Transmitter Precoding with Reduced-Complexity Soft Detection for MIMO Systems

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Abstract— **We present a precoded reduced-complexity soft detection (PRCSD) algorithm for multiple-input multiple-output (MIMO) systems. The linear operations at both transmit and receive sides based on complex Householder transform convert the MIMO channel to be multiple-diagonal, spatially partial-response-like, so that error propagation is alleviated when applying reduced-complexity soft detection at the receiver. The transform results in unitary precoding and feedforward matrices so that neither transmit power boost nor noise enhancement is present. Performance analysis based on pairwise error probability (PEP) shows that PRCSD achieves larger diversity advantage than existing precoding and multiple-beamforming (MB) schemes, which basically attempt to transmit signals through diagonal independent sub-channels and thus may suffer a diversity loss. PRCSD can achieve full diversity as maximum likelihood (ML) detection in some scenarios while reducing complexity significantly.**

Index Terms— **Linear predictive coding.**

I. Introduction

EVER since the emergence of the spatial multiplexing (SM) system known as vertical Bell Laboratories' Layered Space-Time (V-BLAST) [1], intense research works have been directed to find detection schemes to improve the performance upon the original successive interference cancellation (SIC) algorithm while maintaining a practical level of complexity. In [2], we have provided a comprehensive review of existing detection schemes, and proposed a reduced-complexity soft detection (RCSD) algorithm based on the spatially constrained-delay maximum a posterior (MAP) approach as well as soft decision feedback (SDF). RCSD is a generalized SIC scheme which subsumes V-BLAST detection and maximum likelihood (ML) detection as two special cases, and can achieve excellent performance/complexity tradeoff compared with other detection schemes like sphere decoding [3], as discussed in [2].

If channel state information (CSI) at the transmit side (CSI-TX) is available, then some pre-cancellation of interference can be done at the transmitter to assist detection. For example, Tomlinson-Harashima precoding (THP) [4] [5] has been applied by Fisher et al. [6] to multiple-input multiple-output (MIMO) channels to remove the off-diagonal interference, so that very simple detection is required at the receiver. Another approach widely known as beamforming is to apply singular-value decomposition (SVD) of the channel matrix, so that parallel channels (eigenmodes) are generated [7]. The transmit power can be allocated to each eigenmode dynamically according to some criteria (see [8] and the references therein), a method that can be unified into the convex optimization problem as stated in [7].

Most precoding schemes introduced above try to diagonalize the MIMO channel matrix and transmit signals through scalar channels, so that only simple linear detection is required at the receiver. However, from the diversity point of view, these schemes could suffer from significant diversity loss even with power allocation. The present work started from the premise that in practical systems where the receiver can tolerate some additional detection complexity, one should be able to design a linear precoder to generate a banded diagonal, spatially partial-response-like channel, a channel that is well-suited for a reduced-complexity soft detection algorithm. As a consequence, error propagation in RCSD is suppressed while a significant diversity gain is achieved. We shall call this scheme precoded-RCSD (PRCSD). The precoding and feedforward matrix can be obtained by performing complex Householder transform [9] on the MIMO channel matrix. Performance analysis based on the pairwise error probability (PEP) will show that PRCSD achieves excellent diversity advantages compared to existing schemes, and perform comparably to ML detection in some scenarios.

This paper is organized as follows: Section II describes the system under investigation, and PRCSD is proposed in Section III. Performance analysis is given in Section IV, while simulation in Section V validates the analysis. Finally, conclusions are drawn in Section VI.

