NOVEL COMMUNICATION SCHEMES WITH BLIND CHANNEL ESTIMATION IN TDD MIMO SYSTEM

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ABSTRACT

In this paper, we consider practical two-way communication schemes with blind channel estimation in TDD (Time-Division Duplex) MISO (Multi-Input Single-Output) and MIMO (Multi-Input Multi-Output) systems. In MISO system, we show that the blind channel estimation only needs to be done at the transceiver equipped with multiple antennas and this estimated channel can be efficiently utilized for both uplink and downlink communication. In MIMO system, we propose a two-stage communication scheme which begins with uplink transmission using differential space-time block codes and follows with two-way beamforming with blind channel estimation. We demonstrate that the novel schemes possess very good performance through simulations and compare it with training-based schemes.

1. INTRODUCTION

A key factor characterizing the performance of MIMO systems is the amount of channel state information (CSI) that transmitters have. In traditional space-time coding, the transmitter is assumed to have no CSI. On the other hand, if the transmitter has full CSI it can use beamforming which outperforms most space-time codes [1]. If the channel is time-varying, the transmitter needs a scheme to obtain CSI, one way being channel feedback, and a number of recent papers have investigated the effect of channel feedback [2][3]. However, in time-division duplex (TDD) the uplink and downlink channels are in general symmetric if the channel does not vary too rapidly. A channel estimate for the uplink channel can therefore be used by the basestation for beamforming in the dowlink channel, and vice versa. The channel estimates can either be obtained by transmitting training sequences in uplink and downlink, or by using blind channel estimation. In this paper we will study TDD beamforming with blind channel estimation. In a MISO system the uplink channel estimation can be used directly for beamforming in the donwlink. In a MIMO system, we propose a two stage system: firstly the transmitters use differential space-time coding. It then uses the received signal in the first stage to estimate the channels, and uses this for beamforming in the second stage.

We assume that the channels are discrete-time frequency flat block-Rayleigh-fading with each block holding constant for M_T symbol intervals. In this paper, we model the channel coefficient between each transmit and receiver pair as i.i.d complex Gaussian random variables with zero mean and variance 0.5 per real dimension. The transmit power at transceiver *1* and transceiver *2* are P_1 and P_2 respectively. $()^\dagger$, $()^T$ and $()^*$ denote the operations of conjugate transpose, transpose and conjugate respectively.

2. COMMUNICATION SCHEME IN A MISO SYSTEM

Consider a single user TDD MISO system with N antennas at transceiver *1* and 1 antenna at transceiver *2*. Let M_u and M_d be the number of symbol intervals used for uplink (*2* $\rightarrow$ *1*) and downlink (*1* $\rightarrow$ *2*) transmission respectively with $M_T = M_u + M_d$. The received signals sampled in symbol m at transceiver *1* and transceiver *2* are

$$\boldsymbol{r}_1(m) = \boldsymbol{h}x_2(m) + \boldsymbol{n}_1(m), \tag{1}$$

$$r_2(m) = \boldsymbol{h}^T\boldsymbol{x}_1(m) + n_2(m), \tag{2}$$

respectively, where $\boldsymbol{h}$ is a $N \times 1$ channel vector, $\boldsymbol{n}_1(m)$ and $n_2(m)$ are zero mean circular complex Gaussian noise with variance σ_1^2 and σ_2^2 respectively. Since only one antenna is employed at transceiver *2*, the uplink transmission is $x_2(m) = s_2(m)$. Instead at transceiver *1*, beamforming can be applied for the downlink transmission as follows:

$$\boldsymbol{x}_1(m) = \frac{\boldsymbol{h}^*}{\|\boldsymbol{h}\|}s_1(m), \tag{3}$$

where $s_1(m)$ and $s_2(m)$ are differentially coded source data. Due to the transmit power constraint $\|s_2(m)\|^2 = P_2$, the auto-correlation matrix of received signals at transceiver *1* is $\boldsymbol{R}_1 = \boldsymbol{h}\boldsymbol{h}^\dagger P_2 + \sigma_1^2 \boldsymbol{I}$, which can be estimated by

$$\hat{\boldsymbol{R}}_1 \triangleq \frac{1}{M_u}\sum_{m=1}^{M_u}\left\{\boldsymbol{r}_1(m)\boldsymbol{r}_1(m)^\dagger\right\} = \hat{\boldsymbol{U}}_1\hat{\boldsymbol{\Lambda}}_1^2\hat{\boldsymbol{U}}_1^\dagger,$$

