A Data-Bearing Approach for Pilot-Aiding in Space-Time Coded MIMO Systems

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***Abstract*—This paper presents a novel technique for pilot-based channel estimation and data detection by exploiting a null-space property and an orthogonality property of the data bearer and pilot matrices. The data and pilot extraction procedures can be done independently by a simple linear transformation exploiting the null-space property. The maximum-likelihood (ML) receiver employed for data detection and the unconstrained-ML estimator employed for channel estimation can be designed separately by using the orthogonality property. In addition, the linear minimum mean-squared error (LMMSE) channel estimator is also proposed to improve the performance of channel estimation. The simulation results show that, among three data bearer and pilot structures including time-multiplexing (TM)-based, ST-block-code (STBC)-based, and code-multiplexing (CM)-based structures, the CM-based structure shows superior performance over the TM-based and the STBC-based structures in term of the probability of detection error, e.g. BER, for nonquasi-static flat Rayleigh fading channels, while the performances of these three structures are quite close for quasi-static flat Rayleigh fading channels.**

I. Introduction

Multiple-Input Multiple-Output (MIMO) communication systems provide prominent benefits to wireless communications, including enhanced capacity and reliability [1], [2]. Recently, the space-time (ST) codes have been proposed in [3], [4] for MIMO communications, in which the bit error rate (BER) of the communication systems is significantly improved without increasing transmission power by exploiting transmit diversity [3]. A major challenge in wireless ST communications employing a coherent detector is channel state information acquisition. Typically, the channel state information is acquired or estimated by using a pilot or training signal, the known signal transmitted from the transmitter to the receiver. This technique has been widely used because it is feasible to implement, and such a low computational complexity channel estimator can be implemented [5].

Two main pilot-based channel estimation techniques have been widely used in both single-input single-output (SISO) and MIMO systems: the pilot symbol assisted modulation (PSAM) technique and the pilot-embedding technique. In the SISO system, the PSAM technique has been intensively studied in [5] in the presence of fre- quency-nonselective fading channels, and was recently extended to MIMO systems [6], [7]. This technique firstly time-multiplexes a pilot signal into a transmit data stream, and then, at the receiver, this pilot signal is extracted from the received signal for acquiring the channel state information. The disadvantage of this technique is the sparse pilot arrangement that results in poor tracking of channel variation. In addition, the denser the pilot signals, the poorer the bandwidth efficiency.

On the other hand, the pilot-embedding or pilot-superimposed technique, which has been firstly proposed for the SISO systems [8] and for the MIMO systems [9], [10], can be done by adding a sequence of pilot signals directly to the data stream, and such soft-decoding methods, e.g. Viterbi algorithm [8], [10], are employed for channel estimation and data detection. Despite the better bandwidth efficiency of this technique, because it does not sacrifice any separate time slots for transmitting the pilot signal, the disadvantages of this technique are the highly computational complexity of the decoder and the highly computational delay in the channel estimation process given the channels are slowly varying.

Our purpose is to design a novel pilot-aiding approach for ST coded MIMO systems with affordable computational cost and better fast-fading channel acquisition. The basic idea is to simplify channel estimation and data detection processes by taking advantage of the null-space and orthogonality properties of the data-bearer and pilot matrices. The data-bearer matrix is used for projecting the ST data matrix onto the orthogonal subspace of the pilot matrix. By the virtue of the null-space and orthogonality properties, in our proposed data-bearing approach for pilot-aiding, a block of pilot matrix is directly elementwise added into a block of data matrix, that are mutually orthogonal to each other. The benefit that we are able to expect from this approach is better channel estimation performance, since the estimator is able to take into account the channel variation within the transmitted data block. In addition, a low computational complexity channel estimator and an enhance bandwidth efficiency, at least equal to the PSAM technique, are also expected. Throughout this paper, $(\cdot)^H$ denotes the complex-conjugate transpose, $(\cdot)^T$ denotes the transpose, and $(\cdot)^*$ denotes the complex conjugate.

The rest of this paper is organized as follows. In Section II, we present MIMO channel and system models. We propose a data-bearing approach for pilot-aiding in Section III, including the possible data bearer and pilot matrices, the channel estimation, and the data detection. The simulation results are given in Section IV and we conclude this paper in Section V.

