Low Complexity Per-Antenna Rate and Power Control Approach for Closed-Loop V-BLAST

Hairuo Zhuang, *Student Member, IEEE*, Lin Dai, *Member, IEEE*, Shidong Zhou, *Member, IEEE*, and Yan Yao, *Senior Member, IEEE*

***Abstract*—Previous studies have shown that per-antenna rate and power control can greatly increase the data throughput of vertical Bell Labs layered space–time (V-BLAST), while an extra transmit antenna selection can provide additional diversity advantage. In this letter, we combine the transmit antenna selection with power and rate control of each antenna. We derive a simple criterion for minimum bit-error rate (BER) or minimum total transmit power when the data throughput is constant over time. Zero-forcing and zero-forcing successive interference cancellation detections are considered. For practical implementation, we also present a fast algorithm that gives near-optimal performance with very low complexity. Simulation results show that the proposed closed-loop BLAST outperforms the open-loop V-BLAST significantly in terms of BER performance, especially when the antennas exhibit strong fading correlations.**

***Index Terms*—Multiple-input–multiple-output (MIMO), rate and power control, transmit antenna selection, vertical Bell Labs layered space–time (V-BLAST).**

I. Introduction

MULTIPLE-INPUT–multiple-output (MIMO) antenna systems can provide enormous capacity by spatial multiplexing [1]–[4]. Recently, link adaptation techniques, in which transmission parameters such as modulation rate, coding rate, and power are dynamically adapted to the prevailing channel conditions, have been used in conjunction with MIMO techniques to achieve higher spectral efficiency and better transmission quality [5]. In [6], the attainable throughput with rate adaptation only was studied. Extended vertical Bell Labs layered space–time (V-BLAST) with both rate and power adaptation per antenna was proposed in [7] and [8], which show that the open-loop capacity can be achieved with conventional single-dimensional coding using simple per-antenna rate control. And by using power allocation, the capacity can be increased slightly further. At the same time, there are also works focusing on selecting a subset of the available transmit antennas based on some criteria to provide diversity advantage over fading channels [9], [10].

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H. Zhuang was with the Department of Electronic Engineering, Tsinghua University, Beijing 100084, China. He is now with the Department of Electrical and Computer Engineering, University of California at San Diego, San Diego, CA 92037 USA (e-mail: hzhuang@ucsd.edu).

L. Dai is with the Hong Kong University of Science and Technology, Kowloon, Hong Kong (e-mail: eedailin@ust.hk).

S. Zhou and Y. Yao are with the State Key Lab on Microwave and Digital Communications, Tsinghua University, Beijing 100084, China (e-mail: zhousd@wireless.mdc.tsinghua.edu.cn; yaoy@tsinghua.edu.cn).

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In this letter, we adopt the mechanism of both transmit antenna selection and per-antenna rate and power control, and design the criterion from a different perspective. As we know, a constant information rate is often desirable in some practical applications that require real-time transmission. Moreover, the hardware design complexity can be significantly reduced if the total information throughput remains constant. Thus, in this letter, we are motivated to improve the transmission quality of V-BLAST using simple per-antenna rate and power control when the information rate is predetermined. Unlike the extended V-BLAST proposed in [7] and [8], that tends to use all available transmit antennas to approach the open-loop capacity, selecting fewer antennas can, on one hand, increase the detection diversity order while, on the other hand, it does not utilize some of the available degrees of freedom. It is not immediately evident whether this is advantageous or not. Therefore, we derive a per-antenna rate and power control criterion to manage this tradeoff. Given the required spectral efficiency, the criterion judiciously selects a set of antennas and adjusts the rate and power of each one according to the channel status to minimize either bit-error rate (BER) or the total transmit power. Compared with transmit antenna selection schemes [9] that impose the same rate and power on the selected antennas, our scheme can more fully exploit the channel information as both rate and power of each active antenna are adapted to the channel state. Simulations show that our proposed closed-loop BLAST (C-BLAST) outperforms the open-loop V-BLAST significantly, especially when the channel is poorly conditioned (e.g., fading correlations between antennas).

We outline our models and assumptions in Section II, and derive the design criterion in Section III. Simulation results are presented and discussed in Section IV. Finally, Section V contains our concluding remarks.

