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MEASUREMENT BASED EVALUATION OF LOW COMPLEXITY RECEIVERS FOR D-TXAA HSDPA

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ABSTRACT

In 2007, the 3GPP agreed to standardize D-TxAA (Dual Stream Transmit Antenna Array) as the MIMO extension for HSDPA (High Speed Downlink Packet Access). In this work, we examine receiver structures for D-TxAA.

At a high SINR, D-TxAA operates in the dual stream mode, i.e. two independent data streams are transmitted to one user simultaneously. In this mode, our receiver uses equalization and optional interference cancelation. At a low SINR, D-TxAA operates in single stream mode, in which several users can be served simultaneously, and hence an interference aware equalizer is necessary to achieve high performance. Utilizing these receiver structures, we measure the average throughput of D-TxAA in outdoor-to-indoor scenarios in the inner city of Vienna.

1. INTRODUCTION

In the last years, MIMO systems gained more and more attention in standardization committees. Already back in 1999, a multiple transmit antenna option (requiring only one receive antenna) was specified by the 3GPP for UMTS. This so-called closed loop mode with transmit diversity (TxAA) [1] uses strongly quantized precoding at the transmitter to increase the SINR at the user equipment. In 2007, the 3GPP standardized D-TxAA (Dual Stream Transmit Diversity) [2], the first true MIMO extension in UMTS. D-TxAA is downward compatible to TxAA and equals TxAA when the SINR at the user equipment is low. At larger SINRs, D-TxAA switches to dual stream mode and transmits two independently coded HSDPA (High Speed Downlink Packet Access) [3] data streams. An important difference between dual stream and single stream mode is that in the dual stream mode all 15 spreading codes of length 16 are assigned to a single user (the 16-th spreading code is required for transmitting pilot and control information). In single stream mode, the Node B can serve multiple users at the same time distinguished only by different spreading codes. Therefore, in single stream mode the user equipment receiver has to be aware of the interference generated by the data channels of other users.

The switching between single and dual stream mode, the encoding rate, and the modulation alphabet is determined by the CQI (Channel Quality Indicator) feedback. Depending on the channel conditions and the type of receiver implemented, the user equipment calculates an appropriate CQI value in order to achieve a block error ratio of 10 % [1].

In this work, we investigate the performance of two low-complexity receivers for D-TxAA by performing outdoor-to-indoor measurements in the inner city of Vienna. For these measurements we utilize the Vienna MIMO testbed [4], capable of transmitting and receiving arbitrary signals with 5 MHz bandwidth at 2.5 GHz. For comparison, we also measure the performance of the two SIMO systems transmitted from the first and second transmit antenna, respectively.

2. RECEIVER STRUCTURES

In this section we define the system model and describe the two types of D-TxAA receivers implemented. First, we discuss the straightforward MIMO MMSE (Minimum Mean Squared Error) equalizer (Figure 1 with S_I open), second we discuss an interference cancelation receiver (Figure 1 with S_I closed). For the SIMO transmission, we implement an LMMSE equalizer that utilizes both receive antenna signals to jointly detect the transmitted symbols.

2.1 System Model

In the following we shortly review the system model introduced in [5, 6]. Using this system model, we will define the MMSE equalizer for both, single stream ($K = 1$) and dual stream ($K = 2$) mode, and the interference cancelation receiver for the dual stream mode. We define the k-th spread and scrambled chip stream at time instant i as

$$\mathbf{s}_i^{(k)} = \left[s^{(k)}[i], \ldots, s^{(k)}[i - L_\mathrm{h} - L_\mathrm{f} + 2]\right]^\mathrm{T}, \qquad (1)$$

where L_h and L_f are the length of the channel impulse response and the equalizer length, respectively. Thus, the vector $\mathbf{s}_i^{(k)}$ contains the $(L_\mathrm{h} + L_\mathrm{f} - 1)$ recent chips. The k-th chip stream is then weighted by the complex precoding coefficients $w_1^{(k)}$ and $w_2^{(k)}$ at the first and the second transmit antenna, respectively. After that, the signals $\mathbf{p}_i^{(1)}$ and $\mathbf{p}_i^{(2)}$, containing the pilot channels, the synchronization channels, and the control channels, are added to the data chip streams.

