Covariance-Based Linear Precoding

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***Abstract*—This paper extends the self-contained theory of linear precoding to the field of covariance based spatio-temporal downlink processing for direct-sequence *code-division multiple-access* (CDMA) systems and shows the applicability to the release 6 of *high-speed downlink packet access* (HSDPA). To this end, a unifying theory is developed to formulate the three known linear filters, namely, the transmit matched filter, the transmit zero-forcing filter, and the transmit Wiener filter, as optimization problems even in systems, where only covariance knowledge is available at the transmitter. Second, the solutions of these transmit filters are given for such systems with partial *channel state information* (CSI). Finally, it is shown how covariance-based linear precoding can be employed in the new generation CDMA system HSDPA, i.e., how channel estimation on the secondary common pilot channel allows for optimum full rank linear precoding employing only partial CSI.**

***Index Terms*—Code-division multiple-access (CDMA), common pilot channels, covariance-based precoding, high-speed downlink packet access (HSDPA), linear precoding, multiple-input–single-output (MISO) downlink, partial channel state information (CSI), transmit processing.**

I. Introduction

DUE TO THE deorthogonalization of the *code-division multiple-access* (CDMA) sequences by the frequency-selective channel, next-generation CDMA systems will mainly suffer from interference limitation. When the receivers in such systems are *noncooperative* (*broadcast channel*, see, e.g., [1]), i.e., the estimate for the transmit signal is generated without the signals of the other receivers (e.g., downlink of a cellular communication system), receive processing like in [2] is suboptimum. In the case of noncooperative receivers and cooperative transmitters, *transmit processing* has to be used instead, where the data signals are jointly transformed by the precoding filter before they are transmitted. The resulting received signals are already equalized and only simple receive filters are needed, since the channel acts like an "equalizer" of the precoding filter. Similar to linear receive processing [2], we can identify three basic linear transmit filter types: the *transmit matched filter* (TxMF), the *transmit zero-forcing filter* (TxZF), and the *transmit Wiener filter* (TxWF). The TxMF [3]–[5], also known as *prerake* in CDMA systems, maximizes the desired signal portion in the received signal, whereas the TxZF [6]–[11] completely suppresses the interference. The TxWF [12]–[15] finds a tradeoff between the other two linear transmit filters by minimizing the *mean square error* (MSE). With their ability to combat both noise and interference influences linear precoding techniques have proven their great potential for *direct-sequence* (DS)-CDMA systems (e.g., [13]).

Due to the lacking reciprocity of uplink and downlink, the standard assumption that full *channel state information* (CSI) is available at the transmitter cannot be fulfilled in *frequency-division duplex* (FDD) CDMA systems unless CSI is fed back from the receivers to the transmitter. However, we do not follow this approach to avoid the signaling overhead. In fact, we propose to design the precoding filters based on *partial CSI*, i.e., on covariance knowledge. As shown in, e.g., [16]–[18], the covariance matrix can be transformed from the first to the second FDD frequency. Additionally, the underlying *long-term channel properties* are very slowly time varying. Therefore, the long-term channel properties are not affected by Doppler shift and no covariance compensation is needed. The other channel properties, i.e., the fast fading coefficients of the eigenmodes for the different delays, are not known to the transmitter and cannot, thus, be used for the precoder design. To ensure a coherent detection at the receiver, we assume that each of the receivers is equipped with a Rake (see [19]), that is a receive filter matched to the channel and the signature of the signal (i.e., CDMA code sequence).

For a CDMA system with partial CSI, Montalbano *et al.* developed a covariance-based transmit filter in [20] aiming at the suppression of intracell interference. In [21], Foster *et al.* developed a similar transmit filter for *Global System for Mobile Communications* (GSM) in the frequency domain. By maximizing the minimum *signal-to-interference-plus-noise ratio* (SINR), Schubert *et al.* in [22] proposed an optimal scheme that bases on covariance knowledge as well. This optimization cannot be solved analytically though and an iterative procedure of high complexity is necessary [22], [23]. Moreover, the approach has not been extended to frequency-selective channels yet and cannot be applied to systems with *common pilot channels* (CPICH) in general.

Since the transmitter is unable to find an expression for the signal at the receive filter output due to the lack of full CSI, we propose to form a model based on the partial CSI with signals, whose average power is the same as the average power of the respective signals in a model with full CSI. With the model based on partial CSI, we are able to design linear precoder for FDD CDMA systems. To this end, this paper is organized as follows. Section II formulates the exact signal model for the DS-CDMA system from which Section III derives a power equivalent long-term signal model, which describes the relevant signal components independent of the channel coefficients and exclusively basing on covariance knowledge. Building upon

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this design model, Section IV formulates the optimization problems for the partial CSI versions of TxMF, TxZF, and TxWF, and gives the corresponding solutions. Finally, Section V adapts the introduced methods to the next-generation CDMA system HSDPA and proves that linear precoding can be applied to systems employing CPICH channel estimation.

II. System Model and Assumptions

Sparing protocol and channel coding considerations, this paper focuses on signal processing in the downlink of a wireless communication system. The transmitting *base station* (BS) shall be provided with partial CSI and shall be equipped with multiple antenna elements. As the *mobile stations* (MSs) are assumed to be single antenna devices and a total number of K users is assumed to be present in the link, the system classifies as a *multiuser* (MU) *multiple-input–single-output* (MISO) system.

The multiple users are served in CDMA[1] with spreading sequences of length χ. The signal $s_k[n]$ in Fig. 1, thus, represents an upsampled version of the unspread symbol sequence for user k. The multiplication with the spreading code is implemented through the precoding filters $\boldsymbol{p}_k[n]$. The receivers at each of the MSs, which have full CSI, is assumed to consist of a code matched filter and a *maximum ratio combining* (MRC) rake receiver for channel equalization. The output signals $\hat{s}_k[n]$ are, thereafter, downsampled and decided.

The following sections aim at modeling the system in Fig. 1 by linear components with *finite-duration impulse responses* (FIRs), i.e., assuming a sequential data transmission instead of blockwise processing. To this end, Toeplitz structured convolutional matrices will be employed to represent the discrete time convolutions on a chip level.

Throughout this article, variables are denoted in italic script, where vectors and matrices are denoted in bold lower and upper case letters. Furthermore, $s^{\natural}[n]$ denotes the noise-free counterpart of $s[n]$. The accents $\hat{(\cdot)}$ and $\tilde{(\cdot)}$ are used as symbols for estimators based on full CSI and partial CSI, respectively. Moreover, $\mathrm{E}[\cdot]$, $(\cdot)^*$, $(\cdot)^{\mathrm{T}}$, $(\cdot)^{\mathrm{H}}$, and $\otimes$ denote the expectation operator, complex conjugation, transposition, complex conjugate transposition, and the Kronecker product, respectively. Finally, $\mathbf{1}_m$ and $\mathbf{0}_{m,n}$ denote the $m \times m$ identity and the $m \times n$ zero matrix, respectively.

A. Precoding

Major goal of this contribution is the design of spatio-temporal transmit filters. The key idea within is, that the receive processing unit adapts to the channel only, which allows for small low power mobile devices. The precoder of user k thereupon adapts to the resulting combination of channel and receive filter, for which it can employ the N_a spatial and $L+1$ temporal degrees of freedom of its vector valued impulse response

$$\boldsymbol{p}_k[n] = \sum_{l=0}^{L} \boldsymbol{p}_{k,l}\delta[n-l] \quad \in \mathbb{C}^{N_\mathrm{a}}.$$

[1]Note that CDMA is chosen exemplarily and that all following derivations can be applied to any multiple-access scheme.

