation allowing for a moderate spatial cost is to incorporate CDMA, leverage the code resource to mitigate MAI from the distinct-code users, and reserve the spatial resource to tackle MAI among the remaining same-code users. The assumption $M \geq Q_D + NQ_M$, as a result, would not be too severe since Q_D and Q_M merely account for the numbers of the same-code OSTBC and SM terminals. The proposed group-wise detection property remains true when separation of the same-code user's signals is considered.

Remark: It should be noted that, in a CDMA based implementation, the cell-wide synchronization is no longer necessary since the MAI among distinct-code users can be rejected through despreading. As a result, only the data streams transmitted from the same-code users need to aligned for further signal separation: symbol synchronization is required merely within a user subset, namely, the same-code user group, but not for all the user terminals.

V. Low-Complexity Algorithm Implementation

By further exploiting the embedded structure of the MF channel matrix $\mathbf{F}$ and its inverse, this section derives a low-complexity detector realization. Since the ZF and MMSE weighting matrices exhibit essentially the same algebraic structure (namely, both fall within the class $\mathcal{F}(L)$ for some L), the discussions will focus on the ZF case for notational simplicity. In what follows it will be shown that the computation of $\mathbf{F}^{-1}$, which would often dominate the cost, boils down to solving a set of linear equations of relatively small dimensions. Moreover, inverting the "big" $\mathbf{F}$ in the initial stage turns out to be the only direct matrix inversion to be performed; the inverse matrix $\mathbf{F}_i^{-1}$ required at each subsequent stage can be recursively computed based on the parameters available in the previous stage. Flop count analysis is also provided for complexity comparison with other competitive methods.

A. Computation of $\mathbf{F}^{-1}$

Recall from Prop. 4.1 that *every* $P \times P$ block off-diagonal submatrix of $\mathbf{F}^{-1}$ is a $P \times P$ orthogonal design, and each diagonal block is sparse (it is a scalar multiple of $\mathbf{I}_P$). This imposes certain structural redundancy in $\mathbf{F}^{-1}$ that should be carefully tackled for avoiding the computation of duplicate parameters. Indeed, since a $P \times P$ orthogonal design is completely characterized by P independent variables [21], it suffices to determine one column, say, the last one, for each off-diagonal block submatrix; the rest columns can simply be obtained from the last one through some known linear transformations [21]. For each diagonal block, in particular, only one unknown needs to be found. The aforementioned *a priori* structural information shows that, to completely specify $\mathbf{F}^{-1}$, we only need to find its (jP)th columns for $1 \leq j \leq L$: this amounts to solving the linear equations of reduced dimensions

$$\mathbf{FG} = \mathbf{E}, \tag{16}$$

in which $\mathbf{G}$ and $\mathbf{E}$ are $L_T \times L$ matrices whose lth columns are, respectively, the (jP)th columns of $\mathbf{F}^{-1}$ and the identity matrix $\mathbf{I}_{L_T}$. To solve for the unknown $\mathbf{G}$ based on (16), one can further take into account the sparse nature of $\mathbf{G}$. Specifically, for the jth column $\mathbf{g}_j$, we *must* have $g_{i,j} = 0$ for $(j-1)P+1 \leq i \leq jP-1$; the imbedded $P-1$ consecutive zero entries in $\mathbf{g}_j$ come from the last column of the jth $P \times P$ diagonal block of $\mathbf{F}^{-1}$. As a result, only the nonzero entries are to be determined. Moreover, since $\mathbf{F}^{-1}$ is Hermitian, in each $\mathbf{g}_j$, we only have to compute the entries *lying below* $g_{jP-1,j}(=0)$. In summary, for the jth column $\mathbf{g}_j$, only the last $P(L-j)+1$ entries are to be determined. It is noted that the number of unknowns is decreased by an amount of P as the column index j increases to $j+1$ (for the last column $\mathbf{g}_{jP-1}$, only one unknown is yet to be computed).

To incorporate the above structural information for solving (16), let $\mathbf{F} = \mathbf{L}\mathbf{L}^T$, where $\mathbf{L}$ is an $L_T \times L_T$ lower triangular matrix, be a Cholesky factorization [8, p-144] of the matrix $\mathbf{F}$; such a factorization is typically required for inverting a nonsingular square real symmetric matrix [8, p-141]. In terms of columns of $\mathbf{G}$ and $\mathbf{E}$, (16) can thus be equivalently rewritten as

$$\mathbf{L}\mathbf{L}^T\mathbf{g}_j = \mathbf{e}_j, \qquad 1 \leq j \leq L. \tag{17}$$

Since $\mathbf{L}$ is lower triangular and hence $\mathbf{L}^T$ is upper triangular, a standard approach to solve for the unknown column vector $\mathbf{g}_j$ in (17) is the forward and back substitutions [8, p-121]: for each j, it first solves $\mathbf{L}\mathbf{u}_j = \mathbf{e}_j$, where $\mathbf{u}_j \triangleq \mathbf{L}^T\mathbf{g}_j$, by forward substitutions for the intermediate vector $\mathbf{u}_j$ and then backward solves $\mathbf{L}^T\mathbf{g}_j = \mathbf{u}_j$ for the desired unknown $\mathbf{g}_j$. As long as the intermediate vector $\mathbf{u}_j$ is obtained, the back substitution process successively computes the elements in $\mathbf{g}_j$ one after another, starting from the *last* entry (since $\mathbf{L}^T$ is upper triangular). Since the unknowns to be determined in each $\mathbf{g}_j$ all lie below the entry $g_{jP-1,j}(=0)$, the back substitution procedure simply terminates as long as $g_{jP,j}$ is computed; the remaining entries on top of $g_{jP-1,j}$ are redundant since $\mathbf{F}^{-1}$ is Hermitian.

Remark: In the MMSE V-BLAST case, an alternative for computing the initial MMSE weight $\left[\left(2/\sigma_v^2\right)\mathbf{F} + \mathbf{I}_{L_T}\right]^{-1}$ is to decompose $\mathbf{F} = \mathbf{H}_c^T\mathbf{H}_c$ into a sum of L_T rank-one matrices, each being an outer-product of a row of the channel matrix $\mathbf{H}_c$, and then uses the Sherman-Morrison formula to recursively obtain the solution [2, p-1725]. Such a recursive procedure does not seem to be a good choice for the considered structured inverse computation. This is because the recursive formulation, each time incorporating a "piece" of the channel matrix component, renders it difficult to exploit the structural information (seen based on the "whole" channel matrix) for discarding the redundant coefficients for computational reduction. In the proposed algorithm [2] (see (28) in [2, p-1725]), the adopted recursive computation of $\mathbf{F}^{-1}$ at each step needs to perform the PM dimensional matrix and vector operations. After performing PL steps, the total number of flop counts is thus $\frac{7}{2}P^3L^2M + \frac{5}{2}P^2LM$, where M is the number of receive antennas with $M \geq L$. For the proposed method, according to [8, p-144], the number of flop counts[3] of performing the Cholesky factorization of $\mathbf{F}$ is $\frac{1}{3}N^3Q_M^3 + N^2P^2Q_DQ_M^2 + \frac{1}{2}N^2Q_M^2 + NP^2Q_D^2Q_M - 2NP^2Q_DQ_M + 3NPQ_DQ_M + \frac{1}{3}P^2Q_D^3$. The number of flop counts of the forward substitution for solving $\mathbf{u}_j$, $1 \le j \le L$, is $\frac{1}{3}N^3P^2Q_M^3 + \frac{1}{2}N^2P^2Q_DQ_M^2 - N^2PQ_D^2Q_M + \frac{1}{3}NP^2Q_D^3 + \frac{3}{2}NP^2Q_D^2Q_M - \frac{1}{2}NP^2Q_DQ_M^2 - 4NPQ_DQ_M$. To solve $\mathbf{g}_j$, $1 \le j \le L$, by the backward substitution, we will need $\frac{1}{3}N^3P^2Q_M^3 + \frac{3}{2}N^2P^2Q_M^2 - 2N^2PQ_D^2 + \frac{1}{2}NP^2Q_D^2Q_M - 2NP^2Q_DQ_M + NP^2Q_D^3 - \frac{1}{2}P^2Q_D^2$ flop counts. As a result, the total number of flop counts for solving $\mathbf{F}^{-1}$ is

