LMMSE Channel Estimation for MIMO W-CDMA with Out-of-Cell Interference Mitigation

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***Abstract*—An LMMSE channel estimator is proposed for the downlink of a MIMO W-CDMA system. The estimation takes into account intra-cell and out-of-cell interference, as well as receive-side spatial correlation. Efficient numerical methods are applied to solve the resulting large block Toeplitz linear equation sets. Various simplifying assumptions about the prior knowledge of second-order statistics used by the estimator are investigated, providing a performance vs. complexity tradeoff. Link level simulation results of a MIMO HSDPA system show that LMMSE channel estimation together with out-of-cell interference-aware equalization yield performance throughput gains of up to 30% over a correlation based channel estimator.**

I. Introduction

Channel estimation in W-CDMA networks is strongly affected by interference that can be divided into intra-cell and out-of-cell interference[1]. At the base-station usually only a small amount of power (approx. 10%) is dedicated to the pilot channels that can be used for channel estimation. The remaining power is dedicated to all other channels and is thus considered as intra-cell interference for the channel estimator. The intra-cell interference increases with the number of transmit antennas at the base-station since the same spreading codes are reused at each transmit antenna. Out-of-cell interference becomes crucial at the cell edge where the received power of the desired base-station is comparable to the received power from other base-stations.

A good channel estimator performance can therefore only be achieved if the estimator takes all types of interference into account. This can be done by a linear minimum mean-square error (LMMSE) channel estimator based on the second order statistics of the involved signals. However, it turns out that the resulting matrices that have to be inverted become large (5120×5120 for a 2×2 MIMO system with slot based channel estimation) prohibiting real-time implementations.

In this paper we tackle the resulting large matrix inversions by the Akaike algorithm that is especially suitable for large symmetric block Toeplitz matrices. We work under the assumption that three dominant neighboring base stations perturb the estimation process. Having knowledge about their influence on the covariance of the reveived signal increases the estimation quality significantly. Thus, the out-of-cell interference is colored and the covariance matrix is depending on the channel matrices of the interfering base-stations.

The additional covariance information can be further exploited in the LMMSE chip level equalizer which is referred to as type 3i equalizer [1] in 3GPP standards. We compare the performance of the LMMSE channel estimator with low complexity channel estimators namely a correlation-based estimator, and the so called tap-wise LMMSE estimator introduced in [2]. This comparison will offer a benchmark for a performance vs. complexity tradeoff.

The paper is organized as follows. Section II describes our system model which considers intra and inter-cell interference. The LMMSE channel estimator with full covariance properties and a procedure for the exact inversion of the involved covariance matrix are derived in Section III. The out-of-cell interference aware equalizer is introduced in Section IV. Simulation results utilizing a physical layer D-TxAA (Dual-Stream Transmit Adaptive Array) [3, 4] HSDPA (High Speed Downlink Packet Access) simulator [5] based on the setting of Section V are presented in Section VI. Finally, we draw our conclusions in Section VII.

II. System Model

The transmit scheme used for HSDPA MIMO is referred to as D-TxAA [6], which supports transmission of up to two parallel streams which belong to the same user. After independent coding, modulation, spreading, and scrambling of the streams linear spatial pre-coding is performed. The pre-coder consists of a pair of weights, which are chosen orthogonal among the streams aiding the receiver in separating the two streams. After pre-coding, the pilot channels are added on both physical antennas and the signals are transmitted simultaneously over the two transmit antennas.

The inter-cell interference that arises from base-stations outside the serving cell is modeled as follows. The transmit signals of three dominant neighboring base-stations are modeled as white Gaussian with powers relative to the total other-cell interference power. These ratios are referred to as dominant interferer proportion (DIP) ratios [1]. The interfering transmit signals are colored by the wireless multi-path

[1]Out-of-cell, other-cell or inter-cell interference can be used interchangeably.

channels between the neighboring base-stations and the user equipment.

We work on the assumption that there is no spatial correlation at the base-station antennas. This can be justified in practice by the spacing of the antennas at the base station. Therefore, we just consider the channel from transmit antenna n_t to all N_R receive antennas. Thus estimation is done for each transmit antenna separately. The delay spread of the channel is L_h samples, and the $N_\mathrm{R}L_\mathrm{h}$ coefficients to estimate are

$$\mathbf{h}_{n_\mathrm{t}} = \left[h_0^{(1,n_\mathrm{t})} \dots h_{L_\mathrm{h}-1}^{(1,n_\mathrm{t})}, \dots, h_0^{(N_\mathrm{R},n_\mathrm{t})} \dots h_{L_\mathrm{h}-1}^{(N_\mathrm{R},n_\mathrm{t})}\right]^\mathrm{T}.$$

