A ROBUST JOINT LINEAR PRECODER AND DECODER MMSE DESIGN FOR SLOWLY TIME-VARYING MIMO CHANNELS

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ABSTRACT

The joint linear precoder and decoder Minimum Mean Squared Error (MMSE) design represents a low complexity yet powerful solution for spatial multiplexing MIMO systems. Its performance, however, critically depends on the availability of timely Channel State Information (CSI) at both transmitter and receiver. In practice, the latter assumption can be severely challenged, due to channel time variations that lead to outdated CSI at the transmitter. State-of-the-art designs mistakenly use the outdated CSI to design the linear precoder and rely on the receiver to reduce the induced degradation. In this paper, we propose a robust Bayesian joint linear precoder and decoder solution that takes into account the uncertainty on the true channel given the outdated CSI at the transmitter. We finally assess the robustness of our design to channel time variations through Monte-Carlo analysis of the system's MMSE and average Bit-Error Rate (BER) performance.

1. INTRODUCTION

To enable spatial multiplexing MIMO systems, the joint linear precoder and decoder Minimum Mean Squared Error (MMSE) design has been proposed [1, 2]. It is a low complexity yet powerful design for applications, where the channel is slowly varying, such that the Channel State Information (CSI) can be made available at both sides of the transmission link. In fact, the latter design exploits this CSI to optimally allocate power across the transmitted data streams in order to reduce the system's Bit-Error Rate (BER). So far, however, most state-of-the-art contributions assume perfect CSI.

Channel time variations can compromise the availability of such perfect timely CSI at both transmitter and receiver. Such channel variations occur due to the wireless terminal movement or due to the movement of objects in the propagation environment. At the receiver, channel estimation is carried out using the preamble prepended to the data payload. If we omit channel estimation errors and assume not too long bursts, one can reasonably assume perfect timely receiver CSI as data and preamble undergo the same channel. This is not case at the transmitter side. In fact, whether the CSI is acquired through a feedback link from the receiver or through direct estimation using training from the receiver, there will always be a delay between the moment a channel realization is observed and the moment it is actually used by the transmitter. Combined with channel time-variations, this delay inevitably leads to outdated CSI at the transmitter. This outdated CSI is mistakenly used in the linear precoder calculation and leads to a degradation in the system's BER performance.

There is scarce literature on the impact of channel time variations on joint precoder and decoder designs. One such contribution is [10] that proposes 2 zero-forcing receiver solutions to reduce the degradation induced by the outdated CSI at the transmitter. The most relevant one estimates and equalizes for the equivalent channel composed of the true channel and the outdated precoder. This is possible because the pilots are also passed through the outdated precoder. Such a solution merely tries to reduce the degradation induced by the outdated CSI at the transmitter. In this contribution, we propose a solution that takes into account the uncertainty on the true channel due to time variations. Similar Bayesian approaches have been already proposed in other contexts such as beamforming for MISO systems [3, 4] and space-time coded MIMO systems [5, 6]. Such solutions, however, do not apply for spatial multiplexing scenarios. A single contribution [7] proposed a Bayesian approach in the context of joint precoder and decoder design. The latter contribution, however, considers a situation where both transmitter and receiver have the same imperfect CSI due to estimation errors. Hence, the proposed solution does not apply in the context of time-varying channels, where transmitter and receiver have different CSIs.

The rest of the paper is organized as follows: Section 2 introduces the data and outdated CSI models. Based on that, we derive our robust joint precoder and decoder MMSE design in Section 3. In Section 4, the performance improvements enabled by the proposed robust design are assessed. Finally, we draw some conclusions in Section 5.

2. SYSTEM MODEL

2.1. Data model

The spatial multiplexing MIMO system, under consideration, consists of a transmitter and a receiver, equipped with an M_T and M_R-element antenna respectively. At the transmitter, the input symbol stream $s(n)$ is demultiplexed into $p \leq \mathrm{Min}(M_R, M_T)$ independent streams, leading to an equivalent p-dimensional spatial symbol stream $\mathbf{s}(k)$. This spatial symbol stream $\mathbf{s}(k)$ is then

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passed through the linear precoder $\mathbf{T}$ before transmission through the M_T-element transmit antenna at rate $1/T$. At the receiver, the M_R complex baseband outputs from the M_R-element receive antenna sampled at rate $1/T$ are filtered by the linear decoder $\mathbf{R}$. The resulting p output streams conveying the detected spatial symbols $\hat{\mathbf{s}}(k)$ are then multiplexed and demodulated. For a flat-fading MIMO channel, the system equation is then given by:

$$\hat{\mathbf{s}}(k) = \mathbf{RHTs}(k) + \mathbf{Rn}(k) \tag{1}$$

where $\mathbf{n}(k)$ is the M_R-dimensional receive noise vector at time k and $\mathbf{H}$ is the $M_R \times M_T$ flat-fading channel matrix whose entries represent the complex channel gains from each transmit antenna to each receive antenna. In all the following, the time index k is dropped for clarity.