The signals of the two transmit antennas are now transmitted over a frequency selective channel. The link between the n_t-th transmit and the n_r-th receive antenna is modeled by the $L_\mathrm{f} \times (L_\mathrm{h} + L_\mathrm{f} - 1)$ dimensional

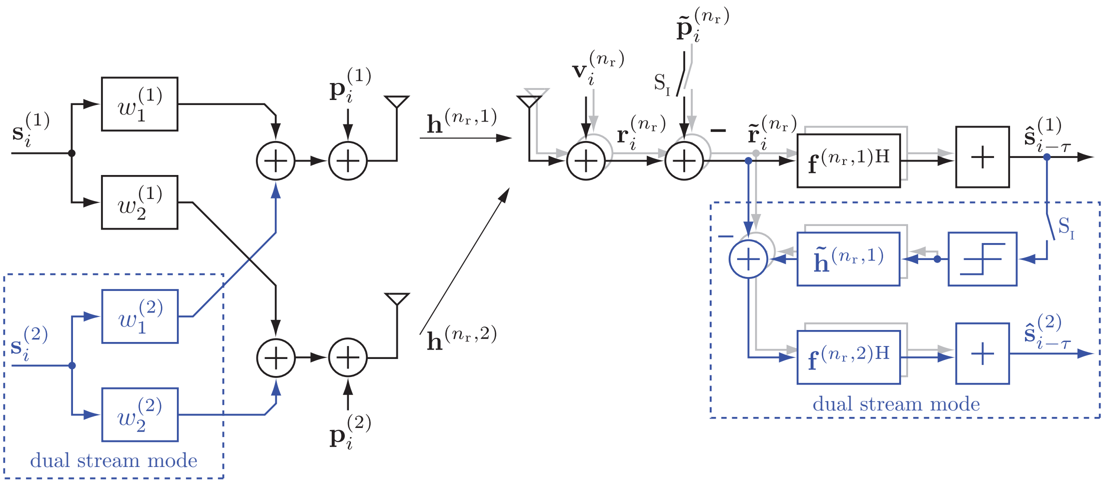

Figure 1: D-TxAA receiver in single stream and dual stream mode, respectively. Interference cancelation is performed when the switch $\mathrm{S_I}$ is closed. We assume without loss of generality that stream one is detected first.

band matrix

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

where the $h_i^{(n_\mathrm{r},n_\mathrm{t})}$ represent the channel impulse response of the n_t-th transmit antenna to the n_r-th receive antenna. The entire frequency selective MIMO channel is modeled by a block matrix $\mathbf{H}$ consisting of $N_\mathrm{R}\times N_\mathrm{T}$ ($N_\mathrm{T}=2$ for TxAA) band matrices defined in (2).

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

By stacking the received signal vectors of all N_R receive antennas

$$\mathbf{r}_i = \left[\mathbf{r}_i^{(1)\mathrm{T}}, \dots, \mathbf{r}_i^{(N_R)\mathrm{T}}\right]^\mathrm{T} \tag{4}$$

we can calculate the receive signal as

$$\mathbf{r}_i = \underbrace{\mathbf{H}\left(\mathbf{W}\otimes\mathbf{I}_{L_h+L_f-1}\right)}_{\mathbf{H}_\mathrm{w}} \begin{bmatrix} \mathbf{s}_i^{(1)} \\ \mathbf{s}_i^{(2)} \end{bmatrix} + \mathbf{H}\begin{bmatrix} \mathbf{p}_i^{(1)} \\ \mathbf{p}_i^{(2)} \end{bmatrix} + \mathbf{v}_i. \tag{5}$$