We consider K consecutive samples of the received signal

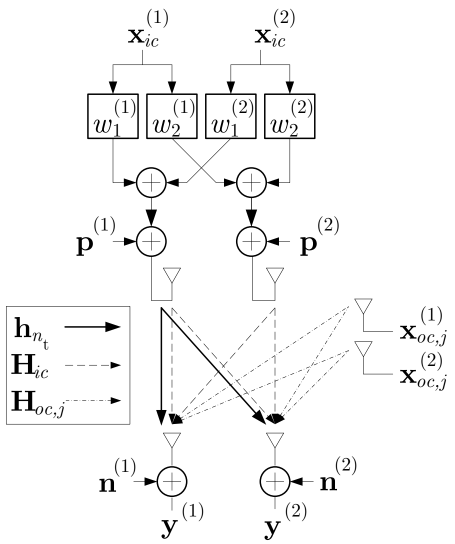

Fig. 1. D-TxAA transmission system with precoding and out-of-cell interference.

starting at time $i + L_\mathrm{h} - 1$ at receive antenna n_r for the channel estimation and store them in the vector $\mathbf{y}^{(n_\mathrm{r})} = [y_{i+L_\mathrm{h}-1}^{(n_\mathrm{r})} \dots y_{i+K+L_\mathrm{h}-2}^{(n_\mathrm{r})}]$. Furthermore, if we define the training matrix $\mathbf{P}_{n_\mathrm{t}}$ containing delay shifted versions of the pilot chip sequence of transmit antenna n_t as columns

$$\mathbf{P}_{n_\mathrm{t}} = \begin{bmatrix} p_{i+L_\mathrm{h}-1}^{(n_\mathrm{t})} & \dots & p_i^{(n_\mathrm{t})} \\ \vdots & & \vdots \\ p_{i+K+L_\mathrm{h}-2}^{(n_\mathrm{t})} & \dots & p_{i+K-1}^{(n_\mathrm{t})} \end{bmatrix}, \tag{1}$$

we can express the received signal $\mathbf{y} = [\mathbf{y}^{(1)}, \dots, \mathbf{y}^{(N_\mathrm{R})}]^\mathrm{T}$ at all receive antennas (see Fig. 1) compactly as

$$\begin{aligned}\mathbf{y} = & \underbrace{(\mathbf{P}_{n_\mathrm{t}} \otimes \mathbf{I}_{N_\mathrm{R}})}_{\text{training } \mathbf{T}} \mathbf{h}_{n_\mathrm{t}} + \underbrace{\mathbf{H}_\mathrm{ic}((\mathbf{W} \otimes \mathbf{I}_K)\mathbf{x}_\mathrm{ic} + \mathbf{p}'_\mathrm{ic})}_{\text{intra}-\text{cell intf.}} + \\ & + \underbrace{\sum_{j=1}^{N_\mathrm{oc}} \mathbf{H}_{\mathrm{oc},j}\mathbf{x}_{\mathrm{oc},j}}_{\text{out}-\text{of}-\text{cell intf.}} + \mathbf{n},\end{aligned} \tag{2}$$

where $\mathbf{x}_\mathrm{ic}$, $\mathbf{x}_{\mathrm{oc},j}$ are the transmitted data streams of the serving base-station and the interfering base-station j respectively. $\mathbf{p}'_\mathrm{ic}$ contains all pilot channels transmitted by the serving base-station, except the pilot of the transmit antenna n_t. $\mathbf{W} = [w_1^{(1)}, w_1^{(2)}; w_2^{(1)}, w_2^{(2)}]$ contains the pre-coding weights of the in-cell data streams and N_oc is the number of dominant interfering base-stations that are taken into account. Furthermore $\mathbf{H}_\mathrm{ic}$ and $\mathbf{H}_{\mathrm{oc},j}$ are Toeplitz block matrices[2] $\mathbf{H}$ which contain the channel impulse responses of the serving base-station and the interfering base-station j, i.e.,

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}^{(1,1)} & \dots & \mathbf{H}^{(1,N_\mathrm{T})} \\ \vdots & \ddots & \vdots \\ \mathbf{H}^{(N_\mathrm{R},1)} & \dots & \mathbf{H}^{(N_\mathrm{R},N_\mathrm{T})} \end{bmatrix}, \tag{3}$$

where $\mathbf{H}^{(n_\mathrm{r},n_\mathrm{t})}$ has the following Toeplitz structure

$$\mathbf{H}^{(n_\mathrm{r},n_\mathrm{t})} = \begin{bmatrix} h_0^{(n_\mathrm{r},n_\mathrm{t})} & \dots & 0 \\ \vdots & \ddots & \vdots \\ h_{L_\mathrm{h}-1}^{(n_\mathrm{r},n_\mathrm{t})} & & h_0^{(n_\mathrm{r},n_\mathrm{t})} \\ \vdots & \ddots & \vdots \\ 0 & \dots & h_{L_\mathrm{h}-1}^{(n_\mathrm{r},n_\mathrm{t})} \end{bmatrix}, \tag{4}$$

with $[h_0^{(n_\mathrm{r},n_\mathrm{t})}, \dots, h_{L_\mathrm{h}-1}^{(n_\mathrm{r},n_\mathrm{t})}]^\mathrm{T}$ the channel impulse response of the link between transmit antenna n_t and receive antenna n_r. $\mathbf{n}$ accounts for the thermal noise generated in the radio frequency frontend of the receiver and interference from sources other than the N_oc base-stations.