2.2. Outdated CSI model

As aforementioned, at time t, the *true* channel realization $\mathbf{H}_t$ is known at the receiver but not at the transmitter. Instead, the transmitter possesses an outdated channel information, $\mathbf{H}_{t-dT}$, corresponding to the channel state dT seconds ago. Under the assumption of dense scattering in the vicinity of both transmitter and receiver, we model the MIMO channel matrix $\mathbf{H}$ as a complex matrix whose entries are i.i.d zero-mean complex Gaussian variables with common variance σ_h^2; $\mathbf{H} \sim \mathcal{N}(\mathbf{0}_{M_R \times M_T}, \sigma_h^2 \mathbf{I}_{M_R M_T})$. $\mathbf{H}_t$ and $\mathbf{H}_{t-dT}$ are basically correlated realizations of the latter channel distribution. Thus, given the outdated CSI $\mathbf{H}_{t-dT}$, we can characterize the unknown current CSI $\mathbf{H}_t$ using the conditional CSI model introduced in [3, 4], as follows:

$$\mathbf{H}_t \sim \mathcal{N}(\rho \mathbf{H}_{t-dT}, \sigma_h^2(1-|\rho|^2)\mathbf{I}_{M_R M_T}) \tag{2}$$

where ρ is the common time-correlation of the i.i.d time-varying MIMO channel coefficients, defined as $\rho = E\{[\mathbf{H}_t]_{i,j}[\mathbf{H}_{t-dT}]_{i,j}^H\}/\sigma_h^2 = \mathcal{R}(dT)$, with $\mathcal{R}(dT)$ depending on the channel time-variation model.

3. A ROBUST JOINT LINEAR PRECODER AND DECODER MMSE SOLUTION

3.1. The state-of-the-art approach

The state-of-the-art approach [10] mistakenly assumes that the CSI available at the transmitter, $\mathbf{H}_{t-dT}$, is perfect. It designs the precoder $\mathbf{T}$ presuming that the receiver has the same CSI and implements the corresponding MMSE decoder $\mathbf{R}$ according to [1, 2]. More specifically, the transmitter designs the linear precoder $\mathbf{T}$, assuming that $\mathbf{T}$ and $\mathbf{R}$ are *jointly* designed to minimize the sum mean squared error subject to fixed average total transmit power P_T constraint as stated in:

$$\begin{cases} \text{Min}_{\mathbf{R},\mathbf{T}} \quad E_{\mathbf{s},\mathbf{n}}\left\{\|\ \mathbf{s} - (\mathbf{RH}_{t-dT}\mathbf{Ts} + \mathbf{Rn})\ \|_2^2\right\} \\ \quad \text{subject to:} \quad E_s \cdot trace(\mathbf{TT}^H) = P_T \end{cases} \tag{3}$$

We assume uncorrelated data symbols of average symbol energy E_s and zero-mean temporally and spatially-white complex Gaussian noise samples with common variance σ_n^2. In reality, the receiver has the timely CSI, $\mathbf{H}_t$, and can form the matched MMSE receiver given by:

$$\mathbf{R} = \left(\mathbf{T}^H\mathbf{H}_t^H\mathbf{H}_t\mathbf{T} + \frac{\sigma_n^2}{E_s}\mathbf{I}_p\right)^{-1}\mathbf{T}^H\mathbf{H}_t^H \tag{4}$$

Clearly, this approach is suboptimal. This is why, capitalizing on the previously introduced outdated CSI model, we next propose an improved joint precoder and decoder MMSE solution that takes into account the uncertainty about the CSI due to channel time variations.

