BER Improved Transmit Power Allocation for D-STTD Systems with QR-Based Successive Symbol Detection

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***Abstract*-We propose a BER improved power allocation scheme for D-STTD systems over i.i.d. Rayleigh fading channels under the QR-based successive detection framework. Instead of relying on BER under a fixed channel realization, the adopted design criterion is the mean BER (assuming there is no inter-layer error propagation) averaged with respect to the channel distribution. Such a design metric has two-fold advantages: (i) It is analytically tractable and is closely related to a block error probability upper bound when inter-layer error propagation occurs, and (ii) There is no need for repeated feedback of the instantaneous channel information. By exploiting a distinctive channel matrix structure unique to D-STTD systems we derive a closed-form approximate upper bound of the considered BER metric; through minimization of this bound an optimal power allocation scheme is obtained. Numerical simulation is used to illustrate the performance of the proposed method.**

Index Terms: **Space-time block codes; successive detection; power allocation; bit error rate; QR decomposition.**

I. INTRODUCTION

Spatially multiplexing multiple groups of orthogonal space-time block coded (STBC) signals is one key approach to realizing high-rate yet high-reliability wireless communications [2], [9], [12]. The double space-time transmit diversity (D-STTD) scheme [6], in which two Alamouti signal groups [1] are simultaneously transmitted, is the building block for such a system configuration. There have been many related research works reported for D-STTD systems, including antenna shuffling to combat channel spatial correlation [6], [8], adaptive modulation [3], and efficient low-complexity receiver designs [4], [5], [9].

QR-based successive symbol detector can strike a bit-error-rate (BER) performance balance between linear equalization and joint maximum-likelihood (ML) decoding. Such a scheme has been widely considered for signal detection in D-STTD and general multi-group STBC systems [2], [4]. This paper addresses the QR based successive signal detection problem for D-STTD systems, focusing on further symbol power loading for improving the BER performance. There have been many plausible performance measures for QR-based successive signal recovery [11], [13]-[15], depending on whether or not inter-layer error propagation is taken into account. The average BER with errorless front-layer decision feedback, although being merely a lower bound of the true mean error rate, remains simple to characterize and, moreover, is closely related to an upper bound of the block error probability when error-propagation occurs [11]: it thus serves as an efficient and meaningful performance metric accounting for the actual error rate outcome. Motivated by this fact and to also guarantee a performance improvement regardless of the instantaneous channel conditions, we propose to design the power loading weights for D-STTD transmission toward minimizing such a mean BER, averaged with respect to the channel distribution. Specifically, by exploiting a distinctive channel matrix structure of the D-STTD transmission we first derive an explicit formula of the associated QR-decomposition. Based on this result, we then derive an approximate closed-from upper bound of the considered BER metric. Through minimizing this bound the power allocation factors are obtained via numerical search. The proposed scheme depends only on the link SNR but not on the instantaneous channel gains: repeated channel state update via feedback is no longer needed. Simulation results show that the QR receiver combined with the proposed power allocation compares favorably with the zero-forcing (ZF) V-BLAST detector [10], in terms of both simulated BER and algorithm complexity.

II. PROBLEM STATEMENT

A. System Model and Basic Assumptions

We consider a D-STTD system with symbol power loading over a flat-fading channel. Following [6] the input-output relation, in terms of block signals, can be described as[a]

$$\mathbf{y}=\underbrace{\begin{bmatrix} h_{11} & h_{12} & h_{13} & h_{14} \\ h_{12}^{*} & -h_{11}^{*} & h_{14}^{*} & -h_{13}^{*} \\ h_{21} & h_{22} & h_{23} & h_{24} \\ h_{22}^{*} & -h_{21}^{*} & h_{24}^{*} & -h_{23}^{*} \end{bmatrix}}_{:=\mathbf{H}} diag\{p_1,\cdots,p_4\}\begin{bmatrix} s_1 & s_2 & s_3 & s_4\end{bmatrix}^T+\mathbf{n}\,, \tag{2.1}$$