II. System Model

The block diagram of our proposed C-BLAST that employs n_T transmit and n_R receive antennas is illustrated in Fig. 1. The source data is first demultiplexed into several independent substreams by a serial-to-parallel converter. These substreams are subsequently coded, modulated separately, and then transmitted simultaneously on the same frequency. The coding, modulation, and average transmit power of each substream are subject to the feedback information. We refer to a combination of specific coding and modulation as a *mode* [5]. Let M_i denote the mode of the ith substream. And the corresponding spectral efficiency is denoted by $R(M_i)$. In particular, $R(M_i) = 0$ means that the ith antenna is not used for transmission. Given the total required spectral efficiency R_t, we define the *mode vector* as $\mathbf{M} \triangleq$

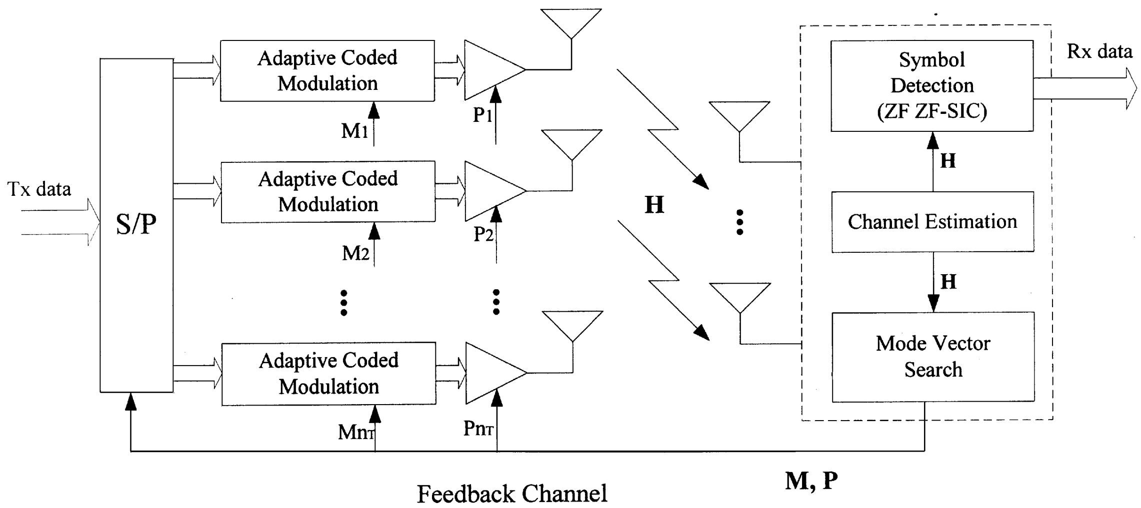

Fig. 1. Block diagram of C-BLAST transmitter and receiver.

$[M_1, M_2, \ldots, M_{n_T}]$ such that $R_t = \sum_{i=1}^{n_T} R(M_i)$. Likewise, for the total transmit power P_t, we define the *power allocation vector* as $\mathbf{P} \triangleq [P_1, P_2, \ldots, P_{n_T}]$ such that $P_t = \sum_{i=1}^{n_T} P_i$, where P_i denotes the average power radiated by the ith transmit antenna.

At the receiver, we assume that the channel is perfectly estimated. Two alternative symbol detection schemes are considered: zero-forcing (ZF) detection and its improved form zero-forcing successive interference cancellation (ZF-SIC) detection [4]. Besides symbol detection, the mode vector $\mathbf{M}$ and the power allocation vector $\mathbf{P}$ are decided according to the criterion derived in Section III. This information is then fed back to adjust the corresponding transmission parameters. We assume that there is neither delay nor error in the feedback channel.

We further assume that the channel is flat fading and quasi-static. The following discrete-time equivalent model is applied:

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n} \tag{1}$$

where $\mathbf{x}$ is an $n_j \times 1$ vector whose jth component represents the signal transmitted from the jth antenna. The received signal is an $n_R \times 1$ vector denoted by $\mathbf{y}$. $\mathbf{n}$ is an $n_R \times 1$ additive white complex Gaussian noise vector with variance σ^2. The channel is represented by an $n_R \times n_T$ matrix $\mathbf{H}$. The channel gains, modeled as zero-mean, unity variance complex Gaussian random variables, remain constant over a frame, but may vary from one frame to another. In the ideal rich scattering environment, the entries of $\mathbf{H}$ are mutually uncorrelated. However, in real scenarios, they may exhibit certain correlations.