3.2. Our proposed approach

We assume that the transmitter knows the conditional distribution of the true channel $\mathbf{H}_t$ and the structure of the receiver $\mathbf{R}$ given by (4). Consequently, instead of the ideal [1] design criterion of (3), the linear precoder $\mathbf{T}$ should be designed to minimize the conditional [2] sum mean squared error given the outdated CSI, subject to a fixed average total transmit power constraint:

$$\begin{cases} \text{Min}_{\mathbf{T}} \quad E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{E_{\mathbf{s},\mathbf{n}}\left\{\|\ \mathbf{s} - (\mathbf{RH}_t\mathbf{Ts} + \mathbf{Rn})\ \|_2^2\right\}\right\} \\ \quad \text{subject to:} \quad E_s \cdot trace(\mathbf{TT}^H) = P_T \end{cases}$$

For tractability, we approach the actual MMSE receiver of (4) by a zero-forcing receiver $\mathbf{R} = \left(\mathbf{T}^H\mathbf{H}_t^H\mathbf{H}_t\mathbf{T}\right)^{-1}\mathbf{T}^H\mathbf{H}_t^H$ while designing $\mathbf{T}$. Consequently, the general MMSE optimization problem can be reduced to:

$$\begin{cases} \text{Min}_{\mathbf{T}} \quad E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{E_{\mathbf{n}}\left\{trace(\mathbf{Rn}(\mathbf{Rn})^H)\right\}\right\} \\ \quad \text{subject to:} \quad E_s \cdot trace(\mathbf{TT}^H) = P_T \end{cases} \tag{6}$$

Resorting to the Lagrange multiplier techniques to solve the above optimization problem, the cost function can be written as:

$$\mathcal{C} = trace\left(E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{\sigma_n^2(\mathbf{T}^H\mathbf{H}_t^H\mathbf{H}_t\mathbf{T})^{-1}\right\} + \lambda E_s\mathbf{TT}^H\right) \tag{7}$$

We have previously stated that, given the outdated CSI, the true channel follows the complex normal distribution of (2). Thus, based on [11], we can identify $\mathbf{T}^H\mathbf{H}_t^H\mathbf{H}_t\mathbf{T}$ as a non-central Wishart distribution. To the best of our knowledge, the calculation of the expectation of its inverse, as required in (7), is still an open problem. The only solutions available are for the simple case where $\mathbf{H}_{t-dT}\mathbf{T}$ is of rank 1, which does not apply here. Consequently, in the following, we investigate an approximate solution. To do so, using the channel distribution of (2), we first instantiate the true channel $\mathbf{H}_t$ as $\mathbf{H}_t = \hat{\mathbf{H}}_{eq} + \Delta$, where $\hat{\mathbf{H}}_{eq} = \rho\mathbf{H}_{t-dT}$ and Δ is the $\mathcal{N}\left(\mathbf{0}_{M_R \times M_T}, \sigma_h^2(1-|\rho|^2)\mathbf{I}_{M_R M_T}\right)$-distributed uncertainty on the true channel given the outdated CSI. The conditional mean in (7) can then be developed into:

$$E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{(\mathbf{T}^H\mathbf{H}_t^H\mathbf{H}_t\mathbf{T})^{-1}\right\} = E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{\left[\underbrace{\mathbf{T}^H\hat{\mathbf{H}}_{eq}^H\hat{\mathbf{H}}_{eq}\mathbf{T}}_{\mathbf{A}} + \underbrace{\mathbf{T}^H(\hat{\mathbf{H}}_{eq}^H\Delta + \Delta^H\hat{\mathbf{H}}_{eq} + \Delta^H\Delta)\mathbf{T}}_{\mathbf{C}}\right]^{-1}\right\} \tag{8}$$

Iteratively using the matrix inversion lemma, the inner inverse can be expressed as:

$$(\mathbf{A}+\mathbf{C})^{-1} = \mathbf{A}^{-1} - \mathbf{A}^{-1}\mathbf{CA}^{-1} + \mathbf{A}^{-1}\mathbf{CA}^{-1}\mathbf{CA}^{-1} - \mathbf{A}^{-1}\mathbf{CA}^{-1}\mathbf{CA}^{-1}\mathbf{CA}^{-1} + \cdots \tag{9}$$

[1] corresponding to the ideal case where both sides of the link have the same perfect timely CSI.

[2] on the true channel.

