Managing the Interference Structure of MIMO HSDPA: A Multi-User Interference Aware MMSE Receiver with Moderate Complexity

Martin Wrulich, *Student Member, IEEE*, Christian Mehlführer, *Member, IEEE*, and Markus Rupp, *Senior Member, IEEE*

***Abstract*—It is known that Wideband Code-Division Multiple Access (W-CDMA) networks are limited by interference more than by any other single effect. Due to the frequency selectivity of the wireless channel, intra-cell interference composed of self- and other-user interference occurs. The intra-cell interference as a result of Transmit Antenna Array (TxAA) HSDPA multi-user transmissions, however, shows a special structure that can be exploited to allow for an efficient interference suppression. The contributions of this article are the following: we (a) propose a suitable system model to derive an intra-cell interference aware Minimum Mean Squared Error (MMSE) equalizer with moderate complexity and investigate its capabilities to suppress the multi-user intra-cell interference, (b) propose solutions to estimate the pre-coding state of the cell both with and without available training, and (c) investigate the resulting throughput performance with physical layer as well as system-level simulations. Our proposed solution can be interpreted as a multi-user extension of the classical MMSE equalizer for HSDPA systems without decoding the undesired users, resulting in only slightly increased complexity.**

***Index Terms*—MIMO, TxAA, HSDPA, W-CDMA, MMSE, interference-awareness, multi-user, blind estimation, pre-coding.**

I. Introduction

High speed downlink packet access (HSDPA) has been standardized as an extension of the Universal Mobile Telecommunications System (UMTS) as a part of the third Generation Partnership Project (3GPP) Release 5 [3]. It is the spectrally most efficient Wideband Code-Division Multiple Access (W-CDMA) system commercially available at the moment. To satisfy the need for higher data rates and new services with the current base station sites, even higher cell capacities and spectral efficiencies have to be achieved. Correspondingly, 3GPP has decided to incorporate Multiple-Input Multiple-Output (MIMO) techniques for the enhancement of Frequency Division Duplex (FDD) HSDPA, with Transmit Antenna Array (TxAA) being one of those [4].

It is well known that the performance of cellular W-CDMA networks is limited by interference more than by any other single effect. In particular, the intra-cell interference caused by the loss of orthogonality between the spreading codes due to the frequency selective channel imposes restrictive throughput constraints. To combat this effect, in HSDPA Minimum Mean Squared Error (MMSE) equalizers are disposed [5], [6]. However, these solutions do not take multi-user operation or the interference structure in the cell into account. In [7], Multi User (MU) interference terms are considered but the authors investigated a non-standard conform pre-coding scheme and the proposed equalizer works on symbol-level as opposed to our chip-level based receiver. Moreover we also take the interference caused by the pilots into account.

In the downlink each receiver needs to detect a *single* desired signal, while experiencing two types of interference. These are caused by the serving Node-B (intra-cell interference) and by a few dominant neighboring Node-Bs (inter-cell interference). In principle, interference mitigation can also be performed at the transmitter, which requires accurate channel state information from all users [8]. This requires lots of signaling and feedback information exchange, which is typically not available in cellular context. Thus, we focus on the receiver side in this article. Handling the interference on the receiver side is still a difficult job, especially in the multi-user case where classical approaches result in complex receiver structures, [9]–[11]. More practical investigations on this subject are also conducted by 3GPP [12], but so far none of these recommendations have been implemented. Given the limited battery capacity of today's handsets, complexity is an important issue [13].

In this paper, we consider the downlink of TxAA HSDPA for which we propose an interference aware receiver that utilizes the spatial structure of the intra-cell interference, i.e., the pre-coding state of the cell. The proposed receiver consists of a linear filter together with an estimator for the pre-coding state. Since in [14] it was shown that interference cancellation on the data channel does not significantly increase the system performance we are not considering Successive Interference Cancellation (SIC) structures here. Furthermore, note that our

Manuscript received April 29, 2009; revised September 7, 2009; accepted December 9, 2009. The associate editor coordinating the review of this paper and approving it for publication was X. Wang.

The authors are with the Institute of Communications and Radio Frequency Engineering, Vienna University of Technology, Austria (e-mail: {mwrulich, chmehl, mrupp}@nt.tuwien.ac.at).

Part of this work is based on two conference publications [1], [2]. Publication [1]: "Interference aware MMSE equalization for MIMO TxAA," in *Proc. IEEE 3rd International Symposium on Communications, Control and Signal Processing (ISCCSP)*, 2008, introduces the concept of the equalizer and investigates its intra-cell interference suppression capabilities. Publication [2]: "Intra-cell interference aware equalization for TxAA HSDPA," in *Proc. IEEE 3rd International Symposium on Wireless Pervasive Computing (ISWPC)*, 2008, then applies the concept to an High-Speed Downlink Packet Access (HSDPA) scenario where we assess the throughput performance.

Digital Object Identifier 10.1109/TWC.2010.04.090612

1536-1276/10$25.00 © 2010 IEEE

interference-aware equalizer does not need to decode the data of the other users like high-complexity MU receivers would have to. The contributions of this article in particular are

(a) we propose an intra-cell interference aware MMSE equalizer with moderate complexity and investigate its capabilities to suppress the interference,
(b) we propose solutions to estimate the pre-coding state of the cell—required for the intra-cell interference aware equalizer—both with and without available training, and
(c) we investigate the throughput performance in physical-layer as well as system-level simulations.

The following notation is used in the paper. Superscripts T, $*$, H and $\#$ stand for matrix transpose, matrix conjugate, matrix transpose conjugate, and matrix pseudo-inverse, respectively. Uppercase and lowercase boldface letters denote matrices and vectors, respectively. The symbols tr$(\cdot)$, det$(\cdot)$, $\|\cdot\|$ and $\otimes$ stand for the trace, the determinant, the Euclidean norm, and the Kronecker product, respectively. Finally, $\mathbb{E}_{\mathbf{v}}\{\cdot\}$ is the expected value with respect to the random vector $\mathbf{v}$ and $\mathbb{E}\{\cdot\}$ with respect to all the random terms inside the braces.

This paper is organized as follows: Section II introduces the transmission scheme for HSDPA that is the origin of our work and defines the necessary system model. In Section III, we derive the interference aware MMSE equalizer and assess its interference-suppression capabilities. Section IV introduces the problem of the pre-coding state estimation and our two proposed solutions, after which we show the throughput performance in physical-layer and system-level simulations in Section V. Finally, we conclude this article in Section VI.

II. Transmit Antenna Array (TxAA) Mode

TxAA has been introduced already with UMTS back in 1999, and now builds the foundation of the spatial multiplexing enhancement in MIMO HSDPA [15]. In contrast to the double stream operation of MIMO HSDPA, TxAA allows for multiple users being served in the downlink simultaneously, which is called *multi-code scheduling*. Multi-code scheduling is well suited to work optimally in terms of the sum-rate throughput and short-term fairness tradeoff [16], and future generation mobile networks implicitly build on similar concepts, e.g. Long Term Evolution (LTE) [17] puts a strong emphasis on multi-user scheduling in its time-frequency downlink frames.

In TxAA HSDPA, every user is assigned a specific number of orthogonal spreading sequences of length 16. Note that HSDPA utilizes multiple parallel spreading sequences per user to increase the spectral efficiency on the wireless link. At a maximum, 15 spreading sequences can be assigned to all users, the 16-th orthogonal spreading sequence is reserved for transmitting the pilot channels and other control channels. However, due to restrictions in the downlink resource allocation signaling, at maximum four users can be served simultaneously. Depending on the link-adaptation feedback in form of Channel Quality Indicator (CQI) values, the scheduler can decide which users should be served in parallel. Furthermore note that the TxAA scheme allows for an arbitrary number of receive antennas being utilized at the User Equipment (UE).

A. System Model

Fig. 1 shows the TxAA transmission scheme for one receive antenna when U users are simultaneously served. For the following, we assume the base-station to be equipped with $N_{\mathrm{T}} = 2$ transmit antennas, and each user u to employ an arbitrary number of receive antennas. Furthermore, we define the spread and scrambled transmit chip stream of user u at time instant i as

$$\mathbf{s}_i^{(u)} \triangleq \left[s_i^{(u)}, \ldots, s_{i-L_{\mathrm{h}}-L_{\mathrm{f}}+2}^{(u)}\right]^{\mathrm{T}}, \tag{1}$$

where L_{h} and L_{f} are the length of the channel impulse response and the equalizer length, respectively. Thus, the vector $\mathbf{s}_i^{(u)}$ contains the $L_{\mathrm{h}} + L_{\mathrm{f}} - 1$ most recently transmitted chips. This notation serves us to represent the convolution of the transmit signal and the frequency selective MIMO channel in vector-matrix notation. Furthermore, note that this chip stream contains the sum of all utilized spreading sequences for user u, thus the multi-code utilization of HSDPA is also represented. We assume that the energy σ_s^2 of the chip stream $\mathbf{s}_i^{(u)}$ of each user u is normalized to one.

Thus, by multiplying $\mathbf{s}_i^{(u)}$ with a factor $\alpha^{(u)}$, the base-station can allocate a certain amount of transmit power to each served user. After the power allocation, the chip streams are weighted by the user-dependent complex pre-coding coefficients $w_1^{(u)}$ and $w_2^{(u)}$ at the first and second transmit antenna, respectively. Our modeling holds for arbitrary pre-coding weights, but we want to note that due to standardization these coefficients are strongly quantized, [4], which we will take advantage of in Section IV. The weighted chip streams of all users are then added to the sequences $\alpha^{(\mathrm{p})}\mathbf{p}_i^{(1)}$ and $\alpha^{(\mathrm{p})}\mathbf{p}_i^{(2)}$, representing the sum of all channels that are transmitted without pre-coding, that is, the Common Pilot CHannel (CPICH), the High-Speed Shared Control CHannel (HS-SCCH), and other signaling channels—to which we will refer to as non-data channels.