II. System Model

The system of our interest is an extension of bit-interleaved coded modulation (BICM) [10] to the multiple-antenna scenario via spatial multiplexing, which will be referred to as BICM-SM. See Fig. 1. We consider N_T transmit and N_R receive antennas, with $N_T \leq N_R$. We shall focus on an $N \times N$ MIMO system, however, since there is always an effective $N \times N$ system resulting after proper signal decomposition, where $N = N_T$. The coded bit sequence, $\mathbf{c} \triangleq [\cdots, c_{k'}, \cdots]$, is interleaved to another bit sequence $\mathbf{d} \triangleq [\cdots, \mathbf{d}_k, \cdots]$, where $\mathbf{d}_k \triangleq [d_k^{1,1}, \cdots, d_k^{1,m}, \cdots, d_k^{N,1}, \cdots, d_k^{N,m}]$. The permutation $\pi : c_{k'} \to d_k^{t,l}$, $1 \leq k' \leq K, 1 \leq t \leq N, 1 \leq l \leq m, 1 \leq k \leq K/mN$, is determined by the interleaving mechanism, where K is the codeword length, and m is the number of bits per symbol. The bit sequence $\mathbf{d}$ is then mapped and multiplexed into a symbol sequence $\mathbf{a}$, which can be further transformed into another sequence $\mathbf{x}$ if CSI-TX is available. Obviously, $\mathbf{x} = \mathbf{a}$ when no CSI-TX is available. At the receiver, an

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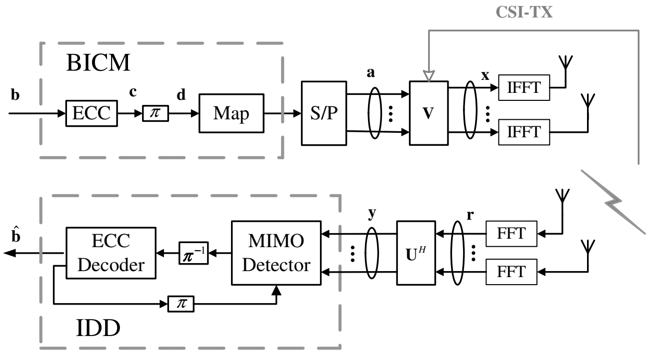

Fig. 1. System model of BICM-SM with IDD, with CSI at TX.

iterative decoding and demodulation (IDD) algorithm with extrinsic information exchanged between the decoder and MIMO detector is employed to improve the performance.

Assuming a flat fading channel, the received signal during the kth time slot can be expressed as

$$\mathbf{r}_k = \sqrt{E_s}\mathbf{H}_k \cdot \mathbf{x}_k + \boldsymbol{\eta}_k \tag{1}$$

where $\boldsymbol{\eta}_k$ is a sequence of Gaussian noise with zero mean and variance $N_0/2$, E_s is the average symbol energy, and the MIMO channel at time k is modelled as $\mathbf{H}_k$ with independent identically-distributed (i.i.d.) Gaussian components. In the sequel the time index k will be omitted unless otherwise noted to avoid confusion. Note that the representation (1) is also applicable to frequency-selective channels, if orthogonal frequency division multiplexing (OFDM) is used to convert the channel into parallel flat fading channels. An OFDM system can be realized through IFFT/FFT operations at the TX and RX, respectively, as included in Fig. 1.

III. Precoding and Detection for MIMO Systems

A RCSD algorithm has been introduced in [2]. By generalizing SIC to multiple layers, the detection process of a particular layer symbol will benefit from other layer symbols through soft decision feedback (SDF). If CSI-TX is available, then some manipulations can be done at both TX and RX to make the channel more suitable for the use of RCSD.

Suppose the channel can be decomposed as $\mathbf{H} = \mathbf{U}\mathbf{S}\mathbf{V}^H$, where both $\mathbf{U}$ and $\mathbf{V}$ are unitary matrices. Then we can apply $\mathbf{V}$ and $\mathbf{U}^H$ at TX and RX, respectively, to generate the effective channel $\mathbf{S}$. Unlike the multiple-beamforming (MB) scheme [7], [13], wherein an effectively diagonal channel is formed, $\mathbf{S}$ is a banded diagonal, spatially partial-response-like channel matrix generated through the complex householder transform with the width of the band is parameterized by $(\tau+1)$. Here, τ determines the complexity/performance trade-off of RCSD [2]. According to [2], the complexity level of RCSD is proportional to $M^{\tau+1}$, with M denoting the symbol constellation size. In this paper, we focus on PRCSD with $\tau > 0$; when $\tau = 0$, PRCSD reduces to the scheme of MB with an understanding that MB is achieved through SVD since $\mathbf{H}$ can not be diagonalized through the householder transform.