Evaluating the conditional expectation of the above expression leads to:

$$E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{(\mathbf{A}+\mathbf{C})^{-1}\right\} = \mathbf{A}^{-1} - \mathbf{A}^{-1}E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{\mathbf{C}\right\}\mathbf{A}^{-1} + \mathbf{A}^{-1}E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{\mathbf{C}\mathbf{A}^{-1}\mathbf{C}\right\}\mathbf{A}^{-1} - \cdots \quad (10)$$

In the latter expression, we shall only keep the terms corresponding to contributions up to the 2^{nd} order statistics of the uncertainty Δ. The latter contributions are contained in the 3 first terms. Consequently, we now explicit these 3 first terms in order to extract the relevant contributions. The first term solely depends on the equivalent channel $\hat{\mathbf{H}}_{eq}$:

$$\mathbf{A}^{-1} = (\mathbf{T}^H\hat{\mathbf{H}}_{eq}^H\hat{\mathbf{H}}_{eq}\mathbf{T})^{-1} \quad (11)$$

Using the fact that the equivalent channel $\hat{\mathbf{H}}_{eq}$ and the uncertainty Δ are uncorrelated as well as the distribution of the uncertainty Δ, the second term in (10) can be reduced to:

$$\mathbf{A}^{-1}E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{\mathbf{C}\right\}\mathbf{A}^{-1} = \sigma_h^2(1-|\rho|^2)M_R\mathbf{A}^{-1}\mathbf{T}^H\mathbf{T}\mathbf{A}^{-1} \quad (12)$$

The relevant part of the third term of (10), that contains only up to 2^{nd} order statistics of the uncertainty Δ, is given by:

$$\mathbf{A}^{-1}E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{\mathbf{C}\mathbf{A}^{-1}\mathbf{C}\right\}\mathbf{A}^{-1} \approx \mathbf{A}^{-1}E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\Big\{\mathbf{T}^H(\hat{\mathbf{H}}_{eq}^H\Delta + \Delta^H\hat{\mathbf{H}}_{eq})\mathbf{T}\mathbf{A}^{-1}\mathbf{T}^H(\hat{\mathbf{H}}_{eq}^H\Delta + \Delta^H\hat{\mathbf{H}}_{eq})\mathbf{T}\Big\}\mathbf{A}^{-1} \quad (13)$$

The previous expression can be further simplified based on the observation that only the cross-products are non-zero:

$$\mathbf{A}^{-1}E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{\mathbf{C}\mathbf{A}^{-1}\mathbf{C}\right\}\mathbf{A}^{-1} \approx \mathbf{A}^{-1}E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\Big\{\mathbf{T}^H(\hat{\mathbf{H}}_{eq}^H\Delta\mathbf{T}\mathbf{A}^{-1}\mathbf{T}^H\Delta^H\hat{\mathbf{H}}_{eq} + \Delta^H\hat{\mathbf{H}}_{eq}\mathbf{T}\mathbf{A}^{-1}\mathbf{T}^H\hat{\mathbf{H}}_{eq}^H\Delta)\mathbf{T}\Big\}\mathbf{A}^{-1} \quad (14)$$

Given an $M_T \times M_T$ matrix $\mathbf{M}$, it is straightforward to establish that:

$$\begin{cases} E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{\Delta\mathbf{M}\Delta^H\right\} = \sigma_h^2(1-|\rho|^2)trace(\mathbf{M})\mathbf{I}_{M_R} \\ E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{\Delta^H\mathbf{M}\Delta\right\} = \sigma_h^2(1-|\rho|^2)trace(\mathbf{M})\mathbf{I}_{M_T} \end{cases} \quad (15)$$

The previous observation can be applied to simplify (14) as follows:

$$\mathbf{A}^{-1}E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{\mathbf{C}\mathbf{A}^{-1}\mathbf{C}\right\}\mathbf{A}^{-1} \approx \sigma_h^2(1-|\rho|^2)\Big(trace(\mathbf{T}\mathbf{A}^{-1}\mathbf{T}^H)\mathbf{A}^{-1}\mathbf{T}^H\hat{\mathbf{H}}_{eq}^H\hat{\mathbf{H}}_{eq}\mathbf{T}\mathbf{A}^{-1} + trace(\hat{\mathbf{H}}_{eq}\mathbf{T}\mathbf{A}^{-1}\mathbf{T}^H\hat{\mathbf{H}}_{eq}^H)\mathbf{A}^{-1}\mathbf{T}^H\mathbf{T}\mathbf{A}^{-1}\Big) \quad (16)$$