III. Rate and Power Control Criterion

We denote the BER of the ith substream after detection by BER_i. Let $\mathrm{BER}_{\max} = \max\{\mathrm{BER}_i | i \in \mathcal{A}\}$. Since the overall BER performance is mainly dictated by the worst substream, in the following, our goal is to find a per-antenna rate and power control criterion that minimizes $\mathrm{BER}_{\max}$ under the constraints of spectral efficiency R_t and transmit power P_t.

We define the *active antenna set* as $\mathcal{A} \triangleq \{i | R(M_i) > 0, \forall i\}$. For ZF detection, $\mathcal{A}$ is an unordered set, since all substreams are detected simultaneously, while for ZF-SIC detection, $\mathcal{A}$ is an ordered set (e.g., $\{1,2,3\} \neq \{2,3,1\}$) whose order corresponds to the detection order. To perform ZF or ZF-SIC detection, a set of nulling vectors is obtained from $\mathbf{H}$ [3], [4]. We denote the nulling vector of the ith substream by $\mathbf{w}_i^{\mathcal{A}} (i \in \mathcal{A})$, which is a function of $\mathcal{A}$. The postdetection signal-to-noise ratio (SNR) of the ith substream is then given by [4]

$$\rho_i = \frac{P_i}{\sigma^2 \left\| \mathbf{w}_i^{\mathcal{A}} \right\|^2}, \quad i \in \mathcal{A}. \tag{2}$$

Thus, the total transmit power can be expressed as

$$\begin{aligned} P_t = &\sigma^2 \sum_{i \in \mathcal{A}} \left(\gamma(M_i, \mathrm{BER}_i) \left\| \mathbf{w}_i^{\mathcal{A}} \right\|^2 R(M_i) \right) \\ \geq &\sigma^2 \sum_{i \in \mathcal{A}} \left(\gamma(M_i, \mathrm{BER}_{\max}) \left\| \mathbf{w}_i^{\mathcal{A}} \right\|^2 R(M_i) \right) \end{aligned} \tag{3}$$

where $\gamma(M, \mathrm{BER})$ is a function of both mode M and BER, representing the E_b/N_0 required in additive white Gaussian noise (AWGN) for the target BER when the mode is M. Since $\gamma(M, \mathrm{BER})$ is generally a complicated expression that depends on specific coding and modulation schemes, directly resolving $\mathrm{BER}_{\max}$ from (3) is extremely difficult. As such, we need some approximation. Obviously, $\gamma(M, \mathrm{BER})$ is monotonously decreasing with BER. And at the same BER level, different modes generally require different E_b/N_0's. This difference can be expressed with a coefficient in terms of mode. At a BER range of interest, if we view this coefficient as a constant regarding BER, we can write it as the product of two decoupled functions

$$\gamma(M, \mathrm{BER}) \approx K(M) \cdot F(\mathrm{BER}) \tag{4}$$

where $K(M)$ is the coefficient in terms of mode and $F(\mathrm{BER})$ is a monotone decreasing function of BER. In some particular cases, $K(M)$ and $F(\mathrm{BER})$ can be obtained analytically. For example, the BER for an AWGN channel with M-ary quadrature amplitude modulation (MQAM) and a coset code, ideal coherent phase detection and maximum-likelihood (ML) decode can be approximated for a wide range of BER by the expression of the following form [14]:

$$\mathrm{BER} \approx C \exp\left(-\frac{1.5\gamma R(M) G_c}{2^{R(M)}-1}\right) \tag{5}$$

where G_c is the coding gain of the coset code and C is a constant. Therefore, we have

$$K(M) = \frac{2^{R(M)+1}-2}{3R(M)} \tag{6}$$

$$F(\mathrm{BER}) = G_c \cdot (\ln C - \ln \mathrm{BER}). \tag{7}$$

In general, as there is no simple closed-form BER approximation like (5), $K(M)$ and $F(\mathrm{BER})$ can be obtained numerically. The decomposition in (4) is usually a tight approximation for a wide range of BER at high SNRs, while at low SNRs, more discrepancy will be introduced as the variation range of BER increases.