The frequency selective channel between the n_{t}-th transmit and the n_{r}-th receive antenna is represented in Fig. 1 by the vector $\mathbf{h}^{(n_{\mathrm{r}},n_{\mathrm{t}})}$, $n_{\mathrm{t}} = 1, 2$, composed of the taps of the channel. For sake of simplicity, let us neglect the non-data channels $\alpha^{(\mathrm{p})}\mathbf{p}_i^{(1)}$ and $\alpha^{(\mathrm{p})}\mathbf{p}_i^{(2)}$ for the moment; they will be included for the main derivation again. Then the multi-user transmission in Fig. 1 can be represented by U virtual antennas, one for each active user, as illustrated in Fig. 2. The resulting equivalent (virtual) channels between user u and receive antenna n_{r} are then given by $\tilde{\mathbf{h}}^{(u,n_{\mathrm{r}})} = w_1^{(u)}\mathbf{h}^{(n_{\mathrm{r}},1)} + w_2^{(u)}\mathbf{h}^{(n_{\mathrm{r}},2)}$, $u = 1, 2, \ldots, U$. From this description it can be seen immediately that the intra-cell interference caused by the other users can be treated like being transmitted over $U-1$ different channels. The classical MMSE equalizer however would be determined only by the pre-coding weights $w_1^{(u)}$ and $w_2^{(u)}$ and does not consider the special structure of the interference. Thus, the degraded transmission scheme of TxAA imposes an interference situation which cannot be handled well by the classical MMSE equalizer that is only matched to the channel of the desired user. This is in contrast to the Single-Input Single-Output (SISO) HSDPA case, where due to the lack of pre-coding, the interference of simultaneously served users is transmitted over the same channel as the one of the

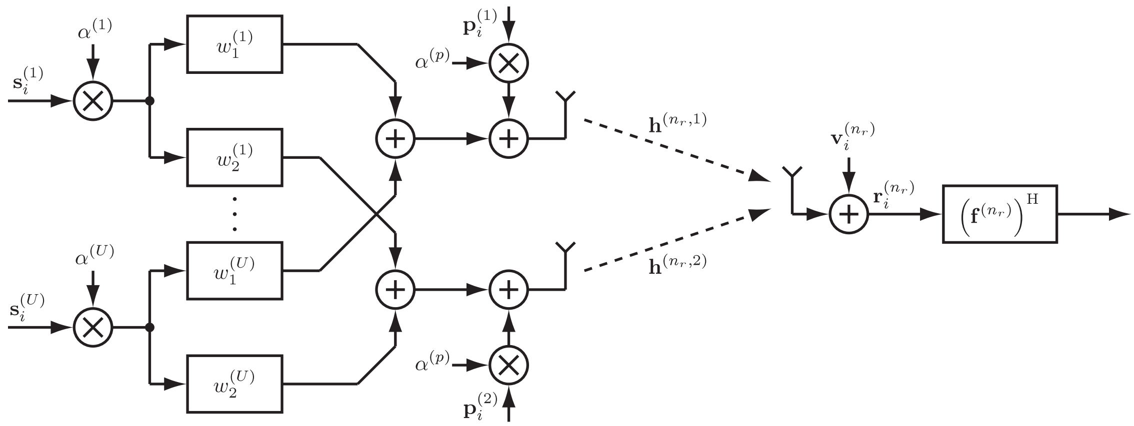


Fig. 1. Multi-user transmission in TxAA, for a total number U of simultaneously served users. The pre-coding is conducted individually for every user. At the receiver, only one receive antenna is depicted, although the scheme allows for an arbitrary number of receive antennas.

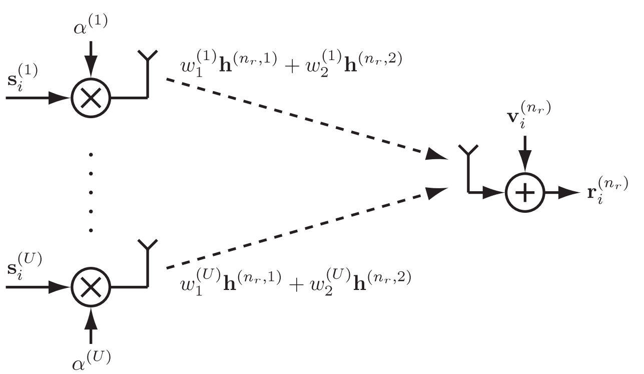


Fig. 2. Equivalent representation of the multi-user TxAA transmission, representing U virtual antennas and U virtual channels.

desired user. Thus, equalization of the desired user's signal also equalizes the signal of the simultaneously served users.

For the deduction of our proposed intra-cell interference aware MMSE equalizer let us define the $L_{\mathrm{f}} \times (L_{\mathrm{h}} + L_{\mathrm{f}} - 1)$ dimensional band matrix modeling the channel between the n_{t}-th transmit and the n_{r}-th receive antenna

$$\mathbf{H}^{(n_{\mathrm{r}},n_{\mathrm{t}})} = \begin{bmatrix} h_0^{(n_{\mathrm{r}},n_{\mathrm{t}})} & \cdots & h_{L_{\mathrm{h}}-1}^{(n_{\mathrm{r}},n_{\mathrm{t}})} & 0 \\ \ddots & & \ddots & \\ 0 & h_0^{(n_{\mathrm{r}},n_{\mathrm{t}})} & \cdots & h_{L_{\mathrm{h}}-1}^{(n_{\mathrm{r}},n_{\mathrm{t}})} \end{bmatrix}. \quad (2)$$

The full frequency-selective MIMO channel can be modeled by a block matrix $\mathbf{H}$ consisting of $N_{\mathrm{R}} \times N_{\mathrm{T}}$ band matrices,

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}^{(1,1)} & \mathbf{H}^{(1,2)} \\ \vdots & \vdots \\ \mathbf{H}^{(N_{\mathrm{R}},1)} & \mathbf{H}^{(N_{\mathrm{R}},2)} \end{bmatrix}, \quad (3)$$

where we explicitly utilized the assumption that TxAA employs only two transmit antennas ($N_{\mathrm{T}} = 2$).

By stacking the received signal vectors of all N_{R} receive antennas

$$\mathbf{r}_i = \left[\left(\mathbf{r}_i^{(1)}\right)^{\mathrm{T}}, \cdots, \left(\mathbf{r}_i^{(N_{\mathrm{R}})}\right)^{\mathrm{T}}\right]^{\mathrm{T}}, \quad (4)$$

and by stacking the transmitted signal vectors of all U users and the vectors $\mathbf{p}_i^{(1)}$ and $\mathbf{p}_i^{(2)}$,

$$\mathbf{s}_i = \left[\left(\mathbf{s}_i^{(1)}\right)^{\mathrm{T}}, \cdots, \left(\mathbf{s}_i^{(U)}\right)^{\mathrm{T}}, \left(\mathbf{p}_i^{(1)}\right)^{\mathrm{T}}, \left(\mathbf{p}_i^{(2)}\right)^{\mathrm{T}}\right]^{\mathrm{T}}, \quad (5)$$

we obtain the compact system description

$$\mathbf{r}_i = \underbrace{\mathbf{H}\left(\mathbf{W}^{(\mathrm{MU})} \otimes \mathbf{I}_{L_{\mathrm{h}}+L_{\mathrm{f}}-1}\right)}_{\mathbf{H}_w} \mathbf{s}_i + \mathbf{v}_i = \mathbf{H}_w \mathbf{s}_i + \mathbf{v}_i. \quad (6)$$

Here, $\otimes$ denotes the Kronecker product, and $\mathbf{v}_i$ is an additive noise vector that can incorporate both the thermal noise and the interference from other base-stations (inter-cell interference). The $2 \times (U+2)$ dimensional matrix $\mathbf{W}$ contains the pre-coding coefficients $w_1^{(u)}$ and $w_2^{(u)}$ of all users, as well as the power coefficients $\alpha^{(u)}$, with $u = 1, 2, \ldots, U$, and is defined as

$$\mathbf{W}^{(\mathrm{MU})} \triangleq \begin{bmatrix} \alpha^{(1)} w_1^{(1)} & \cdots & \alpha^{(U)} w_1^{(U)} & \alpha^{(\mathrm{p})} & 0 \\ \alpha^{(1)} w_2^{(1)} & \cdots & \alpha^{(U)} w_2^{(U)} & 0 & \alpha^{(\mathrm{p})} \end{bmatrix}. \quad (7)$$

This matrix reflects the premise that the non-data channels are not pre-coded, thus the two columns on the right side are specified solely by the single parameter $\alpha^{(\mathrm{p})}$ which controls the total power spent on these channels. In general, we also assume that the power available at the base-station is fully spent, thus the coefficients α are subject to a sum-power constraint

$$\sum_{u=1}^{U} \left(\alpha^{(u)}\right)^2 + 2\left(\alpha^{(\mathrm{p})}\right)^2 = P. \quad (8)$$

Furthermore, we assume the power control to be completely included in the power coefficients $\alpha^{(u)}$ and $\alpha^{(\mathrm{p})}$, which imposes a power constraint on the pre-coding coefficients, $\left\|\mathbf{w}^{(u)}\right\|^2 = 1$, where $\mathbf{w}^{(u)} \triangleq [w_1^{(1)}, w_2^{(2)}]^{\mathrm{T}}$.

