AN EXEMPLARY COMPARISON OF PER ANTENNA RATE CONTROL BASED MIMO-HSDPA RECEIVERS

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ABSTRACT

High-Speed Downlink Packet Access (HSDPA) using Multiple-Input Multiple-Output (MIMO) techniques received increasing attention recently, because MIMO-HSDPA allows capacity enlargement and, therefore, supports the demand of high data services, which are expected to become the dominant source of 3G traffic and of revenue in cellular communications in the next few years. This contribution will discuss one particular MIMO-HSDPA scheme – the Per Antenna Rate Control (PARC) – from the viewpoint of conventional receiver architectures. In addition, the Signal-to-Interference-Ratio (SIR) is introduced as a basic metric for the channel quality indication required in a PARC-based system, by briefly reviewing blind and data-aided SIR estimation methods.

1. INTRODUCTION

HSDPA introduces a set of advanced techniques to improve the efficiency of downlink transmissions for packet-oriented data services and multimedia services in W-CDMA-based networks [1]:

- Adaptive Modulation and Coding (AMC),
- Hybrid Automatic Repeat Request (H-ARQ),
- Fast Scheduling and
- Fast Cell Selection.

All these techniques aim to assist fast adaptations of the transmission to the current channel or network situation. HSDPA is designed as an enhancement for currently deployed WCDMA-based networks; compatibility to former released specifications is maintained. Recently, MIMO technologies have also been proposed for deployment in UTRA mainly to improve data rate [2], we call this approach in the following MIMO-HSDPA.

One major ingredient of MIMO-HSDPA is the so-called *Per Antenna Rate Control (PARC)*. This scheme has been proposed by Lucent Technologies and can be seen as an *enhanced* Vertical-Bell LAboratories layered Space-Time (V-BLAST) system, which achieves spatial multiplexing gain through different data streams on each transmit antenna. Since the BLAST architecture belongs to an open-loop MIMO architecture, the same modulation and coding scheme, as well as equal transmit power, are assigned to all sub-layers.

If the transmitter is equipped with Channel State Information (CSI) and therefore has the principal ability to adjust the antenna data rates independently according to the current channel behaviour, a layered receiver architecture (MMSE with successive interference cancellation) may approach the Shannon capacity limit for an open-loop MIMO link [2]. The Per Antenna Rate Control (PARC) is such a so-called closed-loop MIMO system using Adaptive Modulation and Coding (AMC) for each transmit antenna to make best use of the most recent CSI. The general structure of the PARC transmitter can be seen in Figure 1.

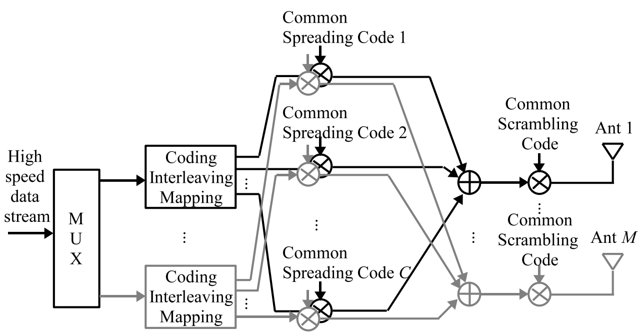

Figure 1: PARC transmitter

The receiver in such a PARC-based system has to measure the channel quality of the received signal from each transmit antenna in the presence of multipath and interference from the other antennas and feeds back side informations - so-called *channel quality indications* (CQI) - to the transmitter. Based on these channel quality indications the transmitter selects the data rate and the associated Modulation and Coding Scheme (MCS) for each antenna individually.

The signal quality seen by the decision unit (symbol decoder) is highly dependent on the performance of the chosen receiver architecture, and not only on the channel. We, therefore, aim to compare the perfomance of well-known receivers at first and introduce signal quality estimation based on the Signal-to-Interference Ratio (SIR) as basic signal quality metric, later.

2. SYSTEM OVERVIEW AND MODEL

The system under consideration is based on the UMTS-FDD HSDPA downlink with M transmit and N receive antennas. The high speed data stream is demultiplexed into M parallel streams and then, according to the specification of HSDPA, 16-QAM or QPSK and an OVSF code with spreading factor SF = 16 are used for each stream.