$$\begin{aligned} C_{initial} = & \frac{2}{3}N^3P^2Q_M^3 + \frac{1}{3}N^3Q_M^3 + \frac{3}{2}N^2P^2Q_DQ_M^2 \\ & + \frac{3}{2}N^2P^2Q_M^2 - N^2PQ_D^2Q_M - 2N^2PQ_D^2 \\ & + \frac{1}{2}N^2Q_M^2 + \frac{1}{3}NP^2Q_D^3 + \frac{3}{2}NP^2Q_D^2Q_M \\ & - 4NP^2Q_DQ_M - NPQ_DQ_M + NP^2Q_D^3 \\ & + \frac{3}{2}NP^2Q_D^2Q_M - \frac{1}{2}NP^2Q_DQ_M^2 + \frac{1}{3}P^2Q_D^3 \\ & - \frac{1}{2}P^2Q_D^2. \end{aligned} \quad (18)$$

It can be seen that the proposed method requires less computation.

B. Recursive Computation of $\mathbf{F}_i^{-1}$

Recall from Section IV that, at the ith stage, the inverse matrix $\mathbf{F}_i^{-1} \triangleq \left(\mathbf{H}_{c,i}^T\mathbf{H}_{c,i}\right)^{-1}$, where $\mathbf{H}_{c,i}$ is obtained by deleting i block(s) of P columns from $\mathbf{H}_c$, is required for determining the optimal detection order and the associated weighting matrix. With $\mathbf{F}^{-1}$ obtained in the initial stage, in what follows we will show how $\mathbf{F}_i^{-1}$ at each stage can be recursively computed based on $\mathbf{F}_{i-1}$ and $\mathbf{F}_{i-1}^{-1}$.

It is noted again that, at the ith processing stage, $1 \le i \le L-1$, the matrix $\mathbf{F}_i = \mathbf{H}_{c,i}^T\mathbf{H}_{c,i}$ is simply obtained from $\mathbf{F}_{i-1}$ $\left(= \mathbf{H}_{c,i-1}^T\mathbf{H}_{c,i-1}\right)$ by deleting one block of P columns and the corresponding indexed block of P rows. Without loss of generality, we may assume that the last column and row blocks of $\mathbf{F}_{i-1}$ are to be deleted; otherwise we can simply permute those to be discarded to the right and bottom ends of $\mathbf{F}_{i-1}$ to fit the prescribed form. As a result, we can partition $\mathbf{F}_{i-1}$ as

$$\mathbf{F}_{i-1} = \left[\begin{array}{c|c} \mathbf{F}_i & \mathbf{B}_{i-1} \\ \hline \mathbf{B}_{i-1}^T & \mathbf{D}_{i-1} \end{array}\right], \quad (19)$$

where $\mathbf{B}_{i-1} \in \mathcal{R}^{(L_T - iP)\times P}$ and $\mathbf{D}_{i-1} = d_{i-1}\mathbf{I}_P$ for some scalar d_{i-1} are to be deleted. Denote by $\bar{\mathbf{F}}_{i-1}$ the $(L_T - iP) \times (L_T - iP)$ principle submatrix of $\mathbf{F}_{i-1}^{-1}$ obtained by deleting its last iP columns and last iP rows; the matrix $\bar{\mathbf{F}}_{i-1}$ is thus available from the $(i-1)$th stage. From (19) and by using the inversion lemma for block matrix[4], $\bar{\mathbf{F}}_{i-1}$ can be expressed in terms of $\mathbf{F}_i$, $\mathbf{B}_{i-1}$, and $\mathbf{D}_{i-1}$ as

$$\begin{aligned} \bar{\mathbf{F}}_{i-1} &= \left(\mathbf{F}_i - \mathbf{B}_{i-1}\mathbf{D}_{i-1}^{-1}\mathbf{B}_{i-1}^T\right)^{-1} \\ &= \left(\mathbf{F}_i - d_{i-1}^{-1}\mathbf{B}_{i-1}\mathbf{B}_{i-1}^T\right)^{-1}, \end{aligned} \quad (20)$$

where the second equality in (20) follows since $\mathbf{D}_{i-1} = d_{i-1}\mathbf{I}_P$. Equation (20) links the matrix $\mathbf{F}_i$, which is to be inverted at the ith step, to the elements of $\mathbf{F}_{i-1}$ and $\mathbf{F}_{i-1}^{-1}$, which are available from the previous iteration. In particular, it follows immediately from (20) that $\mathbf{F}_i = \bar{\mathbf{F}}_{i-1}^{-1} + d_{i-1}^{-1}\mathbf{B}_{i-1}\mathbf{B}_{i-1}^T$, and we can use the matrix inversion lemma [14, p-246] to obtain

$$\begin{aligned} \mathbf{F}_i^{-1} &= \left(\bar{\mathbf{F}}_{i-1}^{-1} + d_{i-1}^{-1}\mathbf{B}_{i-1}\mathbf{B}_{i-1}\right)^{-1} \\ &= \bar{\mathbf{F}}_{i-1} - \bar{\mathbf{F}}_{i-1}\mathbf{B}_{i-1}\left(\mathbf{B}_{i-1}^T\bar{\mathbf{F}}_{i-1}\mathbf{B}_{i-1} + d_{i-1}\mathbf{I}_P\right)^{-1} \\ &\quad \times \mathbf{B}_{i-1}^T\bar{\mathbf{F}}_{i-1}. \end{aligned} \quad (21)$$

From (21), we can see that the computation of the $(L_T - iP) \times (L_T - iP)$ inverse matrix $\mathbf{F}_i^{-1}$ is relieved into inverting the $P \times P$ matrix $\mathbf{B}_{i-1}^T\bar{\mathbf{F}}_{i-1}\mathbf{B}_{i-1} + d_{i-1}\mathbf{I}_P$, which is of a lower dimension. In fact, the efforts to invert this $P \times P$ matrix can be reduced even further. To see this, we need the next result.

Proposition 5.1: Let $\mathbf{B}_{i-1}$ and $\bar{\mathbf{F}}_{i-1}$ be defined in (19) and (20), respectively. Then it follows that $\mathbf{B}_{i-1}^T\bar{\mathbf{F}}_{i-1}\mathbf{B}_{i-1} = \lambda_{i-1}\mathbf{I}_P$ for some scalar λ_{i-1}.