where $\mathbf{y}$ is the received signal vector, $\mathbf{n}$ is a zero-mean complex white Gaussian noise with covariance $N_0\mathbf{I}_4$, $\mathbf{H}$ is the effective channel matrix with h_{ij} being the channel gain between the (j,i)th transmit-receive antenna pair, s_j is the symbol sent through the jth transmit antenna, and p_j is the

a. The symbols $(\cdot)^*$, $(\cdot)^T$, $(\cdot)^H$, $\mathbf{I}_n$, $\det(\cdot)$, and $diag\{\mathbf{x}\}$ denote, respectively, the complex conjugate, transpose, Hermitian, the $n\times n$ identity matrix, determinant, and the diagonal matrix with elements of the vector $\mathbf{x}$ on the main diagonal.

power loading factor for s_j satisfying the power normalization constraint $\sum_{i=1}^{4} p_i^2 = 4$. We assume that the channel gains h_{ij}'s are i.i.d. zero-mean complex white Gaussian with unit-variance.

B. BER of QR-Based Successive Detection

By factorizing $\mathbf{H} = \mathbf{QR}$, where $\mathbf{Q}$ is unitary and $\mathbf{R}$ is upper triangular, and multiplying (2.1) from the left by $\mathbf{Q}^H$, we have

$$\tilde{\mathbf{y}} := \mathbf{Q}^H \mathbf{y} = \mathbf{R} diag\left\{p_1, \cdots, p_4\right\}\left[s_1\ s_2\ s_3\ s_4\right]^T + \mathbf{Q}^H \mathbf{n}. \quad (2.2)$$

Since $\mathbf{R}$ is upper triangular, successive symbol detection through canceling the contributions of the previously detected components can be performed. We assume QPSK modulation is used with average symbol power equal to E_s; the generalization of our results to high-order constellations are straightforward by using related BER expressions in terms of Q function, as in [11]. As long as the symbol in each stage is correctly detected and, thus, there is no layer-wise error propagation, the space-time model (2.2) decouples into four independent links. The resultant average BER, given a channel realization $\mathbf{h}$, is

$$P_{b|\mathbf{h}} := \frac{1}{4}\sum_{i=1}^{4} Q\left(\sqrt{\rho} p_i \left|R_{ii}\right|\right), \quad (2.3)$$

where R_{ii} denotes the ith diagonal entry of $\mathbf{R}$, $\rho := E_s / N_0$ is the signal-to-noise ratio, and $Q(\cdot)$ is the Gaussian tail function. We emphasize that, when inter-layer error propagation occurs, $4P_{b|\mathbf{h}}$ is an upper bound for the block error probability [11]. This implies that, if $P_{b|\mathbf{h}}$ is small, the decision performance can be potentially improved even in the presence of inter-layer error propagation. Motivated by this fact and also to devise a solution irrespective of different channel realizations, we propose to design p_i's by minimizing $P_{b|\mathbf{h}}$ averaged with respect to the channel distribution, i.e.,

$$P_b := \frac{1}{4}\sum_{i=1}^{4}\int_0^{\infty} Q(\sqrt{\rho} p_i \left|R_{ii}\right|) p(\left|R_{ii}\right|) d\left|R_{ii}\right|. \quad (2.4)$$

This is addressed in the next section.

III. Main Results

Based on the well-known Chernoff bound for Q function, we have from (2.4)

$$P_b \le \frac{1}{4}\sum_{i=1}^{4}\int_0^{\infty} \frac{1}{2}\exp\left(-\frac{(\sqrt{\rho} p_i \left|R_{ii}\right|)^2}{2}\right) p(\left|R_{ii}\right|) d\left|R_{ii}\right|. \quad (3.1)$$

The proposed power allocation scheme is based on an explicit (but approximate) formula of the upper bound (3.1). For this we shall first specify the diagonal entries R_{ii}'s in terms of the channel gains h_{ij}'s, $1 \le i \le 4$; this will be done in Section III-A. Based on the established results, in Section III-B we then derive a closed-form expression of the upper bound (3.1).

A. Formulae of R_{ii}'s

Recall from (2.1) that the effective channel matrix $\mathbf{H}$ consists of four Alamouti's blocks [1]. By exploiting this property the formulae of R_{ii}'s can be obtained as follows.