where $\hat{\boldsymbol{U}}_1\hat{\boldsymbol{\Lambda}}_1^2\hat{\boldsymbol{U}}_1^\dagger$ is the eigenvalue decomposition of $\hat{\boldsymbol{R}}_1$. Since only the normalize channel vector is needed for beamfoming scheme, and it can be estimated as the first column vector in $\hat{\boldsymbol{U}}_1$ up to some phase ambiguity, the estimated normalized channel vector can be denoted as $\frac{\hat{\boldsymbol{h}}}{\|\hat{\boldsymbol{h}}\|}e^{j\theta}$, where $\frac{\hat{\boldsymbol{h}}}{\|\hat{\boldsymbol{h}}\|}$ is the estimated normalized channel vector with precision loss due to limited number of observations and θ is the phase ambiguity introduced by blind estimation. Therefore the received signal at transceiver 2 can be written as

$$r_2(m) \quad = \quad \frac{\hat{\boldsymbol{h}}^\dagger}{\|\hat{\boldsymbol{h}}\|}\boldsymbol{h}e^{j\theta}s_1(m) + n_2(m). \qquad (4)$$

2.1. Detection for Downlink Transmission

In order to do the coherent detection, we still need to estimate the unknown complex scalar $\Psi \triangleq \frac{\hat{\boldsymbol{h}}^\dagger}{\|\hat{\boldsymbol{h}}\|}\boldsymbol{h}e^{j\theta}$. For BPSK modulation, an unbiased blind estimator of Ψ (up to a sign ambiguity) is

$$\hat{\Psi} \quad = \quad \sqrt{\frac{1}{M_d P_1}\sum_{m=1}^{M_d} r_2^2(m)}, \qquad (5)$$

where P_1 can be omitted since it doesn't affect the detection for BPSK modulation. To correct the sign ambiguity, the differential BPSK coding is needed at the transmitter. Therefore the coherent detector is

$$\hat{s}_1(m) \quad = \quad sgn\left\{\Re\left(\hat{\Psi}^* r_2(m)\right)\right\}. \qquad (6)$$

2.2. Detection for Uplink Transmission

We can do the maximum ratio combining for the received signal from uplink transmission since the blindly estimated normalized channel vector is already known at transceiver *1*

$$\tilde{r}_1(m) = \frac{\hat{\boldsymbol{h}}^\dagger}{\|\hat{\boldsymbol{h}}\|}e^{j\theta}\boldsymbol{r}_1(m) = \frac{\hat{\boldsymbol{h}}^\dagger}{\|\hat{\boldsymbol{h}}\|}\boldsymbol{h}e^{j\theta}s_2(m) + \tilde{n}_1(m), \quad (7)$$

where we define

$$\tilde{n}_1(m) \triangleq \frac{\hat{\boldsymbol{h}}^\dagger}{\|\hat{\boldsymbol{h}}\|}e^{j\theta}\boldsymbol{n}_1(m). \qquad (8)$$

Obviously, $\tilde{n}_1(m)$ is still zero-mean circular complex gaussian noise. Noticing the similarity of equations (4) and (7), we can apply the same detection scheme on uplink transmission as in the downlink transmission, and the performance at both sides are same if at same received SNR.

3. COMMUNICATION SCHEME IN A MIMO SYSTEM

In a TDD MIMO system which applies N antennas and M antennas at transceiver *1* and transceiver *2* respectively, the received signals at transceiver *1* and transceiver *2* respectively are

$$\boldsymbol{r}_1(m) = \boldsymbol{H}^T\boldsymbol{x}_2(m) + \boldsymbol{n}_1(m), \qquad (9)$$
$$\boldsymbol{r}_2(m) = \boldsymbol{H}\boldsymbol{x}_1(m) + \boldsymbol{n}_2(m), \qquad (10)$$

where $\boldsymbol{H}$ is an $M \times N$ channel matrix. Initially, none of the transceivers know the channel. We therefore propose the following two-stage communication scheme: firstly one of the transceivers transmit the differential space-time block codes. The other transceiver uses the received signal to blindly estimate the channel, and then, in the second stage, uses this for beamforming. After that, the two-way beamforming with blind channel estimation can be implemented. The details are as follows:

3.1. First Stage

Assume the communication begins with uplink transmission, where transceiver *2* transmits differential space-time block codes as proposed in [4], that is, $\boldsymbol{x}_2(m+k)$ is the k-th column of a differential space-time block codeword C_m. For example, as the number of transmitter antennas $N = 2$, the codeword send in block l can be denoted as

$$\boldsymbol{C}_m = \begin{bmatrix} c(2m+1) & -c(2m+2)^* \\ c(2m+2) & c(2m+1)^* \end{bmatrix}, \qquad (11)$$

with

$$\begin{bmatrix} c(2m+1) \\ c(2m+2) \end{bmatrix} = \theta(m)\begin{bmatrix} c(2m-1) \\ c(2m) \end{bmatrix} + \phi(l)\begin{bmatrix} -c(2m)^* \\ c(2m-1)^* \end{bmatrix},$$

where $\theta(m)$ and $\phi(m)$ are mapped from source data sending in block m under a certain mapping pattern. The conjugated auto-correlation matrix of received signals at transceiver *1* is

$$\begin{aligned} \boldsymbol{R}_{1f}^* &= \mathcal{E}\left\{\boldsymbol{r}_1(m)\boldsymbol{r}_1(m)^\dagger\right\}^* \\ &= \mathcal{E}\left\{\boldsymbol{H}^\dagger(C_m C_m^\dagger)^*\boldsymbol{H}\right\} + \sigma_1^2\boldsymbol{I} \\ &= \boldsymbol{H}^\dagger\boldsymbol{H}P_2 + \sigma_1^2\boldsymbol{I}. \end{aligned} \qquad (12)$$

Denote the Singular Value Decomposition (SVD) of channel matrix $\boldsymbol{H}$ as $\boldsymbol{H} = \boldsymbol{U}\boldsymbol{\Lambda}\boldsymbol{V}^{\dagger}$, where $\boldsymbol{\Lambda}$ is a diagonal matrix as $\mathbf{diag}\{\lambda_1, \lambda_2, \cdots, \lambda_{N_0}\}$, therefore the decomposition of $\boldsymbol{R}_{1f}^{*}$ is $\boldsymbol{R}_{1f}^{*} = \boldsymbol{V}(P_2\boldsymbol{\Lambda}^2 + \sigma_1^2\boldsymbol{I})\boldsymbol{V}^{\dagger}$. The detection scheme of the first stage is non-coherent detection as in [4].

3.2. Second Stage

Beamforming can be applied for both uplink and downlink transmissions after the first stage. The transmit signal at transceiver *1* is $\boldsymbol{x}_1(m) = \boldsymbol{v}_1 s_1(m)$, where $\boldsymbol{v}_1$ is the top singular eigenvector of $\boldsymbol{V}$. The received signal at the transceiver *2* is

$$\begin{aligned} \boldsymbol{r}_2(m) &= \boldsymbol{H}\boldsymbol{v}_1 s_1(m) + \boldsymbol{n}_2(m) \\ &= \boldsymbol{u}_1\lambda_1 s_1(m) + \boldsymbol{n}_2(m), \end{aligned} \tag{13}$$

where $\boldsymbol{u}_1$ is the top singular eigenvector of $\boldsymbol{U}$. The auto-correlation matrix of received signals at the transceiver *2* is

$$\boldsymbol{R}_{2s} = \boldsymbol{u}_1\boldsymbol{u}_1^{\dagger}\lambda_1^2 P_1 + \sigma_2^2\boldsymbol{I}. \tag{14}$$

Therefore $\boldsymbol{u}_1$ can be blindly estimated from the auto-correlation matrix also, and the transmit signal at transceiver *2* is $\boldsymbol{x}_2(m) = \boldsymbol{u}_1^{*} s_2(m)$. The received signal at the transceiver *1* is

$$\begin{aligned} \boldsymbol{r}_1(m) &= \boldsymbol{H}^T\boldsymbol{u}_1^{*} s_2(m) + \boldsymbol{n}_1(m) \\ &= \boldsymbol{v}_1^{*}\lambda_1 s_2(m) + \boldsymbol{n}_1(m). \end{aligned} \tag{15}$$

Consequently, the conjugated auto-correlation matrix of received signals at transceiver *1* can be expressed as

$$\boldsymbol{R}_{1s}^{*} = \boldsymbol{v}_1\boldsymbol{v}_1^{\dagger}\lambda_1^2 P_2 + \sigma_1^2\boldsymbol{I}. \tag{16}$$