Proof: For a fixed i, let us define $J \triangleq L - i$. For $1 \le k, l \le J$, denote by $\mathbf{U}_{k,l}$ the (k,l)th $P \times P$ block submatrix of $\bar{\mathbf{F}}_{i-1}$. We drop the index indicating the dependency of J and $\mathbf{U}_{k,l}$ on the number of iteration $i-1$ to simplify notation. Let us write $\mathbf{B}_{i-1} = \left[\mathbf{B}_1^T \cdots \mathbf{B}_J^T\right]^T$, where $\mathbf{B}_k \in \mathcal{O}(P)$; then it follows immediately that

$$\begin{aligned} \mathbf{B}i - 1^T\bar{\mathbf{F}}_{i-1}\mathbf{B}_{i-1} &= \sum_{k,l=1}^{J} \mathbf{B}_k^T\mathbf{U}_{k,l}\mathbf{B}_l \\ &= \sum_{k=1}^{J} \mathbf{B}_k^T\mathbf{U}_{k,k}\mathbf{B}_k \\ &\quad + \sum_{k,l=1,\ k\ne l}^{J} \mathbf{B}_k^T\mathbf{U}_{k,l}\mathbf{B}_l. \end{aligned} \quad (22)$$

Since $\bar{\mathbf{F}}_{i-1} \in \mathcal{F}(J)$, we have by definition $\mathbf{U}_{k,k} = \eta_k\mathbf{I}_k$ for some scalar η_k: the first summation on the right-hand-side of the second equality in (22) thus simplifies as $\sum_{k=1}^{J}\mathbf{B}_k^T\mathbf{U}_{k,k}\mathbf{B}_k = \sum_{k=1}^{J}\eta_k\mathbf{B}_k^T\mathbf{B}_k = \eta\mathbf{I}_P$ for some scalar η. On the other hand, it can be shown that $\mathbf{B}_k^T\mathbf{U}_{k,l}\mathbf{B}_k \in \mathcal{O}(P)$ [21], and this implies $\mathbf{B}_k^T\mathbf{U}_{k,k}\mathbf{B}_l + \mathbf{B}_l^T\mathbf{U}_{l,k}\mathbf{B}_k = \alpha_{k,l}\mathbf{I}_P$ for some scalar $\alpha_{k,l}$. As a result, we have $\sum_{k,l=1,\ k\ne l}\mathbf{B}_k^T\mathbf{U}_{k,l}\mathbf{B}_l = \tilde{\eta}\mathbf{I}_P$ for some scalar $\tilde{\eta}$, and the assertion follows. □

Proposition 5.1 implies

$$\mathbf{B}_{i-1}^T\bar{\mathbf{F}}_{i-1}\mathbf{B}_{i-1} + d_{i-1}\mathbf{I}_P = c_{i-1}\mathbf{I}_P, \quad (23)$$

[3]The low-order terms which are insignificant to the total count are neglected.

[4]Assume that $\mathbf{W} = \begin{bmatrix} \mathbf{X} & \mathbf{Y} \\ \mathbf{Y}^T & \mathbf{Z} \end{bmatrix}$ with $\mathbf{X}$ and $\mathbf{Z}$ invertible. By compatibly partitioning the inverse matrix as $\mathbf{W}^{-1} = \begin{bmatrix} \bar{\mathbf{X}} & \bar{\mathbf{Y}} \\ \bar{\mathbf{Y}}^T & \bar{\mathbf{Z}} \end{bmatrix}$, we then have $\bar{\mathbf{X}} = \left(\mathbf{X} - \mathbf{Y}\mathbf{Z}^{-1}\mathbf{Y}^T\right)^{-1}$.

and, from (21), we reach the following key equation for inverting the matrix $\mathbf{F}_i^{-1}$

$$\mathbf{F}_i^{-1} = \bar{\mathbf{F}}_{i-1} - c_{i-1}^{-1}\bar{\mathbf{F}}_{i-1}\mathbf{B}_{i-1}\mathbf{B}_{i-1}^T\bar{\mathbf{F}}_{i-1}. \qquad (24)$$

Given $\mathbf{F}_{i-1}$ and $\mathbf{F}_{i-1}^{-1}$, equation (24) provides a simple recursive formula for computing $\mathbf{F}_i^{-1}$ without any direct matrix inversion operations. The proposed recursive approach for computing $\mathbf{F}_i^{-1}$ is basically a block based implementation of the method in [2, p-1726] introduced for the conventional symbol-wise V-BLAST algorithm. A distinctive feature of our scheme, nonetheless, is the attractive simplification of (21) to (24), with which the computation of the $P \times P$ inverse matrix $\left(\mathbf{B}_{i-1}^T\bar{\mathbf{F}}_{i-1}\mathbf{B}_{i-1} + d_{i-1}\mathbf{I}_P\right)^{-1}$ in (21) boils down to finding the scalar parameter c_{i-1}^{-1} (this benefits from the usage of orthogonal codes). To roughly assess the associated saving in flop cost, it is noted from (23) that the scalar parameter c_{i-1} is completely determined by the (constant) diagonal entries of the matrix $\mathbf{B}_{i-1}^T\bar{\mathbf{F}}_{i-1}\mathbf{B}_{i-1}$. This implies that the computation of c_{i-1} amounts to evaluating a quadratic form $\mathbf{b}_{i-1}^T\bar{\mathbf{F}}_{i-1}\mathbf{b}_{i-1}$, where $\mathbf{b}_{i-1}$ denotes an arbitrary column of $\mathbf{B}_{i-1}$. The required number of flop counts is no more than $2P^2 + P - 1$, which is substantially small than that required for inverting a $P \times P$ matrix.

C. Complexity Comparisons with Other Multiuser Detection Schemes

This subsection compares the algorithm complexity of the proposed group-wise V-BLAST detector with two typical signal detection schemes for MU ST coded wireless systems, namely, the Naguib's PIC scheme [16] and the Stamoulis's decoupled-based method [19].

It should be noted that the two comparable methods, originally tailored for the MU-STBC environment, rely on the orthogonality property specific to the associated MU channel matrix. In the considered dual-signaling case, this structural requirement is not seen in the dual-mode MU channel matrix $\mathbf{H}_c$ (owing to the presence of high-rate SM streams) but can be "restored" through ST matched filtering (cf. Prop. 3.1). As a result, the two alternative schemes can fit the considered dual-signaling systems based on the MF signal model (8). To evaluate the flops, one should note that, for a dual-signaling system with Q_M SM terminals, there are totally $(NQ_M)^2$ signature blocks of the form $\alpha\mathbf{I}_P$ imbedded in the MF channel matrix $\mathbf{F}$. These diagonal block submatrices will impose certain sparse structure distributed over $\mathbf{F}$, making it difficult to compute the accurate flop counts. We thus assume without loss of generality that the symbol groups of the Q_M SM users are permuted to the top end of the MU symbol vector $\mathbf{s}_c(k)$; in this way all the sparse blocks will cluster in the upper left corner of $\mathbf{F}$ (each $P \times P$ block of the first $PNQ_M \times PNQ_M$ principle submatrix of $\mathbf{F}$ is a scalar multiple of $\mathbf{I}_P$). Such a "centralized" sparse structure of $\mathbf{F}$ can simplify the process of flop evaluations. It is also noted that, for a system with Q_M SM users and Q_D OSTBC users, there are totally $(Q_D + NQ_M)!/[(Q_D)!(NQ_M)!]$ dissimilar group V-BLAST orderings. Due to the imbedded sparse nature of $\mathbf{F}$, the required flop counts will be different for distinct sorting sequences. A reasonable complexity measure of the V-BLAST detector, therefore, would be the mean flop counts, averaged over all the possible orderings, but this turns out to be highly intractable. A sensible approximation to the exact mean flop counts is the average of the minimal and maximal costs. Toward this end, we shall first determine the orderings associated with the two extreme cases. The detection orders allowing for the least computational effort must entail an utmost benefit from the sparse structure in per layer processing. The solutions are therefore the sorting sequences in which exactly all the Q_D OSTBC streams are to be detected in the initial Q_D layers; as such, all the dense blocks in $\mathbf{F}$ resulting from the Q_D diversity users will be removed after the first Q_D iterations and, meanwhile, the sparse structure is retained as much as possible in each processing layer. The detection orders incurring the highest complexity demand, on the other hand, will be the particular choices which decide to recover all the NQ_M SM signals in the initial NQ_M stages. This will leave as many as possible the dense blocks in each $\mathbf{F}_i$ for $i > 1$, hence the smallest possible sparse region to be exploited for computational reduction. The average flop cost of two extreme cases is listed in Table II (the results of the Naguib's and Stamoulis's methods are also obtained in essentially the same manner).