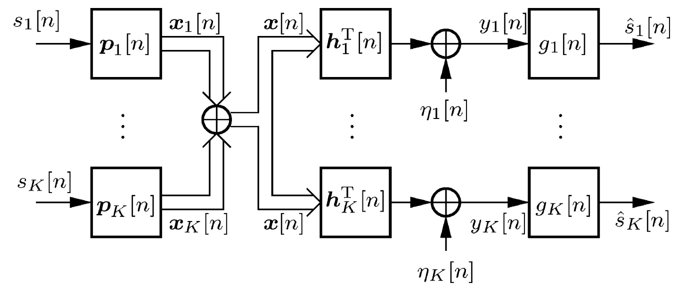

Fig. 1. Block diagram of a MU MISO system.

Here, $\boldsymbol{p}_{k,l} \in \mathbb{C}^{N_\mathrm{a}}$ denotes the lth vector filter coefficient of the precoding filter for user k, L is the temporal order of the transmit filter, and N_a is the number of transmit antennas. This precoding filter is applied to the signal $s_k[n]$, which is obtained as an upsampled[2] version of the symbol sequence, that is $s_k[n]$ results from the symbol sequence with variance σ_s^2 by inserting $\chi - 1$ zeros between two consecutive symbols. Adapting to the combination of channel and receive filter, the filter solutions for $\boldsymbol{p}_k[n]$ will automatically implement the necessary spreading code. Through a discrete-time convolution the linear precoder yields the antenna array transmit signal $\boldsymbol{x}[n]$ through the summation over all K users as

$$\boldsymbol{x}[n] = \sum_{i=1}^{K} \sum_{l=0}^{L} \boldsymbol{p}_{i,l} s_i[n-l].$$

B. Channel Model

The channel model in (2) summarizes the combination of pulse shaping, D/A conversion, and *radio frequency* (RF) components with the spatio-temporal radio channel itself in an equivalent time discrete baseband channel model. Without loss of generality, the number of temporally resolvable channel paths for all users equals $Q+1$.[3] Thus, one channel realization per user consists of $Q+1$ channel vectors $\boldsymbol{h}_{k,q} \in \mathbb{C}^{N_\mathrm{a}}$, $q = 0, \ldots, Q$. With this definition the impulse responses of the channels read as

$$\boldsymbol{h}_k[n] = \sum_{q=0}^{Q} \boldsymbol{h}_{k,q}\delta[n-q] \quad \in \mathbb{C}^{N_\mathrm{a}}.$$

As the channel coefficients are influenced by an extremely large number of complex phenomena, a deterministic model is of prohibitive complexity. The vectors $\boldsymbol{h}_{k,q}$, thus, are modeled as the realizations of correlated stochastic processes, whose characteristics shall be discussed in the sequel. We define the covariance matrix of the qth vector channel coefficient for user k as

$$\boldsymbol{R}_{k,q} = \mathrm{E}\left[\boldsymbol{h}_{k,q}\boldsymbol{h}_{k,q}^{\mathrm{H}}\right] = \sum_{\zeta=1}^{N_\mathrm{a}} \lambda_{k,q,\zeta}\boldsymbol{u}_{k,q,\zeta}\boldsymbol{u}_{k,q,\zeta}^{\mathrm{H}} \tag{1}$$

where $\boldsymbol{U}_{k,q} = [\boldsymbol{u}_{k,q,1}, \ldots, \boldsymbol{u}_{k,q,N_\mathrm{a}}]$ contains the orthonormal eigenvectors $\boldsymbol{u}_{k,q,\zeta}$, $\zeta = 1, \ldots, N_\mathrm{a}$, of the channel covariance

[2]Upsampling and downsampling are not considered explicitely.

[3]Equal number of paths can be assured by adding zero power paths.

matrix[4] $\boldsymbol{R}_{k,q}$ and $\boldsymbol{\Lambda}_{k,q} = \mathrm{diag}(\lambda_{k,q,1},\ldots,\lambda_{k,q,N_\mathrm{a}})$ is a diagonal matrix containing the corresponding eigenvalues. Hence, we can always decompose the channel vector $\boldsymbol{h}_{k,q}$ into a linear combination of the corresponding eigenvectors $\boldsymbol{u}_{k,q,\zeta}$ and obtain a model for the channel to the kth MS

$$\boldsymbol{h}_k[n] = \sum_{q=0}^{Q}\sum_{\zeta=1}^{N_\mathrm{a}} \rho_{k,q,\zeta}\boldsymbol{u}_{k,q,\zeta}\delta[n-q] \quad \in \mathbb{C}^{N_\mathrm{a}} \tag{2}$$

where the coefficients $\rho_{k,q,\zeta}$ of this Karhunen–Loeve decomposition are called *fast fading coefficients*. With this channel model, the above introduced transmit signal $\boldsymbol{x}[n]$, and the complex Gaussian noise model $\eta_k[n] \sim \mathcal{N}_\mathbb{C}(0,\sigma_\eta^2)$, the signal at the kth receiver's input can be formulated as

$$y_k[n] = \sum_{q=0}^{Q}\sum_{\zeta=1}^{N_\mathrm{a}} \rho_{k,q,\zeta}\boldsymbol{u}_{k,q,\zeta}^{\mathrm{T}}\boldsymbol{x}[n-q] + \eta_k[n]. \tag{3}$$

C. Receive Processing

With the assumption of an MRC Rake receiver combined with a code matched filter, the impulse response $g_k[n]$ of the kth receive filter consists of two parts. The first component is the correlation with the complex spreading code $c_k[n]$ of user k by convolving with

$$g_k^{(c)}[n] = \sum_{j=0}^{\chi-1} c_k^*[j]\delta[n+j]$$

where we assume the normalization $\sum_{j=0}^{\chi-1}|c_k[j]|^2 = \chi$. Second, the receiver equalizes the channel with an MRC Rake filter. Due to its single antenna, the receiver has no means to access the spatial domain and the rake filter, thus, is a scalar filter. To determine the corresponding scalar coefficients in an MRC way, the receive filter estimates the channel coefficients employing the pilot channel. Allowing for arbitrary pilot channels, i.e., pilot precoding weights $\boldsymbol{b}_k$, the receive filter can be written as the time-reversed complex conjugate of the resulting scalar pilot channel $\boldsymbol{b}_k^{\mathrm{T}}\boldsymbol{h}_k[n]$

$$g_k^{(\rho)}[n] = \sum_{f=0}^{F}\sum_{\xi=1}^{N_\mathrm{a}} \underbrace{\boldsymbol{b}_k^{\mathrm{H}}\boldsymbol{u}_{k,f,\xi}^*}_{\nu_{k,f,\xi}^*}\rho_{k,f,\xi}^*\delta[n+f]. \tag{4}$$

Note that the order of the channel matched filter $g_k^{(\rho)}[n]$ is $F \leq Q$, i.e., our notation allows the truncation of the pilot channel impulse response (of order Q) to get rid of negligible coefficients. The factors $\nu_{k,f,\xi} = \boldsymbol{b}_k^{\mathrm{T}}\boldsymbol{u}_{k,f,\xi}$ originate from the combination of the beamforming $\boldsymbol{b}_k$ on the pilot channel and the channels' eigencomponents. With these two parts of the receive filter, the formulation of the decision signal $\hat{s}_k[n]$ is obtained as [see (3)]

$$\begin{aligned}\hat{s}_k[n] &= g_k^{(c)}[n] * g_k^{(\rho)}[n] * y_k[n]\\ &= \sum_{j=0}^{\chi-1} c_k^*[j]\sum_{f=0}^{F}\sum_{\xi=1}^{N_\mathrm{a}} \nu_{k,f,\xi}^*\rho_{k,f,\xi}^* y_k[n+f+j]. \end{aligned}\tag{5}$$

[4]As the processes are zero-mean, covariance and correlation are the same. We also assume that the channels of different users and the coefficients of different paths are independent. Thus, $\mathrm{E}[\boldsymbol{h}_{k,q}\boldsymbol{h}_{i,f}^{\mathrm{H}}] = \boldsymbol{0}$ for $k \neq i$ or $q \neq f$.