III. LMMSE Channel Estimator with Full Covariance Properties

In this section, we derive the LMMSE channel estimator and introduce an exact inversion of the covariance matrix. Because of the prohibitive complexity of the LMMSE channel estimator exploiting full covariance properties we introduce simplifying assumptions about these covariance matrices in simulation setting in Section V. The model mismatch introduced through this simplification is analyzed through the performance analysis in Section VI.

A. LMMSE Channel Estimator

LMMSE estimation is a Bayesian method, i.e., it relies on the prior knowledge of the second order statistics (in the form of the covariance matrices, including the information about the power delay profile, spatial correlation...) of the variable to estimate, i.e. $\mathbf{h}_{n_\mathrm{t}}$, the intra-cell and out-of-cell interference, and the thermal noise in (2). If we further define the true covariance of the vector to estimate as

$$\mathbf{R}_{\mathbf{h}_{n_\mathrm{t}}\mathbf{h}_{n_\mathrm{t}}} = \mathrm{E}\left[\mathbf{h}_{n_\mathrm{t}}\mathbf{h}_{n_\mathrm{t}}^\mathrm{H}\right], \tag{5}$$

we can write the received signal covariance as

$$\mathbf{R}_{\mathbf{yy}} = \mathbf{T}\mathbf{R}_{\mathbf{h}_{n_\mathrm{t}}\mathbf{h}_{n_\mathrm{t}}}\mathbf{T}^\mathrm{H} + \mathbf{R}_\mathrm{ic} + \mathbf{R}_\mathrm{oc} + \sigma_\mathrm{n}^2\mathbf{I}, \tag{6}$$

[2]A block Toeplitz matrix is Toeplitz w.r.t. the blocks, i.e., it is composed of submatrices (not necessarily Toeplitz) which form constant block diagonals, whereas a Toeplitz block matrix consists of blocks which are Toeplitz matrices.

where $\mathbf{R}_{\text{ic}} = \mathbf{H}_{\text{ic}}\mathbf{R}_{\text{data,ic}}\mathbf{H}_{\text{ic}}^{\text{H}}$ is the intra-cell covariance matrix given by the second term on the right-hand side of (2). For the calculation we used the fact that $\mathbf{W}\mathbf{W}^{\text{H}} = \mathbf{I}$ and neglected $\mathbf{p}'_{\text{ic}}$. Furthermore, $\mathbf{R}_{\text{oc}} = \sum_j \mathbf{H}_{\text{oc},j}\mathbf{R}_{\text{data,oc},j}\mathbf{H}_{\text{oc},j}^{\text{H}}$ is the covariance of the out-of-cell interference. The noise term $\mathbf{n}$ is assumed Gaussian i.i.d., white, with variance σ_{n}^2 and therefore the LMMSE estimator with full knowledge of the second-order parameters is [7]

$$\begin{aligned}\widehat{\mathbf{h}}_{n_{\text{t}}} =& \mathbf{R}_{\mathbf{h}_{n_{\text{t}}}\mathbf{h}_{n_{\text{t}}}}\mathbf{T}^{\text{H}}\left(\mathbf{T}\mathbf{R}_{\mathbf{h}_{n_{\text{t}}}\mathbf{h}_{n_{\text{t}}}}\mathbf{T}^{\text{H}} + \mathbf{R}_{\text{ic}} + \mathbf{R}_{\text{oc}} + \sigma_{\text{n}}^2\mathbf{I}\right)^{-1}\mathbf{y}\\ =& \left(\mathbf{R}^{-1}_{\mathbf{h}_{n_{\text{t}}}\mathbf{h}_{n_{\text{t}}}} + \mathbf{T}^{\text{H}}(\mathbf{R}_{\text{ic}} + \mathbf{R}_{\text{oc}} + \sigma_{\text{n}}^2\mathbf{I})^{-1}\mathbf{T}\right)^{-1}\\ & \mathbf{T}^{\text{H}}(\mathbf{R}_{\text{ic}} + \mathbf{R}_{\text{oc}} + \sigma_{\text{n}}^2\mathbf{I})^{-1}\mathbf{y},\end{aligned} \tag{7}$$

where the second version is obtained through the Woodbury identity. Note that the apparent contradiction of using the knowledge of $\mathbf{h}_{n_{\text{t}}}$ (through $\mathbf{H}_{\text{ic}}$ in $\mathbf{R}_{\text{ic}}$) itself in the estimator is in practice lifted by the use of a slightly outdated estimate for $\mathbf{H}_{\text{ic}}$. The number of chips used for the channel estimation is $K \gg L_{\text{h}}$. The computational complexity in (7) arises from the operations on the $N_{\text{R}}K \times N_{\text{R}}K$ matrices $\mathbf{R}_{\text{ic}}$ and $\mathbf{R}_{\text{oc}}$ and the tall $N_{\text{R}}K \times N_{\text{R}}L_{\text{h}}$ matrix $\mathbf{T}$.