As Figure 1 shows, the multicode technique can be applied to each stream and, in addition, the OVSF codes can be reused among the streams of different transmit antennas. The resulting signals are multiplied with the same cell-specific primary scrambling code to form the set of M transmitted signal vectors $X_m(k) = [\ldots, x_m(k-1), x_m(k)]^{\mathrm{T}}$.

The frequency selective MIMO channel (with length L) for each spatial path from antenna $m = 1, \ldots, M$ to antenna $n = 1, \ldots, N$ is described through a vector $h_{m,n} = [h_{m,n}(L), h_{m,n}(L-1), \ldots, h_{m,n}(1)]^{\mathrm{T}}$ whose elements are assumed to be uncorrelated, zero mean Gaussian distributed, with their respective variances defined through a single power delay profile across all spatial subchannels.

At first, we examine a system with a single transmit antenna and a single receive antenna. Let $X = [x(1), \ldots, x(k-1), x(k), \ldots]^{\mathrm{T}}$ denote the vector of transmitted data and $Y = [y(1), \ldots, y(k-1), y(k), \ldots]^{\mathrm{T}}$ denote the vector of received samples. Let L_{eq} be the span of the equalizer, measured in units of the chip period. Then a set of L_{eq} received samples is given by

$$\mathbf{y}[k] = \mathrm{diag}(\mathbf{h}^{\mathrm{T}}) \cdot \mathbf{x}[k] + \mathbf{n}[k] \tag{1}$$

where $\mathbf{y}[k] = [y(k - L_{eq} + 1),\ y(k - L_{eq}),\ \cdots,\ y(k-1),\ y(k)]^{\mathrm{T}}$, $\mathbf{x}[k] = [x(k - L_{eq} - L + 2),\ x(k - L_{eq} - L + 1),\ \cdots,\ x(k-1),\ x(k)]^{\mathrm{T}}$ and $\mathbf{n}[k] = [n(k - L_{eq} + 1),\ n(k - L_{eq}),\ \cdots,\ n(k-1),\ n(k)]^{\mathrm{T}}$.

To generalize the signal mode for M transmit and N receive antennas, we can add subscripts to denote the transmitted signal from the m-th transmit antenna and received signal from the n-th receive antenna. At the same time the toeplitz matrices $\mathbf{H}_{m,n}$ describe the frequency selective channel between the m-th transmit antenna and the n-th receive antenna. By stacking the received vectors, we obtain the basic formula for a MIMO scenario

$$\underbrace{\begin{bmatrix} \mathbf{y}_1(k) \\ \vdots \\ \mathbf{y}_N(k) \end{bmatrix}}_{\mathbf{Y}(k)} = \underbrace{\begin{bmatrix} \mathbf{H}_{1,1} & \ldots & \mathbf{H}_{1,M} \\ \vdots & \ddots & \vdots \\ \mathbf{H}_{N,1} & \ldots & \mathbf{H}_{N,M} \end{bmatrix}}_{\mathbf{H}} \underbrace{\begin{bmatrix} \mathbf{x}_1(k) \\ \vdots \\ \mathbf{x}_N(k) \end{bmatrix}}_{\mathbf{X}(k)} + \mathbf{N}(k). \tag{2}$$

3. RECEIVER FOR CDMA

Before we proceed with MIMO-HSDPA receiver architectures conventional approaches for CDMA reception are briefly revisited. *Rake receivers* are commonly used in CDMA systems, especially in W-CDMA systems, mainly because of their numerical stability. However, it is known that the performance of such conventional Rake receivers is severely limited by self- and multiple-access interference through multipath propagation, despite of the beneficial multipath diversity.

Note also, that under HSDPA, a low spreading factor and multipath conditions increase the loss of orthogonality among Orthogonal Variable Spreading Factor (OVSF) channelization codes. In addition, the multicode approach is tightly involved with MIMO-HSDPA (or its counterpart 1X EV-DV) to further increase the downlink data rate with different transmit antennas, while keeping the same spreading codes. This will significantly aggravate the spatial interference. So, the adverse effect of the interference is in all likelihood the dominant factor and the conventional Rake receiver performance is, therefore, diminished. [3].