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II. MIMO Channel and System Models

In this section, we briefly describe the MIMO channel and system models used in this paper. We consider the MIMO communication system with L_t transmit antennas and L_r receive antennas. In general, for a given block index t, a ST symbol matrix $\mathbf{U}(t)$ is an $L_t \times M$ codeword matrix transmitted across the transmit antennas in M time slots. The received symbol matrix $\mathbf{Y}(t)$ at the receiver front-end can be described as follows [10],

$$\mathbf{Y}(t) = \mathbf{H}(t)\mathbf{U}(t) + \mathbf{N}(t), \tag{1}$$

where $\mathbf{H}(t)$ is the $L_r \times L_t$ channel coefficient matrix and $\mathbf{N}(t)$ is the $L_r \times M$ additive complex white Gaussian noise (AWGN) matrix with zero mean and variance $\frac{\sigma^2}{2}\mathbf{I}_{(ML_r \times ML_r)}$ per real dimension. The elements of channel coefficient matrix $\mathbf{H}(t)$ are assumed to be spatially independent, complex Gaussian random variables with zero mean and variance 0.5 per real dimension, which in turn, an independent Rayleigh fading channel is modelled. In this paper, we firstly examine a quasi-static flat Rayleigh fading channel, where $\mathbf{H}(t)$ remains constant over each symbol block but it changes block-by-block independently. Then, we extend our proposed scheme to examine in a nonquasi-static flat Rayleigh fading channel, where $\mathbf{H}(t)$ changes according to a process whose dominant frequency is much faster than $\frac{1}{M}$, meaning that $\mathbf{H}(t)$ is not constant over each symbol block.

III. A Data-Bearing Approach for Pilot-Aiding

In this section, we present a data-bearing approach for pilot-aiding, including the pilot and data extraction procedures, the possible data bearer and pilot matrices, the channel estimation, and the data detection. The data-bearing approach for pilot-aiding firstly directly adds the pilot signal to the ST data and then regard this signal combination as the ST symbol. Our motivation of this approach is to embed the pilot signal onto the ST data in order to capture the variation of the channel at every instant in that ST block for the better channel coefficient estimate. Without loss of generality, we propose the data matrix $\mathbf{Z}(t) \in \mathbb{C}^{L_t \times M}$ as follows,

$$\mathbf{Z}(t) = \mathbf{D}(t)\mathbf{A}, \tag{2}$$

where $\mathbf{D}(t) \in \mathbb{C}^{L_t \times N}$ is the ST data matrix, and $\mathbf{A} \in \mathbb{R}^{N \times M}$ is the data-bearer matrix with N being the number of data time slots. In our implementation, the ST data matrix D(t) is assumed to maintain the energy constraint $\mathrm{E}[\|\mathbf{D}(t)\|^2] = L_t$ with $\|\cdot\|$ being the Frobenius norm. The proposed pilot-aided ST symbol matrix $\mathbf{U}(t)$ can be expressed as follows,

$$\mathbf{U}(t) = \mathbf{Z}(t) + \mathbf{P} = \mathbf{D}(t)\mathbf{A} + \mathbf{P}, \tag{3}$$

where $\mathbf{P} \in \mathbb{R}^{L_t \times M}$ is the pilot matrix. In general, the pilot-aided ST symbol block structure can be demonstrated in Fig.1. The pilot-aided ST symbol block consists of two main parts: data sequences $\{\mathbf{Z}(t)\}_i$ and pilot sequences $\{\mathbf{P}\}_i$, where i stands for a row index, $i = 1, \ldots, L_t$.