Note that the desired value for $\hat{s}_k[m\chi]$ is $s_k[m\chi]$, where $m \in \mathbb{Z}$. Since the receiver performs a downsampling of the receive filter output $\hat{s}_k[n]$, the samples for $n \neq m\chi$ can have arbitrary values.

III. Covariance-Based Design Model

The system model in (5) allows us to formulate the relevant optimization criteria for linear precoding, as presented in [13] and [15], and to obtain optimum precoding solutions. As derived from (5), these solutions though will inherently depend upon all system parameters appearing in (5). As the knowledge about $\rho_{k,q,\zeta}$ is not available to the BS, the resulting solutions are not applicable to systems with only partial CSI. In Section III-B, we therefore shall derive an equivalent system model, that approximates $\hat{s}_k[n]$ in a sense of optimality, as discussed in Section III-A, and does not employ the coefficients $\rho_{k,q,\zeta}$, but bases the system model exclusively on covariance knowledge.

A. Heuristic of the Covariance-Based Design Model

For the development of a suitable heuristic, we recall (1) and (2) to establish a connection between the given covariance matrix and the stochastic variable $\rho_{k,q,\zeta}$

$$\mathrm{E}\left[\boldsymbol{h}_{k,q}\boldsymbol{h}_{k,q}^{\mathrm{H}}\right] = \sum_{\zeta=1}^{N_\mathrm{a}} \mathrm{E}\left[|\rho_{k,q,\zeta}|^2\right]\boldsymbol{u}_{k,q,\zeta}\boldsymbol{u}_{k,q,\zeta}^{\mathrm{H}}$$

which in comparison with the eigendecomposition of the covariance matrix from (1)

$$\mathrm{E}\left[\boldsymbol{h}_{k,q}\boldsymbol{h}_{k,q}^{\mathrm{H}}\right] = \sum_{\zeta=1}^{N_\mathrm{a}} \lambda_{k,q,\zeta}\boldsymbol{u}_{k,q,\zeta}\boldsymbol{u}_{k,q,\zeta}^{\mathrm{H}}$$

proves that the eigenvalue $\lambda_{k,q,\zeta}$ is the second moment $\sigma_{k,q,\zeta}^2 = \mathrm{E}[|\rho_{k,q,\zeta}|^2]$ of the channel coefficient $\rho_{k,q,\zeta}$. We see that covariance knowledge allows to determine the variance of signal components. Consequently, the *heuristic* objective is a *variance true approximation* of all signal components. With this objective, Section III-B derives a power equivalent signal model independent of the coefficients $\rho_{k,q,\zeta}$ unknown to the transmitting BS.

B. Derivation of the Covariance-Based Design Model

The $\rho_{k,q,\zeta}$ independent signal model originates from the second-order moments of all relevant noise-free signal components $\hat{s}_{k,q,\zeta,f}^{\natural}[n]$, i.e., the signal portions at MS k that traveled over the ζth eigencomponent of the qth channel path and the fth rake finger. Note that the noise-free decision signal is

$$\hat{s}_k^{\natural}[n] = \sum_{q=0}^{Q}\sum_{\zeta=1}^{N_\mathrm{a}}\sum_{f=0}^{F} \hat{s}_{k,q,\zeta,f}^{\natural}[n].$$

By rearranging the noise-free part of (5) these signal components are found to be

$$\begin{aligned}\hat{s}_{k,q,\zeta,f}^{\natural}[n] = \sum_{j=0}^{\chi-1}\sum_{\xi=1}^{N_\mathrm{a}}\sum_{i=1}^{K}\sum_{l=0}^{L} & c_k^*[j]\nu_{k,f,\xi}^*\rho_{k,f,\xi}^*\\ &\times \rho_{k,q,\zeta}\boldsymbol{u}_{k,q,\zeta}^{\mathrm{T}}\boldsymbol{p}_{i,l}s_i[n-l-q+f+j].\end{aligned}$$

Aiming for a power equivalent but $\rho_{k,q,\zeta}$ independent representation of these components, the expected value of $|\hat{s}^{\not\eta}_{k,q,\zeta,f}[n]|^2$ is given by

$$\mathrm{E}\left[\left|\hat{s}^{\not\eta}_{k,q,\zeta,f}[n]\right|^2\right] = \mathrm{E}\left[|\rho_{k,q,\zeta}|^2\left|\sum_{\xi=1}^{N_\mathrm{a}}\nu^*_{k,f,\xi}\rho^*_{k,f,\xi}\right|^2\right] \times\mathrm{E}\left[\left|\sum_{j=0}^{\chi-1}\sum_{i=1}^{K}\sum_{l=0}^{L}c^*_k[j]\boldsymbol{u}^{\mathrm{T}}_{k,q,\zeta}\boldsymbol{p}_{i,l}s_i[n-l-q+f+j]\right|^2\right].$$

As the second term already is $\rho_{k,q,\zeta}$ independent, we further on focus on the first term

$$\begin{aligned}\mathrm{E}&\left[|\rho_{k,q,\zeta}|^2\left|\sum_{\xi=1}^{N_\mathrm{a}}\nu_{k,f,\xi}\rho^*_{k,f,\xi}\right|^2\right]\\ &=\sum_{\xi_1=1}^{N_\mathrm{a}}\sum_{\xi_2=1}^{N_\mathrm{a}}\mathrm{E}\left[|\rho_{k,q,\zeta}|^2\nu^*_{k,f,\xi_1}\nu_{k,f,\xi_2}\rho^*_{k,f,\xi_1}\rho_{k,f,\xi_2}\right]\\ &=\begin{cases}2|\nu_{k,f,\zeta}|^2\sigma^4_{k,f,\zeta}+\sum\limits_{\substack{\xi=1\\ \xi\neq\zeta}}^{N_\mathrm{a}}\sigma^2_{k,f,\zeta}|\nu_{k,f,\xi}|^2\sigma^2_{k,f,\xi}, & q=f\\ \sum_{\xi=1}^{N_\mathrm{a}}\sigma^2_{k,q,\zeta}|\nu_{k,f,\xi}|^2\sigma^2_{k,f,\xi}, & \text{else}\end{cases}.\end{aligned}$$