In general, for MIMO-HSDPA, the desired data stream corresponding to a given code and transmit antenna suffers from three types of interferences: the *inter chip interference* (ICI) due to the multipath propagation of the transmitted data stream spread by a single own code, *inter code interference* due to the multipath propagation of the transmitted data streams spread by other codes (either *multi user interference* or *multi code interference*), and *self interference* (SI) from data streams of the desired user (so sharing the same code) but transmitted from other own antennas. Combined *spatial/temporal equalization* and advanced interference suppression/cancellation techniques should be employed to account for all these interferences.

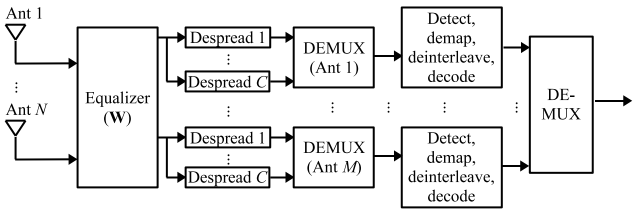

Figure 2: MIMO receiver

Alternatively to the conventional Rake receiver a *weighting matrix* ($\mathbf{W}$) based MIMO receiver is shown in Figure 2. The optimum equalizer is a maximum a posteriori (MAP) equalizer taking adequately into account all types of the mentioned interferences. Such a receiver is not feasible in practice, because of its enormous numerical complexity.

Hence, sub-optimal linear chip-spaced equalization is given more attention, such as:

1. a ZF (Zero-Forcing) equalizer where the weighting matrix is given by $\mathbf{W} = (\mathbf{H}\mathbf{H}^{\mathrm{H}})^{-1}\mathbf{H}^{\mathrm{H}}$ (see Fig. 3),
2. an MMSE (Minimum Mean Square Error) equalizer where $\mathbf{W} = (\mathbf{H}\mathbf{H}^{\mathrm{H}} + {}^{2}_{n}\mathbf{I})^{-1}\mathbf{H}^{\mathrm{H}}$.

After equalization the recovered transmit signals $\hat{\mathbf{x}}(k) = \mathbf{W}^{\mathrm{H}}\mathbf{y}(k)$ will be despreaded, demodulated and decoded.

It is worthwhile to mention here that the popularity of the Rake receiver is mainly due the ease of implementing its basic functional blocks. In contrast, matrix processing based receiver structures can show up numerical instability because of sensitivity to fix-point calculations.

4. CHANNEL QUALITY MEASUREMENT AND INDICATION

In the specification given by [4] the CQI is basically a five bit value, representing an index to a pre-defined table of MCSs, which was designed according to simulation results to guarantee a certain Frame Error Rate (FER) on the HS-PDSCHs (Physical Downlink Shared Channel). This table is specific to the category of the user equipment (UE), which also represents the UE abilities to handle a certain MCS and the number of parallel HS-PDSCH. In addition to the design of such a pre-defined table, a fast and reliable method is required to determine the channel quality, seen by the UE.

The FER or Bit Error Rate (BER) are well-known metrics to characterize the perfomance of a receiver in different channel realizations. On the other hand, these metrics can be used to define a target error rate and to determine, which MCS should be used, in order to achieve the highest possible throughput. Unfortunately, error rates can not be directly measured in a real-life time-variant system, so other metrics must be used.

A proposed metric to determine the channel quality is the estimation of the Signal-to-Interference Ratio (SIR) [5] of the primary Common PIlot CHannel (pCPICH), and if Space-Time Transmit Diversity (STTD) is involved, also on the secondary Common PIlot CHannel (sCPICH). The SIR is used on behalf of the Signal-to-Interference-plus-Noise Ratio (SINR), because interferences are usually dominant.

For a MIMO-HSDPA system exploiting PARC this metric describing the quality of the channel seen by the multiple receive antennas can be deployed as follows: We assume that a pilot channel (pCPICH or one out of multiple sCPICH) is transmitted per antenna, to assist the channel estimation, so it can also be used to estimate the channel quality seen for each transmit antenna. We suppose for the latter, that the data used on the pilot channel are handled by the receiver like any other data channel. The output of the equalizer are the MIMO processed, equalized and despreaded symbols $\tilde{d}_k$ of the corresponding pilot channels for each transmit antenna seperately. Note that for reasons of simplicity we do not augment the symbols by additional indices, like one for different antennas etc..