Substitute (3) into (1), the received symbol matrix $\mathbf{Y}(t)$ in (1) can be rewritten as follows,

$$\mathbf{Y}(t) = \mathbf{H}(t)(\mathbf{D}(t)\mathbf{A} + \mathbf{P}) + \mathbf{N}(t). \tag{4}$$

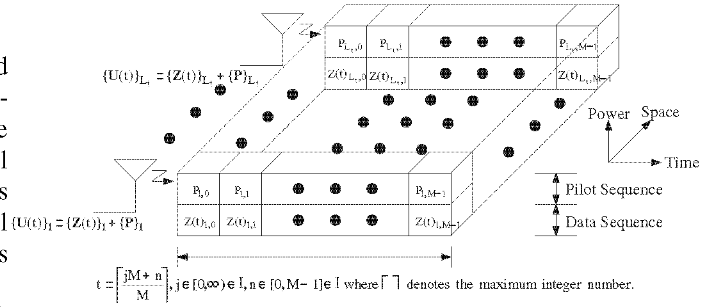

Fig. 1: The pilot-aided ST symbol block structure.

By the data-bearing approach for pilot-aiding, we mean that the data bearer matrix $\mathbf{A}$ and the pilot matrix $\mathbf{P}$ satisfy the following properties:

$$\mathbf{A}\mathbf{P}^T = \mathbf{0} \in \mathbb{R}^{N \times L_t}, \tag{5}$$

$$\mathbf{P}\mathbf{A}^T = \mathbf{0} \in \mathbb{R}^{L_t \times N}, \tag{6}$$

$$\mathbf{A}\mathbf{A}^T = \beta\mathbf{I} \in \mathbb{R}^{N \times N}, \tag{7}$$

$$\mathbf{P}\mathbf{P}^T = \alpha\mathbf{I} \in \mathbb{R}^{L_t \times L_t}, \tag{8}$$

where β is a real-valued data-power factor for controlling the value of data-part power, α is a real-valued pilot-power factor for controlling the value of pilot-part power, $\mathbf{0}$ stands for an all-zero-element matrix, and $\mathbf{I}$ stands for an identity matrix.

From (5)-(8), it is straightforward to verify that the time slots M of the pilot-aided ST symbol matrix $\mathbf{U}(t)$ must satisfy the following equality

$$\mathrm{Rank}(\mathbf{A}) + \mathrm{Rank}(\mathbf{P}) = M. \tag{9}$$

There are three possible different structures of data bearer and pilot matrices, in which the elements of these matrices are real numbers, that satisfy the properties (5)-(8) as follows.

1. Time-Multiplexing (TM)-Based Matrices

The structures of these matrices are given by

$$\mathbf{A} = \sqrt{\beta}\left[\mathbf{0}_{(N \times L_t)}; \mathbf{I}_{(N \times N)}\right],$$
$$\mathbf{P} = \sqrt{\alpha}\left[\mathbf{I}_{(L_t \times L_t)}; \mathbf{0}_{(L_t \times N)}\right], \; M = N + L_t, \tag{10}$$

where ; stands for matrix combining. The TM-based data bearer and pilot matrices are similar to the PSAM concept in [5] and have been used in [7], for instance.

2. ST-Block-Code (STBC)-Based Matrices

The structures of these matrices are given by

$$\mathbf{A} = \sqrt{\beta}\left[\mathbf{0}_{(N \times \tau)}; \mathbf{I}_{(N \times N)}\right],$$
$$\mathbf{P} = \sqrt{\alpha}\left[\mathrm{STBC}_{(L_t \times \tau)}; \mathbf{0}_{(L_t \times N)}\right], \; M = N + \tau, \tag{11}$$

where τ is the time slots used for transmitting one ST block code. The normalized known ST block code [4] is used as the pilot or training information. This kind of data bearer and pilot matrices have been used in [6], for instance.

3. Code-Multiplexing (CM)-Based Matrices

The structures of these matrices are given by

$$\mathbf{A} = \sqrt{\beta}\mathrm{WH}[1:N]_{(N \times M)},$$
$$\mathbf{P} = \sqrt{\alpha}\mathrm{WH}[N+1:M]_{(L_t \times M)}, \; M = N + L_t, \tag{12}$$

where WH$[x:y]$ denotes a submatrix created by splitting the $M \times M$ normalized Walsh-Hadamard matrix [12] starting from x^{th}-row to y^{th}-row. The disadvantage of the CM-based data bearer and pilot matrices is the limitation of a dimension of the Walsh-Hadamard matrix, in which the dimension is proportional to $2^n,\ n \in \mathbb{I}^+$.