Denote $\hat{\boldsymbol{R}}_{1f}^{*}$, $\hat{\boldsymbol{R}}_{2s}$, $\hat{\boldsymbol{R}}_{1s}^{*}$, $\hat{\boldsymbol{v}}_1$,$\hat{\boldsymbol{u}}_1$ as the blind estimates of $\boldsymbol{R}_{1f}^{*}$, $\boldsymbol{R}_{2s}$, $\boldsymbol{R}_{1s}^{*}$, $\boldsymbol{v}_1$,$\boldsymbol{u}_1$ respectively, due to the phase ambiguity introduced by SVD, the received signal at transceiver *1* and transceiver *2* can be transformed as

$$\tilde{\boldsymbol{r}}_2(m) = \hat{\boldsymbol{u}}_1^{\dagger}\boldsymbol{H}\hat{\boldsymbol{v}}_1 e^{j\theta_1} s_1(m) + \tilde{\boldsymbol{n}}_2(m) \tag{17}$$
$$\tilde{\boldsymbol{r}}_1(m) = \hat{\boldsymbol{v}}_1^{T}\boldsymbol{H}^T\hat{\boldsymbol{u}}_1^{*} e^{j\theta_2} s_2(m) + \tilde{\boldsymbol{n}}_1(m), \tag{18}$$

where θ_1 and θ_2 denote the phase ambiguities of the downlink and uplink respectively. Similarly, by using differential BPSK coding on $s_1(m)$ and $s_2(m)$, we can apply the same coherent detection scheme as in MISO system.

4. COMPARISON WITH TRAINING-BASED SCHEME

Training-based channel estimation are widely applied in the practical communication system. To benchmark the performance of the proposed blind scheme, we compare it with a Maximum Likelihood (ML) training-based scheme. Suppose that L symbols are used for the training and the communication begins from uplink transmission, we can stack the signals in (9) as

$$\boldsymbol{Y}_1 = \begin{bmatrix} \boldsymbol{r}_1(1) & \boldsymbol{r}_1(2) & \cdots & \boldsymbol{r}_1(L) \end{bmatrix}^T, \tag{19}$$
$$\boldsymbol{X}_2 = \begin{bmatrix} \boldsymbol{x}_2(1) & \boldsymbol{x}_2(2) & \cdots & \boldsymbol{x}_2(L) \end{bmatrix}^T, \tag{20}$$

where $\boldsymbol{X}_2$ is the training sequence and $\boldsymbol{Y}_1$ is the received signals in the training period. Then the ML estimation of the channel matrix $\boldsymbol{H}$ is

$$\hat{\boldsymbol{H}} = \left(\boldsymbol{X}_2^{\dagger}\boldsymbol{X}_2\right)^{-1}\boldsymbol{X}_2^{\dagger}\boldsymbol{Y}_1. \tag{21}$$

The left top singular eigenvector can be obtained at transceiver *1* by decomposing $\hat{\boldsymbol{H}}$ and then be used for the downlink beamforming. To coherently detect the signal for the downlink transmission, we need to estimate the right top singular eigenvector of $\boldsymbol{H}$ at transceiver *2*. It can be estimated by ML-training or by blind estimation as in the second stage of our proposed scheme. The scheme which uses the ML-training in both uplink and downlink will be called *"full training"*. The scheme which uses ML-training only in the uplink can be regarded as an expansion of our blind scheme with a training initiation, and will be called *"half training"*.

In the MIMO case, the training based schemes and our blind scheme may seem similar, as all schemes are composed of two stages. But is should be emphasized that in our blind scheme useful information is communicated between the two transceivers during the first stage, as opposed to a pure training stage. To make a fair comparison, both BER and rate should be considered. Since part of the time is used for the pure training, higher constellation (or lower coding rate) is need for the transmission of information bits in order to keep the same overall rate. Although blind channel estimation is weaker than training based channel estimation when an equal number of symbols are used, it is difficult to tell which is better if blind channel estimation uses more symbols than the training based channel estimation. Therefore we compare the proposed scheme with *"full training"* and *"half training"* schemes through simulations.

5. SIMULATIONS EXAMPLES

In our simulations, differentially BPSK modulation is applied for the proposed blind schemes. In the MIMO case, we transmit the differential Alamouti's code [4] for the uplink transmission in the first stage. The BER performance of the proposed blind schemes for TDD MISO and TDD MIMO system are shown in Fig. 1 and Fig. 2 respectively. We can see that, when 100 symbols are used to do blind estimation, the proposed blind scheme only has less than 2dB loss compared to ideal beamforming with complete CSI at the transmitter. We can also see that the proposed blind scheme is clearly superior to non-coherent space-time codes [4], even when only 10 symbols are used for blind channel estimation.