Substituting $\mathbf{x} = \mathbf{V}\mathbf{a}$ into (1) and letting $\mathbf{y} = \mathbf{U}^H\mathbf{r}$, a portion of the received signal is defined as:

$$\mathbf{y}^t = \sqrt{E_s}\mathbf{S}^t \cdot \mathbf{a}^t + \mathbf{n}^t, \quad \tau \le t \le N \tag{2}$$

where $\mathbf{y}^t = [y_{t-\tau}, y_{t-\tau+1} - s_{(t-\tau+1)(t+1)}\hat{a}_{t+1}, \cdots, y_t - \sum_{i=t+1}^{\min(t+\tau,N)} s_{ti}\hat{a}_i]^T$, $\mathbf{a}^t = [a_{t-\tau}, \cdots, a_t]^T$, $\mathbf{n}^t = [n_{t-\tau}, \cdots, n_t]^T$, $\mathbf{n} = \mathbf{U}^H\boldsymbol{\eta}$, and $\mathbf{S}^t$ is the corresponding sub-matrix of $\mathbf{S}$:

$$\mathbf{S}^t = \begin{bmatrix} s_{(t-\tau)(t-\tau)} & \cdots & s_{(t-\tau)t} \\ & \ddots & \vdots \\ \mathbf{0} & & s_{tt} \end{bmatrix}. \tag{3}$$

It can be seen that error propagation will be reduced since the current-layer symbol estimate will not cause interference in some of the upper layers.

The well-known householder transform [9] needs to be modified to a complex version, but the details of the modification are not shown here due to the space constraint. This linear precoding happens to be part of the SVD process, which shapes $\mathbf{H}$ into a banded diagonal form to speed up the process of finding its eigenvalues. However, the purpose of precoding here is to make the effective channel more suitable for RCSD processing. From the complexity point of view, the proposed PRCSD scheme has lower complexity than beamforming at the transmitter, but requires higher complexity at the receiver due to the need for nonlinear joint detection.

IV. Performance Analysis

The extrinsic soft output of MIMO detector, i.e., $f(\tilde{d}_k^{t,l}) \triangleq \ln P(d_k^{t,l} = \tilde{d}_k^{t,l})$, for the lth bit within the kth symbol at the tth layer, is computed by (4), as shown in the top of the next page, where $\mathcal{A}(\tilde{d}_k^{t,l})$ denotes the set of all possible $\mathbf{a}_k^t$ whose corresponding bit $d_k^{t,l} = \tilde{d}_k^{t,l} \in \{0, 1\}$, and $P_a(\tilde{d}_k^{i,j})$ is the *a priori* probability of $d_k^{i,j}$ being $\tilde{d}_k^{i,j}$. Note that according to our notational convention, $d_k^{t,l}$ is a random variable (RV) whereas $\tilde{d}_k^{t,l}$ represents a particular realization of the RV.

A. Uncoded Case

In this case, bit-interleaving is not used and the uncoded bits are directly mapped to constellation symbols and demultiplexed to different antennas for transmission. The decision based on (4) can be made memoryless in time and the a priori bit probabilities are assumed to be equal. Therefore, (4) can be approximated and simplified to (omitting the time index k)

$$f(\tilde{d}^{t,l}) \approx \max_{\tilde{\mathbf{a}}^t \in \mathcal{A}(\tilde{d}^{t,l})} \left(-\frac{1}{N_0} \left\| \mathbf{y}^t - \sqrt{E_s}\mathbf{S}^t\tilde{\mathbf{a}}^t \right\|^2 \right). \tag{5}$$