Within, we employed the independent stochastic nature of ρ_{k,q,ξ_1} and ρ_{k,f,ξ_2} for $q\neq f$ or $\xi_1\neq\xi_2$ and introduced the fourth-order moment of a complex Gaussian variable $\mathrm{E}[|\rho_{k,f,\zeta}|^4]=2\sigma^4_{k,f,\zeta}$. Introducing the variable

$$\kappa=\begin{cases}2|\nu_{k,f,\xi}|^2, & \text{for } (q=f)\wedge(\xi=\zeta)\\ |\nu_{k,f,\xi}|^2, & \text{else}\end{cases}$$

simplifies the upcoming notations.[5] With this substitution, the above expectation reads

$$\mathrm{E}\left[|\rho_{k,q,\zeta}|^2\left|\sum_{\xi=1}^{N_\mathrm{a}}\nu_{k,f,\xi}\rho^*_{k,f,\xi}\right|^2\right]=\sum_{\xi=1}^{N_\mathrm{a}}\kappa\sigma^2_{k,q,\zeta}\sigma^2_{k,f,\xi}.$$

This way, it is possible to express the power of all relevant signal components based on partial CSI, i.e., independent of $\rho_{k,q,\zeta}$. Thus, a power equivalent long-term model can be formulated for the noise-free signal component $\tilde{s}^{\not\eta}_{k,q,\zeta,f}[n]$, that traveled over the ζth eigenspace of the qth channel path and the fth rake receiver coefficient as

$$\tilde{s}^{\not\eta}_{k,q,\zeta,f}[n]=\sum_{j=0}^{\chi-1}c^*_k[j]\frac{\nu^*_{k,f,\zeta}}{|\nu_{k,f,\zeta}|}\sqrt{\sum_{\xi=1}^{N_\mathrm{a}}\kappa\sigma^2_{k,q,\zeta}\sigma^2_{k,f,\xi}}\,\boldsymbol{u}^{\mathrm{T}}_{k,q,\zeta}\times\sum_{i=1}^{K}\sum_{l=0}^{L}\boldsymbol{p}_{i,l}s_i[n-l-q+f+j]. \quad (6)$$

The phase shift $\nu^*_{k,f,\zeta}/|\nu_{k,f,\zeta}|$ is necessary to account for the phase errors introduced by the rake receiver due to the precoding of the pilot channel [see (4)]. Computing the variance of the long-term signal $\tilde{s}^{\not\eta}_{k,q,\zeta,f}[n]$ and comparing it with the expression for $\mathrm{E}[|\hat{s}^{\not\eta}_{k,q,\zeta,f}[n]|^2]$ derived above, easily shows, that the found $\rho_{k,q,\zeta}$ independent representation indeed is variance true, i.e., both signals have the same power.

Like the noise-free signal portion $\hat{s}^{\not\eta}_k[n]$, we also have to split up the noise contained in the estimate $\hat{s}_k[n]$ [cf. (5)]

$$\hat{\eta}_k[n]=\sum_{j=0}^{\chi-1}c^*_k[j]\sum_{f=0}^{F}\sum_{\xi=1}^{N_\mathrm{a}}\nu^*_{k,f,\xi}\rho^*_{k,f,\xi}\eta_k[n+f+j].$$

We do so by constructing the long-term equivalent of the noise

$$\tilde{\eta}_k[n]=\sum_{q=0}^{Q}\sum_{\zeta=1}^{N_\mathrm{a}}\sum_{f=0}^{F}\tilde{\eta}_{k,q,\zeta,f}[n]$$

whose variance has to be the same as the variance of the original noise portion $\hat{\eta}_k[n]$

$$\mathrm{E}\left[|\tilde{\eta}_k[n]|^2\right]=\mathrm{E}\left[|\hat{\eta}_k[n]|^2\right].$$

To end up with a simple long-term model, we assume that the $\tilde{\eta}_{k,q,\zeta,f}[n]$ are mutually uncorrelated. Consequently, we have

$$\mathrm{E}\left[|\tilde{\eta}_k[n]|^2\right]=\sum_{q=0}^{Q}\sum_{\zeta=1}^{N_\mathrm{a}}\sum_{f=0}^{F}\mathrm{E}\left[|\tilde{\eta}_{k,q,\zeta,f}[n]|^2\right].$$

Since the $\rho_{k,f,\xi}$ are mutually uncorrelated, and we assume white noise, i.e., $\mathrm{E}[\eta_k[n]\eta^*_k[n+\nu]]=\sigma^2_\eta\delta[\nu]$, we find

$$\mathrm{E}\left[|\hat{\eta}_k[n]|^2\right]=\sum_{j=0}^{\chi-1}|c_k[j]|^2\sum_{f=0}^{F}\sum_{\xi=1}^{N_\mathrm{a}}|\nu_{k,f,\xi}|^2\sigma^2_{k,f,\xi}\sigma^2_\eta.$$

Combining the last two results, we can follow that we get a variance true representation of the noise, when choosing:

$$\tilde{\eta}_{k,q,\zeta,f}[n]=\sqrt{\frac{\chi}{N_\mathrm{a}(Q+1)}}\sum_{\xi=1}^{N_\mathrm{a}}|\nu_{k,f,\xi}|\sigma_{k,f,\xi}\eta_{k,q,\zeta,f}[n] \quad (7)$$

with white $\eta_{k,q,\zeta,f}[n]\sim\mathcal{N}_\mathbb{C}(0,\sigma^2_\eta)$. Note that the $\eta_{k,q,\zeta,f}[n]$ are mutually uncorrelated.

As the expression in (6) is linear in all parameters, we can formulate the long-term model in vector matrix notation

$$\tilde{s}_{k,q,\zeta,f}[n]=\sum_{i=1}^{K}\boldsymbol{p}^{\mathrm{T}}_i\boldsymbol{X}_{k,q,\zeta,f}\boldsymbol{s}_i[n]+\tilde{\eta}_{k,q,\zeta,f} \quad (8)$$

where $\boldsymbol{p}_i\in\mathbb{C}^{N_\mathrm{a}(L+1)}$ is obtained by stacking the $L+1$ filter vectors $\boldsymbol{p}_{i,l}\in\mathbb{C}^{N_\mathrm{a}}$. The vector $\boldsymbol{s}_i[n]\in\mathbb{C}^{L+Q+F+\chi}$, on the other hand, contains all relevant samples of the ith upsampled data signal

$$\boldsymbol{s}_i[n]=\left[s_i[n+F+\chi-1],\ldots,s_i[n-L-Q]\right]^{\mathrm{T}}.$$

Note that the matrix $\boldsymbol{X}_{k,q,\zeta,f}\in\mathbb{C}^{N_\mathrm{a}(L+1)\times L+Q+F+\chi}$ depends on long-term parameters only

$$\boldsymbol{X}_{k,q,\zeta,f}=\frac{\nu^*_{k,f,\zeta}}{|\nu_{k,f,\zeta}|}\sqrt{\sum_{\xi=1}^{N_\mathrm{a}}\kappa\sigma^2_{k,q,\zeta}\sigma^2_{k,f,\xi}}\,\boldsymbol{V}_{k,q,\zeta,f}\boldsymbol{C}_k \quad (9)$$

[5]Note that κ should be indexed by q, f, ζ, and ξ. For the sake of notational convenience, we leave this specification to the variances always following κ.

where $\boldsymbol{C}_k \in \mathbb{C}^{L+Q+F+1 \times L+Q+F+\chi}$ is a Toeplitz matrix representing the receiver-side correlation with the spreading code

$$\boldsymbol{C}_k = \begin{bmatrix} c_k^*[\chi-1] & c_k^*[\chi-2] & \cdots & 0 & 0 \\ 0 & c_k^*[\chi-1] & & 0 & 0 \\ \vdots & & \ddots & & \vdots \\ 0 & 0 & \cdots & c_k^*[1] & c_k^*[0] \end{bmatrix}.$$

The matrix $\boldsymbol{V}_{k,q,\zeta,f} \in \mathbb{C}^{N_\mathrm{a}(L+1) \times L+Q+F+1}$ is given as

$$\boldsymbol{V}_{k,q,\zeta,f} = [\boldsymbol{0}_{L+1 \times F+q-f}, \boldsymbol{1}_{L+1}, \boldsymbol{0}_{L+1 \times Q+f-q}] \otimes \boldsymbol{u}_{k,q,\zeta}$$

and we define $\boldsymbol{X}$ by stacking $\boldsymbol{X}_{k,q,\zeta,f}$ horizontally for $\{k,q,\zeta,f\} = \{1,0,1,0\}, \ldots, \{K,Q,N_\mathrm{a},F\}$.