Proposition 3.1: Let us partition the channel matrix $\mathbf{H}$ in (2.1) into four 2×2 submatrices as $\mathbf{H} = \begin{bmatrix} \mathbf{H}_1 & \mathbf{H}_2 \\ \mathbf{H}_3 & \mathbf{H}_4 \end{bmatrix}$, and let $\mathbf{H} = \mathbf{QR}$ be an associated QR-decomposition. Then the diagonal entries R_{ii}, $1 \le i \le 4$, in $\mathbf{R}$ are given by

$$R_{11} = R_{22} = \sqrt{\det(\mathbf{H}_1) + \det(\mathbf{H}_3)}, \quad (3.2)$$

and

$$R_{33} = R_{44} = \sqrt{\frac{\det(\mathbf{H})}{\det(\mathbf{H}_1) + \det(\mathbf{H}_3)}}. \quad (3.3)$$

[*Proof*]: The proof is based on an explicit expression for $\mathbf{Q}$ and $\mathbf{R}$ constructed according to [7], and the detailed derivation is relegated to the Appendix. □

B. Upper Bound of P_b in Closed-Form

As $R_{11} = R_{22}$ and $R_{33} = R_{44}$ (cf. (3.2) and (3.3)), we can rewrite P_b in (3.1) into

$$P_b \le \frac{1}{8}\left\{\underbrace{\sum_{i=1}^{2}\int_0^{\infty}\exp\left(-\frac{(\sqrt{\rho} p_i \left|R_{11}\right|)^2}{2}\right) p(\left|R_{11}\right|) d\left|R_{11}\right|}_{:=\varepsilon_1} + \underbrace{\sum_{i=3}^{4}\int_0^{\infty}\exp\left(-\frac{(\sqrt{\rho} p_i \left|R_{33}\right|)^2}{2}\right) p(\left|R_{33}\right|) d\left|R_{33}\right|}_{:=\varepsilon_2}\right\}. \quad (3.4)$$

In what follows we will derive analytic expressions for ε_1 and ε_2; this in turn yields a closed-form upper bound of P_b.

i) Analytic Form of ε_1: By definitions of $\mathbf{H}_1$ and $\mathbf{H}_3$, we have $\det(\mathbf{H}_1) = \left|h_{11}\right|^2 + \left|h_{21}\right|^2$ and $\det(\mathbf{H}_3) = \left|h_{12}\right|^2 + \left|h_{22}\right|^2$, which together with (3.2) imply

$$2\left|R_{11}\right|^2 = 2\left|h_{11}\right|^2 + 2\left|h_{21}\right|^2 + 2\left|h_{12}\right|^2 + 2\left|h_{22}\right|^2. \quad (3.5)$$

Since $2\left|h_{ij}\right|^2$ is a central chi-square random variable with degrees-of-freedom equal to two, $2\left|R_{11}\right|^2$ is thus central chi-square distributed with degrees-of-freedom equal to eight, with the probability density function (PDF) given by

$$f(2\left|R_{11}\right|^2) = \frac{(2\left|R_{11}\right|^2)^{4-1}}{2^4\Gamma(4)}\exp(-\left|R_{11}\right|^2), \quad (3.6)$$

where $\Gamma(\bullet)$ is the Gamma function. By performing a change of variable the PDF of $\left|R_{11}\right|$ is obtained as

$$f(\left|R_{11}\right|) = \frac{2\left|R_{11}\right|^{2(4-1)+1}}{(4-1)!}\exp(-\left|R_{11}\right|^2). \quad (3.7)$$