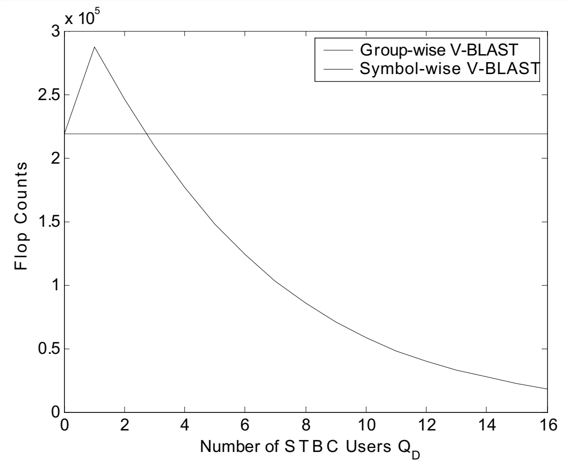

Fig. 2. Flop counts of the proposed group-wise V-BLAST detection as a function of Q_D for an MU dual-signaling system, with $Q = 16$, $Q_M = Q - Q_D$ and $N = 2$.

Example 5.1: Considering the QPSK modulation (i.e., $D = 2$) and choosing $N = 2$ (thus $P = K = 2$), $Q_D = Q_M = 2$, $L = 6$, we have the result: $C_{Naguib} \approx 4336$, $C_{Stamoulis} \approx 6648$ and $C_{Group} \approx 1083$, and this yields the approximate ratio of complexity of $4 : 6 : 1$ for Naguib's two-step method, Stamoulis's method and proposed group-wise V-BLAST method. As we can see, the proposed solution requires less computation.

VI. Two-Stage Processing of Dual-Mode Signals

Let us consider a dual-signaling platform with Q users, each equipped with $N = 2$ transmit antennas (Alamouti's code for diversity). In case that all the Q cell users are in the SM mode, there will be totally $NQ = 2Q$ transmitted symbols during one signaling period, and one can use the conventional symbol-wise V-BLAST detector to separate the $2Q$ coupled symbol streams. Assume that there are $1 \leq Q_D \leq Q$ users

TABLE II

FLOP COUNTS OF DIFFERENT DETECTION METHODS (D: CONSTELLATION SIZE).

Method	Flop Counts
Group-Wise V-BLAST	$2N^3P^2Q_M^3+\frac{1}{6}N^3PQ_M^3+\frac{1}{3}N^3Q_M^3+\frac{19}{6}N^2P^2Q_DQ_M^2+N^2P^2Q_M^2-N^2PQ_D^2Q_M-2N^2PQ_D^2+N^2PQ_DQ_M^2+\frac{3}{2}N^2PQ_M^2+\frac{4}{3}NP^2Q_D^3+\frac{11}{2}NP^2Q_D^2Q_M-\frac{1}{2}NP^2Q_DQ_M^2-12NP^2Q_DQ_M+\frac{1}{3}NPQ_D^2Q_M+\frac{4}{3}P^2Q_D^3$
Stamoulis's Method	$\frac{25}{12}N^4P^2Q_M^4+\frac{23}{3}N^3P^2Q_DQ_M^3-\frac{19}{6}N^3P^2Q_M^3+\frac{25}{2}N^2P^2Q_D^2Q_M^2-\frac{23}{2}N^2P^2Q_DQ_M^2-\frac{5}{4}N^2P^2Q_M^2+\frac{7}{4}N^2PQ_M^2+\frac{25}{6}NP^2Q_D^3Q_M-\frac{23}{2}NP^2Q_D^2Q_M-\frac{7}{2}NP^2Q_DQ_M+\frac{25}{12}P^2Q_D^4-\frac{23}{6}P^2Q_D^3$
Naguib's 2-Step Method	$2\left(2^D\right)N^2P^2Q_M^2+2\left(2^D\right)N^2PQ_M^2+4\left(2^D\right)NP^2Q_DQ_M+5\left(2^D\right)NPQ_DQ_M+2\left(2^D\right)P^2Q_D^2+5N^3P^2Q_M^3-2N^3PQ_M3+12N^2P^2Q_DQ_M^2-5N^2P^2Q_M^2-3N^2PQ_DQ_M^2+\frac{9}{2}N^2PQ_M^2+11NP^2Q_D^2Q_M-10NP^2Q_DQ_M-3NPQ_D^2Q_M+4P^2Q_D^3$
Two-Stage V-BLAST	$3N^2P^2Q_DQ_M^2+\frac{1}{2}N^2PQ_DQ_M^2+\frac{7}{2}N^2Q_M^3+3NP^2Q_D^2Q_M-6NP^2Q_DQ_M+\frac{1}{2}NPQ_D^2Q_M+P^2Q_D^3$

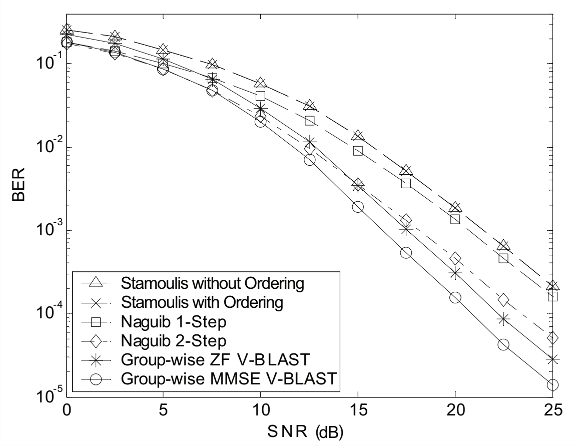

Fig. 3. Average BER performances of various detection schemes as a function of SNR for a four-user MU-STBC system, with $N = 2$ and $M = 4$. The Alamouti's code with 8-PSK modulation is used for each user.