IV. Transmit Filter Solutions

Aiming at the optimization of the spatio-temporal transmit filters $\boldsymbol{p}_k$, the objective functions as proposed in [4], [7], and [13] will be formulated by means of (8).

A. Transmit Matched Filter (TxMF)

The optimization objective of the matched filtering principle is to maximize the power of the desired signal component at the receiver's output. This power is denoted as the cross correlation between the signals $\tilde{s}_{k,q,\zeta,f}[n]$ in (8) and the desired signal component at chip number n

$$w_{k,q,\zeta,f}[n] = \begin{cases} \pi_{k,\zeta,f} s_k[n], & \text{for } (q=f) \wedge (n=\chi m) \\ 0, & \text{else} \end{cases}$$

where $m \in \mathbb{Z}$ and

$$\pi_{k,\zeta,f} = \frac{\nu_{k,f,\zeta}^*}{|\nu_{k,f,\zeta}|} \sqrt{\sum_{\xi=1}^{N_\mathrm{a}} \kappa \sigma_{k,f,\zeta}^2 \sigma_{k,f,\xi}^2}. \tag{10}$$

The desired signal $w_{k,q,\zeta,f}[n]$ is constructed such that signal components propagating over nonmatching channel paths and rake fingers are suppressed, since these signal components add up incoherently. Moreover, the property of $s_k[n]$ to be the upsampled symbol sequence for user k is taken into account.

We define the vectors $\boldsymbol{w}_k[n], \tilde{\boldsymbol{s}}_k[n] \in \mathbb{C}^{N_\mathrm{a}(Q+1)(F+1)}$ by stacking the samples of the desired signal $w_{k,q,\zeta,f}[n]$ [see (10)] and the long-term equivalent $\tilde{s}_{k,q,\zeta,f}[n]$ [see (8)] of the estimate for $q,\zeta,f = \{0,1,0\}, \ldots, \{Q,N_\mathrm{a},F\}$, respectively. Moreover, let $\boldsymbol{e}_\mu \in \{0,1\}^{L+Q+F+\chi}$ be the $(F+\chi)$th column of $\boldsymbol{1}_{L+Q+F+\chi}$, i.e., the vector that selects the chip of interest from the impulse response of the complete system of precoder, channel, rake, and code correlation $s_k[m\chi] = \boldsymbol{s}_k^\mathrm{T}[m\chi]\boldsymbol{e}_\mu$. Note that $\mathrm{E}[\tilde{s}_{k,q,\zeta,f}[m\chi] s_k^*[m\chi]] = \boldsymbol{p}_k^\mathrm{T} \boldsymbol{X}_{k,q,\zeta,f} \boldsymbol{e}_\mu$. With these conventions, the TxMF optimization problem reads[6]

$$\{\boldsymbol{p}_{\mathrm{MF},1}, \ldots, \boldsymbol{p}_{\mathrm{MF},K}\} = \underset{\{\boldsymbol{p}_1,\ldots,\boldsymbol{p}_K\}}{\mathrm{argmax}} \sum_{k=1}^{K} \mathrm{Re}\{\mathrm{E}[\boldsymbol{w}_k^\mathrm{H}[m\chi]\tilde{\boldsymbol{s}}_k[m\chi]]\}$$
$$\text{s.t.:} \quad \sum_{n=m\chi}^{(m+1)\chi-1} \mathrm{E}\left[\|\boldsymbol{x}[n]\|_2^2\right] = E_\mathrm{tr}. \tag{11}$$

[6]Note that the $\mathrm{Re}\{\cdot\}$ operator equivalently can be replaced by $|\cdot|$.

Note that only the samples of the receive filter output for $n = m\chi$ with $m \in \mathbb{Z}$ are of interest, since the receiver performs a downsampling by χ due to the employed CDMA with spreading factor χ. Additionally, note that E_tr denotes the total transmit power for the symbols of all K users, i.e., the transmit power of χ chips. As shown in [15], the solution to above TxMF optimization can be expressed as

$$\boldsymbol{p}_{\mathrm{MF},k} = \beta_\mathrm{MF} \sum_{f=0}^{F} \sum_{\zeta=1}^{N_\mathrm{a}} \pi_{k,\zeta,f} \boldsymbol{X}_{k,f,\zeta,f}^* \boldsymbol{e}_\mu,$$
$$\beta_\mathrm{MF} = \sqrt{\frac{E_\mathrm{tr}}{\sum_{k=1}^{K} \sigma_s^2 \left\| \sum_{f=0}^{F} \sum_{\zeta=1}^{N_\mathrm{a}} \pi_{k,\zeta,f} \boldsymbol{X}_{k,f,\zeta,f}^* \boldsymbol{e}_\mu \right\|_2^2}}. \tag{12}$$

From this result and with the definition of $\boldsymbol{X}_{k,q,\zeta,f}$ in (9), it can be shown that the impulse response of the TxMF is

$$\boldsymbol{p}_{\mathrm{MF},k}[n] = \beta_\mathrm{MF} \sum_{l=0}^{L_\mathrm{MF}} c_k[l] \delta[n-l] \sum_{f=0}^{F} \sum_{\zeta=1}^{N_\mathrm{a}} |\pi_{k,\zeta,f}|^2 \boldsymbol{u}_{k,f,\zeta}$$

where $L_\mathrm{MF} = \min(L, \chi-1)$. We observe that each of the TxMF's coefficients is the same vector weighted by $c_k[l]$. Since the rake receiver already maximizes the desired signal power in the temporal domain, the TxMF can only contribute to this goal by spatial processing. The result in (12), therefore, can be decomposed into a temporal filter $c_k[n]$ matching the correlation with the spreading code at the receiver and a pure spatial filter, i.e., a beamformer.