Due to the overall decision feedback structure of (P)RCSD, the error rate performance will be limited by the error probability of the bits in the sub-channel being decoded first, i.e., $P(d^{N,l} \to \hat{d}^{N,l}|\mathbf{S}^N), 1 \le l \le m$, where $d^{N,l}$ and $\hat{d}^{N,l}$ represent opposite bits. Assuming $\mathbf{a}^N$ and $\hat{\mathbf{a}}^N$ differ only at a_N, the Chernoff bound of this PEP can be written as [11]

$$P(d^{N,l} \to \hat{d}^{N,l}|\mathbf{S}^N) \le \exp\left(-\frac{E_s}{4N_0} \cdot d_{min}^2 \cdot q_N^2 \right) \tag{6}$$

where d_{min}^2 is the minimum Euclidean distance between two constellation points, and $q_t^2, 1 \le t \le N$, is the squared norm of the right-most column of $\mathbf{S}^t$, i.e., $q_t^2 = \left\| [s_{(t-\tau)t}, \cdots, s_{tt}]^T \right\|^2$. Based on the computations of $\mathbf{S}^t$, we know $q_N^2 \sim \chi^2_{2(\tau+1)}$,

$$f(\tilde{d}_k^{t,l}) = \ln \sum_{\tilde{\mathbf{a}}_k^t \in \mathcal{A}(\tilde{d}_k^{t,l})} \exp\left(-\frac{1}{N_0}\left\|\mathbf{y}_k^t - \sqrt{E_s}\mathbf{S}_k^t \tilde{\mathbf{a}}_k^t\right\|^2 + \sum_{\substack{(i,j)=(t-\tau,1)\\(i,j)\neq(t,l)}}^{(t,m)} \ln P_a(\tilde{d}_k^{i,j})\right) \tag{4}$$

TABLE I

DISTRIBUTION OF $q_{k,t}^2$, AND THE VALUE OF δ_t

	RCSD		PRCSD	
	$1\leq t\leq \tau+1$	$\tau+1<t\leq N$	$1\leq t\leq\tau$	$\tau<t\leq N$
$q_{k,t}^2$	χ_{2N}^2	$\chi^2_{2(N-t+\tau+1)}$	χ_{2N}^2	$\chi^2_{2(2N-2t+\tau+1)}$
δ_t	N	$N-t+\tau+1$	N	$2N-2t+\tau+1$

for both RCSD and PRCSD schemes. Thus, the unconditional upper bound is obtained as

$$\begin{aligned} P(d^{N,l} \to \hat{d}^{N,l}) &\leq E_{\mathbf{S}^N}\left[\exp\left(-\frac{E_s}{4N_0}\cdot d_{min}^2 \cdot q_N^2\right)\right] \\ &= \left(1+\frac{E_s}{4N_0}d_{min}^2\right)^{-(\tau+1)}. \end{aligned} \tag{7}$$

It is seen that in terms of diversity advantage [11], which is indicated by the exponent of (7), both RCSD and PRCSD achieve the same order of $(\tau+1)$. However, PRCSD has less error propagation, which will manifest to some extent as coding advantage, as shown through simulation results next. The presence of other possible error events $(\mathbf{a}^N, \hat{\mathbf{a}}^N)$ having more than one symbol error would not change the diversity order in (7), since q_N^2 would still be $\chi^2_{2(\tau+1)}$-distributed (albeit with a potentially larger variance).

B. BICM-SM

This is the system depicted in Fig. 1. Now applying IDD, the a priori bit probabilities in (4) can be updated from the extrinsic output of the decoder. To gain better insight, we also assume that a "genie demapper" [12] is deployed, which has perfect a priori knowledge, i.e., $P_a(d_k^{t,l} = \tilde{d}_k^{t,l}) = 1$ for a particular $\tilde{d}_k^{t,l}$. In practice, this can be reasonably achieved given a sufficient number of IDD iterations. Then (4) again simplifies to

$$f(\tilde{d}_k^{t,l}) = -\frac{1}{N_0}\left\|\sqrt{E_s}\mathbf{S}_k^t(\mathbf{a}_k^t - \tilde{\mathbf{a}}_k^t) + \mathbf{n}_k^t\right\|^2 \tag{8}$$