B. *Exact Inversion Exploiting the Covariance Matrix Structure*

We denote the $N_{\text{R}}K \times N_{\text{R}}K$ Toeplitz block (TB) matrix $\mathbf{R}_{\text{TB}} = \mathbf{R}_{\text{ic}} + \mathbf{R}_{\text{oc}} + \sigma_{\text{n}}^2\mathbf{I}$ in (7). In the following, we consider a system with two transmit and receive antennas, i.e., $N_{\text{T}} = N_{\text{R}} = 2$ for simplicity. We further denote the four $K \times K$ submatrices of $\mathbf{R}_{\text{TB}}$ as

$$\mathbf{R}_{\text{TB}} = \begin{bmatrix} \mathbf{R}_{\text{AA}} & \mathbf{R}_{\text{BB}} \\ \mathbf{R}_{\text{CC}} & \mathbf{R}_{\text{DD}} \end{bmatrix}. \tag{8}$$

$\mathbf{R}_{\text{TB}}$ is Hermitian but not Toeplitz since the main diagonal is not constant. The submatrices $\mathbf{R}_{\text{AA}}$ and $\mathbf{R}_{\text{DD}}$ are Hermitian and Toeplitz with real valued main diagonals. Compared to the overall dimension of $\mathbf{R}_{\text{AA}}, \ldots, \mathbf{R}_{\text{DD}}$, only a few, i.e., L_{h} off-diagonals contain nonzero entries (see Fig. 3). The submatrices $\mathbf{R}_{\text{BB}}$ and $\mathbf{R}_{\text{CC}}$ are not Hermitian but Toeplitz. Furthermore, $\mathbf{R}_{\text{BB}}^{\text{H}} = \mathbf{R}_{\text{CC}}$.

In the following we propose a procedure that can deal with the complexity of the covariance matrix inversion and provides a computationally feasible solution to derive the exact $\widehat{\mathbf{h}}_{n_{\text{t}}}$. The key element of the procedure is an iterative algorithm for the inversion of a block Toeplitz (BT) matrix introduced by Akaike in 1973 [8] and a permutation of $\mathbf{R}_{\text{TB}}$ that turns it into a block Toeplitz matrix $\mathbf{R}_{\text{BT}}$. The desired matrix inversions in (7) can be rewritten as an equivalent system of linear equations as follows

$$\begin{aligned}\mathbf{R}_{\text{TB}}^{-1}\mathbf{T} &= \mathbf{X}_\alpha & \qquad \mathbf{R}_{\text{TB}}^{-1}\mathbf{y} &= \mathbf{x}_\beta\\ \Leftrightarrow \mathbf{R}_{\text{TB}}\mathbf{X}_\alpha &= \mathbf{T} & \qquad \Leftrightarrow \mathbf{R}_{\text{TB}}\mathbf{x}_\beta &= \mathbf{y},\end{aligned} \tag{9}$$

where we are seeking for $\mathbf{X}_\alpha$ and $\mathbf{x}_\beta$ equivalently. We permute $\mathbf{R}_{\text{TB}}$ by pre- and post-multiplication with the matrix $\mathbf{P} = [\mathbf{I}_K \otimes [1,0]^{\text{T}}, \mathbf{I}_K \otimes [0,1]^{\text{T}}]$, i.e., $\mathbf{R}_{\text{BT}} = \mathbf{P}\mathbf{R}_{\text{TB}}\mathbf{P}^{\text{T}}$. The $N_{\text{T}}K \times N_{\text{T}}K$ matrix $\mathbf{P}$ is structured as depicted in Fig. 2 with ones indicated by the black boxes. The permutation gathers the

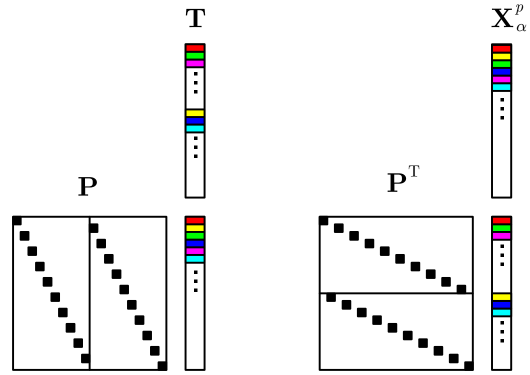

Fig. 2. Impact of $\mathbf{P}$ on the rows of $\mathbf{T}$ and $\mathbf{X}_\alpha^p$.