A. Channel Estimation

The channel estimation of our proposed data-bearing approach for pilot-aiding can be achieved by first simply post-multiplying the received symbol matrix $\mathbf{Y}(t)$ in (4) by the transpose of the pilot matrix $\mathbf{P}^T$ for extracting the pilot part. By using (5) and (8), and dividing the result by α, thus yielding

$$\frac{\mathbf{Y}(t)\mathbf{P}^T}{\alpha} = \mathbf{H}(t) + \frac{\mathbf{N}(t)\mathbf{P}^T}{\alpha}. \tag{13}$$

Let us define $\dot{\mathbf{Y}}(t) = \frac{\mathbf{Y}(t)\mathbf{P}^T}{\alpha}$ and $\dot{\mathbf{N}}(t) = \frac{\mathbf{N}(t)\mathbf{P}^T}{\alpha}$. In the sequel, we need $\dot{\mathbf{Y}}(t)$ and the parameters to be represented as column vectors. Denote $\mathbf{y}(t) \triangleq \mathbf{vec}(\dot{\mathbf{Y}}(t))$, $\mathbf{n}(t) \triangleq \mathbf{vec}(\dot{\mathbf{N}}(t))$, and $\mathbf{h}(t) \triangleq \mathbf{vec}(\dot{\mathbf{H}}(t))$, where $\mathbf{vec}(\cdot)$ denotes vectorizing conversion. The pilot-projected received symbol matrix $\dot{\mathbf{Y}}(t)$ in (13) can be rewritten as follows,

$$\mathbf{y}(t) = \mathbf{h}(t) + \mathbf{n}(t), \tag{14}$$

where $\mathbf{n}(t) = \frac{1}{\alpha}(\mathbf{I} \otimes \mathbf{N}(t))\mathbf{vec}(\mathbf{P}^T)$ with $\otimes$ being the Kronecker product. The second-order statistics of the pilot-projected noise vector $\mathbf{n}(t)$ are determined as follows,

$$\boldsymbol{\mu}_{\mathbf{n}(t)} = \frac{1}{\alpha}\mathrm{E}[(\mathbf{I} \otimes \mathbf{N}(t))\mathbf{vec}(\mathbf{P}^T)] = \mathbf{0}_{(L_t L_r \times 1)}, \tag{15}$$

$$\begin{aligned} \mathbf{V}_{\mathbf{n}(t)} &= \tfrac{1}{\alpha^2}\mathrm{E}[\mathbf{n}(t)\mathbf{n}^H(t)] = \tfrac{\sigma^2}{2\alpha^2}\mathrm{Diag}\left(B_i\right), \\ B_i &= \textstyle\sum_{j=1}^{M} |P_{i,j}|^2 \mathbf{I}_{(L_r \times L_r)},\ i \in \{1, \ldots, L_t\}, \end{aligned} \tag{16}$$

where $\boldsymbol{\mu}_{\mathbf{n}(t)}$ and $\mathbf{V}_{\mathbf{n}(t)}$ stand for the mean vector and the covariance matrix of the pilot-projected noise vector $\mathbf{n}(t)$ per real dimension, respectively, $P_{i,j}$ is the i^{th}-row j^{th}-column element of the pilot matrix $\mathbf{P}$, Diag$(\cdot)$ stands for the diagonal matrix created by concatenating submatrices $B_i,\ i \in 1, \ldots, L_t$, into the diagonal elements.

From (8), it can be shown that $\sum_{j=1}^{M} |P_{i,j}|^2 = \alpha,\ \forall i$. Hence, we can rewrite (16) as follows,

$$\mathbf{V}_{\mathbf{n}(t)} = \frac{\sigma^2}{2\alpha}\mathbf{I}_{(L_t L_r \times L_t L_r)} \text{ per real dimension.} \tag{17}$$