With (3.7), the summand in ε_1 becomes

$$\begin{aligned}
&\int_0^\infty \exp\left(-\frac{(\sqrt{\rho}p_i|R_{11}|)^2}{2}\right)\frac{2|R_{11}|^{2(4-1)+1}}{(4-1)!}\exp(-|R_{11}|^2)d|R_{11}| \\
&= \frac{2}{(4-1)!}\int_0^\infty |R_{11}|^{2(4-1)+1}\exp\left[-|R_{11}|^2(1+\frac{\rho p_i^2}{2})\right]d|R_{11}| \qquad (3.8)\\
&\underset{(a)}{=} \frac{2}{(4-1)!}\cdot\frac{(4-1)!}{2(1+\frac{\rho p_i^2}{2})^4} = (1+\frac{\rho p_i^2}{2})^{-4},
\end{aligned}$$

where (a) is obtained by performing integration with respect to $|R_{11}|$. Based on (3.8), a closed-form expression for ε_1 is

$$\varepsilon_1 = \sum_{i=1}^{2}(1+\frac{\rho p_i^2}{2})^{-4}. \qquad (3.9)$$

ii) Analytic Approximation of ε_2: We shall note that the closed-form expression of ε_1 in (3.9) hinges entirely chi-square nature of $2|R_{11}|^2$. Such a property, however, no longer holds for $2|R_{33}|^2$, since according to (3.3) straightforward manipulations show

$$\begin{aligned}
2|R_{33}|^2 &= 2\cdot\frac{\det(\mathbf{H})}{\det(\mathbf{H}_1)+\det(\mathbf{H}_3)} \\
&= 2\cdot\frac{\det(\mathbf{H}_1)\det(\mathbf{H}_4-\mathbf{H}_3\mathbf{H}_1^{-1}\mathbf{H}_2)}{\det(\mathbf{H}_1)+\det(\mathbf{H}_3)} \\
&= 2\cdot\frac{\det(\mathbf{H}_1)\left[\det(\mathbf{H}_4)+\frac{\det(\mathbf{H}_3)\det(\mathbf{H}_2)-2C}{\det(\mathbf{H}_1)}\right]}{\det(\mathbf{H}_1)+\det(\mathbf{H}_3)} \\
&= 2\cdot\frac{\det(\mathbf{H}_1)\det(\mathbf{H}_4)+\det(\mathbf{H}_3)\det(\mathbf{H}_2)-2C}{\det(\mathbf{H}_1)+\det(\mathbf{H}_3)}
\end{aligned} \qquad (3.10)$$

where

$$\begin{aligned}
C = \mathrm{Re}\{&h_{23}[h_{21}^*(h_{11}h_{13}^*+h_{14}h_{12}^*)+h_{22}^*(h_{12}h_{13}^*-h_{14}h_{11}^*)] \\
&+h_{24}[h_{21}^*(h_{11}h_{14}^*-h_{13}h_{12}^*)+h_{22}^*(h_{12}h_{14}^*+h_{13}h_{11}^*)]\}
\end{aligned}. \qquad (3.11)$$

It seems quite formidable to analytically characterize the exact PDF of $2|R_{33}|^2$ based on (3.10). To sidestep this difficulty, we will instead seek for an approximate PDF via curve-fitting to the simulated density of $2|R_{33}|^2$. Through intensive numerical test and curve fitting procedures (details omitted due to space limitation) it is found that the true density of $2|R_{33}|^2$ is well approximated by the following Gamma PDF:

$$f_{gamma}(x\mid k,\theta) = \left[\frac{x^{k-1}}{\theta^k\Gamma(k)}\right]\exp\left(-\frac{x}{\theta}\right), \text{ with } k=2 \text{ and } \theta=2. \qquad (3.12)$$

Figure 1 shows the computed histogram of $2|R_{33}|^2$ and the proposed approximate Gamma PDF (3.12); the proposed analytic approximation (3.12) is seen to well predict the simulated results. With the aid of (3.12), an approximate PDF of $|R_{33}|$ is in turn found as

$$f(|R_{33}|) \approx \frac{2|R_{33}|^{2(2-1)+1}}{(2-1)!}\exp(-|R_{33}|^2). \qquad (3.13)$$