Here, $\otimes$ symbolizes the Kronecker product, the matrix $\mathbf{W}$ is the precoding matrix at the transmitter, and $\mathbf{v}_i$ is additive noise. If single stream transmission is performed, $\mathbf{W}$[1] is given by

$$\mathbf{W} = \begin{bmatrix} w_1^{(1)} & 0 \\ w_2^{(1)} & 0 \end{bmatrix} \tag{6}$$

and in the dual stream mode by

$$\mathbf{W} = \begin{bmatrix} w_1^{(1)} & w_1^{(2)} \\ w_2^{(1)} & w_2^{(2)} \end{bmatrix}. \tag{7}$$

2.2 MMSE Equalizer

By minimizing the quadratic error between the transmitted data chip stream and the equalized data chip stream, the MMSE equalizer is directly obtained from the system model in Equation (5). By assuming that the transmitted data sequences and the noise are white, we can calculate the equalizer filter coefficients [7, 8, 5, 6] as

$$\mathbf{f}^{(k)} = \left(\mathbf{H}_\mathrm{w}\mathbf{H}_\mathrm{w}{}^\mathrm{H} + \frac{\sigma_v^2}{\sigma_s^2}\mathbf{I}\right)^{-1}\mathbf{H}_\mathrm{w}\mathbf{e}_{\tau_k,2(L_\mathrm{h}+L_\mathrm{f}-1)}. \tag{8}$$

Here, $\mathbf{f}^{(k)} = \left[\mathbf{f}^{(1,k)\mathrm{T}}, \dots, \mathbf{f}^{(N_R,k)\mathrm{T}}\right]^\mathrm{T}$ defines the N_R equalization filters (see also Figure 1) for the k-th data chip stream. The vector $\mathbf{e}_{\tau_k,2(L_\mathrm{h}+L_\mathrm{f}-1)}$ is a zero vector of length $2(L_\mathrm{h}+L_\mathrm{f}-1)$ with a single one at position

$$\tau_k = \tau + (k-1)(L_\mathrm{h}+L_\mathrm{f}-1) \quad k=1\dots K. \tag{9}$$

After applying the equalizer $\mathbf{f}^{(k)}$ to the receive signal, we obtain an estimate $\mathbf{s}_{i-\tau}^{(k)}$ of the k-th data chip sequence. The variable τ specifies the delay of the equalized signal and has to fulfill $\tau \geq L_\mathrm{h}$ due to causality.

2.3 Interference Cancelation Receiver

The interference cancelation receiver is shown in Figure 1 (switch $\mathrm{S_I}$ closed). This receiver first cancels the deterministic interference

$$\tilde{\mathbf{p}}_i^{(n_\mathrm{r})} = \mathbf{H}\begin{bmatrix} \mathbf{p}_i^{(1)} \\ \mathbf{p}_i^{(2)} \end{bmatrix} \tag{10}$$

of the pilot and synchronization channels. After that, the data stream with larger post equalization SINR

[1] Note that this matrix $\mathbf{W}$ can be slightly modified to also consider the interference generated by the data streams of other users, thus making the MMSE equalizer to this system model an interference aware equalizer. See [5] for more details.

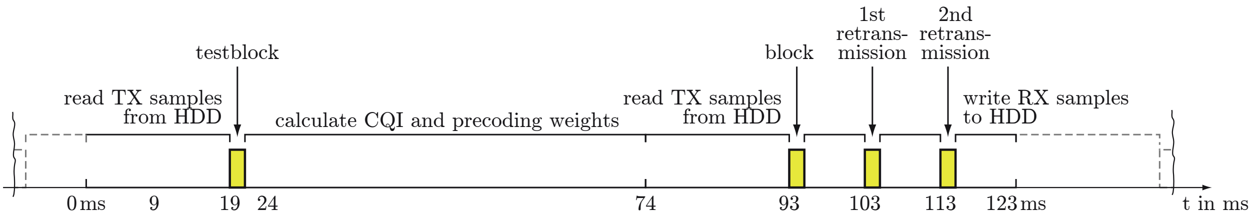

Figure 2: Testbed timing.