Obviously, the pilot-projected noise vector $\mathbf{n}(t)$ is the complex white Gaussian vector, hence, the log-likelihood function $\ln(p(\mathbf{y}(t)|\mathbf{h}(t)))$ is given by [13]

$$\begin{aligned} \ln(p(\mathbf{y}(t)|\mathbf{h}(t))) &= \ln\left(\tfrac{1}{\pi^{L_r L_t} det(\mathbf{V}_{\mathbf{n}(t)})}\right) \\ &-(\mathbf{y}(t) - \mathbf{h}(t))^H \mathbf{V}_{\mathbf{n}(t)}^{-1}(\mathbf{y}(t) - \mathbf{h}(t)). \end{aligned} \tag{18}$$

1) Unconstrained Maximum-Likelihood (ML) Channel Estimator: The maximum-likelihood estimator [13] maximizes the log-likelihood function $\ln(p(\mathbf{y}(t)|\mathbf{h}(t)))$ as follows,

$$\hat{\mathbf{h}}(t) = \max_{\mathbf{h}(t)}\{\ln(p(\mathbf{y}(t)|\mathbf{h}(t)))\}. \tag{19}$$

Differentiating (18) and equating the result to zero, we have the maximum-likelihood estimator as follows,

$$\hat{\mathbf{h}}(t) = \mathbf{y}(t) \text{ or } \hat{\mathbf{H}}(t) = \dot{\mathbf{Y}}(t). \tag{20}$$

It is worth noticing that our ML estimator is the pilot-projected received vector (or matrix) $\mathbf{y}(t)$ (or $\mathbf{H}(t)$) itself.

2) Linear Minimum Mean-Squared Error (LMMSE) Channel Estimator: We further improve the performance of the unconstrained ML channel estimator in (20) by employing the L-tap LMMSE channel interpolation. The L-tap LMMSE channel estimator can be expressed as follows,

$$h_{j,i}^{LMMSE}(t) = \mathbf{w}_{j,i}^H \hat{\mathbf{h}}_{j,i}^L(t), \tag{21}$$

where $h_{j,i}^{LMMSE}(t)$ denotes the j^{th}-row i^{th}-column element of the LMMSE-estimated channel matrix, $\mathbf{w}_{j,i} = [w_{j,i}(0) \cdots w_{j,i}(L-1)]^T$ denotes the L-tap finite impulse response (FIR) linear filter's weight vector, and $\hat{\mathbf{h}}_{j,i}^L(t) = [\hat{h}_{j,i}(t) \cdots \hat{h}_{j,i}(t-L+1)]^T$ denotes the L-element input vector constructed from the j^{th}-row i^{th}-column element of the ML-estimated channel matrix in (20) taking values in the block interval $[t-L+1, t]$. The optimization criterion, assuming the channels are wide-sense stationary (WSS), for the L-tap LMMSE channel estimator is given by

$$J(\mathbf{w}_{j,i}) = \arg\min_{\mathbf{w}_{j,i}} \mathrm{E}\left[\|h_{j,i}(t) - \mathbf{w}_{j,i}^H \hat{\mathbf{h}}_{j,i}^L(t)\|^2\right], \tag{22}$$

where $h_{j,i}(t)$ denotes the j^{th}-row i^{th}-column element of the true channel matrix $\mathbf{H}(t)$ in (1).

Differentiating (22) with respect to $\mathbf{w}_{j,i}^*$ and equating the result to zero, the optimum LMMSE weight vector $\mathbf{w}_{j,i}^{opt}$ is given by

$$\mathbf{w}_{j,i}^{opt} = \mathbf{R}_{\hat{\mathbf{h}}_{j,i}^L(t)}^{-1} \mathbf{p}_{\hat{\mathbf{h}}_{j,i}^L(t)}, \tag{23}$$

where $\mathbf{R}_{\hat{\mathbf{h}}_{j,i}^L(t)} = \mathrm{E}[\hat{\mathbf{h}}_{j,i}^L(t)\hat{\mathbf{h}}_{j,i}^{H_L}(t)]$ and $\mathbf{p}_{\hat{\mathbf{h}}_{j,i}^L(t)} = \mathrm{E}[h_{j,i}^*(t)\hat{\mathbf{h}}_{j,i}^L(t)]$. According to (14), (17), (20), and the uncorrelatedness of the channel and noise coefficients, the autocorrelation matrix $\mathbf{R}_{\hat{\mathbf{h}}_{j,i}^L(t)}$ can be further simplified as follows,