Based on (3.13), the summand in ε_2 can be approximated by

$$\begin{aligned}
&\int_0^\infty \exp\left(-\frac{(\sqrt{\rho}p_i|R_{33}|)^2}{2}\right)\frac{2|R_{33}|^{2(2-1)+1}}{(2-1)!}\exp(-|R_{33}|^2)d|R_{33}| \\
&= \frac{2}{(2-1)!}\int_0^\infty |R_{33}|^{2(2-1)+1}\exp\left[-|R_{33}|^2(1+\frac{\rho p_i^2}{2})\right]d|R_{33}| \qquad (3.14)\\
&= \frac{2}{(2-1)!}\cdot\frac{(2-1)!}{2(1+\frac{\rho p_i^2}{2})^2} = (1+\frac{\rho p_i^2}{2})^{-2},
\end{aligned}$$

and hence

$$\varepsilon_2 \approx \sum_{i=3}^{4}(1+\frac{\rho p_i^2}{2})^{-2}. \qquad (3.15)$$

Combining (3.4), (3.9), and (3.15), P_b in (3.4) can thus be (approximately) upper bounded by

$$P_b \le \frac{1}{8}\left\{\sum_{i=1}^{2}(1+\frac{\rho p_i^2}{2})^{-4}+\sum_{i=3}^{4}(1+\frac{\rho p_i^2}{2})^{-2}\right\}. \qquad (3.16)$$

We thus propose to design the power loading factors (p_1,p_2,p_3,p_4) toward minimizing the average BER upper bound in (3.16), subject to the power normalization constraint $\sum_{i=1}^4 p_i^2 = 4$. As the cost function is highly nonlinear in p_i's, there do not seem to exist closed-form optimal solutions. Instead, the problem is solved via numerical search (e.g., by using **fmincom** in MATLAB Optimization Toolbox).

VI. PERFORMANCE

We compare the proposed approach with four other schemes, namely, linear ZF receiver, Stamoulis's decoupled signal recovery scheme [9], QR receiver without power loading, and the ZF V-BLAST detector [10], in terms of simulated BER. The proposed power loading factors via minimizing the closed-form bound in (3.16) are obtained by fmincom in MATLAB Optimization Toolbox. The results are shown in Figure 2. As we can see, the proposed method does outperform the QR receiver without power loading: there is about a 2 dB gain in the moderate-to-high SNR regime. Also, our method compares favorably with the ZF V-BLAST detector when SNR is above 20 dB. In terms of algorithm complexity, the ZF V-BLAST receiver involves signal ordering and pseudo-inverse computations; the total flop cost is 576 multiplications and 484 additions. The QR receiver involves mainly a QR decomposition which calls for 262 multiplications and 112 additions: it is thus more computationally efficient compared with the V-BLAST based solution.

APPENDIX: PROOF OF PROPOSITION 3.1

Since $\mathbf{H}_i^H\mathbf{H}_i = \mathbf{H}_i\mathbf{H}_i^H = \det(\mathbf{H}_i)\mathbf{I}_2$, through manipulations it can be shown that

$$\underbrace{\begin{bmatrix} \mathbf{H}_1 & \mathbf{H}_2 \\ \mathbf{H}_3 & \mathbf{H}_4 \end{bmatrix}^H}_{=\mathbf{H}^H} \cdot \underbrace{\frac{1}{\sqrt{\det(\mathbf{H}_1)+\det(\mathbf{H}_3)}} \begin{bmatrix} \mathbf{H}_1 & -\mathbf{H}_3^H \\ \mathbf{H}_3 & \mathbf{H}_3^{-H}\mathbf{H}_1^H\mathbf{H}_3^H \end{bmatrix}}_{:=\mathbf{Q}_1}$$

$$= \underbrace{\begin{bmatrix} \sqrt{\det(\mathbf{H}_1)+\det(\mathbf{H}_3)}\mathbf{I}_2 & \mathbf{0} \\ \dfrac{\mathbf{H}_2^H\mathbf{H}_1+\mathbf{H}_4^H\mathbf{H}_3}{\sqrt{\det(\mathbf{H}_1)+\det(\mathbf{H}_2)}} & \dfrac{-\mathbf{H}_2^H\mathbf{H}_3^H+\mathbf{H}_4^H\mathbf{H}_3^{-H}\mathbf{H}_1^H\mathbf{H}_3^H}{\sqrt{\det(\mathbf{H}_1)+\det(\mathbf{H}_3)}} \end{bmatrix}}_{:=\mathbf{G}}, \quad (A.1)$$