where $\mathbf{a}_k^t$ and $\tilde{\mathbf{a}}_k^t$ are two symbol sets with the corresponding bits differing only at $d_k^{t,l}$. As in (6), the Chernoff bound of the bit error probability that the bit $d_k^{t,l}$ is erroneously taken as $\hat{d}_k^{t,l}$ is

$$P(d_k^{t,l} \to \hat{d}_k^{t,l}|\mathbf{S}_k^t) \leq \exp\left(-\frac{E_s}{4N_0}\cdot d_{min}^2 \cdot q_{k,t}^2\right) \tag{9}$$

where $q_{k,t}^2 = \left\|[s_{(t-\tau)t}, \cdots, s_{tt}]_k^T\right\|^2$ is still the squared norm of the right-most column of $\mathbf{S}_k^t$. The distribution of $q_{k,t}^2$ is as shown in Table I.

Consider the worst case scenario where two bit sequences $\mathbf{d}$ and $\hat{\mathbf{d}}$ have the minimum Hamming distance d_H. After some similar manipulations to obtain the Chernoff bound, the conditional PEP of $(\mathbf{d}, \hat{\mathbf{d}})$ can be expressed as

$$\begin{aligned} P(\mathbf{d}\to\hat{\mathbf{d}}|\mathbf{S}) &= P\left(\sum_{k,t,l}^{d_H} f(\hat{d}_k^{t,l}) > \sum_{k,t,l}^{d_H} f(d_k^{t,l})\right) \\ &\leq \exp\left(-\frac{E_s}{4N_0}\cdot d_{min}^2 \cdot \sum_{k,t}^{d_H} q_{k,t}^2\right) \end{aligned} \tag{10}$$

where the interleaver is assumed to spread the coded bits so that distinct k' of $(\mathbf{c}, \hat{\mathbf{c}})$ result in distinct indices (k,t) of $(\mathbf{d}, \hat{\mathbf{d}})$, so the summation in (10) is conducted over d_H items.

1) Quasi-static Channel: If the channel is assumed to be static for the whole codeword length, i.e., $\mathbf{H}_k$ is constant irrespective of k, then only N out of all $q_{k,t}^2$'s in (10) are distinct RVs. By again omitting the time index k, we obtain the PEP of $(\mathbf{d}, \hat{\mathbf{d}})$ as

$$\begin{aligned} P(\mathbf{d}\to\hat{\mathbf{d}}) &\leq E_{\mathbf{S}}\left[\exp\left(-\frac{E_s}{4N_0}\cdot d_{min}^2\cdot\sum_{t=1}^{N}\alpha_t q_t^2\right)\right] \\ &\leq \left(\frac{E_s}{4N_0}\right)^{-\sum_{t=1}^N \delta_t} \cdot \prod_{t=1}^{N}\left(\alpha_t d_{min}^2\right)^{-\delta_t} \end{aligned} \tag{11}$$

where α_t denotes the multiplicity of q_t^2 in (10), $\sum_{t=1}^N \alpha_t = d_H$. The exponent δ_t is computed based on the distribution of $q_{k,t}^2$, and is also shown in Table I.

From (11), it is seen that the diversity advantage in this case is $\beta = \sum_{t=1}^N \delta_t$, and the other product term in (11) can be considered as the coding advantage. Furthermore, assuming distinct bits are spread over all layers by the interleaver so that $\alpha_t \geq 1$ for $1\leq t\leq N$, the diversity advantages achieved by RCSD and PRCSD are $\beta_{RCSD} = \frac{N^2-\tau^2+N-\tau+2N\tau}{2}$ and $\beta_{PRCSD} = N^2$, respectively. It is seen that PRCSD always achieves the full diversity of N^2, whereas RCSD achieves the diversity order between $\frac{N(N+1)}{2}$, that of V-BLAST detection when $\tau=0$, and N^2, that of ML when $\tau = N-1$. It has been shown that BICM with MB could also possibly achieve the full diversity [13], if the similar assumption of distinct bits being spread over all eigenmodes by the interleaver holds. However, it will be shown through simulation that BICM-MB is still inferior to PRCSD in terms of coding advantage.