III. Interference-Aware MMSE Equalization

Having our system model being specified in Section II, we are now able to derive the resulting MMSE equalizer. Without loss of generality we assume in the following that the sequence of user one is to be reconstructed. The MMSE equalizer coefficients can be calculated by minimizing the quadratic cost function [18]

$$J(\mathbf{f}) = \mathbb{E}\left\{\left|\mathbf{f}^{\mathrm{H}}\mathbf{r}_i - s_{i-\tau}^{(1)}\right|^2\right\}, \tag{9}$$

with τ specifying the delay of the equalized signal, fulfilling $\tau \geq L_{\mathrm{h}}$ due to causality. In this work, we assume the channel to be known at the receiver site.

This cost function minimizes the distance between the equalized chip stream and the transmitted chip stream. In Equation (9), the vector $\mathbf{f}$ defines N_{R} equalization filters,

$$\mathbf{f} = \left[\left(\mathbf{f}^{(1)}\right)^{\mathrm{T}}, \cdots, \left(\mathbf{f}^{(N_{\mathrm{R}})}\right)^{\mathrm{T}}\right]^{\mathrm{T}}. \tag{10}$$

Each filter $\mathbf{f}^{(n_{\mathrm{r}})} = \left[f_0^{(n_{\mathrm{r}})}, \cdots, f_{L_{\mathrm{f}}-1}^{(n_{\mathrm{r}})}\right]^{\mathrm{T}}$ has a length of L_{f}.

The desired interference-aware equalizer then can be obtained by minimizing the cost function, which can be performed by differentiating Equation (9) with respect to $\mathbf{f}^*$ [19], evaluating the expectation operation and setting the derivative equal to zero:

$$\frac{\partial J}{\partial \mathbf{f}^*} = 0 \Rightarrow \mathbf{f} = \sigma_s^2 \left(\mathbf{H}_w \mathbf{R}_{ss} \mathbf{H}_w^{\mathrm{H}} + \mathbf{R}_{vv}\right)^{-1} \mathbf{H}_w \mathbf{e}_\tau. \tag{11}$$

The matrices $\mathbf{R}_{ss}$ and $\mathbf{R}_{vv}$ are the signal and noise covariance matrices, respectively, and the vector $\mathbf{e}_\tau$ is a zero vector of length $(U+2)(L_{\mathrm{h}} + L_{\mathrm{f}} - 1)$ with a single one at position τ.

If the transmitted data signals of the users are uncorrelated with equal power σ_s^2, the covariance matrix $\mathbf{R}_{ss}$ becomes $\sigma_s^2\mathbf{I}$, and if we assume the noise vector $\mathbf{v}_i$ white with variance σ_v^2, the noise covariance matrix becomes $\sigma_v^2\mathbf{I}$. Without loosing generality, we can assume the signal covariance σ_s^2 to be equal to one, because the individual transmit powers of the users are determined by $\alpha^{(u)}$. The variance σ_v^2 is composed of the thermal noise and the power received from the other base-stations. Note that if the receiver shall take the structure of the inter-cell interference into account, effort has to be put to obtain an accurate estimation of the covariance matrix $\mathbf{R}_{vv}$.

Since this equalizer considers the interference of all users in the cell due to the full knowledge of the matrix $\mathbf{W}^{(\mathrm{MU})}$, we call it intra-cell interference aware MMSE equalizer. The standard equalizer is a special case of our solution and neglects the interference from other users, which we consequently call single-user (SU) equalizer in the following. It can be calculated from Equation (11) by using the single-user pre-coding weight matrix of rank one

$$\mathbf{W}^{(\mathrm{SU})} = \begin{bmatrix} \alpha^{(1)} w_1^{(1)} \\ \alpha^{(1)} w_2^{(1)} \end{bmatrix} \mathbf{e}_1^{\mathrm{T}}, \tag{12}$$

instead of the multi-user matrix $\mathbf{W}^{(\mathrm{MU})}$. Here, $\mathbf{e}_1$ is a zero column-vector of length $U+2$ and a one at the first position. If only a single user is receiving data in the cell, both equalizers are very similar with the only difference being that the intra-cell interference aware equalizer also considers the interference generated by the non-data channels.

Note that the structure of the derived linear equalizer would in principle allow for an extension to decision-feedback receivers. However, if the interference caused by the other users is sought to be cancelled, their data has to be estimated and thus the computational complexity of the resulting receiver would be increased significantly. Furthermore, decision feedback receivers suffer from impractical delay constraints in W-CDMA systems and error propagation.

A. Interference Suppression

Having derived the solution of the interference-aware MMSE equalizer, we want to analytically assess its interference-suppression capabilities as well as its performance bounds. To do so, we adapted the model of [20] which describes the post-equalization and despreading Signal to Interference and Noise Ratio (SINR) for arbitrary linear receivers in a multi-stream closed loop MIMO Code-Division Multiple Access (CDMA) system. The remaining intra-cell interference after equalization—for a specific channel realization and utilization of pre-coding vectors—generated by the desired user and all other active users is explicitly given by

$$P_{\text{intra}} = \sum_{\substack{m=0 \\ m\neq\tau}}^{L_{\mathrm{h}}+L_{\mathrm{f}}-2} \left|\mathbf{f}^{\mathrm{H}}\boldsymbol{\gamma}_m^{(1)}\right|^2 + \sum_{u=2}^{U}\sum_{m=0}^{L_{\mathrm{h}}+L_{\mathrm{f}}-2}\left|\mathbf{f}^{\mathrm{H}}\boldsymbol{\gamma}_m^{(u)}\right|^2, \tag{13}$$

with $\boldsymbol{\gamma}_m^{(u)}$ denoting the m-th column of the user dedicated channel matrix

$$\mathbf{H}^{(u)} = \mathbf{H}\left(\left[\alpha^{(u)} w_1^{(u)}, \alpha^{(u)} w_2^{(u)}\right]^{\mathrm{T}} \otimes \mathbf{I}_{L_{\mathrm{h}}+L_{\mathrm{f}}-1}\right), \tag{14}$$

where we also assumed perfect channel and noise power estimation at the receiver. The utilization of the pre-coding vectors in the cell also defines the "pre-coding state of the cell", for which a rigorous definition will follow in Definition IV.1 in Section IV.

Since Equation (13) depends on the current realization of the pre-coding state—in particular the pre-coding choices of the interfering users—we cannot use it directly to evaluate the performance of our proposed equalizer for general conditions. Thus, we approximate the remaining intra-cell interference by its expectation over the pre-coding choices of the other users.[1] Given a number U of active users in the cell, the remaining intra-cell interference after equalization then becomes

$$\mathbb{E}_w\{P_{\text{intra}}\} = \underbrace{\sum_{\substack{m=0 \\ m\neq\tau}}^{L_{\mathrm{h}}+L_{\mathrm{f}}-2} \left|\mathbf{f}^{\mathrm{H}}\boldsymbol{\gamma}_m^{(1)}\right|^2}_{f_{\text{self}}} + \sum_{u=2}^{U} \underbrace{\mathbb{E}_w\left\{\sum_{m=0}^{L_{\mathrm{h}}+L_{\mathrm{f}}-2}\left|\mathbf{f}^{\mathrm{H}}\boldsymbol{\gamma}_m^{(u)}\right|^2\right\}}_{f_{\text{other}}}, \tag{15}$$

with f_{self} and f_{other} denoting the determinative factors for the self- and other-user intra-cell interference remaining after equalization, and $\mathbb{E}_w\{\cdot\}$ being the expectation with respect to the pre-coding coefficients $w_1^{(u)}, w_2^{(u)}$ of the other users.

As already mentioned, the 3GPP specifies a quantized codebook of possible pre-coding vectors [4] with the UE being responsible for evaluating and signaling the pre-coding vector that leads to the best pre-equalization SINR. It is important to note that the CQI feedback has to be evaluated jointly with

[1] Note that the pre-coding of the desired user is adapted according to the current channel state and is thus known at the UE end. This is the minimum information required to evaluate any suitable equalizer.

TABLE I
SIMULATION PARAMETERS FOR FIG. 3.

Parameter	Value
simulated slots	1000
slot time	2/3 ms
receive antennas N_R	2
pre-coding codebook	3GPP TxAA [4]
equalizer span L_f	40 chips
equalizer delay τ	20 chips
pre-coding delay	11 slots
mobile speed	3 km/h
channel profile	ITU PedB [22]
active users U	4
fading model	improved Zheng model [23], [24]

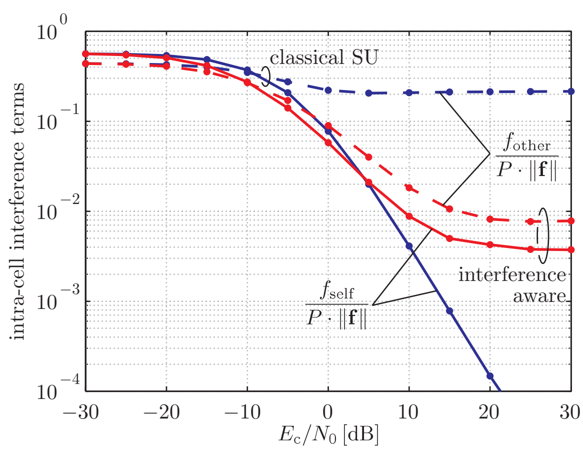


Fig. 3. Intra-cell interference terms f_{self} and f_{other} for the classical SU equalizer and the proposed interference-aware equalizer, assuming perfect knowledge of the cell's pre-coding state.

the pre-coding to obtain good throughput results [14]. Given the pre-coding codebook of [4], in [21] we were also able to show that in Double Transmit Antenna Array (D-TxAA)—even if multi-code scheduling would be implemented—all spatial dimensions are utilized. Thus, an equalizer has no possibility to exploit information about the pre-coding state of the cell to suppress the intra-cell interference.