otherwise choosing the OSTBC mode. To realize a two-fold diversity gain for the Q_D OSTBC links, the receiver has to spend two time periods for data buffering, during which there are totally $4(Q-Q_D)+2Q_D$ independent symbols sent from all the users: $4(Q-Q_D)$ are from the SM users and $2Q_D$ are from OSTBC users. Since $4(Q-Q_D)+2Q_D > 2Q$ whenever $1 \leq Q_D \leq Q$, the presence of Q_D users switching to the OSTBC mode does augment the data processing dimension (to an excess of $2(Q-Q_D)$ over the full high-rate case). For $Q = 16$, Fig. 2 shows the required flop counts of the group-wise V-BLAST detector as Q_D increases from 0 to 16. The $Q_D = 0$ case, with the most compact data model, serves as the inherent benchmark cost. As the figure shows, the proposed group-wise V-BLAST scheme tends to prevent significant excessive flop counts over the $Q_D = 0$ case, and there is even a save in the counts for $Q_D \geq 3$. It is noted that the count reaches the peak when $Q_D = 1$, and monotonically decreases with Q_D. This is not unexpected, and is indeed true in general, since the excess data processing dimension over the benchmark case is $2(Q-Q_D)$, which attains the maximal when $Q_D = 1$. For the full diversity case ($Q = Q_D = 16$), in which the number of symbols to be detected is the same as the full high-rate case ($= 2Q$), the proposed method achieves largest flop count reduction. Based on these observations, the "joint" processing of the dual-mode signal could entail more excess flops over the benchmark performance when a few OSTBC users are present. In such a case, a plausible approach for complexity reduction is to first detect the OSTBC streams, since they may potentially be more robust. By removing the contributions of the detected OSTBC symbols from the data model, one can resort to a symbol-wise V-BLAST algorithm to recover the remaining SM streams over a relatively small data model dimension. Since the detection order in this way may violate the actual optimal sorting, such a "two-stage" processing strategy can reduce computation at the expense of a possible performance drop. The computational cost of such an approach is also included in Table II, based on which the amount of flop reduction with respect to the original method can be found as

$$\begin{aligned} C_\Delta \triangleq\ & (\text{flop counts of original appraoch}) \\ & -(\text{flop counts of two-stage approach}) \\ =\ & 2N^3P^2Q_M^3+\frac{1}{6}N^3PQ_M^3+\frac{1}{6}N^2P^2Q_DQ_M^2 \\ & +N^2P^2Q_M^2-N^2PQ_D^2Q_M-2N^2PQ_D^2 \\ & +\frac{1}{2}N^2PQ_DQ_M^2+\frac{3}{2}N^2PQ_M^2-\frac{7}{2}N^2Q_M^3 \\ & +\frac{4}{3}NP^2Q_D^3+\frac{5}{2}NP^2Q_DQ_M-\frac{1}{2}NP^2Q_DQ_M^2 \\ & -6NP^2Q_DQ_M-\frac{1}{6}NPQ_D^2Q_M+\frac{1}{3}P^2Q_D^3. \end{aligned} \quad (25)$$

Given the system parameters considered in Example 5.1, the two-stage approach reduces about a 47% computational complexity associated with the original method with optimal ordering.

VII. SIMULATION RESULTS

In this section we use several numerical simulations to illustrate the performance of the proposed group-wise V-BLAST detector. Each user's channel is quasi-static: it remains constant over each packet of 100 symbol blocks and independently varies between packets. Also, perfect channel knowledge at the receiver is assumed. In all simulations, the transmit power of all users are set to be equal; the number of receive antennas M is set to meet the minimal requirement, i.e., $M = L = Q_D + NQ_M$.

A. MU-STBC Systems with Alamouti's Code

This simulation considers the special MU-STBC case (with Alamouti's code) and evaluates the performances of the proposed method with the Stamoulis's method [19] and the

Naguib's method [16], both being tailored for the MU-STBC systems. Fig. 3 compares the detection performances of the three families of methods in a four-user system over i.i.d. Rayleigh fading channels (with 8-PSK modulation) in terms of the average bit error rates (BERs) (averaged over the four detected streams). The results show that the group-wise MMSE V-BLAST attains a better performance. The Naguib's two-step method [16] yields a comparable performance when SNR is low but it degrades in the medium-to-high SNR regime. It is noted that, in the Naguib's two-step method, the detection accuracy in the second stage hinges entirely on the reliability of all the initial signal estimates. Accordingly, each user's symbol stream, detected via the PIC based mechanism, would be subject to an essentially equal risk of error propagation resulting from the incorrect decisions in the initial stage. The V-BLAST solution, on the other hand, can provide a layer-wise increase in receive diversity to limit the effect of possible decision error leakage: this would account for the superiority average performance over the Naguib's approach. The Stamoulis's method [19], with or without ordering, gives a relatively poor performance among the three families; this is because, as compared with the other two alternatives, the decoupled-based formulation also does not induce an increase in receive diversity as the iteration goes on. Detailed comparisons of the three competitive methods for MU-STBC systems, in terms of algorithm operations and complexities, are referred to [24].

B. Symbol Detection in Dual-Signaling Systems

We consider a system with four users, two experiencing correlated channels described by the Ricean model [4] and the other two being over i.i.d. Rayleigh fading channels (two transmit antennas are placed at each user terminal) The Alamouti and SM schemes, respectively, are used for data transmission over Ricean and Rayleigh channels; the symbol constellations used are QPSK for SM terminals and 16-QAM for OSTBC terminals so that the bit data rates are the same. The first experiment compares the performances of the above three detection schemes in the dual-signaling environment. Fig. 4 (a) (b), respectively, show the average BER of the two detected OSTBC users (the κ-factors in the two Ricean channels are both set to be $\kappa = 10$, which corresponds to a medium eigenvalue spread = 5.3) and the two SM users. Compared with Fig. 3, we can see that the BER curves in the dual-singling case exhibit similar tendency as in the MU-STBC systems; the proposed MMSE group-wise V-BLAST achieves the best performance as long as SNR is above 10 dB. It is noted that the performance of the proposed two-stage processing scheme (for further computational reduction) is comparable to that of the Naguib's two-step method; however, it incurs a 3 dB loss in SNR at BER = 10^{-3} for OSTBC users and $1.5 \sim 2$ dB for SM users as compared with the original method (with optimal ordering). The second experiment simulates the BER performances of three representative detection schemes (the ordered Stamoulis's method, the Naguib's two-step method, and the proposed group-wise MMSE V-BLAST) at different channel correlation tendencies. For the two Ricean channels, we consider five different Ricean κ-factors: 0, 1, 10, 30, and