(see [6] for a detailed description on the calculation of this SINR) is equalized first[2]. The equalized chip stream is then descrambled, despread, and hard decided to the nearest symbol constellation point. This reconstructed symbol stream is then again spread, scrambled and convoluted with the "virtual" channel $\tilde{\mathbf{h}}^{(n_r,1)} = w_1^{(1)}\mathbf{h}^{(n_r,1)} + w_2^{(1)}\mathbf{h}^{(n_r,2)}$. The virtual channel $\tilde{\mathbf{h}}^{(n_r,1)}$ is given by the true channel coefficients and the precoding coefficients. The reconstructed interference can now be subtracted from the receive signal $\tilde{\mathbf{r}}_i^{(n_r)}$ and a second equalizer yields the second data chip stream.

3. MEASUREMENT SETUP

For the measurements we utilize the Vienna MIMO testbed, described in detail in [4] and [9]. As transmit antennas we used basestation antennas mounted 16 m above rooftop. The receiver was placed at two different positions indoors in the same building.

The feedback required by HSDPA for adapting the transmit signal to the actual channel condition is implemented as follows.

- At first, one HSDPA frame (the testblock) is transmitted, see Figure 2, at time index 19 ms.
- This HSDPA frame is then evaluated in a "mini-receiver" which performs synchronization, receive filtering, channel and noise estimation, and an estimation of the post equalization SINR. The post equalization SINR is calculated for all four possible precoding vectors specified in D-TxAA. The SINR values are then mapped to CQI values using an optimized CQI mapping table[3]. This method leads to four CQI values (and corresponding transport block sizes)—one for each possible precoding vector. Now, we select that precoding vector that corresponds to the largest transport block size and thus the largest potential throughput. The calculation of the CQI value and the precoding weight takes about 50 ms.
- The CQI value and the precoding weights are signalled back to the transmitter of the testbed using a LAN connection.
- The transmitter then transmits an HSDPA subframe that was pre-generated and stored on a harddisk. This subframe has been encoded with the CQI value, and weighted by the precoding coefficients, signalled by the receiver. Since in HSDPA, HARQ (Hybrid Automatic Retransmission Request) is immanent for the system performance, we also transmit the first and second retransmission with a delay of about 10 ms. These three subframes are stored on a harddisk at the receiver for later, off-line evaluation.

This procedure was carried out for a SIMO transmission at the first transmit antenna, a SIMO transmission at the second transmit antenna, and the MIMO D-TxAA transmission[4]. These three transmissions were repeated at 16 different transmit attenuator values leading to different average receive SNRs. Averaging over small scale fading was achieved by repeating everything at 3840 receive antenna positions, equally spaced distributed over an area of $4\times4\,\lambda$.

4. MEASUREMENT RESULTS

The results presented in the following are evaluated in terms of physical layer throughput. The "instantaneous" throughput at one transmit attenuator value and one receive antenna position is calculated as

$$D = \frac{\mathrm{TBS}}{k \cdot \mathrm{TTI}}. \tag{11}$$

Here, TBS denotes the Transport Block Size, i.e. the number of bits transmitted in one HSDPA subframe. The TTI (Transmission Time Interval) here is chosen equal to 2 ms. The value k corresponds to the number of transmissions required for correctly receiving this subframe. If the subframe is not received correctly after two retransmissions (i.e. $k = 3$), the instantaneous throughput is set to zero. The average throughput in one scenario is now calculated by averaging over all 3840 receive antenna positions.

4.1 SIMO System

The measured data throughput in two outdoor-to-indoor scenarios is shown in Figures 3 and 5. In both scenarios, we measure a slightly larger throughput of the 1×2 system when transmitting over the first transmit antenna. We also see that the throughput curves of the SIMO transmissions saturate at about 11 Mbit/s although a Category 14 user equipment should be able to achieve about 19 Mbit/s. The reason for this is that the three largest CQI values (64-QAM with a coding rate >0.7) are never decoded correctly within a single transmission, so these CQI values always require at least one

[2] In the following we assume without loss of generality that the first data chip stream has the larger SINR.