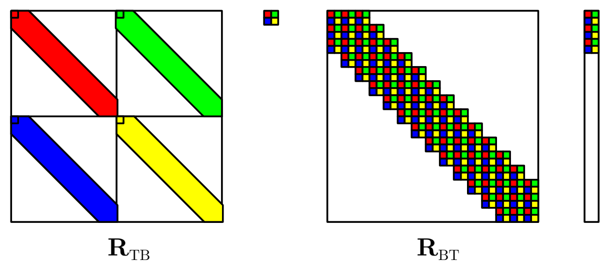

Fig. 3. Permutation of the covariance matrix.

nonzero elements of $\mathbf{R}_{\text{TB}}$ along a broader main diagonal of $\mathbf{R}_{\text{BT}}$. Thus, we read out elements of $\mathbf{R}_{\text{AA}}, \ldots, \mathbf{R}_{\text{DD}}$ which have relatively the same coordinates and put them into the submatrices which are placed into $\mathbf{R}_{\text{BT}}$ (see Fig. 3). The permutation leads to a Hermitian block Toeplitz matrix $\mathbf{R}_{\text{BT}}$ which is composed of submatrices $\dot{\mathbf{R}}_k$ of basic dimension 2×2 with a structure given by

$$\mathbf{R}_{\text{BT}} = \begin{bmatrix} \dot{\mathbf{R}}_0 & \dot{\mathbf{R}}_1^{\text{H}} & \cdots & \dot{\mathbf{R}}_{K-1}^{\text{H}} \\ \dot{\mathbf{R}}_1 & \dot{\mathbf{R}}_0 & & \dot{\mathbf{R}}_{K-2}^{\text{H}} \\ \vdots & & \ddots & \\ \dot{\mathbf{R}}_{K-1} & \dot{\mathbf{R}}_{K-2} & \cdots & \dot{\mathbf{R}}_0 \end{bmatrix}. \tag{10}$$

The first block column contains all information about the covariance matrix. Furthermore $\dot{\mathbf{R}}_k = \mathbf{0}\,\forall |k| \geq L_{\text{h}}$.
Using the established permutation in (9) leads to

$$\begin{aligned}\mathbf{P}^{\text{T}}\mathbf{R}_{\text{BT}}\mathbf{P}\mathbf{X}_\alpha &= \mathbf{T} & \qquad \mathbf{P}^{\text{T}}\mathbf{R}_{\text{BT}}\mathbf{P}\mathbf{x}_\beta &= \mathbf{y}\\ \Leftrightarrow \mathbf{R}_{\text{BT}}\underbrace{\mathbf{P}\mathbf{X}_\alpha}_{\mathbf{X}_\alpha^p} &= \underbrace{\mathbf{P}\mathbf{T}}_{\mathbf{T}^p} & \qquad \Leftrightarrow \mathbf{R}_{\text{BT}}\underbrace{\mathbf{P}\mathbf{x}_\beta}_{\mathbf{x}_\beta^p} &= \underbrace{\mathbf{P}\mathbf{y}}_{\mathbf{y}^p}\\ \Leftrightarrow \mathbf{R}_{\text{BT}}\mathbf{X}_\alpha^p &= \mathbf{T}^p & \qquad \Leftrightarrow \mathbf{R}_{\text{BT}}\mathbf{x}_\beta^p &= \mathbf{y}^p.\end{aligned} \tag{11}$$

We can solve (11) for $\mathbf{X}_\alpha^p$ column-wise and for $\mathbf{x}_\beta^p$ with the iterative block Toeplitz matrix inversion algorithm. The Akaike algorithm needs the first block column $[\dot{\mathbf{R}}_0, \ldots, \dot{\mathbf{R}}_{K-1}]^{\text{T}}$ and the desired column of $\mathbf{T}^p$ or $\mathbf{y}^p$ as input parameters. Then we have to permute the rows of $\mathbf{X}_\alpha^p$ and $\mathbf{x}_\beta^p$ in order to get $\mathbf{X}_\alpha$ and $\mathbf{x}_\beta$ since $\mathbf{X}_\alpha = \mathbf{P}^{\text{T}}\mathbf{X}_\alpha^p$ (see Fig. 2) and $\mathbf{x}_\beta = \mathbf{P}^{\text{T}}\mathbf{x}_\beta^p$ equivalently. Finally, we get $\widehat{\mathbf{h}}_{n_{\text{t}}} = (\mathbf{R}^{-1}_{\mathbf{h}_{n_{\text{t}}}\mathbf{h}_{n_{\text{t}}}} + \mathbf{T}^{\text{H}}\mathbf{X}_\alpha)^{-1}\mathbf{T}^{\text{H}}\mathbf{x}_\beta$.