Let us now utilize the analytical description of the post-equalization intra-cell interference to assess the interference suppression capabilities and the theoretical performance bounds of the proposed equalizer. Table I lists the simulation parameters we applied to evaluate the capability of our proposed equalizer to suppress the intra-cell interference caused by the other active users in the cell. Fig. 3 shows the performance in terms of the intra-cell interference suppression capabilities, both for the self interference f_{self} and the other-user interference f_{other}, assuming perfect knowledge of the cell's pre-coding state. We normalized the two coefficients by the total received interference power, i.e., since the channel is by assumption already normalized to one, by dividing by the norm of the equalizer coefficients $\|\mathbf{f}\|$ and the total transmitted intracell power P. It can be observed that our proposed equalizer is able to outperform the classical SU by significantly reducing the interference term f_{other} of the other users. The

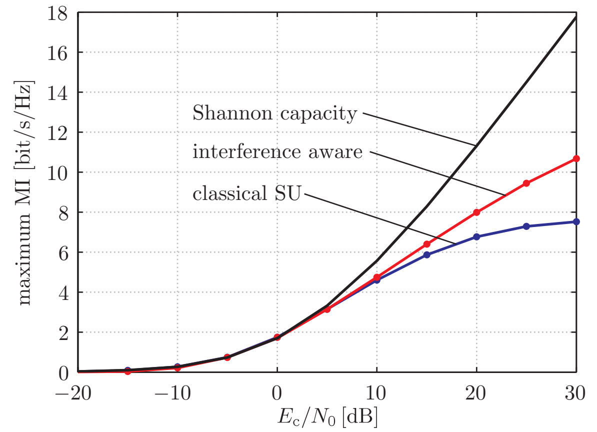


Fig. 4. Comparison of the maximum MI performances of the classical SU equalizer and the proposed interference-aware equalizer when assuming Gaussian post-equalization interference.

self interference term f_{self} on the other hand becomes larger around 5 dB E_c/N_0. As specified by the cost function in Equation (9) the interference aware equalizer minimizes the overall interference. Fig. 3 illustrates that at higher E_c/N_0 the equalizer sacrifices self interference cancellation performance for the sake of a lower overall intra-cell interference.

Based on these results, it is also possible to evaluate bounds for the spectral efficiency. With the expected intra-cell interference given by Equation (13), and considering the desired signal power being proportional to $|\mathbf{f}^{\mathrm{H}}\boldsymbol{\gamma}_\tau^{(1)}|^2$ [21], the equivalent downlink data transmission channel including the equalization can be represented as an SISO Additive White Gaussian Noise (AWGN) channel. The corresponding Signal to Noise Ratio (SNR) thus is given by

$$\mathrm{SNR} = \frac{\mathrm{SF}\left|\mathbf{f}^{\mathrm{H}}\boldsymbol{\gamma}_\tau^{(1)}\right|^2}{f_{\text{self}} + f_{\text{other}} + N_0}, \tag{16}$$

when considering only intra-cell interference and with SF denoting the spreading factor of the High-Speed Physical Downlink Shared CHannel (HS-PDSCH). This corresponds to a single-cell scenario. Based on this SNR, the maximum Mutual Information (MI) can be evaluated as $\mathrm{MI} = \log_2(1 + \mathrm{SNR})$, which denotes an upper bound if perfect (Gaussian) coding would by utilized. Also note that the distribution of the post-equalization interference in practice is not necessarily Gaussian. Nevertheless, the so derived maximum MI can serve as a figure of merit to assess the performance gain achievable by the interference-aware equalizer. Fig. 4 shows the maximum MI for the classical SU equalizer and the interference-aware equalizer together with the Shannon channel capacity of a $N_T \times N_R = 2 \times 2$ channel. It can be observed that the proposed equalizer offers huge potential performance gains in the higher E_c/N_0 region.

B. Complexity

Given the solution to the interference aware MMSE equalization in Equation (11), it is interesting to assess the additional complexity that is needed to compute our proposed filter. Assuming uncorrelated transmit sequences with constant power, i.e. $\mathbf{R}_{ss} = \sigma_s^2 \mathbf{I}$, the additional complexity can be evaluated analytically with the product $\mathbf{H}_w \mathbf{R}_{ss} \mathbf{H}_w^{\mathrm{H}} \propto \mathbf{H}_w \mathbf{H}_w^{\mathrm{H}}$ that is needed to compute the required inverse for the evaluation of $\mathbf{f}$. By writing

$$\mathbf{H}_w \mathbf{H}_w^{\mathrm{H}} = \mathbf{H} \left(\mathbf{W}^{(\mathrm{MU})} \left(\mathbf{W}^{(\mathrm{MU})} \right)^{\mathrm{H}} \otimes \mathbf{I}_{L_{\mathrm{h}}+L_{\mathrm{f}}-1} \right) \mathbf{H}^{\mathrm{H}}, \quad (17)$$

it can easily be seen that the additional cost of our proposed simulator is only determined by the larger matrix multiplication of $\mathbf{W}^{(\mathrm{MU})} \left(\mathbf{W}^{(\mathrm{MU})} \right)^{\mathrm{H}}$ instead of $\mathbf{W}^{(\mathrm{SU})} \left(\mathbf{W}^{(\mathrm{SU})} \right)^{\mathrm{H}}$, which is low compared to the cost of e.g. the inverse.

In particular, if we assume matrix multiplications to be of order $\mathcal{O}\{K^3\}$ for squared matrices, the added complexity of considering the multi-user pre-coding matrix is of order $\mathcal{O}\{(U+2)^3\}$. The matrix inverse however is of order $\mathcal{O}\{(N_{\mathrm{R}} L_{\mathrm{f}})^3\}$, and since U is typically much smaller than L_{f} in practical systems, the multiplication of the multi-user pre-coding matrix is negligible. Compared to standard chip-level MMSE equalizers [5], [14], the proposed interference-aware equalizer does not suffer of a significant increase in complexity.

IV. Pre-Coding State Estimation

In TxAA HSDPA, the MIMO channel is estimated by utilizing the CPICH, similar to UMTS. To be able to calculate the receive filter for the data channel, i.e. the HS-PDSCH, however, the mobile needs to know (a) the power offset of the individual HS-PDSCH compared to the CPICH, and (b) the pre-coding coefficients that the base-station applied for the transmission. The power offset (a) is signaled by higher layers, [25], and the pre-coding coefficients (b) of all simultaneous transmissions are signaled on the according HS-SCCH, [26], [27], where every active user has its own channel. This unfortunately makes things difficult for our proposed equalizer, because the HS-SCCHs are scrambled with user-specific scrambling sequences[2], thus making it impossible to monitor the pre-coding state of the other users.

In order to overcome this problem, three different solutions are possible

(a) change the signaling scheme in the HS-SCCH such that all active users know about the whole pre-coding state in the cell,
(b) include some training data in the HS-PDSCHs of the users to estimate the pre-coding state, or
(c) blindly estimate the pre-coding state,

whereas (a) is an obvious solution that needs no further explanation. Thus, in the following, we will shortly discuss a possible solution for (b), but our main focus is dedicated to the blind estimation (c), because it can be implemented without changes in the transmission standard.

The principal estimation problem is the following. According to Fig. 1, every active user can have his own pre-coding coefficient pair $\{w_1^{(u)}, w_2^{(u)}\}$ and his own power factor $\alpha^{(u)}$, where only the coefficients dedicated to the user himself are known (cf. the discussion above). In addition, it is also not known how many users, i.e. the number U, are currently active. Estimation problems of this kind can be investigated within the framework of random set theory [10], leading to optimum Bayesian Maximum Likelihood (ML) estimators. However, these solutions require a joint estimation of the data sequence and the pre-coding state which is typically computationally very complex and thus disadvantageous for battery powered mobile devices. Accordingly, we will restrict ourselves to classical approaches with reasonable complexity.

As already mentioned, according to [4], the pre-coding codebook that is utilized in practice is strongly quantized. In addition, several users may share the same pre-coding vector. Let us denote the codebook of possible pre-coding vectors by $\mathcal{W} = \{\mathbf{v}_1, \ldots, \mathbf{v}_{|\mathcal{W}|}\}$, $\mathbf{w}^{(u)} \in \mathcal{W}$, with $|\mathcal{W}|$ defining its cardinality. Furthermore, let us denote the set of users being served with the same pre-coding vector as $\mathcal{U}_k = \{u : \mathbf{w}^{(u)} = \mathbf{v}_k\}$, with $k = 1, \ldots, |\mathcal{W}|$ denoting the codebook index of the corresponding pre-coding vector. Note that $\bigcup_{k=1}^{|\mathcal{W}|} \mathcal{U}_k$ does not necessarily have to be equal $\mathcal{W}$ and that in case no user utilizes the pre-coding vector $\mathbf{v}_k$, the corresponding set is empty $\mathcal{U}_k = \emptyset$. With these definitions, the pre-coding state can be defined as follows.

Definition IV.1. *The pre-coding state of an HSDPA cell is defined as*

$$\mathcal{P} \triangleq \left\{ \tilde{\alpha}^{(1)}, \cdots, \tilde{\alpha}^{(k)}, \cdots, \tilde{\alpha}^{(|\mathcal{W}|)}, \alpha^{(p)} \right\}, \quad (18)$$

where each $\tilde{\alpha}^{(k)}$ denotes the total power used to transmit on a particular pre-coding vector $\mathbf{v}_k$,

$$\tilde{\alpha}^{(k)} = \sqrt{\sum_{u \in \mathcal{U}_k} \left(\alpha^{(u)} \right)^2}, \quad (19)$$

and $\alpha^{(p)}$ denotes the power coefficient of the non-data channels.