Recalling (11) and the fact that $trace(\mathbf{YZ}) = trace(\mathbf{ZY})$, the third term of (10) can finally be reduced to:

$$\mathbf{A}^{-1}E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{\mathbf{C}\mathbf{A}^{-1}\mathbf{C}\right\}\mathbf{A}^{-1} \approx \sigma_h^2(1-|\rho|^2)\Big(trace(\mathbf{T}\mathbf{A}^{-1}\mathbf{T}^H)\mathbf{A}^{-1} + p\mathbf{A}^{-1}\mathbf{T}^H\mathbf{T}\mathbf{A}^{-1}\Big)$$

Based on the above calculations, (10) can be expressed as:

$$E_{\mathbf{H}_t|\mathbf{H}_{t-dT}}\left\{(\mathbf{A}+\mathbf{C})^{-1}\right\} \approx \mathbf{A}^{-1} + \sigma_h^2(1-|\rho|^2)\Big[(p-M_R)\mathbf{A}^{-1}\mathbf{T}^H\mathbf{T}\mathbf{A}^{-1} + trace(\mathbf{T}\mathbf{A}^{-1}\mathbf{T}^H)\mathbf{A}^{-1}\Big] \quad (17)$$

Let $\mathbf{T} = \mathbf{U}_T\Sigma_T\mathbf{V}_T^H$ be the singular value decomposition of the precoder $\mathbf{T}$. On the one hand, it is clear that $\mathbf{V}_T$ does not alter the cost function of (7) so it can be simply set to identity. On the other hand, state-of-the-art literature shows that, given the equivalent channel $\hat{\mathbf{H}}_{eq} = \mathbf{U}_{t-dT}\hat{\Sigma}_{eq}\mathbf{V}_{t-dT}^H$, the optimal transmit strategy is to beamform into the eigenmodes of the mean channel i.e $\mathbf{U}_T = \mathbf{V}_{t-dT}$. The remaining unknown is the optimal power allocation policy, given by Σ_T, that we subsequently determine through minimizing the cost function of (7). To do so, we replace $\mathbf{T} = \mathbf{V}_{t-dT}\Sigma_T$ and $\mathbf{A}^{-1} = (\hat{\Sigma}_{eq}^2\Sigma_T^2)^{-1}$ in (17). This allows us to explicit the cost-function of (7) in Σ_T:

$$\mathcal{C} \approx \sigma_n^2 trace\Big[\sigma_h^2(1-|\rho|^2)(p-M_R)\hat{\Sigma}_{eq}^{-4}\Sigma_T^{-2} + \Big(1+\sigma_h^2(1-|\rho|^2)trace(\hat{\Sigma}_{eq}^{-2})\Big)\hat{\Sigma}_{eq}^{-2}\Sigma_T^{-2} + \frac{\lambda E_s}{\sigma_n^2}\Sigma_T^2\Big] \quad (18)$$

We then differentiate the previous cost-function with respect to Σ_T, what results into:

$$\frac{d\mathcal{C}}{d\Sigma_T} \approx \sigma_n^2\Big[-\sigma_h^2(1-|\rho|^2)(p-M_R)\hat{\Sigma}_{eq}^{-4}\Sigma_T^{-3} - \Big(1+\sigma_h^2(1-|\rho|^2)trace(\hat{\Sigma}_{eq}^{-2})\Big)\hat{\Sigma}_{eq}^{-2}\Sigma_T^{-3} + \frac{\lambda E_s}{\sigma_n^2}\Sigma_T\Big] \quad (19)$$

Finally, setting to zero the previous differential, enables us to identify the optimal Σ_T, solution to the optimization problem formulated in (6), as follows:

$$\Sigma_T = \Bigg(\frac{\sigma_n^2}{\lambda E_s}\Big[\sigma_h^2(1-|\rho|^2)(p-M_R)\hat{\Sigma}_{eq}^{-4} + \Big(1+\sigma_h^2(1-|\rho|^2)trace(\hat{\Sigma}_{eq}^{-2})\Big)\hat{\Sigma}_{eq}^{-2}\Big]^+\Bigg)^{1/4} \quad (20)$$

where λ is the Lagrange multiplier to be calculated to satisfy the power constraint.