B. Transmit Zero-Forcing Filter (TxZF)

The TxZF attempts to completely suppress the interference and allows for a β-scaled version of the desired signal component [see (10)]. The remaining degrees of freedom are used to maximize the power of the received signal components via minimizing β^{-2} (cf. e.g., [15])

$$\{\boldsymbol{p}_{\mathrm{ZF},1}, \ldots, \boldsymbol{p}_{\mathrm{ZF},K}\} = \underset{\{\boldsymbol{p}_1,\ldots,\boldsymbol{p}_K\}}{\mathrm{argmin}} \ \beta^{-2}$$
$$\text{s.t.:} \quad \tilde{s}_{k,q,\zeta,f}[m\chi] = \beta w_{k,q,\zeta,f}[m\chi] \quad \text{and}$$
$$\sum_{n=m\chi}^{(m+1)\chi-1} \mathrm{E}\left[\|\boldsymbol{x}[n]\|_2^2\right] = E_\mathrm{tr} \tag{13}$$

where $k \in \{1,\ldots,K\}$, $q \in \{0,\ldots,Q\}$, $\zeta \in \{1,\ldots,N_\mathrm{a}\}$, and $f \in \{0,\ldots,F\}$. With the method of Lagrangian multipliers, we get the TxZF solution (see [15])

$$\boldsymbol{p}_{\mathrm{ZF},k} = \sqrt{\frac{E_\mathrm{tr}}{\sum_{k=1}^{K} \sigma_s^2 \boldsymbol{\pi}_k^\mathrm{T} \boldsymbol{X}^\dagger \boldsymbol{X}^{\dagger,\mathrm{H}} \boldsymbol{\pi}_k^*}} \boldsymbol{X}^{\dagger,\mathrm{T}} \boldsymbol{\pi}_k, \tag{14}$$

where $(\cdot)^\dagger$ denotes the Moore–Penrose pseudoinverse. The $\boldsymbol{X}$ and $\boldsymbol{\pi}_k$ are defined in Section III-B. Due to the necessary pseudoinversion, ill conditioned scenarios might cause extremely large denominators and, thus, very small values for β can occur. The signal, therefore, is received free of interference but with extremely low power making it very sensitive towards noise adulterations in these settings. Nevertheless, even partial CSI at the transmitter allows for interference suppression as the plots in Section IV-D confirm.

C. Transmit Wiener Filter (TxWF)

The objective of the TxWF is the modified MSE. As known from the full CSI counterpart (e.g., [15]), the variance of the difference between the β^{-1}-weighted estimate and the desired signal is minimized with respect to the spatio-temporal filter coefficients. Accounting for the transmit power constraint, the optimization problem, thus, reads as

$$\{\boldsymbol{p}_{\mathrm{WF},1},\ldots,\boldsymbol{p}_{\mathrm{WF},K}\} = \underset{\{\boldsymbol{p}_1,\ldots,\boldsymbol{p}_K\}}{\operatorname{argmin}} \sum_{k=1}^{K} \mathrm{E}\left[\left\|\boldsymbol{w}_k[m\chi] - \beta^{-1}\tilde{\boldsymbol{s}}_k[m\chi]\right\|_2^2\right]$$
$$\text{s.t.:} \sum_{n=m\chi}^{(m+1)\chi-1} \mathrm{E}\left[\|\boldsymbol{x}[n]\|_2^2\right] = E_{\mathrm{tr}}. \quad (15)$$

The solution to this optimization is obtained in [15] and can be written as

$$\boldsymbol{p}_{\mathrm{WF},k} = \beta_{\mathrm{WF}}\left(\boldsymbol{X}^*\boldsymbol{X}^{\mathrm{T}} + \gamma\frac{\sigma_\eta^2}{E_{\mathrm{tr}}}\mathbf{1}\right)^{-1}\boldsymbol{X}^*\boldsymbol{\pi}_k.$$
$$\beta_{\mathrm{WF}} = \sqrt{\frac{E_{\mathrm{tr}}}{\sum_{k=1}^{K}\sigma_s^2\boldsymbol{\pi}_k^{\mathrm{H}}\boldsymbol{X}^{\mathrm{T}}\left(\boldsymbol{X}^*\boldsymbol{X}^{\mathrm{T}} + \gamma\frac{\sigma_\eta^2}{E_{\mathrm{tr}}}\mathbf{1}\right)^{-2}\boldsymbol{X}^*\boldsymbol{\pi}_k}}. \quad (16)$$

For notational brevity, we introduced the scalar

$$\gamma = \chi\sum_{k=1}^{K}\sum_{f=0}^{F}\sum_{\xi=1}^{N_{\mathrm{a}}}|\nu_{k,f,\xi}|^2\sigma_{k,f,\xi}^2. \quad (17)$$

The matrix $\boldsymbol{X}$ and the vectors $\boldsymbol{\pi}_k, k = 1,\ldots,K$, are defined in Section III-B. Note that $\boldsymbol{X}$ is a sparse matrix. Therefore, the matrix inversion necessary for the computation of (16) can be computed very efficiently.

D. Numerical Evaluation

This section evaluates the proposed methods in terms of uncoded *bit-error rate* (BER). Basis shall be a channel model, with equally strong paths separated by one chip duration, which is denoted as *power delay profile 1* (PDP1). Moreover, the *directions of departure* (DOD) are independent and identically uniformly distributed within the sector from −60° to 60° [*spatial channel model 1* (SCM1)]. Note that the paths of one user have the same DOD, and we assume that all path channels have rank-1 covariance matrices. As a consequence of above assumptions, the used pilot precoding $\boldsymbol{b}_k$ is of no relevance within this section.

1) Comparison of Covariance Based Precoders: The plot in Fig. 2 compares the three presented linear precoding schemes. The results visualize that linear precoding can base on covariance knowledge and that such a precoding clearly outperforms a single transmitter *without precoding*. All linear precoding solutions (TxMF, TxZF, and TxWF) result in operable techniques, which in this setting decrease the BER below 10^{-2}. Among the three precoding schemes, the TxWF performs best, as it finds the optimum tradeoff between power and interference oriented processing. We see that the TxWF converges to the TxMF for low and to the TxZF for very high SNR values.

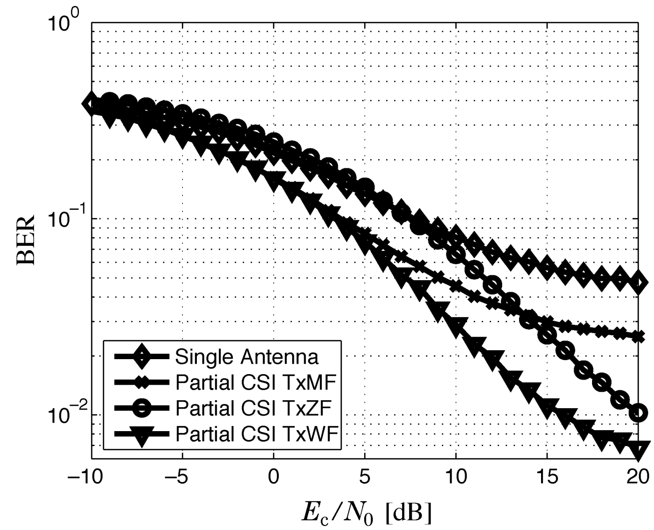

Fig. 2. Comparison of schemes based on covariance channel knowledge.

TABLE I
SYSTEM PARAMETERS 1

Parameter		Fig. 2	Fig. 3	Fig. 4
Tx Filter Order	L	4	4	0
Channel Order	Q	1	2	0
Rx Filter Order	F	1	2	0
Spreading Factor	χ	4	4	1
User Number	K	4	4	1, 2, 6
# Tx Antennas	N_{a}	4	4	8
# Rx Antennas	—	1	1	1
Path Powers	$\rho_{k,q,\zeta}$	PDP1	PDP1	—
Spatial Channel	—	SCM1	SCM1	SCM3

2) Comparison With Full CSI Precoders: In a slightly different setting (cf. Table I), the covariance-based precoding schemes are compared with the linear precoders that rely on full and instantaneous CSI. For sake of comparability, these schemes as well operate with a Rake receiving MS. As can be observed in Fig. 3, the lack of instantaneous knowledge leads to a power drawback of several decibels (dB) at BERs of 10^{-1}, as well as to a lack of diversity, as the steepness of the TxWF and the TxZF curves indicate. Still the covariance-based schemes allow the suppression of interfering components, for which the covariance-based TxZF and the covariance-based TxWF even outperform the TxMF with full CSI for high SNR.