By substituting (4) into (3), we have

$$F(\mathrm{BER}_{\max}) \leq \frac{P_t}{\sigma^2 \sum_{i\in\mathcal{A}} \left(\left\|\mathbf{w}_i^{\mathcal{A}}\right\|^2 K(M_i) R(M_i)\right)}. \tag{8}$$

Note that $F(\mathrm{BER}_{\max})$ is monotonously decreasing with $\mathrm{BER}_{\max}$, and equality holds if and only if $\mathrm{BER}_1 = \mathrm{BER}_2 = \cdots = \mathrm{BER}_n = \mathrm{BER}_{\max}$. Therefore, we arrive at our final criterion

$$\widehat{\mathcal{A}}, \widehat{\mathbf{M}} = \arg\min_{\mathcal{A},\mathbf{M}} \sum_{i\in\mathcal{A}} \left(\left\|\mathbf{w}_i^{\mathcal{A}}\right\|^2 K(M_i) R(M_i)\right). \tag{9}$$

And the corresponding power allocation vector $\mathbf{P} = [P_1, P_2, \ldots, P_n]$ satisfies

$$P_i = \begin{cases} P_t \dfrac{\left\|\mathbf{w}_i^{\mathcal{A}}\right\|^2 K(M_i) R(M_i)}{\sum_{k\in\mathcal{A}} \left(\left\|\mathbf{w}_k^{\mathcal{A}}\right\|^2 K(M_k) R(M_k)\right)}, & i \in \mathcal{A} \\ 0, & i \notin \mathcal{A} \end{cases} \tag{10}$$

which comes straight from the fact that the BERs of all substreams should be equal.

If maintenance of a fixed target BER is required, it is straightforward to calculate the minimum transmit power P_t by (8). Thus, the price for the constant data throughput at constant BER is the variation of total transmit power. Alternatively, if constant transmit power is required, BER may vary as the channel changes. This may be acceptable when there are less stringent requirements for the target BER so that the BER variation falls into acceptable ranges. We also note that the minimum value given in (9) quantifies the channel quality. An abnormally large value indicates serious channel degradation and may result in unacceptable BER performance or extremely high transmit power, even using our criterion. In this case, some other measures are required, such as reduction of information rate or a simple transmission cutoff. Such strategies may depend on practical systems and will not be discussed here.

Since all possible ordering must be considered when the criterion is based on ZF-SIC, an exhaustive search using the criterion in (9) needs to test $n_T \sum_{k=1}^{n_T} 1/k!$ possible antenna sets to find the optimal active antenna set, which is prohibitive for practical implementation. Fortunately, we find that the benefit brought about by ordering is rather small. This coincides with the conclusion in [7] that when ideal rate adaptation is adopted, ordering is not necessary. Therefore, at the cost of slight performance loss, we greatly reduce the complexity by using a fixed detection order (e.g., detecting according to the order of antenna indexes). In this case, the number of possible active antenna sets is reduced to $2^{n_T}-1$, which is as many as that based on ZF detection. Since the complexity is now growing exponentially, it is still too complicated to be conducted in real time. Thus, we provide a search algorithm that can further reduce $2^{n_T}-1$ to a number less than n_T with additional slight performance loss. The whole algorithm can be described compactly through the recursive procedure as follows.

initialization:

$$\mathcal{A}^{(n_T)} = \{1, 2, \ldots, n_T\}$$
$$\texttt{metric}^{(n_T+1)} = +\infty \quad k \leftarrow n_T$$

recursion:

obtain the nulling vector set $\{\mathbf{w}_i^{\mathcal{A}^{(k)}} | i \in \mathcal{A}\}$
find mode vector $\mathbf{M}^{(k)}$ that minimizes $\texttt{metric}^{(k)} = \sum_{i\in\mathcal{A}^{(k)}} \left(\left\|\mathbf{w}_i^{\mathcal{A}^{(k)}}\right\|^2 K(M_i) R(M_i)\right)$
if $\texttt{metric}^{(k)} > \texttt{metric}^{(k+1)}$ the algorithm ends, output $\mathcal{A}^{(k+1)}$ and $\mathbf{M}^{(k+1)}$ as the final results
else if $k = 1$ the algorithm ends, output $\mathcal{A}^{(1)}$ and $\mathbf{M}^{(1)}$ as the final results
else $\mathcal{A}^{(k-1)} \leftarrow \mathcal{A}^{(k)} - \{i\}$ where $i = \arg\max_{j\in\mathcal{A}^{(k)}} \left\|(\mathbf{H}^{\mathcal{A}^{(k)}})_j^+\right\|$ and $(\mathbf{H}^{\mathcal{A}^{(k)}})_j^+$ is the jth row of the pseudoinverse of $\mathbf{H}^{\mathcal{A}^{(k)}}$

$$k \leftarrow k-1$$

repeat the above recursion.