[3] The CQI mapping table was optimized to achieve about 10% block error ratio, as specified in the standard [10, Section 6A.2].

[4] Note that the precoding weight calculation obviously is only performed for the MIMO transmission, in the SIMO transmission it is sufficient to determine the post equalization SINR and map it to a CQI value.

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Table 1: HSDPA System Parameters.

Parameter	**Value**
CPICH $E_\mathrm{c}/I_\mathrm{or}$	$-10\,\mathrm{dB}$
SCH/PCCPCH $E_\mathrm{c}/I_\mathrm{or}$	$-10\,\mathrm{dB}$
User equipment capability (MIMO)	10M
User equipment capability (SIMO)	14
Channel realizations	3840
Maximum transmit power	40 dBm
Center frequency	2.5 GHz

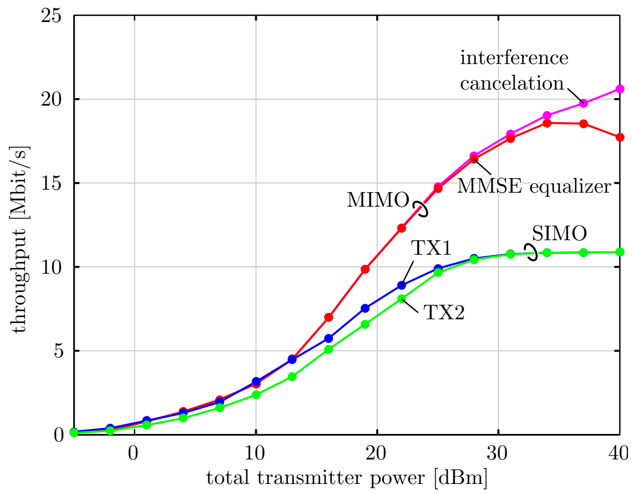

Figure 3: Average throughput of 1×2 and 2×2 transmissions at the first receiver position.

retransmission. Since this would lead to a BLER larger than 50 % we had to exclude these three CQI values from the CQI mapping table.

The above mentioned incorrect decoding within a single transmission is caused by the lack of cancelation of the pilot and synchronization channels, i.e. a too simple receiver. Especially the synchronization channel has a large impact on the data channels since it is transmitted without spreading and scrambling. Interference cancelation of these deterministic channels can substantially increase the performance at large SINRs [11].

The BLER (Block Error Ratio) of the SIMO system is shown in Figures 4 and 6. A BLER between 2 % and 10 % is achieved over a large dynamic range of about 30 dB. This is in good agreement to the standard which demands a BLER smaller than 10 % [10, Section 6A.2].

4.2 MIMO System

At low transmit power, the MIMO system (with both receivers) achieves the same throughput (see Figures 3 and 5) as the SIMO transmission over the better transmit antenna. At large transmit power, the MMSE equalizer shows a strange behavior—the throughput drops with increasing transmit power. This is due to a saturation of the post equalization SINR caused by the interference of the deterministic pilot and synchronization channels, and the interference of the second data stream.

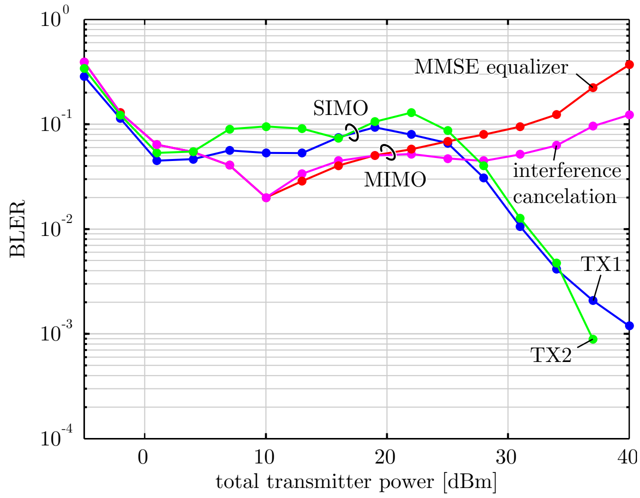

Figure 4: Average BLER of 1×2 and 2×2 transmissions at the first receiver position.