$$\mathbf{R}_{\hat{\mathbf{h}}_{j,i}^L(t)} = \mathbf{R}_{\mathbf{h}_{j,i}^L(t)} + \frac{\sigma^2}{\alpha}\mathbf{I}_{L \times L}, \tag{24}$$

where $\mathbf{R}_{\mathbf{h}_{j,i}^L(t)} = \mathrm{E}[\mathbf{h}_{j,i}^L(t)\mathbf{h}_{j,i}^{H_L}(t)]$ and $\mathbf{h}_{j,i}^L(t) = [h_{j,i}(t) \cdots h_{j,i}(t-L+1)]^T$. In a similar way to (24), the cross-correlation vector $\mathbf{p}_{\hat{\mathbf{h}}_{j,i}^L(t)}$ can be further simplified as follows,

$$\mathbf{p}_{\hat{\mathbf{h}}_{j,i}^L(t)} = \mathrm{E}[h_{j,i}^*(t)\mathbf{h}_{j,i}^L(t)] = \mathbf{p}_{\mathbf{h}_{j,i}^L(t)}. \tag{25}$$

Substituting (24) and (25) into (23), and then substituting (23) into (21), the L-tap LMMSE channel estimator can be rewritten as

$$h_{j,i}^{LMMSE}(t) = \left[(\mathbf{R}_{\mathbf{h}_{j,i}^L(t)} + \frac{\sigma^2}{\alpha}\mathbf{I}_{L \times L})^{-1}\mathbf{p}_{\mathbf{h}_{j,i}^L(t)}\right]^H \hat{\mathbf{h}}_{j,i}^L(t). \tag{26}$$

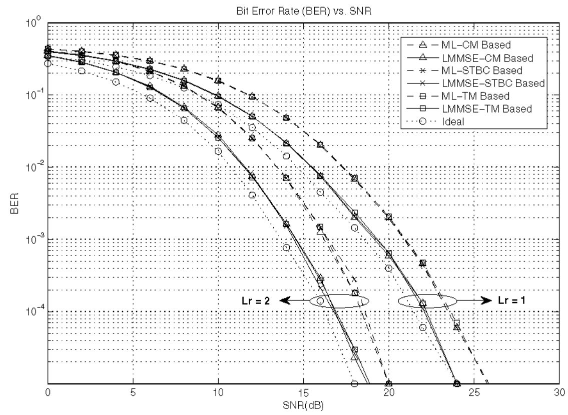

Fig. 2: The graph of BER in quasi-static flat Rayleigh fading channels.

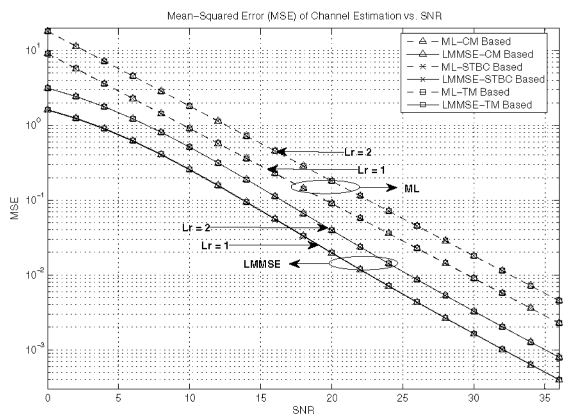

Fig. 3: The graph of MSE of the channel estimation in quasi-static flat Rayleigh fading channels.

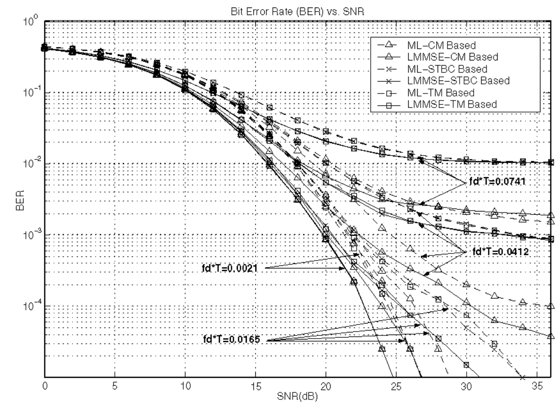

Fig. 4: The graph of BER when $L_r = 1$ in nonquasi-static flat Rayleigh fading channels.

