Generalized Mutual Information Based 
LTE-Advanced Uplink MIMO Receiver Analysis

Eckhard Ohlmer and Gerhard Fettweis 
Vodafone Chair Mobile Communications Systems, Technische Universität Dresden, Germany
Email: {eckhard.ohlmer, gerhard.fettweis}@ifn.et.tu-dresden.de

Abstract—The low peak-to-average power ratio of SC-FDMA, employed in the 3GPP-LTE-Advanced uplink, comes at the cost of tremendously increased receiver complexity as compared to OFDM. In this work, we investigate the achievable rate of four low-complexity, mismatched receiver designs based on a recently developed generalized mutual information framework. The receiver performance is analyzed depending on the channel length, subcarrier mapping and spatial correlation. Results may guide the system designer to choose a receiver, matched to the expected operation conditions. Numerical results stress that receivers which treat a MIMO channel with memory as if it was memoryless may be well suited for narrow bandwidths while additional interference cancelation components may be required when using large bandwidths.

I. INTRODUCTION

Single-carrier frequency division multiple access (SC-FDMA) has been selected in the 3GPP-LTE-Advanced uplink [1] rather than orthogonal frequency division multiple access (OFDMA) for its better peak-to-average power ratio. Single- and multiuser multiple-input multiple-output (MIMO) transmission on top enable high-data rate communication. When analyzing MIMO SC-FDMA transmission it turns out to be equivalent to a MIMO inter-symbol interference (ISI) channel, i.e., a channel with memory. For such channels, the complexity of the optimal maximum likelihood sequence detector, implemented in terms of a Viterbi decoder, grows exponentially in the number of antennas and the number of channel taps. Implementing an optimal receiver is, therefore, a challenging task.

Suboptimal lower-complexity receivers have been devised based on frequency domain equalization (FDE) to mitigate spatial- and inter-symbol interference (ISI), followed by time domain detection. As a consequence of FDE, the receiver noise will be spatially and temporally correlated. In particular, the temporal correlation is not accounted for by time domain detection, i.e., a channel with memory is treated as if it was memoryless. Examples of such approximations comprise linear receivers [2] or sphere search receivers [3]. Of course, the receiver performance will degrade if the memoryless channel assumption is violated which would be the case for increasing delay spreads. Other receiver types beyond these two [2], [3] are possible as well. For the wireless system designer it is important to know in advance what to expect from different receiver designs under certain operation conditions, in order to draw a decision which receiver scheme shall be actually implemented. To ease such a decision we developed a framework to compute achievable rates of low-complexity, mismatched receiver designs [4]. It allows to compute the achievable rate in terms of generalized mutual information (GMI) [5] corresponding to a particular simplified receiver detection metric. This framework is especially meaningful if adaptive coding and modulation can be employed in order to avoid transmission errors which, indeed, is the case in current cellular standards such as 3GPP-LTE-Advanced.

In this paper we apply this framework in order to assess the impact of channel length (i.e. channel memory), contiguous or distributed subcarrier mapping, and spatial correlation on the performance of four different MIMO SC-FDMA receiver designs. These designs are derived from the optimal receiver by introducing increasingly simplifying assumptions on the receiver detection metric.

The remainder of this work is structured as follows: The SC-FDMA system setup is introduced and analyzed in Section II. Different low-complexity receiver designs are derived from the optimal receiver in Section III. Their performance comparison follows in Section IV. Section V concludes the paper.

Notation: \( \mathcal{N}_c(\mu, \Phi) \) denotes the complex normal distribution (mean \( \mu \), covariance \( \Phi \)). Normal (\( a \)) and boldface (\( \mathbf{a} \)) letters denote scalars and vectors or matrices respectively. \( p(\cdot) \), \( P(\cdot) \), \( \mathbb{E}[\cdot] \), and \( I(\cdot) \) denote a probability, a probability density function (pdf), the expectation with regard to random vector \( a \), and the mutual information, \( I_{T_1} \). \( (\cdot)^T \) and \( (\cdot)^H \) denote the identity matrix of size \( T \times T \), the matrix transpose and the hermitian operator, respectively. \( \mathbf{F}_M \in \mathbb{C}^{M \times M} \) denotes a Fourier matrix with elements \( [\mathbf{F}_M]_{m1,m2} = \exp(-j2\pi m_1 m_2/M) \).