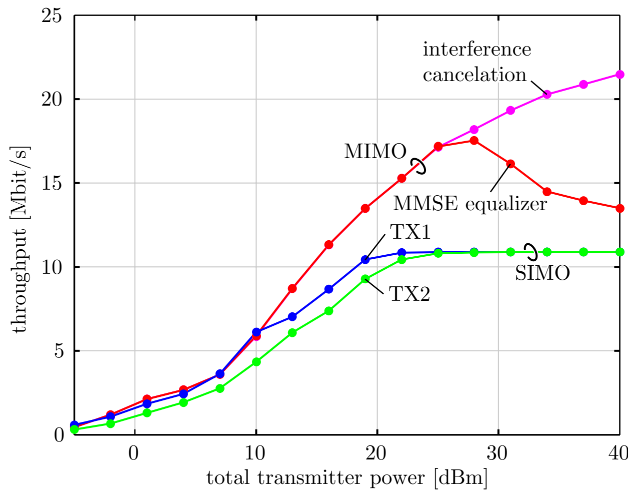

Figure 5: Average throughput of 1×2 and 2×2 transmissions at the second receiver position.

The saturation of the post equalization SINR leads to a subsequent selection of too large CQI values that cannot be decoded correctly without retransmission. The interference cancelation receiver that cancels the deterministic channels and the first data stream before detecting the second stream does not have this problem and is able to achieve larger SINR values and thus larger throughput. In order to be able to decode also the largest CQI values, a D-TxAA receiver therefore necessarily has to employ interference cancelation .

The BLER (Block Error Ratio) of the MIMO system is shown in Figures 4 and 6. Also here, a BLER between 2 and 10 % is achieved over a large dynamic range of about 30 dB. The straightforward MMSE equalizer shows at large transmit power increased block error ratios causing the throughput loss noticed above.

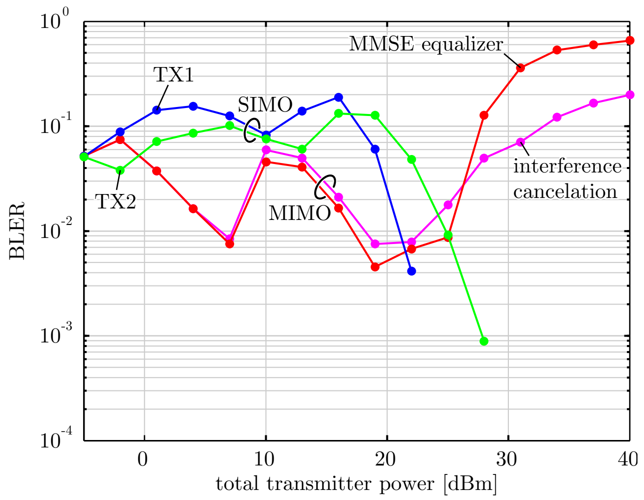

Figure 6: Average BLER of 1×2 and 2×2 transmissions at the second receiver position.

5. CONCLUSIONS

In this work, we presented throughput and BLER measurement results of outdoor-to-indoor HSDPA transmissions. The SIMO measurements showed that a Category 14 user equipment cannot decode the three largest CQI values without receiving a retransmission. A more complex receiver that uses interference cancelation of the deterministic (i.e. pilot and synchronization) channels is necessary to decode the largest CQI values within a single transmission. The MIMO measurements showed similar results, interference cancelation is absolutely necessary to achieve high data throughput at large CQI values.

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