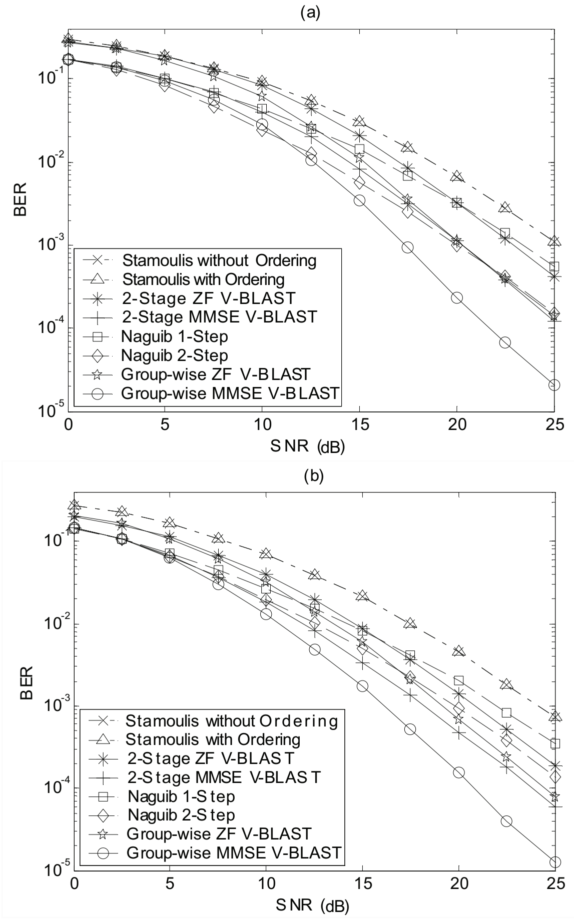

Fig. 4. BER performance of various detection schemes as a function of SNR for a four-user dual-signaling system, with $Q_D = Q_M = 2$, $N = 2$, $M = 6$ and $\kappa = 10$. 16-QAM and QPSK modulations are used for OSTBC signaling and SM signaling, respectively. (a) Average BER of the two detected OSTBC users. (b) Average BER of the two detected SM users.

100 (large κ implies severe correlations; the extreme selections $\kappa = 0$ and $\kappa = 100$, respectively, render the channel to be independent fading and almost light-of-sight). Figs. 5 (a) (b) show the average BER of the two classes of detected streams for different κ factors. It can be seen that the performances in all cases deteriorate as κ increases. This is not unexpected since the OSTBC scheme may lose the diversity gain over the correlated channels, and large κ factor thus incurs large BER degradation. The performance drop of the SM scheme may result from the increased amount of error propagation due to the poorly detected OSTBC streams caused by the loss in diversity gain over low-rank channels. The figure also shows that, for a fixed SNR, the proposed group-wise MMSE V-BLAST seems to incur less BER spread as κ increases. This could benefit from the V-BLAST mechanism, in which the detect-and-cancel process induces more receive diversity and improves detection accuracy in each layer, leading to better average performance against severe spatial correlation.

VIII. Conclusion

Co-channel interference mitigation in multiuser space-time wireless systems is of great importance for maintaining a good link quality. The originality of the presented study is the

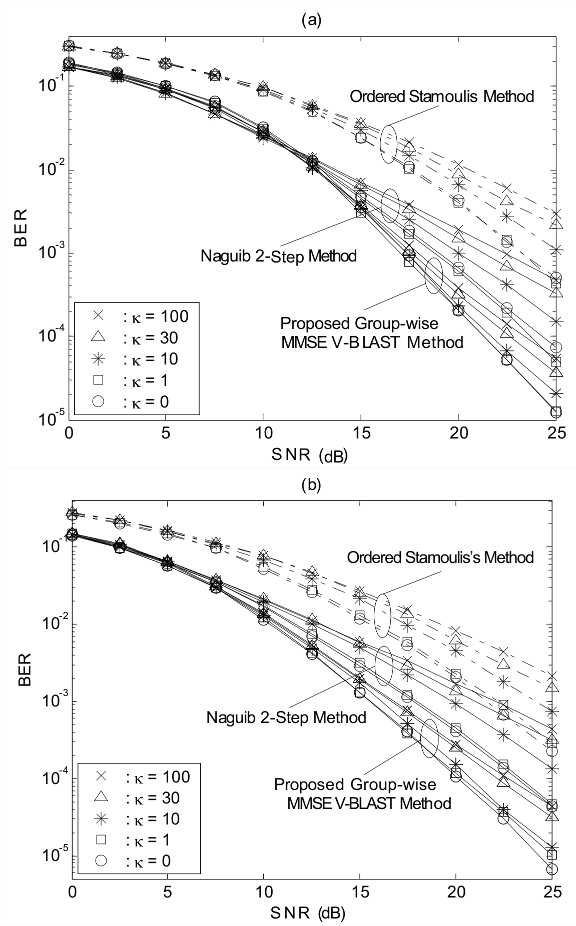

Fig. 5. BER performance of three detection schemes in a four-user dual-signaling system for different Ricean κ-factors, with $Q_D = Q_M = 2$, $N = 2$ and $M = 6$. 16-QAM and QPSK modulations are used for OSTBC signaling and SM signaling, respectively. (a) Average BER of the two detected OSTBC users. (b) Average BER of the two detected SM users.

investigation of the impact of orthogonal ST block codes on the V-BLAST based detection in a multiuser dual-signaling environment. The proposed group-wise detection property, based on exploiting the block-orthogonal structure imbedded in the matched-filtered channel matrix, potentially reduces computations and the overall decoding load. Our flop count analysis shows that, with a medium number of OSTBC users in the cell, the complexity of the proposed solution is comparable to that of the conventional symbol-wise V-BLAST algorithm in a pure SM signaling environment: our solution is thus an attractive receiver candidate for a dual-signaling platform. The distinctive structure of the channel matrix also leads to a low-complexity detector realization, which affords relatively low computational cost when compared with existing methods. Numerical simulations demonstrate the effectiveness of the proposed V-BLAST based solution: it compares favorably with existing reported interference mitigation schemes for multiuser space-time coded systems.

APPENDIX I
PROOF OF PROPOSITION 3.1

Assume that real-valued constellations with unit-rate codes ($K = P$) are used for STBC terminal. Part *(1)* of the proposition has been shown in [11]. It only needs to prove *(2)* and *(3)*.

Proof of (2): We first note from (3) and (4) that the effective signal component (priori to matched filtering) of the stream from the nth transmit antenna of the pth SM user is $\tilde{\mathbf{H}}_{p,n}\tilde{\mathbf{A}}_p^{(n)}$, where $\tilde{\mathbf{H}}_{p,n} \in \mathcal{R}^{2MP\times 2P}$ denotes the matrix consisting of the $(2P(n-1)+l)$th columns of $\tilde{\mathbf{H}}_p$ (see (4)), for $1 \le l \le 2P$, $\tilde{\mathbf{A}}_p^{(n)} \in \mathcal{R}^{2P\times P}$ denotes the real-valued ST modulation matrix of the nth antenna at the pth SM user as defined in (4). Accordingly, the MF coupling signature between this stream and that from the dth antenna of the qth SM user is $\left(\tilde{\mathbf{A}}_p^{(n)}\right)^T \tilde{\mathbf{H}}_{p,n}^T\tilde{\mathbf{H}}_{q,d}\tilde{\mathbf{A}}_q^{(d)}$, whose (i,j)th entry, $1 \le i,j \le P$, is directly computed as

$$\begin{aligned} f_{p,q}^{(i,j)}(n,d) &= \sum_{k=1}^{P}\left(\tilde{\mathbf{a}}_{p,i}^{(k)}(n)\right)^T \bar{\mathbf{H}}_{p,n}^T\bar{\mathbf{H}}_{q,d}\tilde{\mathbf{a}}_{q,j}^{(k)}(d) \\ &= \sum_{k=1}^{K}\mathrm{Re}\left\{\left(a_{p,i}^{(k)}(n)\right)^* \mathbf{h}_{p,n}^H\mathbf{h}_{q,d}a_{q,j}^{(k)}(d)\right\} \\ &= \mathrm{Re}\left\{\sum_{m=1}^{M}\left(h_{p,n}^{(m)}\right)^* \mathbf{a}_{p,i}^H(n)\mathbf{a}_{q,j}(d)h_{q,d}^{(m)}\right\} \end{aligned} \tag{26}$$