Obviously, under the worst condition, in which the optimal set contains only one antenna, the algorithm only requires testing n_T possible active antenna sets. Detailed simulations show that for most common system and channel configurations, our fast algorithm (with fixed ordering for ZF-SIC) only results in less than 0.4-dB degradation in the minimum metric, compared with the optimal exhaustive search (with optimal ordering for ZF-SIC). This loss is quite acceptable, considering the significant reduction in computation intensity.

IV. Simulation Results

We consider an uncoded system with four transmit antennas and six receive antennas. ZF-SIC with index-order detection and the fast search algorithm given above are adopted at the receiver in C-BLAST. At each transmit antenna, only two modes are adopted: $m_1 = \{\text{uncoded QPSK}\}$ and $m_2 = \{\text{uncoded 16QAM}\}$. Obviously, $R(m_1) = 2$ and $R(m_2) = 4$. Then $K(m_1) = 1$ and $K(m_2) = 2.5$, which are obtained analytically from (6). The total spectral efficiency R is constrained to be 8 b/s/Hz. For comparison purposes, the

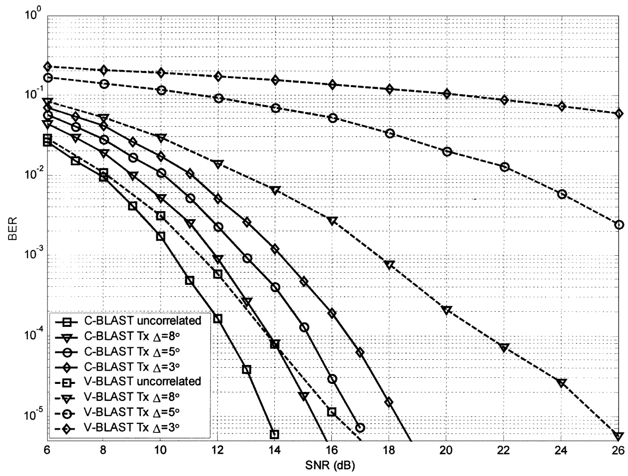

Fig. 2. Performance comparison between C-BLAST and V-BLAST with different angle spread at the transmitter.

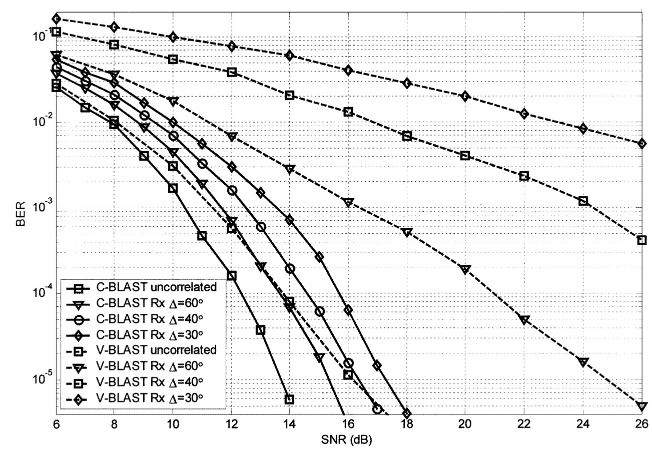

Fig. 3. Performance comparison between C-BLAST and V-BLAST with different angle spread at the receiver.

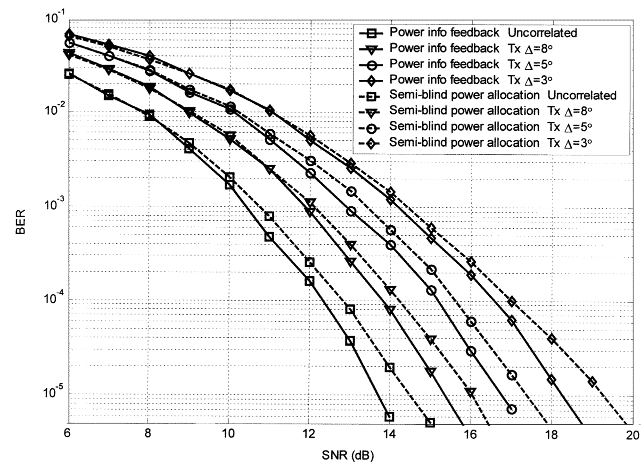

Fig. 4. Performance comparison between C-BLAST with power info feedback and C-BLAST with semiblind power allocation with different angle spread at the transmitter.