A. Training Sequence based Pre-Coding State Estimation

Although the focus of this article lies in the blind pre-coding state estimation, we developed a simple training based pre-coding state estimator as a performance bound for the blind estimator. Note that if a training based estimation should be used in HSDPA the 3GPP standard would have to be altered. Before going into the details on the training sequence based estimator, we want to note that the input-output relation in Equation (6) can be rewritten as

$$\mathbf{r} = \mathbf{H} \begin{bmatrix} \mathbf{S}\mathbf{D}_1 \\ \mathbf{S}\mathbf{D}_2 \end{bmatrix} \boldsymbol{\alpha} + \mathbf{v}, \quad (20)$$

where we omitted the time index i for sake of notational clarity. Here, the matrices $\mathbf{D}_1$ and $\mathbf{D}_2$ contain the pre-coding coefficients of the active users,

$$\mathbf{D}_1 = \mathrm{diag}\left\{ w_1^{(1)}, \cdots, w_1^{(U)}, 1, 0 \right\} \quad (21)$$

$$\mathbf{D}_2 = \mathrm{diag}\left\{ w_2^{(1)}, \cdots, w_2^{(U)}, 0, 1 \right\}, \quad (22)$$

and the matrix $\mathbf{S}$ is the re-arranged transmit vector s,

$$\mathbf{S} = \left[\mathbf{s}_i^{(1)}, \cdots, \mathbf{s}_i^{(U)}, \mathbf{p}_i^{(1)}, \mathbf{p}_i^{(2)} \right]. \quad (23)$$

The vector $\boldsymbol{\alpha}$ finally lists the power coefficients $\boldsymbol{\alpha} = \left[\alpha^{(1)}, \cdots, \alpha^{(U)}, \alpha^{(\mathrm{p})}, \alpha^{(\mathrm{p})} \right]^{\mathrm{T}}$.

[2]The scrambling sequence in HSDPA is a function of the user identification number, known only to the base-station and the particular user.

Considering the special structure of our problem in Equation (6) and (20), some redundancy can be observed. In particular, if two interfering users u_1 and u_2 utilize the same pre-coding vector $\mathbf{w}^{(k)}$, the input-output relation can also be represented by only one user utilizing $\mathbf{w}^{(k)}$ with $\sqrt{(\alpha^{(u_1)})^2 + (\alpha^{(u_2)})^2}$ as power coefficient. Furthermore, note that we assume the non-data channels at the two transmit antennas to deploy the same power coefficient $\alpha^{(\mathrm{p})}$. Thus, it is sufficient to estimate only the pre-coding state $\mathcal{P}$. For the interference-aware MMSE equalizer, this is sufficient to be able to effectively suppress the intra-cell interference.

Given the definition of the pre-coding state $\mathcal{P}$ in Equation (18), the vector $\tilde{\boldsymbol{\alpha}} = \left[\tilde{\alpha}^{(1)}, \cdots, \tilde{\alpha}^{(|\mathcal{W}|)}, \alpha^{(\mathrm{p})}\right]^{\mathrm{T}}$ has to be estimated. Let us define the estimation error as

$$C = \left\| \tilde{\boldsymbol{\alpha}} - \hat{\tilde{\boldsymbol{\alpha}}} \right\|_2^2, \tag{24}$$

with $\hat{\tilde{\boldsymbol{\alpha}}}$ denoting the estimate of $\tilde{\boldsymbol{\alpha}}$. In case that training data is available, e.g. at the beginning of each transmission frame in HSDPA, a possible estimator is given by the Least Squares (LS) solution. Let us set $\tilde{\mathbf{D}}_1 = \mathrm{diag}\left\{ w_1^{(1)}, \cdots, w_1^{(|\mathcal{W}|)} \right\}$ and $\tilde{\mathbf{D}}_2 = \mathrm{diag}\left\{ w_2^{(1)}, \cdots, w_2^{(|\mathcal{W}|)} \right\}$. In addition, let us assume that the base-station provides orthogonal training sequences $\mathbf{t}^{(k)}$ for every pre-coding vector $\mathbf{w}^{(k)}$ that can be formed into the training matrix $\mathbf{S}_{\mathrm{T}} = \left[\mathbf{t}^{(1)}, \cdots, \mathbf{t}^{(|\mathcal{W}|)}\right]$. Then the LS estimator of $\boldsymbol{\theta} = \left[\tilde{\alpha}^{(1)}, \cdots, \tilde{\alpha}^{(|\mathcal{W}|)}\right]^{\mathrm{T}}$ (not including $\alpha^{(\mathrm{p})}$) is given by [28]

$$\hat{\boldsymbol{\theta}}_{\mathrm{LS}} = \mathrm{Re}\left\{ \left(\mathbf{H} \begin{bmatrix} \mathbf{S}_{\mathrm{T}} \tilde{\mathbf{D}}_1 \\ \mathbf{S}_{\mathrm{T}} \tilde{\mathbf{D}}_2 \end{bmatrix} \right)^{\#} \mathbf{r} \right\}^{+}, \tag{25}$$

where the real-value operator $\mathrm{Re}\{\cdot\}^{+}$ ensures the coefficients to be real valued and positive, even in the low SNR regime. The coefficient $\alpha^{(\mathrm{p})}$ can be calculated with the help of the sum power constraint in Equation (8). Using the sum power constraint in Equation (8), the augmented LS estimate of $\tilde{\boldsymbol{\alpha}}$ is given by $\hat{\tilde{\boldsymbol{\alpha}}}_{\mathrm{LS}} = [\hat{\boldsymbol{\theta}}_{\mathrm{LS}}^{\mathrm{T}}, P - \hat{\boldsymbol{\theta}}_{\mathrm{LS}}^{\mathrm{T}} \hat{\boldsymbol{\theta}}_{\mathrm{LS}}]^{\mathrm{T}}$.

B. Blind Pre-Coding State Estimation

Blind estimation is generally a quite challenging task, in particular in a multi-user context. Typical approaches treat the unknown inputs—in our case the unknown transmit data $\mathbf{s}$—as *nuisance parameters* that the estimator has to cope with in order to supply blind estimates of the parameters of interest.

The ML principle provides a systematic way for deducing the Minimum Variance Unbiased (MVU) estimator, maximizing the joint likelihood function $f_{\mathbf{r}}(\mathbf{r}; \tilde{\boldsymbol{\alpha}}, \mathbf{s})$ [29]. As discussed in [30], [31], there exists a number of possibilities to avoid the joint estimation of all parameters, i.e. $\tilde{\boldsymbol{\alpha}}$ and $\mathbf{s}$. The *unconditional* or stochastic ML criterion models the vector of nuisance parameters as a random vector and maximizes the marginal of the likelihood function conditioned to $\mathbf{s}$. Unfortunately, the unconditional ML estimator is generally unknown, because the expectation with respect to $\mathbf{s}$ cannot be solved in closed form. However, in the low SNR regime, the unconditional likelihood function becomes quadratic in the observation with independence of the statistical distribution of the nuisance parameters. Nevertheless, this estimator class is generally difficult to solve and works only reasonably well in the low SNR regime [29]. This fact motivated research in second-order estimators, e.g. the *conditional* ML criterion, that models the nuisance parameters as deterministic unknowns and maximizes the compressed likelihood function. Unfortunately, this estimator class would require the matrix $\mathbf{H}_w$ to be tall [31], which in our signal model is not the case.

Another class is the *Gaussian* ML that models the nuisance parameters as Gaussian random variables in order to obtain an analytical solution for the expectation in $\mathbb{E}_{\mathbf{s}}\left\{ f_{\mathbf{r}|\mathbf{s}}(\mathbf{r}|\mathbf{s}; \tilde{\boldsymbol{\alpha}}) \right\}$, [32]. This assumption seems to fit naturally into our system model because due to multi-code operation many different transmit symbols are added up, likely resulting in a nearly Gaussian distribution. Although there are also other approaches to blindly estimate $\tilde{\boldsymbol{\alpha}}$, e.g. [33], we decided in favor of this estimator class.

According to [32], the Gaussian ML estimator is the one minimizing the nonlinear cost function

$$\begin{aligned} \Lambda_{\mathrm{GML}}(\boldsymbol{\alpha}) &= \mathrm{tr}\Big(\ln \mathbf{R}(\boldsymbol{\alpha}) + \mathbf{R}^{-1}(\boldsymbol{\alpha})\, \hat{\mathbf{R}} \Big) \\ &= \ln \det\left[\mathbf{R}(\boldsymbol{\alpha})\right] + \mathrm{tr}\Big[\mathbf{R}^{-1}(\boldsymbol{\alpha})\, \hat{\mathbf{R}} \Big], \end{aligned} \tag{26}$$

with $\hat{\mathbf{R}} = \mathbf{r}\mathbf{r}^{\mathrm{H}}$ denoting the sample covariance matrix, and

$$\mathbf{R}(\boldsymbol{\alpha}) = \mathbf{H}_w \mathbf{H}_w^{\mathrm{H}} + \sigma_v^2 \mathbf{I} \tag{27}$$

being its expected value as a function of $\boldsymbol{\alpha}$. Note however, that this would require knowledge of the number of users U to be utilized in estimation. To overcome this problem, let us point out that $\mathbf{W}^{(\mathrm{MU})}$ from Equation (7) can be rewritten as $\mathbf{W}^{(\mathrm{MU})} = \mathbf{W}_{\mathrm{CB}} \, \mathrm{diag}\{\boldsymbol{\alpha}\}$, with

$$\mathbf{W}_{\mathrm{CB}} \triangleq \begin{bmatrix} w_1^{(1)} & \cdots & w_1^{(U)} & 1 & 0 \\ w_2^{(1)} & \cdots & w_2^{(U)} & 0 & 1 \end{bmatrix}. \tag{28}$$

Then—keeping in mind that we want to estimate the pre-coding state $\mathcal{P}$—we can replace $\mathbf{W}^{(\mathrm{MU})}$ in the context of Equation (27) by

$$\tilde{\mathbf{W}} = \left[\mathbf{w}^{(1)}, \cdots, \mathbf{w}^{(|\mathcal{W}|)}\right] \mathrm{diag}\left\{ \underbrace{\left[\tilde{\alpha}^{(1)}, \cdots, \tilde{\alpha}^{(|\mathcal{W}|)}\right]^{\mathrm{T}}}_{\boldsymbol{\theta}} \right\}, \tag{29}$$

thus leading to

$$\mathbf{R}(\boldsymbol{\theta}) = \mathbf{H} \left(\tilde{\mathbf{W}} \tilde{\mathbf{W}}^{\mathrm{H}} \otimes \mathbf{I} \right) \mathbf{H}^{\mathrm{H}} + \sigma_v^2 \mathbf{I}, \tag{30}$$

and the new cost function $\tilde{\Lambda}_{\mathrm{GML}}(\boldsymbol{\theta})$. This means that the corresponding power coefficients of every possible pre-coding vector will be estimated, and thus no knowledge of the number of active users is needed anymore. Also note that as in the training based estimation, $\alpha^{(\mathrm{p})}$ can be calculated by making use of the sum power constraint in Equation (8).