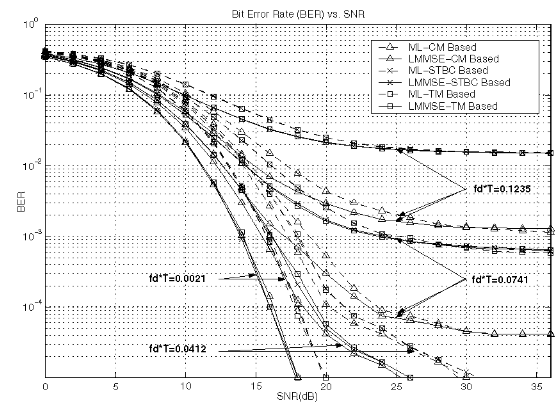

Fig. 5: The graph of BER when $L_r = 2$ in nonquasi-static flat Rayleigh fading channels.

B. Data Detection

We further describe the data detection procedure. Firstly, the data part of the received symbol matrix $\mathbf{Y}(t)$ is extracted by post-multiplying the received symbol matrix $\mathbf{Y}(t)$ by the transpose of the data bearer matrix $\mathbf{A}^T$. By using (6) and (7), and dividing the result by β, the data-bearer-projected received symbol matrix is given by

$$\frac{\mathbf{Y}(t)\mathbf{A}^T}{\beta} = \mathbf{H}(t)\mathbf{D}(t) + \frac{\mathbf{N}(t)\mathbf{A}^T}{\beta}. \tag{27}$$

Let us define $\ddot{\mathbf{Y}}(t) = \frac{\mathbf{Y}(t)\mathbf{A}^T}{\beta}$ and $\ddot{\mathbf{N}}(t) = \frac{\mathbf{N}(t)\mathbf{A}^T}{\beta}$. We can rewrite (27) as follows,

$$\ddot{\mathbf{Y}}(t) = \mathbf{H}(t)\mathbf{D}(t) + \ddot{\mathbf{N}}(t). \tag{28}$$

The maximum-likelihood receiver is employed for decoding the transmitted ST data matrix $\mathbf{D}(t)$ by using the estimated channel coefficient obtained in either (20) or (26) as the channel state information. The maximum-likelihood receiver computes the decision matric and decides the codeword that minimizes this decision matric [3],

$$\{\hat{d}_t^i\} = \min_{\{d_t^i\}} \left\{ \sum_{t=1}^{N} \sum_{j=1}^{L_r} |y_t^j - \sum_{i=1}^{L_t} \hat{h}_{j,i} d_t^i|^2 \right\}, \\ \forall d_t^i,\ i \in \{1,\dots,L_t\},\ t \in \{1,\dots,N\}, \tag{29}$$

where y_t^j denotes the j^{th}-row t^{th}-column element of the data-bearer-projected received symbol matrix $\ddot{\mathbf{Y}}(t)$, $\hat{h}_{j,i}$ denotes the j^{th}-row i^{th}-column element of the estimated channel coefficient matrix, and $\hat{d}_t^i$ denotes the i^{th}-row t^{th}-column element of the estimated ST data matrix $\hat{\mathbf{D}}(t)$.

IV. Simulation Results

In this section, we demonstrate the performance of the data-bearing approach for pilot-aiding. Without loss of generality, we examine a 4×4 orthogonal ST block code of [4]. The setting parameters of our experiments are: the normalized pilot-aided ST symbol power is 1 watt/pilot-aided ST symbol block; the time slots are 8 time slots/pilot-aided ST symbol block. The data part's power is constantly allocated 50% and the pilot part's power is constantly allocated 50% of the normalized pilot-aided ST symbol power. In addition, 4-PSK modulation is employed in these experiments and a number of taps of the LMMSE channel estimator is 3.

A. Quasi-Static Flat Rayleigh Fading Channels

In this situation, the element of the channel coefficient matrix $\mathbf{H}(t)$ in (1) are taken from the normalized time-varying channel which is modelled as Jakes' model [14], where $fd * T = 0.08$ (fast fading) with fd being the Doppler's shift and T being the symbol period.