where $\tilde{\mathbf{a}}_{p,i}^{(k)}(n) \triangleq \left[\mathrm{Re}\left\{a_{p,i}^{(k)}(n)\right\}\ \mathrm{Im}\left\{a_{p,i}^{(k)}(n)\right\}\right]^T \in \mathcal{R}^2$ with $a_{p,i}^{(k)}(n)$ being the nth entry of $\mathbf{a}_{p,i}^{(k)}$, (the kth column of the ST modulation matrix $\mathbf{A}_{p,i}$),

$$\bar{\mathbf{H}}_{p,n} \triangleq \begin{bmatrix} \mathrm{Re}\{\mathbf{h}_{p,n}\} & -\mathrm{Im}\{\mathbf{h}_{p,n}\} \\ \mathrm{Im}\{\mathbf{h}_{p,n}\} & \mathrm{Re}\{\mathbf{h}_{p,n}\} \end{bmatrix} \in \mathcal{R}^{2M\times 2}, \tag{27}$$

with $\mathbf{h}_{p,n} \triangleq \left[h_{p,n}^{(1)},\ldots,h_{p,n}^{(M)}\right]^T \in \mathcal{C}^M$ being the nth column of $\mathbf{H}_p$, and $\mathbf{a}_{p,i}^T(n)$ is the nth row of $\mathbf{A}_{p,i}$. Recall that $\mathbf{A}_{p,i}$ is simply a $P\times P$ zero matrix except that the rth row is equal to $\mathbf{e}_s^T$, where $i = (r-1)P + s$ with $1 \le r \le N$, $1 \le s \le P$, and $\mathbf{e}_s$ is the sth standard unit vector in $\mathcal{R}^P$. As a result, it is easy to check that for $1 \le i,j \le P$, $\mathbf{a}_{p,i}^H(n)\mathbf{a}_{q,j}(d) = 1$ if $i = j$, but $\mathbf{a}_{p,i}^H(n)\mathbf{a}_{q,j}(d) = 0$ whenever $i \ne j$. The above analysis thus shows that, for $p,q \in \mathcal{S}_M$, we always have $\left(\tilde{\mathbf{A}}_p^{(n)}\right)^T \tilde{\mathbf{H}}_{p,n}^T\tilde{\mathbf{H}}_{q,d}\tilde{\mathbf{A}}_q^{(d)} = \alpha\mathbf{I}_P$ for some scalar α.

Proof of (2): It suffices to verify that the MF coupling signature between the pth STBC stream and the SM stream transmitted from the nth antenna of the qth SM terminal is a $P\times P$ orthogonal design. It is easy to see that the effective signal coupling matrix is $\tilde{\mathbf{A}}_p^T\tilde{\mathbf{H}}_p^T\tilde{\mathbf{H}}_{q,n}\tilde{\mathbf{A}}_q^{(n)}$, where $\tilde{\mathbf{A}}_p$ is the real-valued ST modulation matrix of the pth user, with the (i,j)th entry, $1 \le i,j \le P$, being computed as

$$\begin{aligned} f_{p,q}^{(i,j)}(n) &= \sum_{k=1}^{P}\left(\tilde{\mathbf{a}}_{p,i}^{(k)}\right)^T \bar{\mathbf{H}}_p^T\bar{\mathbf{H}}_{q,n}\tilde{\mathbf{a}}_{q,j}^{(k)}(n) \\ &= \mathrm{Re}\left\{\sum_{m=1}^{M}\left(\mathbf{h}_p^{(m)}\right)^H \mathbf{A}_{p,i}\mathbf{a}_{q,j}(n)h_{q,n}^{(m)}\right\} \end{aligned} \tag{28}$$

where $\left(\mathbf{h}_p^{(m)}\right)^T$ is the mth row of $\mathbf{H}_p$, and

$$\bar{\mathbf{H}}_p \triangleq \begin{bmatrix} \mathrm{Re}\{\mathbf{H}_p\} & -\mathrm{Im}\{\mathbf{H}_p\} \\ \mathrm{Im}\{\mathbf{H}_p\} & \mathrm{Re}\{\mathbf{H}_p\} \end{bmatrix} \in \mathcal{R}^{2M\times 2N}. \tag{29}$$

TABLE III

SUMMARY OF STRUCTURES OF THE MF CROSS-COUPLING MATRIX $\mathbf{F}_{p,q}$ FOR COMPLEX-VALUED CONSTELLATIONS.

	$N = 2 (K = 2)$	$N = 3$ or $4 (K = 8)$
$p, q \in \mathcal{S}_D$	$p = q$: $\mathbf{F}_{q,q} = \alpha_q \mathbf{I}_4$ $p \neq q$: $\mathbf{F}_{p,q} \in \mathcal{O}(4)$	$p = q$: $\mathbf{F}_{q,q} = \alpha_q \mathbf{I}_8$ $p \neq q$: $\mathbf{F}_{p,q} \in \mathcal{U}(8)$ where $\mathcal{U}^{(1,1)}(8) = \mathcal{U}^{(2,2)}(8) \in \mathcal{O}(4)$ $\mathcal{U}^{(1,2)}(8) = \mathcal{U}^{(2,1)}(8) = \mathbf{O}_4$
$p, q \in \mathcal{S}_M$	$p = q$: $\mathbf{F}_{q,q} \in \mathcal{R}^{8\times 8}$ $\mathbf{F}_{q,q}^{(n,n)} = \alpha_{q,n}\mathbf{I}_4$ $\mathbf{F}_{q,q}^{(n,d)} \in \mathcal{V}(4)$, $1 \le n, d \le 2$ $p \neq q$: $\mathbf{F}_{p,q} \in \mathcal{R}^{8\times 8}$ $\mathbf{F}_{p,q}^{(n,d)} \in \mathcal{V}(4)$, $1 \le n, d \le 2$ where $\mathcal{V}^{(1,1)}(4) = \mathcal{V}^{(2,2)}(4) = c_1\mathbf{I}_2$ $\mathcal{V}^{(1,2)}(4) = -\mathcal{V}^{(2,1)}(4) = c_2\mathbf{I}_2$	$p = q$: $\mathbf{F}_{q,q} \in \mathcal{R}^{16N\times 16N}$ $\mathbf{F}_{q,q}^{(n,n)} = \alpha_{q,n}\mathbf{I}_{16}$ $\mathbf{F}_{q,q}^{(n,d)} \in \mathcal{V}(16)$, $1 \le n, d \le N$ $p \neq q$: $\mathbf{F}_{p,q} \in \mathcal{R}^{16N\times 16N}$ $\mathbf{F}_{p,q}^{(n,d)} \in \mathcal{V}(16)$, $1 \le n, d \le N$ where $\mathcal{V}^{(1,1)}(16) = \mathcal{V}^{(2,2)}(16) = c_1\mathbf{I}_8$ $\mathcal{V}^{(1,2)}(16) = -\mathcal{V}^{(2,1)}(16) = c_2\mathbf{I}_8$
$p \in \mathcal{S}_D$, $q \in \mathcal{S}_M$	$\mathbf{F}_{p,q} \in \mathcal{R}^{4\times 8}$ $\mathbf{F}_{p,q}^{(1,n)} \in \mathcal{O}(4)$, $1 \le n \le 2$	$\mathbf{F}_{p,q} \in \mathcal{R}^{8\times 16N}$ $\mathbf{F}_{p,q}^{(1,n)} \in \mathcal{D}(8,16)$, $1 \le n \le N$, where $\mathcal{D}^{(1,1)}(8,16) = \mathcal{D}^{(1,2)}(8,16) = \mathcal{D}^{(2,3)}(8,16)$ $= -\mathcal{D}^{(2,4)}(8,16) \in \mathcal{O}(4)$ $\mathcal{D}^{(2,2)}(8,16) = \mathcal{D}^{(1,3)}(8,16) = \mathcal{D}^{(1,4)}(8,16)$ $= -\mathcal{D}^{(2,1)}(8,16) \in \mathcal{O}(4)$