and

$$\mathbf{G} \cdot \underbrace{\begin{bmatrix} \mathbf{I}_2 & \mathbf{0} \\ \mathbf{0} & \dfrac{\left(-\mathbf{H}_2^H\mathbf{H}_3^H+\mathbf{H}_4^H\mathbf{H}_3^{-H}\mathbf{H}_1^H\mathbf{H}_3^H\right)^H}{\sqrt{\det(-\mathbf{H}_2^H\mathbf{H}_3^H+\mathbf{H}_4^H\mathbf{H}_3^{-H}\mathbf{H}_1^H\mathbf{H}_3^H)}} \end{bmatrix}}_{:=\mathbf{Q}_2}$$

$$= \underbrace{\begin{bmatrix} \sqrt{\det(\mathbf{H}_1)+\det(\mathbf{H}_3)} & \mathbf{0} \\ \dfrac{\mathbf{H}_2^H\mathbf{H}_1+\mathbf{H}_4^H\mathbf{H}_3}{\sqrt{\det(\mathbf{H}_1)+\det(\mathbf{H}_3)}} & \dfrac{\sqrt{\det(-\mathbf{H}_2^H\mathbf{H}_3^H+\mathbf{H}_4^H\mathbf{H}_3^{-H}\mathbf{H}_1^H\mathbf{H}_3^H)}}{\sqrt{\det(\mathbf{H}_1)+\det(\mathbf{H}_3)}} \end{bmatrix}}_{:=\mathbf{R}^H} \quad (A.2)$$

It is straightforward to verify that $\mathbf{Q}_1$ in (A.1) is unitary; since $-\mathbf{H}_2^H\mathbf{H}_3^H+\mathbf{H}_4^H\mathbf{H}_3^{-H}\mathbf{H}_1^H\mathbf{H}_3^H$ remains an Alamouti block [7], $\mathbf{Q}_2$ in (A.2) is also unitary. Combining (A.1) and (A.2) we have $\mathbf{H}^H\mathbf{Q}_1\mathbf{Q}_2=\mathbf{R}^H$, and hence $\mathbf{H}=\mathbf{QR}$ with $\mathbf{Q}=\mathbf{Q}_1\mathbf{Q}_2$ is an associated QR decomposition. We finally note that

$$\begin{aligned} \det(\mathbf{H}) &= \det\left(\mathbf{QR}\right) = \det\left(\mathbf{R}\right) \\ &= \sqrt{\det(-\mathbf{H}_2^H\mathbf{H}_3^H+\mathbf{H}_4^H\mathbf{H}_3^{-H}\mathbf{H}_1^H\mathbf{H}_3^H)}, \end{aligned} \quad (A.3)$$

where the second equality holds since $\mathbf{Q}$ is unitary and the last equality follows by definition of $\mathbf{R}$ in (A.2). The assertion thus follows from (A.2) and (A.3). □

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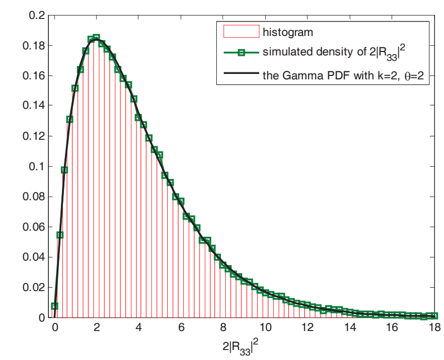

Figure 1. Simulated density of $2\left|R_{33}\right|^2$ and the associated approximation via Gamma PDF with $k=2$ and $\theta=2$.

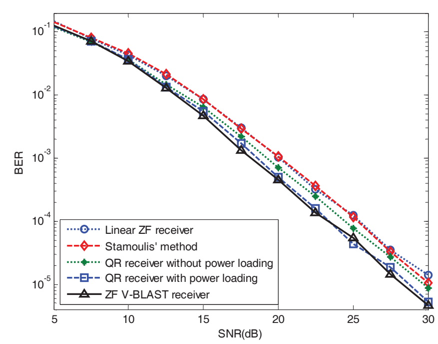

Figure 2. BER performances of D-STTD systems with different receivers (QPSK modulation).

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