4. PERFORMANCE RESULTS

In this section, we illustrate the improvements, in terms of average MMSE and BER, that our robust joint linear precoder and decoder MMSE design offers over state-of-the-art designs that simply ignore the fact that the CSI at the transmitter is outdated. In order to do that, we use the well-known Jakes model [8] to instantiate a realistic outdated CSI model based on (2). Under the assumption of isotropic scattering and moving terminal, this model describes the time-correlation function as $\mathcal{R}(dT) = J_0(2\pi f_D dT)$, where J_0 is the zero-th order Bessel function of the first kind and f_D is the Doppler frequency. In our simulations, we use the parameters which have been standardized for this model in the context of indoor WLANs [9] as it is a potential application for joint linear precoder and decoder designs. In particular, we consider a Doppler frequency of 50 Hz. Finally, as aforementioned, the receiver possesses and uses perfect timely CSI to form the MMSE receiver of (4). We further consider the case of a $(4, 4)$ MIMO set-up with 2 QPSK-modulated data streams. Figures 1 and 2 respectively plot this set-up's average MMSE and BER performances, for a normalized delay $f_D dT = 2$ corresponding to a time-correlation $\rho = 0.15$. Clearly, our robust joint linear precoder and decoder MMSE design exhibits a lower average MMSE and consequently a lower average BER performance, when compared to the state-of-the-art design. More specifically, Figure 2 shows that our robust

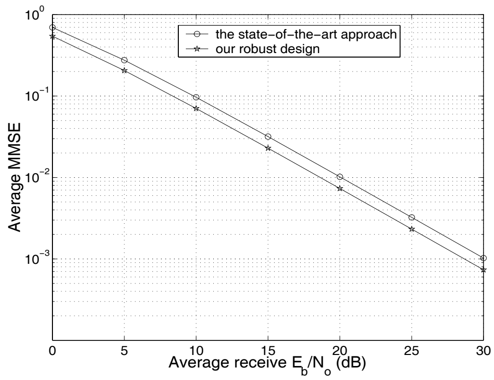

Fig. 1. *Average MMSE comparison for a (4,4) MIMO set-up with 2 streams at spectral efficiency of* 4 *bits/s/Hz and* $\rho = 0.15$

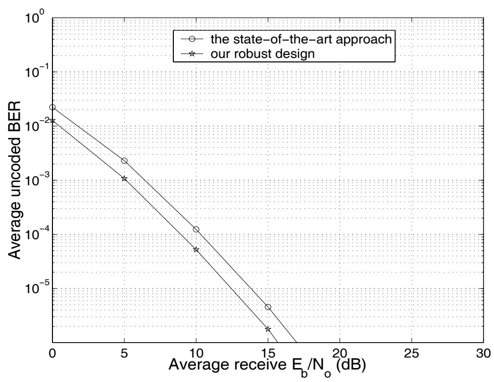

Fig. 2. *Average uncoded BER comparison for a (4,4) MIMO set-up with 2 streams at spectral efficiency of* 4 *bits/s/Hz and* $\rho = 0.15$

design offers a 1.5 dB SNR gain over the state-of-the-art design at average $BER = 10^{-3}$. Figure 3 further compares the MMSE performance of our robust design to that of the state-of-art design over a large range of delays at a fixed average receive $E_b/N_o = 20$ dB. For delays up to 10^{-1} corresponding to time correlations larger than 0.9, both designs exhibit the same average MMSE. However, as the delay increases and the time correlation gets low, our design clearly outperforms the state-of-the-art design. This is due to the fact that our robust design takes into account the uncertainty around the true channel, due to channel time variations, in the design of the precoder **T**. To do so, our robust design requires only the additional estimation of the time correlation ρ.

5. CONCLUSIONS

In this paper, we have proposed a robust linear precoder and decoder MMSE design that exploits the knowledge of the conditional channel distribution, given the available outdated CSI. We have also shown that our robust approach outperforms the state-of-the art approach in terms of MMSE and BER in slowly time-varying scenarios.

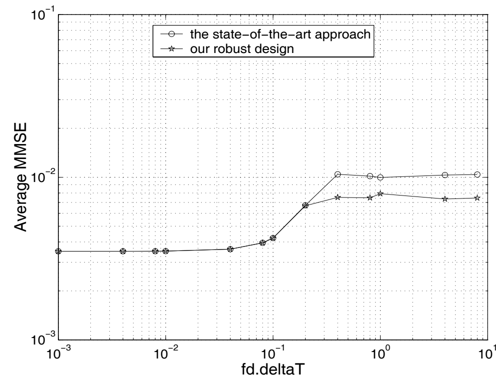

Fig. 3. *Average MMSE versus time for a (4,4) MIMO set-up with 2 streams at spectral efficiency of* 4 *bits/s/Hz and* $E_b/N_o = 20$ *dB*

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