3) Comparison With SINR Balancing: Comparing the proposed covariance-based schemes with the SINR balancing scheme [22], Fig. 4 plots the uncoded BER for flat fading[7] ($Q = L = F = 0$) and spreading factor $\chi = 1$. According to SCM3 the path DODs are chosen randomly from a uniform distribution $\mathcal{U}([-60°, 60°])$. Moreover, paths are subject to an angular spread with Laplacian distribution of variance 2°. The depicted results show that the performance loss with respect to the SINR optimal scheme in this setting is negligible. Especially, in the BER region around 10^{-1}, that is of interest when applying forward error correcting codes, both schemes show

[7]The SINR balancing scheme has not yet been extended to frequency-selective channels.

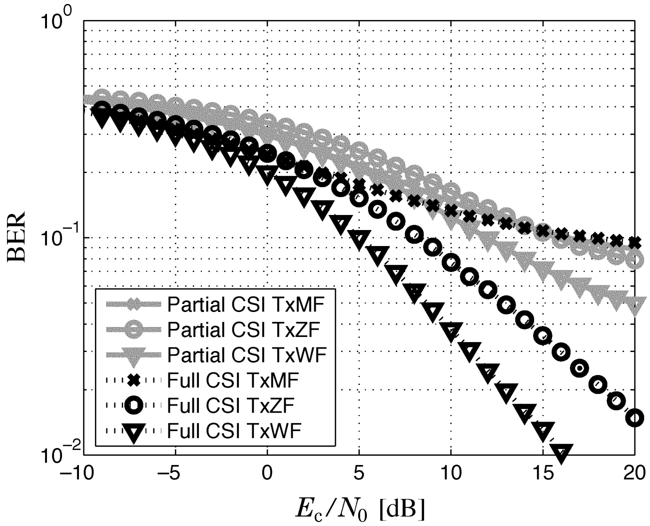

Fig. 3. Comparison of covariance-based precoding and full CSI precoding.

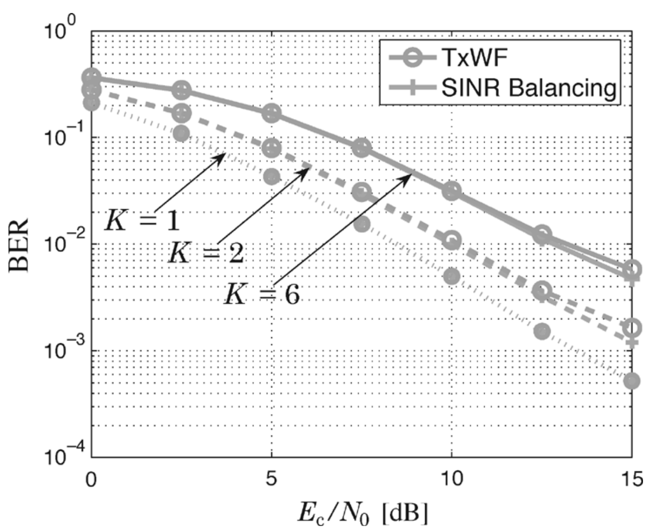

Fig. 4. Comparison with the SINR balancing solution.

identical performance in all load settings. Thus, the proposed solution provides a favorable alternative to the iterative scheme in [22], as it provides a closed solution of low complexity and can be applied in frequency-selective settings which are of major interest for CDMA systems.

V. Application to HSDPA

With HSDPA, a downlink mode was introduced that is capable of delivering high data rates to a small number of users. With the extension to multiantenna BSs, MISO techniques became a relevant topic for HSDPA. The following sections, thus, adapt the generic approaches from Section IV to this prominent next-generation CDMA system.

A. Secondary-Common Pilot Channel (S-CPICH) Estimation

The receiving MS in general is given two possibilities to perform its channel estimation. First, the *dedicated control channel* (DCCH) is used to transmit pilot sequences. Efficiency considerations only allow for very low power on these control channels, as the resulting signaling overhead multiplies with the number of users. Thus, a second alternative was introduced, providing a *common pilot channel* (CPICH). The training sequence on this channel is used by all MSs in the downlink, for which it can be granted a significantly larger amount of transmit power. The remainder of this paper will focus on CPICH estimation as it yields drastically better results.

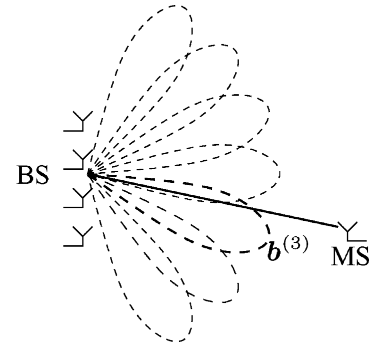

Fig. 5. S-CPICH channel estimation.

Though, a single CPICH does not provide sufficient means for channel estimation in MISO systems. Compensating the need for additional pilot channels by the arrays antenna gain, the S-CPICH was introduced. As depicted in Fig. 5, the S-CPICH transmits the different pilot sequences through different beamforming weights. These beamforming vectors are chosen from a grid of equidistant array steering vectors $\boldsymbol{b}^{(i)}$

$$\boldsymbol{b}_k = \boldsymbol{b}^{(j_k)}, \quad \text{where } j_k = \underset{i}{\operatorname{argmax}}\, \boldsymbol{u}_{k,0,1}^{\mathrm{H}} \boldsymbol{b}^{(i)}. \tag{18}$$

The MS, thus, adapts its MRC filter coefficient to the resulting channel substitute component $\nu_{k,f,\xi}\rho_{k,f,\xi}$

$$g_k^{(\rho)}[n] = \sum_{f=0}^{F} \sum_{\xi=1}^{N_{\mathrm{a}}} \left(\boldsymbol{b}^{(j_k),\mathrm{T}} \boldsymbol{u}_{k,f,\xi} \right)^* \rho_{k,f,\xi}^* \delta[n+f].$$

Presuming covariance knowledge at the transmitter the vectors $\boldsymbol{u}_{k,f,\xi}$ are as well available to the BS, as the S-CPICH beamforming weights $\boldsymbol{b}^{(j_k)}$. Thus, the BS can precompute the adulterations $\nu_{k,f,\xi}$ and include them into the covariance-based signal models from (8).

B. Linear Precoding in HSDPA

Beside this special choice for $\boldsymbol{b}_k$ the HSDPA setting introduces the following two simplifications.

1) Space-Only Processing: The use of long scrambling codes as standardized for HSDPA can be modeled by time variant spreading codes, changing every symbol period. Thus, spatio-temporal transmit filtering is of prohibitively high complexity. Consequently, at this stage, the considerations are reduced to post scrambling spatial transmit processing with a separate spreading prior to scrambling. The resulting filters, thus, are suboptimum versions of what is generally possible.

2) Rank-1 Processing: To further decrease the computational burden of the proposed methods, we simplify the *design model* to one spatial component per temporal path. Thus, the

TABLE II
SYSTEM PARAMETERS 2

Parameter		Fig.6	Fig.7
Tx Filter Order	L	0	0
Channel Order	Q	3	3
Rx Filter Order	F	3	3
Spreading Factor	χ	16	16
User Number	K	4	4
Scrambling	—	on	on
# Tx Antennas	N_{a}	8	8
# Rx Antennas	—	1	1
# S-CPICH Beams	N_{b}	8	8
Path Powers	$\rho_{k,q,\zeta}$	PDP2	PDP2
Spatial Channel	—	SCM2	SCM3

transmitter assumes rank-1 channel paths. Again, note that full approach linear precoding can do even better.