performance of the original open-loop V-BLAST with optimal detection ordering is also presented. Uncoded BER is obtained by averaging large volumes of channel realizations, while $\mathrm{SNR} = P_t/\sigma^2$ is fixed over time. We adopt the correlated channel model described in [11]–[13]. Linear arrangement of the antenna array is assumed at both the transmitter and the receiver, with the spacing between the neighboring antennas being $d_t = 4\lambda$ and $d_r = 0.5\lambda$, respectively. We also assume the "broadside" case as defined in [11], and the incoming waves are uniformly distributed in the angle spread Δ [12].

Fig. 2 shows the uncoded BER performance comparison between C-BLAST and V-BLAST. We assume that the receive antennas are mutually uncorrelated, while the correlation between transmit antennas varies with the angle spread Δ. As shown in the figure, C-BLAST outperforms V-BLAST significantly at high SNRs, as well as medium SNRs. Even when all transmit antennas are mutually uncorrelated, C-BLAST still gives about 1-dB gain over V-BLAST at a BER of 10^{-3}. In this scenario, C-BLAST tends to choose all four antennas with quaternary phase-shift keying (QPSK) modulation in most situations, as V-BLAST does. The relatively lower performance gain is achieved mainly due to the more efficient power allocation for each antenna. With the decrease of angle spread, the correlation between transmit antennas increases and the performance of V-BLAST degrades quickly to an unacceptable level, while the C-BLAST scheme still maintains fairly good performance. When $\Delta = 3^\circ$, C-BLAST almost always chooses the two antennas with the maximal separation, each of which adopts 16QAM modulation. This ties in with our intuition, since the two maximally separated antennas show the least correlation. However, V-BLAST always imposes the same data rate on all four antennas. The two neighboring antennas exhibit strong correlation and thus, these data streams can hardly be separated.

The effect of receive antenna correlations on C-BLAST and V-BLAST is depicted in Fig. 3, where the transmit antennas are assumed uncorrelated while the correlation between receive antennas varies with the angle spread. Similar to the effect of transmit antenna correlation, we observe that the impact of the correlation at the receiver on V-BLAST is significant, while C-BLAST still maintains quite acceptable performance. Similar observations can be found in systems with other antenna configurations.

Finally, we will evaluate the effect of imperfect power allocation with an extreme case, in which the mode vector $\mathbf{M}$ is fed back without the power allocation vector $\mathbf{P}$. The transmitter makes the power allocation decision based solely on $\mathbf{M}$. We refer it as a semiblind power allocation strategy. For example, we may simply allocate power proportional to the spectral efficiency, namely

$$P_i = P_t \frac{R(M_i)}{R_t}, \quad 1 \leq i \leq n_T. \tag{11}$$

Fig. 4 shows simulation results of this semiblind power allocation strategy with the same channel scenarios as in Fig. 2. With the reference to the curves with perfect power allocation feedback, it can be observed that the semiblind power allocation scheme does suffer from imperfect power allocation. However, the degradation is slight and within an acceptable range. For example, only less than 0.5-dB degradation is found at a BER of 10^{-3}. Therefore, with a slight performance loss, we greatly reduce the amount of feedback data, as the accurate power allocation requires a large amount of data.

V. Conclusion

In this letter, we focused on improving the transmission quality of V-BLAST by transmit antenna selection with rate and power control of each antenna. We derived a simple criterion that aims at minimizing either BER or the total transmit power while keeping the total data throughput unchanged. We also presented a near-optimal search algorithm with low complexity, which is more practical for implementation. Simulation results showed that the proposed C-BLAST outperformed V-BLAST significantly in terms of BER performance, especially in the presence of fading correlation between antennas. In our letter, we only derived the criterion based on ZF or ZF-SIC. However, some superior detection schemes, such as minimum mean-square error detection, are more preferable in some practical implementations. Deriving criteria based on such schemes is a topic for future work.

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