IV. Out-of-Cell Interference Aware Equalizer

Chip-level LMMSE equalizers capable of out-of-cell interference mitigation are considered in [1] and referred to as type 3i receivers. Following the notation of [9] we define the received chip stream at time instant i as

$$\mathbf{r}_i^{(n_\mathrm{r})} = [r_i^{(n_\mathrm{r})}, \ldots, r_{i-L_\mathrm{h}-L_\mathrm{f}+2}^{(n_\mathrm{r})}], \tag{12}$$

where L_h and L_f are the length of the channel impulse response and the equalizer length, respectively. Thus, the vector contains the $(L_\mathrm{h} + L_\mathrm{f} - 1)$ recent chips received at antenna n_r. We can express the received signal $\mathbf{r}_i = [\mathbf{r}_i^{(1)}, \ldots, \mathbf{r}_i^{(n_\mathrm{r})}]^\mathrm{T}$ at all N_R receive antennas compactly as

$$\mathbf{r}_i = \underbrace{\mathbf{H}_\mathrm{ic}^\mathrm{T}(\mathbf{W} \otimes \mathbf{I}_{L_\mathrm{h}+L_\mathrm{f}-1})}_{\mathbf{H}_\mathrm{w}} \mathbf{x}_\mathrm{ic} + \sum_{j=1}^{N_\mathrm{oc}} \mathbf{H}_{\mathrm{oc},j}^\mathrm{T} \mathbf{x}_{\mathrm{oc},j} + \mathbf{n}. \tag{13}$$

In (13) we neglected the pilot, data control and synchronization channels. The MMSE equalizer coefficients can be calculated by minimizing the quadratic cost function for a given τ

$$\mathbf{J}(\mathbf{f}) = \mathrm{E}\left[|\mathbf{f}^\mathrm{H}\mathbf{r}_i - x_{\mathrm{ic},i-\tau}|^2\right].$$

In [1] the equalizer coefficients can be found as

$$\begin{aligned}\mathbf{f} =&(\mathbf{H}_\mathrm{w}\mathbf{R}_\text{data,ic}\mathbf{H}_\mathrm{w}^\mathrm{H} + \sum_{j=1}^{N_\mathrm{oc}} \mathbf{H}_{\mathrm{oc},j}^\mathrm{T}\mathbf{R}_{\text{data,oc},j}(\mathbf{H}_{\mathrm{oc},j}^\mathrm{T})^\mathrm{H} + \sigma_\mathrm{n}^2\mathbf{I})^{-1}\cdot \\ &\cdot \mathbf{H}_\mathrm{w}\mathbf{e}_\tau, \end{aligned} \tag{14}$$

with $\mathbf{e}_\tau$ a zero vector of length $(L_\mathrm{h} + L_\mathrm{f} - 1)$ with a single "one" at position τ.

V. Simulation Setting and Model Mismatch

In this section, the setting for a D-TxAA HSDPA link level simulation is presented. We simulate the experienced throughput of a user that gets served with data with a fixed transport format according to the fixed reference channel (FRC) for D-TxAA introduced in [10]. The precoding coefficients $\mathbf{W}$ of the user are adaptively adjusted in order to maximize the power received of the first stream. We consider subframe-wise channel estimation, i.e. $K = 7680$. The power of the transmit signal of the neighboring base-stations are chosen according to the DIP which defines the ratio of the power of a given interfering base-station $\sigma_{\mathrm{oc},j}^2$ over the total other-cell interference power I_oc [1]. This can be written as $\mathrm{DIP}_j = \sigma_{\mathrm{oc},j}^2 / I_\mathrm{oc}$ where $I_\mathrm{oc} = \sum_j \sigma_{\mathrm{oc},j}^2 + \sigma_\mathrm{n}^2$. DIP values are chosen as $[0.4,\ 0.2,\ 0.1]$ according to the median DIP values in [1]. Our simulation settings are shown in detail in Table I. The channel model and the simplifications of the covariance terms are introduced below.

TABLE I
Simulation parameters.

Parameter	Value
Desired user CQI	FRC
CPICH $E_\mathrm{c}/I_\mathrm{or}$	$-10\,\mathrm{dB}$
SCH/PCCPCH $E_\mathrm{c}/I_\mathrm{or}$	$-12\,\mathrm{dB}$
User equipment capability	10M
Rx spatial correlation	$\rho = 0.5$
Power delay profile	Pedestrian B [11]
UE speed	$3\,\mathrm{km/h}$
$I_\mathrm{or}/I_\mathrm{oc}$	$5\,\mathrm{dB}$
OCNS	on
Receivers	LMMSE equalizer
	LMMSE type 3i equalizer