With these simplifying assumptions, the long-term signal model can be formulated as [cf. (6) and (7)]

$$\tilde{s}_{k,q,1,f}[n] = \sqrt{\kappa}\sigma_{k,f,1}\sigma_{k,q,1}\boldsymbol{u}_{k,q,1}^{\mathrm{T}} \sum_{i=1}^{K}\sum_{l=0}^{L} \boldsymbol{p}_{i,l} s_i[n-l-q+f] + \frac{1}{\sqrt{Q+1}}|\nu_{k,f,1}|\sigma_{k,f,1}\eta_{k,q,1,f}[n] \quad (19)$$

which due to the linearity in all components, can also be expressed as [cf. (8)]

$$\tilde{s}_{k,q,1,f}[n] = \sum_{i=1}^{K} \boldsymbol{p}_i^{\mathrm{T}} \boldsymbol{X}'_{k,q,f} \boldsymbol{s}_i[n] + \tilde{\eta}_{k,q,f}[n] \quad (20)$$

where $\boldsymbol{X}'_{k,q,f}$ only depends on $\sigma_{k,q,1}$ and $\nu_{k,q,1}$, $\forall k,q$. With this notation, the optimization results from Section IV remain valid and the linear precoders for the HSDPA setting can be obtained by replacing the matrix $\boldsymbol{X}$ and the vector $\boldsymbol{\pi}_k$ in (12), (14), and (16) by a matrix $\boldsymbol{X}'$ comprising the $\boldsymbol{X}'_{k,q,f}$, $\forall k,q,f$ and a suitable vector $\boldsymbol{\pi}_k$ (cf. Section III-B).

C. Numerical Evaluation

This section uses the numerical simulation of an HSDPA downlink to evaluate the above methods. The resulting uncoded BER is chosen as the relevant metric. As shown before (cf. [24]), the drawn conclusions hold for coded BER and block error rate as well.[8] The system is modeled as block fading in long-term and short-term properties, where the channel coefficients are assumed constant for the duration of one slot and the covariance matrix changes every frame, i.e., every 30th slot. The parameter details can be obtained from Table II. Within, SCM2 has constant DODs with 2° angular spread at 7.5°, 15°, 22.5°, and 30° for the four users. The different temporal paths of each user have identical spatial channels. SCM3 is known from Section IV-D3. Moreover, PDP2 is exponentially decreasing with 3 dB per path, PDP1 sees equal powers for all paths.

At first, let us compare the different approaches for one covariance realization. This way, users 1 and 3 have DODs

[8]The fairness issues introduced by the use of sum criteria like the MSE is of higher significance in these settings though.

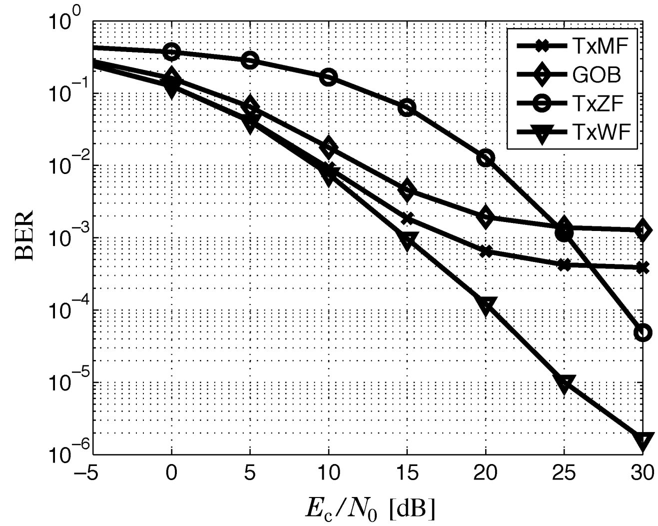

Fig. 6. Comparison of different S-CPICH schemes.

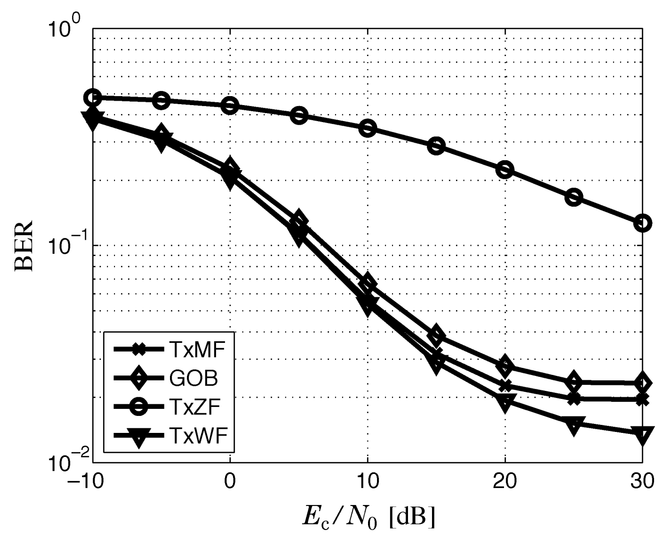

Fig. 7. Mean performance of the S-CPICH schemes.

identical with the center angle of the corresponding S-CPICH beam, while users 2 and 4 have DODs right in between two such beams. The resulting BER performances are plotted in Fig. 6. The evaluation compares the three linear precoding proposals made, with the use of the S-CPICH *grid of beam* (GOB), that employs the pilot beamforming weights for data transmission, i.e., $\boldsymbol{p}_k = \boldsymbol{b}_k$. The simulation result shows the superiority of the TxWF. It outperforms all other schemes for all SNR, because it recovers the full antenna gain and suppresses the interference when reasonable. The other two linear precoding schemes each share one of these two advantages against the GOB. Note that the GOB approach in this setting is not at all able to provide uncoded BER of less than 10^{-3} as the system in this range is interference limited and the GOB has no means to tackle interference.

With the channel model SCM3, this section evaluates the presented methods over a large number of randomly selected spatial channels. The resulting BER curves are shown in Fig. 7. While the performance of the TxZF significantly suffers from the channel realizations with ill conditioned situations, TxMF and TxWF sustain the advantages from the previous results. As argued for every realization separately, their performance benefits from the recoverage of the full antenna gain against the GOB approach. Moreover, the interference suppression capability of the TxWF accounts for an additional performance enhancement, resulting in a distinctively lower level of saturation and a power advantage of 1.2 dB at a BER of 10^{-1}.

VI. Conclusion

Recapitulating, the power equivalent derivation of a covariance-based signal model for the investigated DS-CDMA system allowed for the formulation of the three linear precoding objectives "power of the desired signal component," "interference free transmission," and "modified mean square error" independently of the channel coefficients. The resulting optimum solutions proved, that FDD counterparts for all three linear precoders exist and that linear precoding, in general, as well as MSE minimization and interference suppression in detail are possible in FDD DS-CDMA settings. Moreover, the adaptation of the derived solutions to the next-generation CDMA system HSDPA, educed a general solution for the handling of S-CPICHs and managed to overcome the problem of full rank linear precoding. The resulting algorithms outperform the state of the art approach for every covariance realization.

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