The graph of BER of the pilot-aided MIMO system compared with the ideal-channel MIMO system when 1 and 2-receive antennas are employed is shown in Fig.2. Notice that, at BER = 10^{-4}, the SNR differences between the ideal-channel and the ML channel estimator are about 2.3 dB, whereas the SNR differences between the ideal-channel and the LMMSE channel estimator are about 0.5 dB, for both the 1 and 2-receive antenna schemes. In addition, the SNR differences between the ML and LMMSE channel estimators are about 1.8 dB.

The graph of MSE of the channel estimation of the pilot-aided MIMO system when 1 and 2-receive antennas are employed is shown in Fig.3. Notice that the MSE of the ML channel estimator is larger than the LMMSE channel estimator, and all three kinds of data bearer and pilot structures yield the same MSE. It is worth noticing that the LMMSE channel estimator performs better than the ML channel estimator because of the higher accurate channel estimate, as shown in Fig.3.

B. Nonquasi-Static Flat Rayleigh Fading Channels

In this situation, we consider the situation where the channel coefficient matrix is not constant over the ST symbol block. Without loss of generality, we give an example where the channel coefficient matrix changes twice within one ST symbol block, i.e. there exists $\mathbf{H}_1(t)$ and $\mathbf{H}_2(t)$ in the t^{th}-block ST symbol matrix.

1) 1-Receive Antenna Scheme: The graph of BER of the pilot-aided MIMO system when $fd*T$ are 0.0021 (slow fading), 0.0165, 0.0412, and 0.0741 (fast fading) is shown in Fig.4. Notice that, when Doppler's shifts are small, e.g. $fd * T = 0.0021$, the probability of detection error of three kinds of data bearer and pilot structures are quite the same; however, when Doppler's shifts are getting larger, the CM-based structure is much better than the TM- and STBC-based structures, where the error floors of the CM-based structure are much lower than the TM- and STBC-based structures. Since the nonquasi-static flat Rayleigh fading channel is the severe situation, there exists error floors that increase significantly as the Doppler's shift increases.

2) 2-Receive Antenna Scheme: The graph of BER of the pilot-aided MIMO system when $fd * T$ are 0.0021 (slow fading), 0.0412, 0.0741, and 0.1235 (fast fading) is shown in Fig.5. Similarly to the 1-receive antenna scheme, the CM-based structure is much better than the TM- and STBC-based structures.

Obviously, the CM-based structure performs much better than the TM- and STBC-based structures because it takes into account both of the channel submatrices $\mathbf{H}_1(t)$ and $\mathbf{H}_2(t)$ in the channel estimation, whereas the other two structures exploit either of them according to their structures. Furthermore, the LMMSE channel estimator performs better than the ML channel estimator in a low SNR region, in which the AWGN is the major factor that causes the detection error; however, in a high SNR region, the channel mismatch plays a major role in causing the detection error resulting in the comparable error floors for the LMMSE and ML channel estimators. In addition, the 2-receive antenna scheme is more robust to the fast fading channel than the 1-receive antenna scheme.

V. Conclusion

In this paper, we have proposed the data-bearing approach for pilot-aiding for joint data detection and channel estimation in ST coded MIMO systems, including the necessary properties for the data bearer and pilot matrices, the pilot and data extraction procedures, the possible data bearer and pilot matrices, the channel estimation, and the data detection. Obviously, our data-bearing approach for pilot-aiding subsumes the classical pilot-based channel estimation, e.g. PSAM, and it reveals that, in the quasi-static flat Rayleigh fading channels, the performance of three kinds of data bearer and pilot structures, i.e. the TM-based, STBC-based, CM-based structures, are quite the same. Furthermore, in the nonquasi-static flat Rayleigh fading channels, the CM-based structure shows superior performance over the TM- and STBC-based structures especially for high Doppler's shift scenarios, where the error floors of the former are much lower than the other two. Due to the page limitation, the interested readers could further explore the performance analysis of our data-bearing approach for pilot-aiding and the optimum power allocation for the data and pilot parts in [15].

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