A. Spatially Correlated Channel Model

The channel model is a spatially correlated time-varying channel model with adjustable Doppler spectrum. The variation in time can be controlled through the maximum velocity of the user. The model introduces correlation in the lag and the receive space whereas no correlation between transmit antennas is considered. The absence of spatial correlation at the base station antennas can be justified by the spacing of the antennas. The joint receive and power delay profile (PDP) correlation can be controlled with $\mathbf{R}_{\mathbf{h}_{n_\mathrm{t}}\mathbf{h}_{n_\mathrm{t}}}$, i.e.,

$$\mathbf{R}_{\mathbf{h}_{n_\mathrm{t}}\mathbf{h}_{n_\mathrm{t}}} = \underbrace{\begin{bmatrix} 1 & \rho \\ \rho & 1 \end{bmatrix}}_{\mathbf{R}_\mathrm{Rx}} \otimes \underbrace{\mathbf{R}_\mathrm{PDP}}_{\text{PDP corr.}}, \tag{15}$$

with ρ the correlation parameter for the receive antennas of the receive spatial correlation $\mathbf{R}_\mathrm{Rx}$ and $\mathbf{R}_\mathrm{PDP}$ the correlation of the PDP. If we set $\rho = 0$ we model uncorrelated receive antennas whereas $\rho = 1$ introduces fully spatial receive correlation. We used a diagonal PDP correlation matrix $\mathbf{R}_\mathrm{PDP}$ and didn't introduce correlation between the lags, i.e., $\mathbf{R}_\mathrm{PDP} = \mathrm{diag}(\mathrm{PDP})$, with PDP set according to Pedestrian B.

B. Model Mismatch

We study the following simplifications of the covariance term $R_{\mathbf{h}_{n_\mathrm{t}}\mathbf{h}_{n_\mathrm{t}}}$:

A: Real model, $R_{\mathbf{h}_{n_\mathrm{t}}\mathbf{h}_{n_\mathrm{t}}} = \mathbf{R}_\mathrm{Rx} \otimes \mathrm{diag}(\mathrm{PDP})$,
B: Known spatial correlation, PDP ignored: $R_{\mathbf{h}_{n_\mathrm{t}}\mathbf{h}_{n_\mathrm{t}}} = \mathbf{R}_\mathrm{Rx} \otimes \frac{1}{L_\mathrm{h}}\mathbf{I}_{L_\mathrm{h}}$,
C: No covariance information: $R_{\mathbf{h}_{n_\mathrm{t}}\mathbf{h}_{n_\mathrm{t}}} = \mathbf{I}_{N_\mathrm{R}\cdot L_\mathrm{h}}$,

where $\mathbf{R}_\mathrm{Rx}$ is a 2×2 Hermitian matrix representing spatial receive correlation, and the matrix $\mathrm{diag}(\mathrm{PDP})$ contains the channel power delay profile on its diagonal. For the covariance term $\mathbf{R}_\mathrm{TB} = \mathbf{R}_\mathrm{ic} + \mathbf{R}_\mathrm{oc} + \sigma_\mathrm{n}^2\mathbf{I}$, the following simplifications are considered:

1: Knowledge of outdated estimate $\mathbf{H}_\mathrm{ic}$ and perfect knowledge of $\mathbf{H}_{\mathrm{oc},j}$ and $\sigma_{\mathrm{oc},j}^2$ $\forall j$: $\mathbf{R}_\mathrm{TB} = E_\mathrm{c}\mathbf{H}_\mathrm{ic}\mathbf{H}_\mathrm{ic}^H + \sum_j \sigma_{\mathrm{oc},j}^2 \mathbf{H}_{\mathrm{oc},j}\mathbf{H}_{\mathrm{oc},i}^\mathrm{H} + \sigma_\mathrm{n}^2\mathbf{I}$,
2: Knowledge of $\mathbf{H}_\mathrm{ic}$, $\mathbf{R}_\mathrm{Rx}$ and $\sigma_{\mathrm{oc},j}^2$ $\forall j$: $\mathbf{R}_\mathrm{TB} = E_\mathrm{c}\mathbf{H}_\mathrm{ic}\mathbf{H}_\mathrm{ic}^\mathrm{H} + \sum_j \mathbf{R}_\mathrm{Rx} \otimes \sigma_{\mathrm{oc},j}^2\mathbf{I} + \sigma_\mathrm{n}^2\mathbf{I}$,
3: Knowledge of E_c, $\sigma_{\mathrm{oc},j}^2$ $\forall j$ and $\mathbf{R}_\mathrm{Rx}$: $\mathbf{R}_\mathrm{TB} = \mathbf{R}_\mathrm{Rx} \otimes E_\mathrm{c}\mathbf{I} + \sum_j \mathbf{R}_\mathrm{Rx} \otimes \sigma_{\mathrm{oc},j}^2\mathbf{I} + \sigma_\mathrm{n}^2\mathbf{I}$,
4: Knowledge of E_c, $\sigma_{\mathrm{oc},j}^2$ $\forall j$: $\mathbf{R}_\mathrm{TB} = E_\mathrm{c}\mathbf{I} + \sum_j \sigma_{\mathrm{oc},j}^2\mathbf{I} + \sigma_\mathrm{n}^2\mathbf{I} = E_\mathrm{c}\mathbf{I} + I_\mathrm{oc}\mathbf{I}$.