2) Fast Fading Channel: In this case all $q_{k,t}^2$ in (10) are independent RVs, and the time index k should not be omitted. We obtain the PEP of $(\mathbf{d}, \hat{\mathbf{d}})$ as

$$\begin{aligned} P(\mathbf{d}\to\hat{\mathbf{d}}) &\leq E_{\mathbf{S}}\left[\exp\left(-\frac{E_s}{4N_0}\cdot d_{min}^2\cdot\sum_{k,t}^{d_H} q_{k,t}^2\right)\right] \\ &\leq \left(\frac{E_s}{4N_0}\right)^{-\sum_{t=1}^N \alpha_t\cdot\delta_t} \cdot \prod_{t=1}^{N}\left(d_{min}^2\right)^{-\alpha_t\delta_t} \end{aligned} \tag{12}$$

where α_t and δ_t have the same definition as in the quasi-static case, since the distribution of $q_{k,t}^2$ is independent of k.

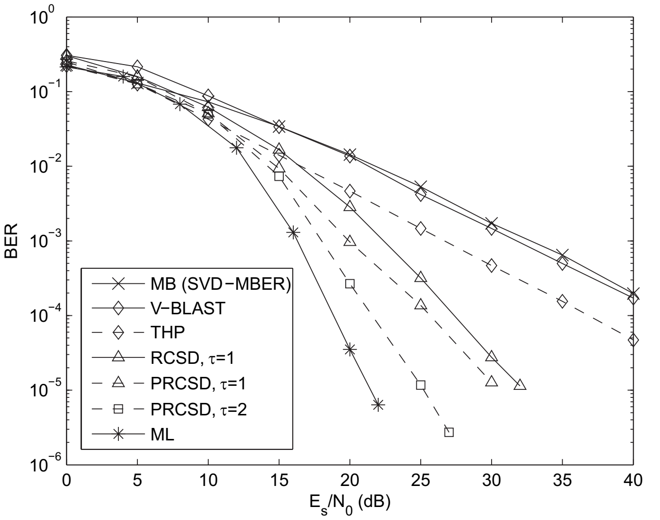

Fig. 2. Performance of detection schemes for uncoded system, flat fast fading channel.

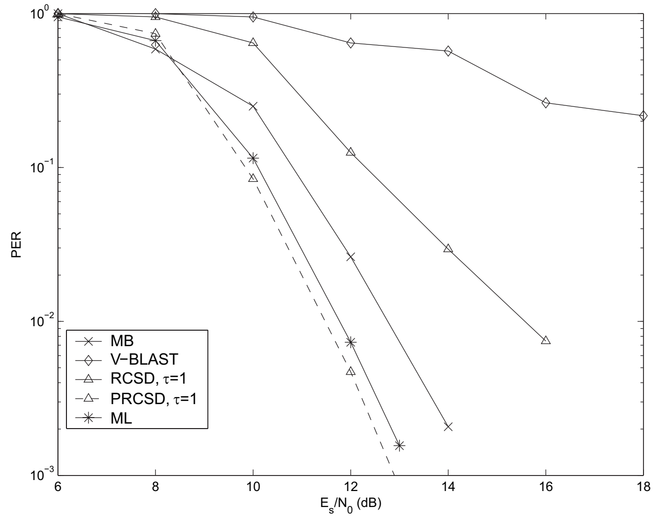

Fig. 3. Performance of detection schemes for BICM-SM system, flat quasi-static fading channel.