The minimum of $\tilde{\Lambda}_{\mathrm{GML}}$ can for example be found by means of iterative or time-recursive scoring methods [19], [32] based on the recursion

$$\hat{\boldsymbol{\theta}}_{k+1} = \hat{\boldsymbol{\theta}}_k + \mathbf{J}_{\mathrm{GML}}^{-1}\left(\hat{\boldsymbol{\theta}}\right) \boldsymbol{\nabla}_{\mathrm{GML}}\left(\mathbf{r}; \hat{\boldsymbol{\theta}}\right), \tag{31}$$

with $\mathbf{J}_{\mathrm{GML}}^{-1}\left(\hat{\boldsymbol{\theta}}\right)$ and $\boldsymbol{\nabla}_{\mathrm{GML}}\left(\mathbf{r};\hat{\boldsymbol{\theta}}\right)$ being the Fisher information matrix and the gradient respectively in the Gaussian case. Alternatively, any other known efficient optimization technique like for example Sequential Quadratic Programming (SQP) methods [34] can be used.

When utilizing the blind pre-coding state estimator, the overall complexity of the interference-aware equalizer is of course increased. The computational complexity order of one step in the optimization process can be approximated by the complexity of evaluating the cost function,

$$2\left[\mathcal{O}\left\{N_{\mathrm{R}}L_{\mathrm{f}}\left(L_{\mathrm{h}}+L_{\mathrm{f}}-1\right)^{2}\left(U+2\right)N_{\mathrm{T}}\right\}+\mathcal{O}\left\{\left(N_{\mathrm{R}}L_{\mathrm{f}}\right)^{2}\right\}+\mathcal{O}\left\{\left(N_{\mathrm{R}}L_{\mathrm{f}}\right)^{3}\right\}\right], \tag{32}$$

where we again assumed a matrix multiplication complexity order of $\mathcal{O}\left\{K^{3}\right\}$. If an optimization algorithm based on Newton iterations is utilized, the number of required iterations can also coarsely be bounded by $\log_2 \log_2 1/\varepsilon$ with ε specifying the desired accuracy, [35]. In practice we observed a convergence of the utilized algorithm in around eight iterations. The blind pre-coding estimation thus increases the computational complexity much more significantly than the changes in the MMSE equalizer structure.

C. Estimator Performance

Fig. 5 shows the estimation error C for different E_{c}/N_0 values both for the LS estimator and the second-order blind estimator. The simulation parameters are the same as in Table I, except that we simulated the performance for one and for two receive antennas, respectively, which we denoted by $N_{\mathrm{T}} \times N_{\mathrm{R}} = 2 \times 2$ and 2×1. For the LS estimation, we used Hadamard sequences of length 64 for the training. To refine the estimation, we furthermore took advantage of the knowledge that the coefficients $\tilde{\alpha}^{(i)}$ have to be real-valued and strictly positive, as well as that the sum power constraint of Equation (8) cannot be exceeded. Fig. 5 shows that the training based estimator works reasonably well from $-20dBE_{\mathrm{c}}/N_0$ on, and that the performance saturates at around 10 dB. The blind estimator is not able to deliver similar results, with an operation range starting approximately at $-10dBE_{\mathrm{c}}/N_0$. For both estimator classes, the availability of a second receive antenna is beneficial for the pre-coding state estimation. However, for the blind pre-coding state estimator the gain is even more dramatic. Let us note that the (poor) performance of the blind estimation is still sufficient for our equalizer, as we show in Section V.

V. Throughput Evaluation

We split our performance evaluation into two different parts, (a) physical-layer simulations for a fixed transmission setup of TxAA HSDPA, and (b) system-level simulations with adaptive feedback and scheduling. Each simulation approach has a different focus, with the physical-layer simulations covering channel encoding and decoding, W-CDMA processing, as well as channel estimation in detail. On the other hand, system-level simulations represent a whole HSDPA network, with adaptive feedback, scheduling and Radio Resource Control (RRC) algorithms. For the following results we also assumed that channel and noise power are perfectly known at the receiver.

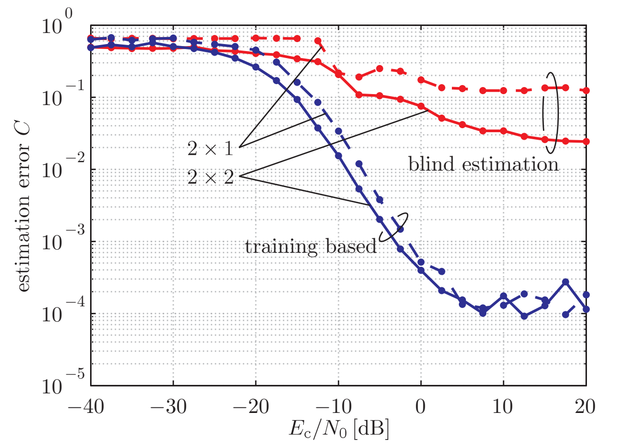


Fig. 5. Estimation error C of the LS and the second-order blind estimator versus E_{c}/N_0, see Equation (24).

TABLE II
Simulation Parameters for Physical-Layer Simulations.

Parameter	Value
active users U	4
desired user CQI	13
interfering HS-PDSCH $E_{\mathrm{c}}/I_{\mathrm{or}}$	$[-6, -8, -10]$ dB
interfering user CQIs	$[16, 11, 8]$
interfering user pre-coding	$[1, \frac{1}{\sqrt{2}}(1-j)]$, $[1, \frac{1}{\sqrt{2}}(-1+j)]$, $[1, \frac{1}{\sqrt{2}}(-1-j)]$
pre-coding codebook	3GPP TxAA [4]
CPICH $E_{\mathrm{c}}/I_{\mathrm{or}}$	-10 dB
other non-data channel $E_{\mathrm{c}}/I_{\mathrm{or}}$	-12 dB
UE capability class	6
channel profile	ITU PedA, PedB
UE speed	3 km/h

A. Physical-Layer Simulation Results

We conducted physical-layer simulations utilizing a standard compliant W-CDMA simulator. The simulation assumptions in Table II correspond to a cell in which four users are receiving data simultaneously. User 1 is moving through the cell and obtains the pre-coding coefficients as adaptively requested, according to the definition in the standard [4]. We assume the three interfering users to be stationary, thus their pre-coding coefficients and transmit power do not change. In these simulations we assume that all users are always scheduled with the same CQI value, i.e., no link adaptation besides the pre-coding. Dynamic pre-coding for all users will be considered in the system-level simulations. Also we again defined the MIMO antenna scenarios by $N_{\mathrm{T}} \times N_{\mathrm{R}} = 2 \times 2$ and 2×1, respectively.

The achieved data throughput of User 1 in a Pedestrian A (PedA) and Pedestrian B (PedB) environment is plotted in Fig. 6 and Fig. 7, respectively. In both scenarios, the Interference Aware (IA) equalizer with perfect knowledge[3]

[3]For these investigations we did not consider any overhead for the perfect pre-coding state knowledge, as is necessary if the pre-coding state is signaled.

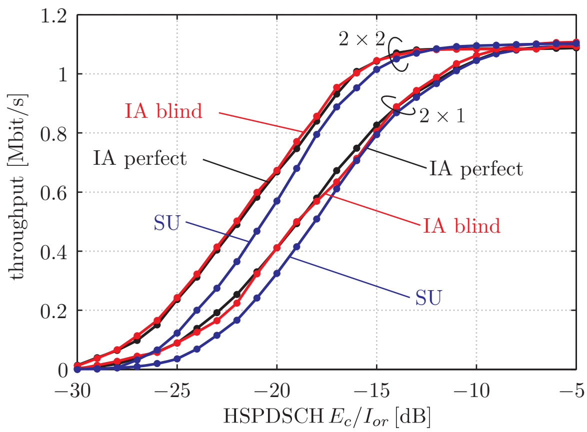


Fig. 6. Throughput of desired user in a spatially uncorrelated ITU PedA channel at CQI 13, corresponding to a maximum throughput of 1.14 Mbit/s.