VI. Simulation Results

Fig. 4 depicts the simulated data throughput over the spatially correlated channel. The correlation based estimator performs worst, since it does not consider interference at all. The tap-wise LMMSE estimator [2] with perfect channel

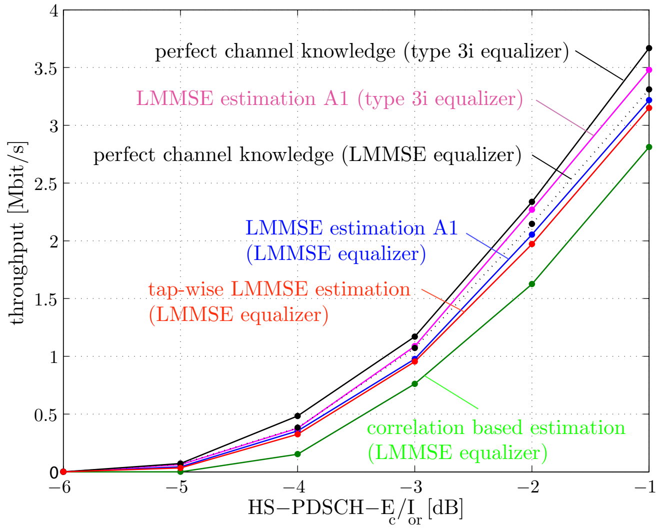

Fig. 4. Throughput of the D-TxAA scenario.

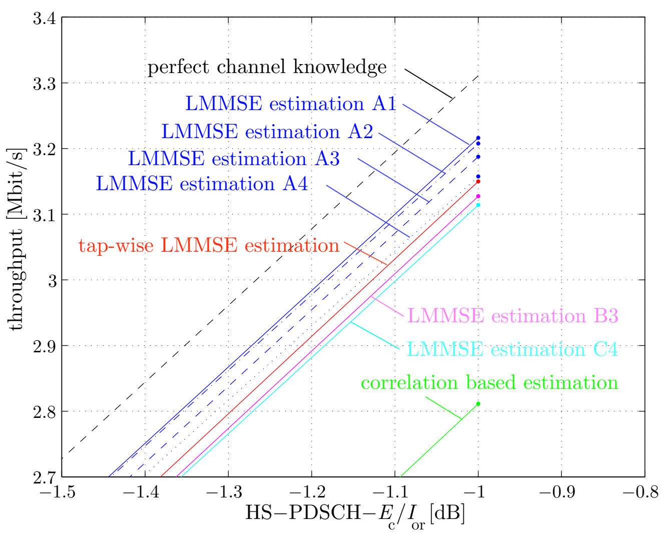

Fig. 5. Comparison of the simplifying assumptions of the covariance matrix with LMMSE equalizer.

autocorrelation estimation already outperforms the correlation based estimator by about 18% at E_c/I_{or}=-2 dB. The LMMSE estimator A1 slightly outperforms the tap-wise LMMSE since it takes into account the colored noise from intra-cell data and out-of-cell interference. The performance gain comes with the cost of a substantially increased computational complexity introduced by the exact covariance matrix inversion. Exploiting the full knowledge about the out-of-cell and intra-cell interference in the LMMSE equalizer (LMMSE estimator A1 with type 3i equalizer) further increases the performance. Fig. 5 depicts the impact of the simplifying assumptions of the covariance matrix on the throughput (LMMSE equalizer). The LMMSE channel estimators A1, A2, A3 and A4 outperform the tap-wise LMMSE estimator. The increase of knowledge about intra-cell and out-of-cell covariance leads to incremental performance gains when changing from LMMSE version A4 to A1. B3 and C4 perform worse than the tap-wise LMMSE

TABLE II
THROUGHPUT GAIN IN % OVER THE CORRELATION BASED ESTIMATOR (AT E_c/I_{or}=-2 dB).

Channel estimator	Gain
Perfect CSI (type 3i equalizer)	42 %
LMMSE A1 (type 3i equalizer)	36 %
Perfect CSI	30 %
LMMSE A1	24 %
Tap-wise LMMSE, PDP perf.	18 %
correlation-based	0 %

estimator since B3 only takes into account the receive spatial correlation and assumes a uniform power delay profile, whereas D4 has no covariance information. Table II summarizes the throughput gains of the various channel estimators over the correlation-based estimator at E_c/I_{or}=-2 dB.

VII. CONCLUSIONS

A procedure for the inversion of large covariance matrices was derived, and applied to LMMSE channel estimation in the context of MIMO HSDPA. The impact of various simplifying assumptions about the structure of the covariance matrices on the estimation performance and the comparison with low complexity channel estimators was investigated by link level simulations. LMMSE channel estimation in conjunction with type 3i equalizers can provide throughput gains of up to 30% over a correlation based channel estimator.

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