Similar to the proposal in [7] the Euclidean distance

$$ {}_k = |\tilde{d}_k - d_k|^2 \tag{3}$$

of the k-th received symbol $\tilde{d}_k$ and the k-th transmitted symbol d_k can be used as metric for channel quality. Under high SIR conditions the metric small and for low SIR conditions high.

Basically, this metric is motivated by the simple system model

$$\tilde{d}_k = d_k + i_k, \tag{4}$$

where the power of interference i_k is given by

$$I = \mathrm{E}\{|i_k|^2\} = \mathrm{E}\{|\tilde{d}_k - d_k|^2\}. \tag{5}$$

Provided that d_k and i_k are uncorrelated, it follows

$$\tilde{D} = \mathrm{E}\{|\tilde{d}_k|^2\} = \underbrace{\mathrm{E}\{|d_k|^2\}}_{=D} + \underbrace{\mathrm{E}\{|i_k|^2\}}_{=I}, \tag{6}$$

so that

$$\mathrm{SIR}_{\text{data-aided}} = 10\log\left(\frac{\mathrm{E}\{|\tilde{d}_k|^2\}}{\mathrm{E}\{|\tilde{d}_k - d_k|^2\}} - 1\right). \tag{7}$$

Note that this postprocessed SIR does not necessarily correspond to the SIR of the channel, because it depends on the equalizer performance and quality of synchronisation. In any case it represents the signal quality seen by the decision unit (symbol decoder), which is of major relevance in a PARC system to correctly choose the MCS.

Because knowledge of the transmitted data is required, we will refer this method as *data-aided*. If the channel is not perfectly equalized, e.g. for imperfect phase reconstruction of the data symbols received, we expect a loss of performance of this data aided SIR estimator.

Opposite to the data-aided approach is a blind method which exploits only the statistical signal properties and requires therefore almost no knowledge about the transmitted signal. In order to sketch the underlying idea assume that the sCPICH and pCPICH are QPSK-modulated, so that the despreaded and equalized output $\tilde{d}_k$ shows a *constant magnitude* in the undistorted case. For that reason the squared expectation of the magnitude

$$(\mathrm{E}\{|\tilde{d}_k|\})^2 = \tilde{D}_\mathrm{m} \tag{8}$$

equals the power $\mathrm{E}\{|\tilde{d}_k|^2\}$ of $\{\tilde{d}_k\}$. Hence, fluctuations of the constant magnitude orginate from the distortions i_k, so that ${}^2_{|\tilde{d}_k|}$ becomes a measure for

$$I = {}^2_{|\tilde{d}_k|} = \mathrm{E}\{|\tilde{d}_k|^2\} - \mathrm{E}\{|\tilde{d}_k|\}^2. \tag{9}$$

Dividing $\tilde{D}_\mathrm{m}$ and I yields

$$\frac{\tilde{D}_\mathrm{m}}{I} = \frac{(\mathrm{E}\{|\tilde{d}_k|\})^2}{\mathrm{E}\{|\tilde{d}_k|^2\} - \mathrm{E}\{|\tilde{d}_k|\}^2}, \tag{10}$$

which is basically the Squared Signal-to-Variance (SNV) estimator given in [6],

$$\frac{D}{I} = \frac{(\frac{1}{K}\ {}^{K-1}_{k=0} |\tilde{d}_k|)^2}{\frac{1}{K}\ {}^{K-1}_{k=0} |\tilde{d}_k|^2 - (\frac{1}{K}\ {}^{K-1}_{k=0} |\tilde{d}_k|)^2}, \tag{11}$$

with K the number of symbols in the moving averaging window. Note that this estimator

$$\mathrm{SIR}_{\text{blind}} = 10\log\left(\frac{D}{I}\right) \tag{12}$$

relies on the received data only so that it is reasonable to call it *blind*. A similar SIR estimator of blind type is given in [8]. Note that for clarity of presentation and limited space we will focus on the presented data-aided and blind SIR estimators only.