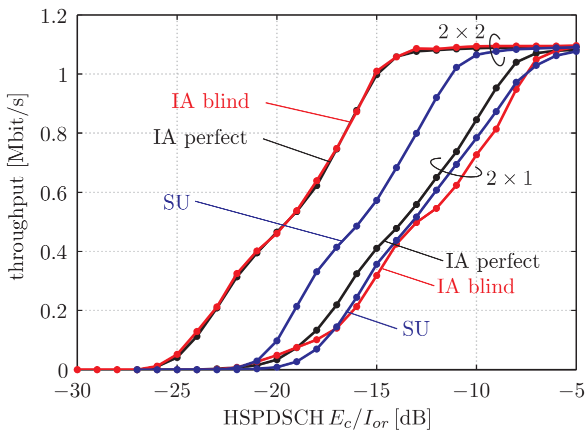


Fig. 7. Throughput of desired user in a spatially uncorrelated ITU PedB channel at CQI 13, corresponding to a maximum throughput of 1.14 Mbit/s.

of the pre-coding state significantly outperforms the single user equalizer. If the pre-coding state of the cell is blindly estimated, the performance of the MMSE equalizer nearly approaches the performance when $\mathcal{P}$ is perfectly known and two receive antennas are available. Here, $E_\mathrm{c}/I_\mathrm{or}$ denotes the ratio of the average transmit energy per chip (E_c) to the total transmit power spectral density (I_or).

The gain in the PedB channel in Fig. 7 is much larger than the gain in the PedA channel which has a much shorter maximum delay spread. This is caused by the larger loss of orthogonality in the PedB environment and the subsequently larger post equalization interference. In the 2×1 case, the equalizer applying the blind pre-coding state estimation looses compared to the equalizer with perfect knowledge of $\mathcal{P}$, which is a result of the considerably worse estimator performance when only one receive antenna is available, see Fig. 5. The fact that the performance loss is greater in the PedB channel is due to its significantly larger delay spread, which represents a more challenging environment for the equalizer making it more sensitive to estimation errors in the pre-coding state.

For larger number of receive antennas, the simulation results show larger performance gains. Especially in Fig. 7, the pre-coding state estimation improves significantly in the 2×2 case, thus closing the gap to the throughput performance of the equalizer utilizing perfect knowledge of $\mathcal{P}$. The interference aware equalizer can effectively utilize the spatial information to suppress the interfering signals. The largest performance increase of the interference aware equalizer was found for the 2×2 PedB environment with 4 dB. We also conducted a set of simulations with varying equalizer length L_f and different CQI and pre-coding setups, but the same conclusions as observable in Fig. 6 and Fig. 7 were drawn, thus we did not include these results in the paper. The key parameters are the delay spread and the number of receive antennas.

TABLE III
SIMULATION PARAMETERS FOR SYSTEM-LEVEL SIMULATIONS.

Parameter	**Value**
simultaneously active users U	4
transmitter frequency	1.9 GHz
base-station distance	1000 m
total power available at Node-B	20 W
CPICH power	0.8 W
power of other non-data channels	1.2 W
spreading codes available for HSDPA	15
macro-scale pathloss model	urban micro [36]
scheduler	round robin
stream power loading	uniform
users in the cell	25
cell deployment	layout type 1 [37]
pre-coding codebook	3GPP TxAA [4]
UE capability class	10
equalizer span	40 chips
feedback delay	11 slots
channel profile	ITU PedA, PedB
UE speed	3 km/h, random direction
simulation time	25 000 slots, each 2/3 ms

B. System-Level Results

To assess the performance on network level, we also conducted a set of system-level simulations with the simulator described in [20], [21]. The simulation assumptions in Table III correspond to a 19 site scenario with a homogenous network load in which the multi-code scheduler serves four active users simultaneously. All 25 simulated users are moving through the cell with random directions, adaptively reporting their CQI and pre-coding feedback according to their capability class [15]. The feedback delay was set to eleven slots. The estimation of the pre-coding state of the cell for the MU interference-aware equalizer was performed blindly, as described in Section IV-B.

The distributions of the SINR for the PedA and PedB channel, averaged over all active users in the cell, are plotted in Fig. 8. It can be observed that the IA equalizer is able to deliver significantly higher SINRs for PedB channels. In the PedA environment, the gain is negligible.

Fig. 9 shows the average sector throughput. The IA equalizer outperforms the classical (SU) equalizer significantly, again with more remarkable gains in the PedB environment—up to 11.7%. Similar to the physical-layer simulation results, for both channels the equalizer is able to utilize the advantage

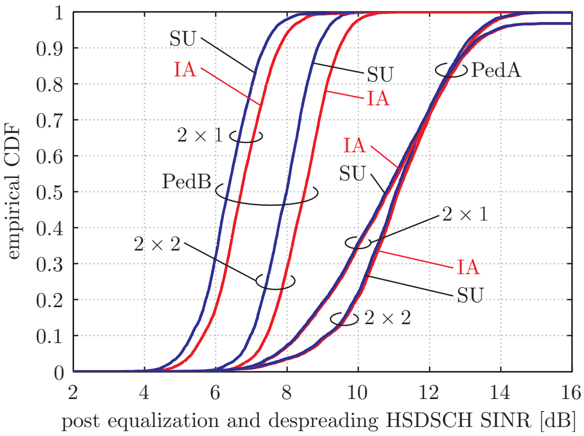


Fig. 8. Empirical CDFs of the post equalization and despreading SINR of the HS-DSCH.

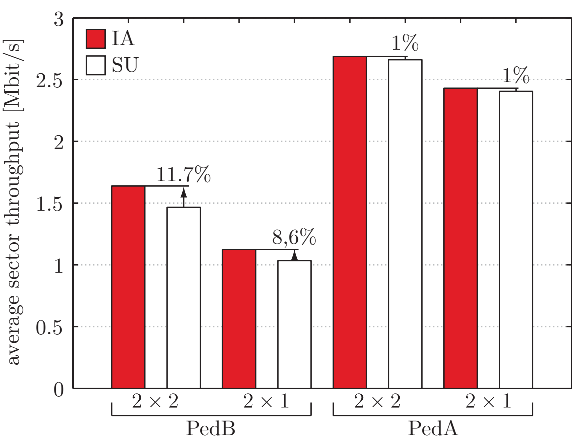


Fig. 9. Average sector throughput results in the network specified by Table III.

of multiple receive antennas (2×2) to advance the pre-coding state estimation, and thus the throughput performance compared to the single receive antenna case (2×1).

VI. Conclusions

In this article, we present a system model for TxAA HSDPA that takes the structure of the intra-cell interference in case of multi-code scheduling into account. The consideration of all simultaneously served users in the derivation of the MMSE equalizer leads to an interference aware equalizer which has only slightly increased complexity compared to the SU MMSE equalizer. Simulations show greatly reduced post equalization interference for the interference aware equalizer. We furthermore identify a description for the pre-coding state of an HSDPA cell, which allows us to formulate and test training-based and blind estimators. Finally, we evaluate the performance gain by means of physical-layer and system-level simulations. The results show that the interference aware MMSE equalizer performance gain increases with the frequency selectivity of the channel and the number of receive antennas. Both physical-layer as well as system-level simulations identify our proposed receiver structure as superior to the classical approach.

Acknowledgment

This work has been funded by the Institute of Communications and Radio-Frequency Engineering, mobilkom austria AG, and the Christian Doppler Laboratory for Wireless Technologies for Sustainable Mobility. The views expressed in this paper are those of the authors and do not necessarily reflect the views within mobilkom austria AG.

References

[1] M. Wrulich, C. Mehlführer, and M. Rupp, "Interference aware MMSE equalization for MIMO TxAA," in *Proc. IEEE 3rd International Symposium on Communications, Control and Signal Processing (ISCCSP)*, 2008, pp. 1585–1589.

[2] C. Mehlführer, M. Wrulich, and M. Rupp, "Intra-cell interference aware equalization for TxAA HSDPA," in *Proc. IEEE 3rd International Symposium on Wireless Pervasive Computing*, 2008, pp. 406–409.

[3] H. Holma and A. Toskala, *WCDMA for UMTS – Radio Access For Third Generation Mobile Communications*, 3rd ed. John Wiley & Sons, Ltd, 2005.

[4] Technical Specification Group Radio Access Network, "Multiple-input multiple-output UTRA," 3rd Generation Partnership Project (3GPP), Tech. Rep. TR 25.876 Version 7.0.0, Mar. 2007.

[5] M. Melvasalo, P. Janis, and V. Koivunen, "MMSE equalizer and chip level inter-antenna interference canceler for HSDPA MIMO systems," in *Proc. IEEE 63rd Vehicular Technology Conference Spring (VTC)*, vol. 4, 2006, pp. 2008–2012.

[6] S. Shenoy, M. Ghauri, and D. Slock, "Receiver designs for MIMO HSDPA," in *Proc. IEEE International Conference on Communications (ICC)*, 2008, pp. 941–945.

[7] H. Zhang, M. Ivrlac, J. A. Nossek, and D. Yuan, "Equalization of multiuser MIMO high speed downlink packet access," in *Proc. IEEE 19th International Symposium on Personal, Indoor and Mobile Radio Communications (PIMRC)*, 2008, pp. 1–5.

[8] H. Zhang and H. Dai, "Cochannel interference mitigation and cooperative processing in downlink multicell multiuser mimo networks," *EURASIP J. Wireless Commun. Netw.*, vol. 2, pp. 222–235, Dec. 2004.

[9] F. Petre, M. Engels, A. Bourdoux, B. Gyselinckx, M. Moonen, and H. D. Man, "Extended MMSE receiver for multiuser interference rejection in multipath DS-CDMA channels," in *Proc. IEEE VTS 50th Vehicular Technology Conference Fall (VTC)*, vol. 3, Sep. 1999, pp. 1840–1844.

[10] E. Biglieri and M. Lops, "Multiuser detection in a dynamic environment—part I: user identification and data detection," *IEEE Trans. Inf. Theory*, vol. 53, no. 9, pp. 3158–3170, Sep. 2007.

[11] E. Virtej, M. Lampinen, and V. Kaasila, "Performance of an intra- and inter-cell interference mitigation algorithm in HSDPA system," in *Proc. IEEE 67th Vehicular Technology Conference Spring (VTC)*, 2008, pp. 2041–2045.