In (28), $\mathbf{a}_{q,j}^T(n)$ is the nth row of $\mathbf{A}_{q,j}$, whereas the OSTBC ST modulation matrix $\mathbf{A}_{p,i}$ is directly determined by the construction of the orthogonal codewords (see [21]). For the $p = 2$ with real-valued constellation case (assuming $n = 1$ without loss of generality), we have

$$\mathbf{A}_{p,1} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \quad \mathbf{A}_{p,2} = \begin{bmatrix} 0 & 1 \\ -1 & 0 \end{bmatrix}; \quad \mathbf{A}_{q,1} = \begin{bmatrix} 1 & 0 \\ 0 & 0 \end{bmatrix}, \quad \mathbf{A}_{q,2} = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix}; \tag{30}$$

From (28)-(30), it can be easily checked that $f_{p,q}^{(1,1)}(1) = f_{p,q}^{(2,2)}(1)$ and $f_{p,q}^{(2,1)}(1) = -f_{p,q}^{(1,2)}(1)$, which immediately implies that $\tilde{\mathbf{A}}_p^T \tilde{\mathbf{H}}_p^T \tilde{\mathbf{H}}_{q,n} \tilde{\mathbf{A}}_q^{(n)}$ belongs to $\mathcal{O}(2)$. For $P \in \{3, 4\}$, the results can be shown by first identifying the respective families of $\mathbf{A}_{p,i}$'s and $\mathbf{a}_{q,j}^T(n)$'s and by following the same procedures.

APPENDIX II
MATCHED-FILTERED CHANNEL MATRIX: COMPLEX-VALUED CONSTELLATION CASE

We present the analogue results of Prop. 3.1 when the complex-valued symbols are used. It is noted that, in this case, the available code rate depends on the number of transmit antennas N [21]: full-rate code for $N = 2$ (hence $K = P = 2$), and half-rate code for $N = 3$ or 4 ($K = 8$, $P = 4$). The results are summarized in Table III. The derivations are to first identify in each case the ST modulation matrices; each signature matrix is then computed to verify the result (the detail is referred to [12]. Note that in Table III, $\mathcal{A}^{(i,j)}$ is the (i, j)th block submatrix of $\mathcal{A}$ with proper matrix dimension.

APPENDIX III
PROOF OF LEMMA 4.1

The proof is based on a crucial fact about the orthogonal designs [21]. Specifically, it can be checked by analytic computations that, if $\mathbf{M}_1, \mathbf{M}_2 \in \mathcal{O}(P)$, then so are $\mathbf{M}_1 + \mathbf{M}_2$ and $\mathbf{M}_1\mathbf{M}_2$, that is,

Fact 1: The set $\mathcal{O}(P)$ is closed under addition and multiplication. Moreover, for any $\mathbf{M}_1 \in \mathcal{O}(P)$, it is easy to see that $\mathbf{M}_1 + \mathbf{M}_1^T = \gamma \mathbf{I}_P$ for some γ.

Based on Fact 1, the result can be shown by induction on L.

For the $L = 1$ case, the result is obvious since $\mathbf{F} = \alpha \mathbf{I}_P$. Assume that the result is true for an arbitrary $L > 1$, that is, $\mathbf{F} \in \mathcal{F}(L)$ implies $\mathbf{F}^{-1} \in \mathcal{F}(L)$ for such an L. We have to check that $\mathbf{F}^{-1} \in \mathcal{F}(L+1)$ whenever $\mathbf{F} \in \mathcal{F}(L+1)$. To see this, let us partition an arbitrary $\mathbf{F} \in \mathcal{F}(L+1)$ as

$$\mathbf{F} = \left[\begin{array}{c|c} \mathbf{A} & \mathbf{B} \\ \hline \mathbf{B}^T & \mathbf{D} \end{array}\right], \tag{31}$$

where $\mathbf{A} \in \mathcal{R}^{PL\times PL}$, $\mathbf{B} \in \mathcal{R}^{PL\times P}$, and $\mathbf{D} \in \mathcal{R}^{P\times P}$. We note that, since $\mathbf{F} \in \mathcal{F}(L+1)$, we have *(a)* $\mathbf{A} \in \mathcal{F}(L)$ and hence $\mathbf{A}^{-1} \in \mathcal{F}(L)$ by assumption, *(b)* $\mathbf{D} = c\mathbf{I}_P$ for some scalar c, and *(c)* if we write $\mathbf{B} = \left[\mathbf{B}_1^T \cdots \mathbf{B}_L^T\right]^T$, where $\mathbf{B}_i \in \mathcal{R}^{P\times P}$, then we have $\mathbf{B}_i \in \mathcal{O}(P)$. Let us similarly write

$$\mathbf{F}^{-1} = \left[\begin{array}{c|c} \bar{\mathbf{A}} & \bar{\mathbf{B}} \\ \hline \bar{\mathbf{B}}^T & \bar{\mathbf{D}} \end{array}\right], \tag{32}$$

where $\bar{\mathbf{A}} \in \mathcal{R}^{PL\times PL}$, $\bar{\mathbf{B}} \in \mathcal{R}^{PL\times P}$, and $\bar{\mathbf{D}} \in \mathcal{R}^{P\times P}$. To show that $\mathbf{F}^{-1} \in \mathcal{F}(L+1)$, it suffices to check that *(1)* $\bar{\mathbf{A}} \in \mathcal{F}(L)$, *(2)* $\bar{\mathbf{B}} = \left[\bar{\mathbf{B}}_1^T \cdots \bar{\mathbf{B}}_L^T\right]^T$, where $\bar{\mathbf{B}}_i \in \mathcal{R}^{P\times P}$, is such that each $\bar{\mathbf{B}}_i \in \mathcal{O}(P)$, and *(3)* $\bar{\mathbf{D}} = d\mathbf{I}_P$ for some scalar d. Properties *(1)-(3)* can be shown based on the inversion formula for block matrices, and by exploiting Fact 1. The detail is referred to [24].

APPENDIX IV
GROUP-WISE V-BLAST DETECTION: COMPLEX-VALUED CONSTELLATION CASE

The group-wise V-BLAST detection for complex-valued constellation case can similarly be established if we observe from Table III that the MF channel matrix $\mathbf{F}$ does consist of orthogonal design block submatrices. By going through essentially the same arguments as what we have done in the real symbol case, we can similarly derive a block based ZF/MMSE V-BLAST detector ($2P$ real symbols are detected for an STBC user and $2K$ real symbols for an antenna of an SM user per iteration). Since the derivations are basically the same as those in the real symbol case, we will not go them through but only refer the detail to [12].

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