5. SIMULATION RESULTS

In this section, simulation results are presented. Firstly, we compare the performances of three receiver structures – Rake, ZF and MMSE. A 2×2 MIMO scenario is assumed and 16 QAM and QPSK are utilized to modulate the signals of each transmit antenna. The power delay profile of the channel between each transmit and receive antenna pair is chosen as a chip-spaced version of 3GPP case 1 channel with Rayleigh fading coefficients. The spreading factor equals 16 and $C = 1$, 4 and 10 codes are chosen for different multicode scenarios. The Space-Time (ST) Rake architecture consists of the conventional *Maximum Ratio Combining* (MRC) Rake receiver and a succeeding spatial demultiplexing module. Fig. 3 and Fig. 4 demonstrate the performance of different receiver structures in different scenarios. Here we assume perfect channel estimation and the equalizers operate at chip rate (3.84 Mchips/s). The two figures highlight that even without multicode technique (i.e. $C = 1$), low spreading gain and self interference result in a significant error floor for the Rake receiver. As expected, the ZF and MMSE receiver clearly outperform the ST Rake receiver.

On the other hand, for a multicode scenario, we find that with increasing number of codes, the edge of the MMSE receiver becomes negligible. Hence, *inter code interferences* are the dominant part among all interferences, especially when the number of codes is high, i.e. $C = 10$.

The numerical result of SIR estimation is shown in Fig. 5. Observe that the Rake receiver goes into saturation first, while the ZF and MMSE behave rather similar. Note also that the principal behaviour of SIR saturation is in well agreement with the error floor phenomenon shown in the previous two figures. For this reason the estimated SIR reflects the changes of the BERs adequately so that the selected decision metric seems to be useful in a PARC system for MCS selection.

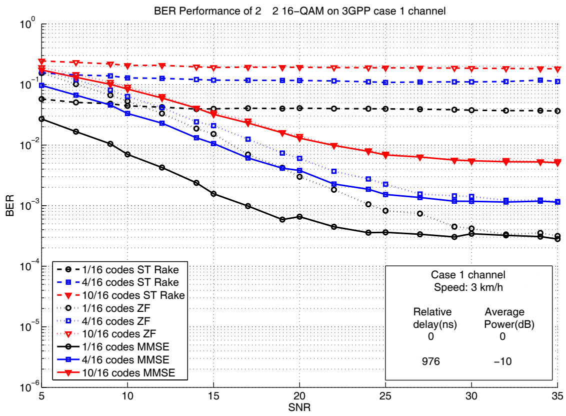

Figure 3: BER vs. SNR for ST-Rake, TD-ZF and TD-MMSE, 16 QAM used

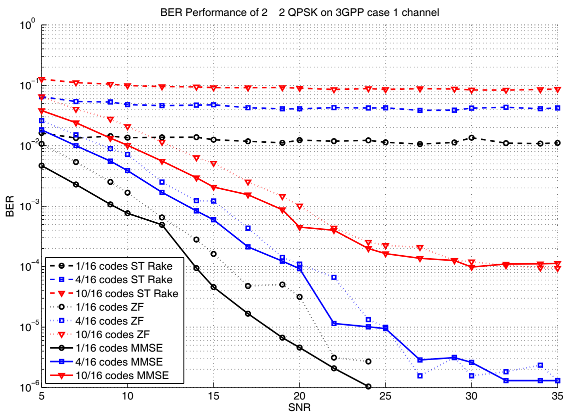

Figure 4: BER vs. SNR for ST-Rake, TD-ZF and TD-MMSE, QPSK used

6. CONCLUSION

In this paper, we addressed the PARC MIMO-HSDPA scheme and compared the performances of three conventional different receiver structures in different scenarios. In addition, we presented two different SNR estimation methods – blind method and data-aided method. Simulation show that even the blind SNR estimation already principally reflects the required behaviour for adequate MCS-selection. Future work aims for embedding the blind SIR estimation into a complete time-variant system in order to gain more meaningful results for practical deployment.

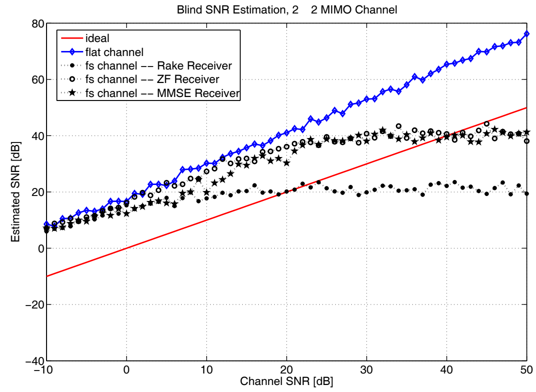

Figure 5: Estimated SNR vs. SNR for ST-Rake, TD-ZF and TD-MMSE, blind method used

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