[12] Technical Specification Group Radio Access Network, "Feasibility study on interference cancellation for UTRA FDD user equipment (UE)," 3rd Generation Partnership Project (3GPP), Tech. Rep. TR 25.963 Version 7.0.0, Apr. 2007.

[13] L. Mailaender, "Linear MIMO equalization for CDMA downlink signals with code reuse," *IEEE Trans. Wireless Commun.*, vol. 4, no. 5, pp. 2423–2434, Sep. 2005.

[14] C. Mehlführer, S. Caban, M. Wrulich, and M. Rupp, "Joint throughput optimized CQI and precoding weight calculation for MIMO HSDPA," in *Proc. 42nd Asilomar Conference on Signals, Systems and Computers*, Oct. 2008.

[15] Technical Specification Group Radio Access Network, "Physical layer procedures (FDD)," 3rd Generation Partnership Project (3GPP), Tech. Rep. TS 25.214 Version 7.4.0, Mar. 2007.

[16] D. I. Kim and S. Fraser, "Two-best user scheduling for high-speed downlink multicode CDMA with code constraint," in *Proc. IEEE Global Telecommunications Conference (GLOBECOM)*, vol. 4, 2004, pp. 2569–2663.

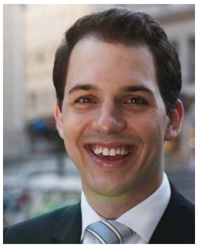
[17] Technical Specification Group Radio Access Network, "Evolved universal terrestrial radio access (E-UTRA); LTE physical layer – general description," 3rd Generation Partnership Project (3GPP), Tech. Rep. TS 36.201 Version 8.3.0, Mar. 2009.

[18] J. Proakis, *Digital Communications*, 4th ed. McGraw-Hill Science Engineering, Aug. 2000.

[19] S. Haykin, *Adaptive Filter Theory*, 4th ed. Prentice-Hall, 2002.

[20] M. Wrulich, S. Eder, I. Viering, and M. Rupp, "Efficient link-to-system level model for MIMO HSDPA," in *Proc. IEEE 4th Broadband Wireless Access Workshop*, 2008.

[21] M. Wrulich and M. Rupp, "Computationally efficient MIMO HSDPA system-level modeling," *EURASIP J. Wireless Communications and Networking*, vol. 2009, Article ID 382501, 14 pages, 2009. doi:10.1155/2009/382501.
[22] Members of ITU, "Recommendation ITU-R M.1225: Guidelines for evaluation of radio transmission technologies for IMT-2000," International Telecommunication Union (ITU), Tech. Rep., 1997.
[23] Y. Zheng and C. Xiao, "Simulation models with correct statistical properties for rayleigh fading channels," *IEEE Trans. Commun.*, vol. 51, no. 6, pp. 920–928, June 2003.
[24] T. Zemen and C. Mecklenbräuker, "Time-variant channel estimation using discrete prolate spheroidal sequences," *IEEE Trans. Signal Process.*, vol. 53, no. 9, pp. 3597–3607, Sep. 2005.
[25] Technical Specification Group Radio Access Network, "Radio link control (RLC) protocol specification," 3rd Generation Partnership Project (3GPP), Tech. Rep. TS 25.322 Version 8.4.0, Mar. 2009.
[26] ——, "Multiplexing and channel coding (FDD)," 3rd Generation Partnership Project (3GPP), Tech. Rep. TS 25.212 Version 8.5.0, Mar. 2009.
[27] ——, "Medium access control (MAC) protocol specification," 3rd Generation Partnership Project (3GPP), Tech. Rep. TS 25.321 Version 7.0.0, Mar. 2006.
[28] T. K. Moon and W. C. Stirling, *Mathematical Methods and Algorithms for Signal Processing*. Prentice-Hall, 2000.
[29] S. M. Kay, *Fundamentals of Statistical Signal Processing. Estimation Theory.* Prentice-Hall, 1993, vol. 1.
[30] B. Ottersten, M. Viberg, and T. Kailath, "Analysis of subspace fitting and ML techniques for parameter estimation from sensor array data," *IEEE Trans. Signal Process.*, vol. 40, no. 3, pp. 590–600, Mar. 1992.
[31] J. Villares and G. Vázquez, "Second-order parameter estimation," *IEEE Trans. Signal Process.*, vol. 53, no. 7, pp. 2408–2420, July 2005.
[32] ——, "The gaussian assumption in second-order estimation problems in digital communications," *IEEE Trans. Signal Process.*, vol. 55, no. 10, pp. 4994–5002, Oct. 2007.
[33] P. Stoica and N. Arye, "MUSIC, maximum likelihood, and Cramer-Rao bound," *IEEE Trans. Acoust., Speech, Signal Process.*, vol. 37, no. 5, pp. 720–741, May 1989.
[34] M. J. D. Powell, *A Fast Algorithm for Nonlinearly Constrained Optimization Calculations*. Springer Berlin / Heidelberg, 1978, vol. 630, pp. 144–157.
[35] S. Boyd and L. Vandenberghe, *Convex Optimization*. Cambridge University Press, 2004.
[36] D. J. Cichon and T. Kürner, *COST 231 – Digital Mobile Radio Towards Future Generation Systems*. COST, 1998, ch. 4.
[37] Technical Specification Group Radio Access Network, "Spatial channel model for multiple input multiple output (MIMO) simulations," 3rd Generation Partnership Project (3GPP), Tech. Rep. TS 25.996 Version 7.0.0, Jun. 2007.



Martin Wrulich was born in 1980 in Klagenfurt, Carinthia. He received his Dipl.-Ing. degree from Vienna University of Technology in March 2006, which was awarded by the "Diplomarbeitspreis der Stadt Wien". Furthermore, he received three times the scholarship for excellent academic performance from the faculty of electrical engineering and information technology. He started his doctoral studies in the area of system-level modeling and network optimization at the institute of communications and radio-frequency engineering in Vienna as a project assistant for Prof. Rupp in a research cooperation between Vienna University of Technology and mobilkom austria AG. In 2009, he received a Marshallplan scholarship to join the group of Prof. Paulraj at Stanford University, after which he finished his Ph.D. "sub-auspiciis presidentis." He also acted as a TPC member for the IEEE broadband wireless access workshop in 2009. His current research interests are system-level modeling, resource allocation optimization, and cooperative communications.

Christian Mehlführer was born in Vienna, Austria, in 1979. In 2004 he received his Dipl.-Ing. degree in electrical engineering from the Vienna University of Technology. Besides his diploma studies he worked part time at Siemens AG where he performed integration tests of GSM carrier units. After finishing his diploma thesis on implementation and real-time testing of space-time block codes at the Institute of Communications and Radio-Frequency Engineering, Vienna University of Technology, for which he received the Vodafone Förderpreis 2006 (together with Sebastian Caban), he started his doctoral thesis at the same institute. In 2009, he finished his Ph.D. about measurement-based performance evaluation of WiMAX and HSDPA with summa cum laude. His research interests include experimental investigation of MIMO systems, UMTS HSDPA, WLAN (802.11), WiMAX (802.16), and the upcoming 3GPP LTE system.

Markus Rupp received his Dipl.-Ing. degree in 1988 at the University of Saarbruecken, Germany and his Dr.-Ing. degree in 1993 at the Technische Universitaet Darmstadt, Germany, where he worked with Eberhardt Haensler on designing new algorithms for acoustical and electrical echo compensation. From November 1993 until July 1995, he had a postdoctoral position at the University of Santa Barbara, California with Sanjit Mitra where he worked with Ali H. Sayed on a robustness description of adaptive filters with impact on neural networks and active noise control. From October 1995 until August 2001 he was a member of Technical Staff in the Wireless Technology Research Department of Bell-Labs at Crawford Hill, NJ, where he worked on various topics related to adaptive equalization and rapid implementation for IS-136, 802.11 and UMTS. Since October 2001 he is a full professor for Digital Signal Processing in Mobile Communications at the Vienna University of Technology where he founded the Christian-Doppler Laboratory for Design Methodology of Signal Processing Algorithms in 2002 at the Institute for Communciations and RF Engineering. He served as Dean from 2005-2007. He was associate editor of IEEE TRANSACTIONS ON SIGNAL PROCESSING from 2002-2005, is currently associate editor of EURASIP JOURNAL OF ADVANCES IN SIGNAL PROCESSING, and EURASIP JOURNAL ON EMBEDDED SYSTEMS. He is elected AdCom member of EURASIP since 2004 and serving as president of EURASIP from 2009-2010. He authored and co-authored more than 300 papers and patents on adaptive filtering, wireless communications, and rapid prototyping, as well as automatic design methods.

Reference:

M. Wrulich, C. Mehlführer, and M. Rupp, “Managing the interference structure of MIMO HSDPA: A multi-user interference aware MMSE receiver with moderate complexity,” IEEE Transactions on Wireless Communications, vol. 9, no. 4, pp. 1472–1482, Apr. 2010, doi: 10.1109/TWC.2010.04.090612. [Online]. Available: http://publik.tuwien.ac.at/files/PubDat_180743.pdf

BibTeX:

```
@article{Wrulich_TWC2010,
  author =        {M. Wrulich and C. Mehlf\"uhrer and M. Rupp},
  title =         {Managing the Interference Structure of {MIMO} {HSDPA}: A Multi-User Interference
Aware {MMSE} Receiver with Moderate Complexity},
  journal =       {IEEE Transactions on Wireless Communications},
  volume =        9,
  number =        4,
  year =          2010,
  month =         apr,
  pages =         {1472--1482},
  ISSN =          {1536-1276},
  doi =           {10.1109/TWC.2010.04.090612},
  url =           {http://publik.tuwien.ac.at/files/PubDat_180743.pdf},
}
```