For the ML-training based schemes, we use space-time block codes with orthogonal design [5] in the first stage. In order to make a simple, fair comparison, we assume that half of the symbols are used as training for the ML-training based schemes, and that subsequently QPSK modulation (as opposed to BPSK for the proposed blind scheme) is applied for the transmission of information bits in order to keep the same overall rate as the proposed blind scheme. We compare the proposed blind scheme with ML-training based schemes in 2×2 MIMO system. The result is shown in Fig. 3. It is readily seen that our blind scheme overwhelmingly outperforms the *"full training"* scheme in both stages. The performance of the *"half training"* scheme is only slightly worse than our propose blind scheme in the second stage since both of them use the same blind estimation in the second stage. However, the non-coherent scheme still has about 2dB gain than the *"half training"* scheme in the first stage even for the long block ($Mu = 100$).

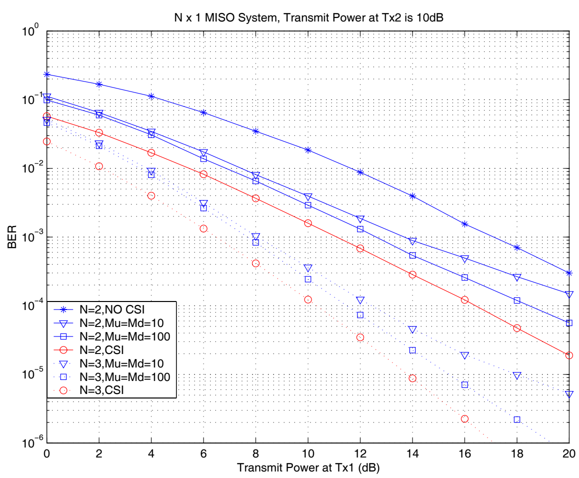

Fig. 1. BER of MISO system equipped N transceivers at Tx1. The variance of the noise is $\sigma_1^2 = \sigma_2^2 = 1$.

6. REFERENCES

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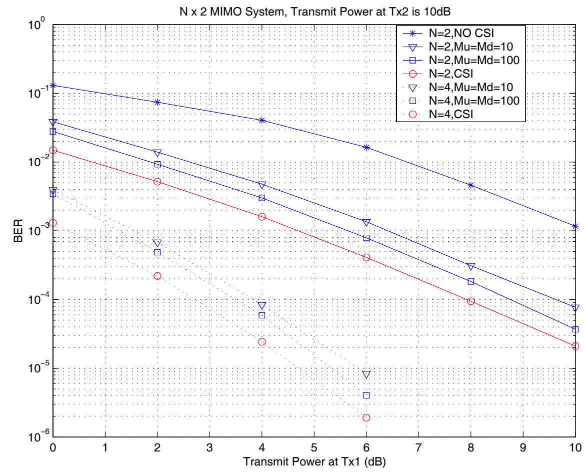

Fig. 2. BER of MIMO system equipped N transceivers at Tx1. The variance of the noise is $\sigma_1^2 = \sigma_2^2 = 1$.

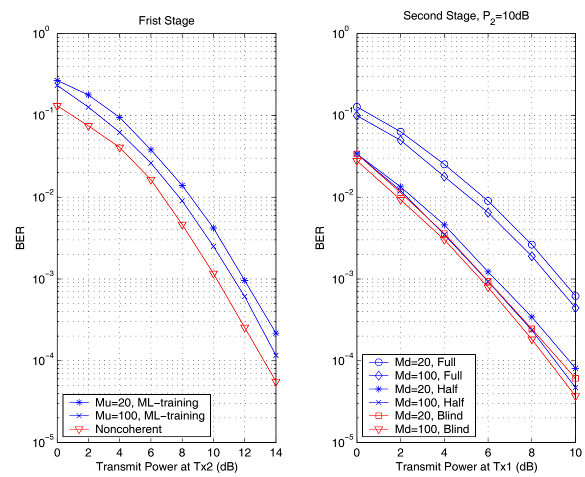

Fig. 3. Comparison with the ML-training schemes in 2×2 MIMO system. The variance of the noise is $\sigma_1^2 = \sigma_2^2 = 1$.

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