Accordingly, the diversity advantage achieved by (P)RCSD is $\beta = \sum_{t=1}^{N} \alpha_t \delta_t$, and α_t's are any set of non-negative integers satisfying $\sum_{t=1}^{N} \alpha_t = d_H$. For a special case of $\alpha_t = \frac{d_H}{N}$, which is also assumed in [13], we have $\beta = \frac{d_H}{N} \sum_{t=1}^{N} \delta_t$, and then $\beta_{RCSD} = \frac{d_H(N^2-\tau^2+N-\tau+2N\tau)}{2N}$ and $\beta_{PRCSD} = Nd_H$, respectively.

Extending the analysis to the frequency-selective channel will produce the similar result as (12), since when OFDM is applied the multiplicative signal model of (1) can still be employed with $\mathbf{H}_k$ denoting the equivalent channel matrix in the frequency-domain for the kth sub-carrier, and $\mathbf{H}_k$ can be assumed to be i.i.d. when the channel is highly frequency-selective. This has been investigated in [13], but the approximation of the singular-value distribution there is a bit optimistic; the expected level of diversity advantage may not be achieved in practice. Surely the achieved diversity order should be limited by the maximum diversity available, i.e., $\beta = \min\{\sum_{t=1}^{N} \alpha_t \delta_t, N^2 L\}$, where L denotes the order of time or frequency selectivity of the channel. In the following simulations, we assume there is abundant diversity available, so $\beta = \sum_{t=1}^{N} \alpha_t \delta_t$ is always possible to achieve.

V. Simulation Results

In this section simulation results are presented to verify the performance analysis results and compare the proposed PRCSD with some other schemes discussed in Section III. We focus on a 4×4 system with 16-QAM, for both uncoded case and the coded case with a standardized rate-1/2 (133, 171) convolutional code (CC). In the coded case, a pseudo-random interleaver of [14] with size 1024 is used for PRCSD. To be consistent, we still use the same interleaver used in [13] for BICM-MB, but we find that the performance of BICM-MB will not change much by switching to the interleaver of [14].

A. Uncoded Case

The decision is made memoryless at the output of the detector, so the diversity advantage is not affected regardless of whether the fast or the quasi-static fading channel is considered. Here we choose the fast fading channel. In Fig. 2, the simulated error rate performance of the schemes under comparison is shown. It is seen that V-BLAST detection, THP, and MB with power allocation under the minimum-BER (MBER) criterion, all have similar diversity advantages as expected, whereas THP has some coding gain over V-BLAST detection due to its absence of error propagation, and the other two show almost indistinguishable performance. RCSD and PRCSD with $\tau = 1$ achieve larger diversity advantage, while PRCSD also outperforms RCSD with a 2 dB gain due to reduced error propagation. Increasing to $\tau = 2$, the performance of PRCSD approaches that of ML ($\tau = 3$), clearly showing the performance/complexity tradeoff of PRCSD with different τ's.

B. BICM-SM, Flat Quasi-static Channel

In this case, a total diversity order of 16 is available. While in practice, designing an interleaver to guarantee the spreading mechanism as implied in (10) is unrealistic, we find that the interleaver of [14] can serve as such a desired interleaver in average sense. In Fig. 3, we show the packet error rate (PER), which is a more reasonable measure in quasi-static cases. As expected, RCSD with $\tau = 1$ achieves better diversity advantage than V-BLAST, and PRCSD has better diversity advantage than RCSD (both with $\tau = 1$) while outperforming the latter with a 4 dB coding gain at the PER of 10^{-2}. The schemes of PRCSD, ML, and MB achieve the same full diversity advantage, whereas PRCSD outperforms MB by about 1.5 dB coding gain. With CSI-TX being utilized, it is not surprising to see PRCSD showing comparable performance (even slightly better here) to ML, which would be optimal when no CSI-TX is available.

C. BICM-SM, Flat Fast Channel

The minimum Hamming distance of the CC (133, 171) is $d_H = 10$, and we assume that the channel fading varies every symbol period, so there is always enough diversity in time

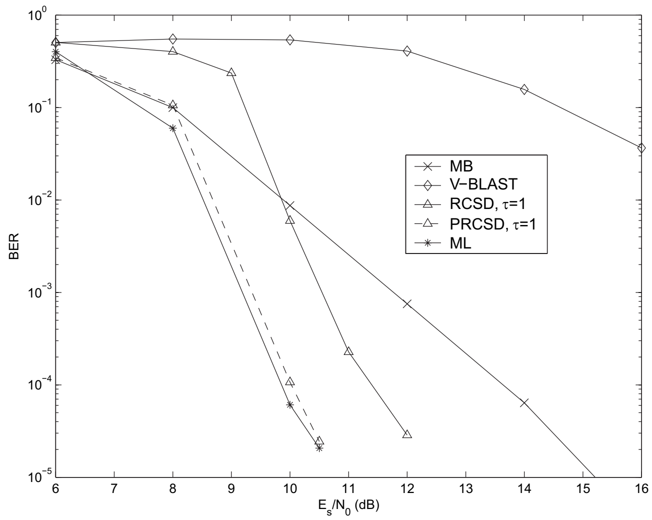

Fig. 4. Performance of detection schemes for BICM-SM system, flat fast fading channel.

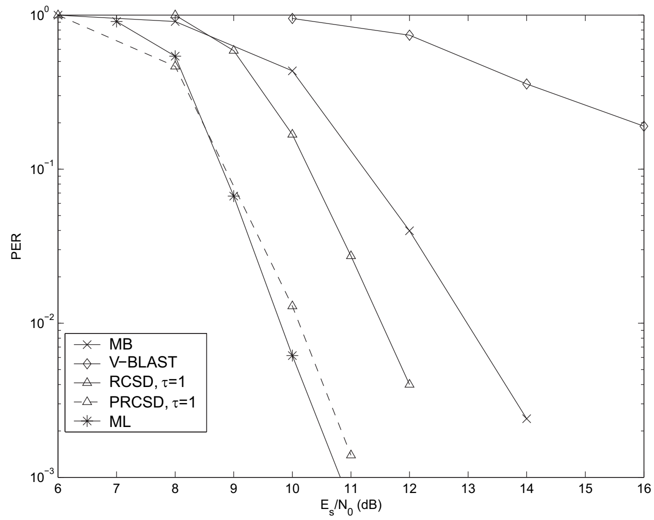

Fig. 5. Performance of detection schemes for BICM-SM system, frequency-selective block fading channel.

as well as in space to exploit. V-BLAST is expected to have the same diversity order as RCSD with $\tau = 0$, but due to its hard decision feedback (HDF) error propagation could be severe so the genie demapper assumption does not hold. In practice, V-BLAST detection may achieve smaller diversity than expected. Fig. 4 shows that PRCSD still outperforms RCSD with 1.5 dB and achieves comparable performance to ML detection. The MB scheme is supposed to achieve higher diversity than PRCSD and ML, if the assumptions of [13] are satisfied. However, the approximation about the singular-value distribution in [13] is somewhat optimistic, and the interleaver of [13] can not guarantee all the eigenmodes to be used either; so the actually achieved diversity and performance of MB are much worse.

D. BICM-SM, Frequency-selective Block Fading Channel

The frequency-selective channel is assumed to have an exponential power profile with the *rms* delay spread of 150 ns (corresponding to $L \approx 16$), and the channel is static for one OFDM symbol period. Fig. 5 shows similar relative performances of those detection schemes as in Fig. 4. PRCSD has coding advantage over RCSD, achieves higher diversity advantage than MB, and eventually performs comparably to ML.

VI. Conclusion

A linear transmitter precoding scheme is proposed to shape the MIMO channel to be banded-diagonal, spatially partial-response-like, so that RCSD can be applied at the receiver with reduced error propagation, achieving excellent performance while maintaining practical detection complexity. The operation can be realized through the complex Householder transform of the channel matrix, which results in unitary precoding and feedforward matrices so that neither transmit power boost nor noise enhancement is present. Performance analysis based on PEP shows that the PRCSD achieves higher diversity advantage than existing precoding and multiple-beamforming schemes, and can achieve full diversity as ML detection in some scenarios. Simulation results have validated the analysis.

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