II. TRANSMISSION SETUP

A. System Model

We investigate a MIMO SC-FDMA system with \( T \) transmit antennas and \( R \) receive antennas, sketched in Fig. 1. The transmit antennas may either belong to multiple users or to a single user. The important property is the transmission of a single codeword per antenna as shown in Fig. 1. The rate of each of the \( T \) codewords can be adapted individually based on rate feedback which is assumed to match the achievable rate of a particular receiver design perfectly. At transmit antenna \( t \), the encoded symbol vector \( \mathbf{x}_t = [x_{t1}, \ldots, x_{tM}]^T \) constitutes a time-domain SC-FDMA symbol, comprising \( M \) complex valued symbols with zero mean and variance \( \sigma^2_t/T \). Alternatively,
Fig. 1: MIMO SC-FDMA transmission setup with MMSE-FDE and per-antenna rate control based on feedback

we define the vector symbol $x[m] = [x_1[m], \ldots, x_T[m]]^T$ to be transmitted from all $T$ antennas at time instant $m$.

At antenna $t$, the SC-FDMA symbol $\tilde{x}_t$ is transformed into frequency domain by an M-point discrete Fourier transform (M-DFT), followed by a mapping to $M$ out of $N \geq M$ subcarriers of an OFDM symbol (SC-Map). The mapping can either be contiguous or distributed, using every $Q$-th subcarrier. The time domain signal is computed using an N-point inverse DFT (N-IDFT), followed by prepending a cyclic prefix of length $N_F$ and transmission over the wireless multi-path MIMO channel with impulse response (CIR) $h_{r,t} \in \mathbb{C}^{N \times 1}$ between transmit-receive antenna pair $t,r$. Note that the cyclic prefix turns the linear convolution with the CIR into a circular convolution. Each CIR is assumed to be non-amplifying, i.e., $\mathbb{E}[(\sum_{m=0}^{N-1} |h_{r,t}[n]|^2)^2] = 1$. Only the first $L = N_F + 1$ complex Gaussian distributed channel coefficients are assumed to be non-zero. Spatial correlation, caused, e.g., by small antenna spacings, is modeled using the Kronecker correlation model [6] $\mathbb{E}[h_{r_1,t_1}[n] h_{r_2,t_2}[n]] = \rho_{rx}^{(r_1 - r_2)^2} \rho_{tx}^{(t_1 - t_2)^2}$. We choose equal transmitter and receiver correlation $\rho_{tx} = \rho_{rx} = \rho$.

At the receiver, white Gaussian noise $\tilde{y}_r \in \mathbb{C}^{N \times 1}$ with variance $\sigma_r^2$ is added at antenna $r$. The received signal $y$ at all antennas $\tilde{y}_t \in \mathbb{C}^{N \times 1}$ and $h_{r,t} \in \mathbb{C}^{N \times N}$ denote the received signal at antenna $r$ and the channel impulse response $h_{r,t}$ in each of its rows.

The received signal is transformed into frequency domain by an N-DFT, followed by selecting the $M$ used subcarriers. A minimum-mean-square-error (MMSE) filter $W[m] \in \mathbb{C}^{T \times R}$ is applied at each of the $M$ subcarriers in order to mitigate spatial interference and frequency selectivity, i.e., ISI. The $m$-th filter reads

$$W[m] = H^H[m] \left( H[m] H^H[m] + \frac{\sigma_r^2}{\sigma_x^2} M N_T I_R \right)^{-1}.$$  \(2\)

$H[m] \in \mathbb{C}^{R \times T}$ denotes the frequency domain channel matrix. Its elements are derived from the CIR as follows $[H_{r,t}[1], \ldots, H_{r,t}[N]]^T = F_N \tilde{h}_{r,t}$. The $t$-th filter output at all $M$ used subcarriers is transformed into the time domain by an M-IDFT yielding the time domain signal $\tilde{y}_t \in \mathbb{C}^{M \times 1}$. Similar to (1), the end-to-end transmission equation including FDE reads

$$\begin{bmatrix} \tilde{y}_1 \\ \vdots \\ \tilde{y}_T \end{bmatrix} = \begin{bmatrix} \tilde{h}_{1,1} & \cdots & \tilde{h}_{1,T} \\ \vdots & \ddots & \vdots \\ \tilde{h}_{T,1} & \cdots & \tilde{h}_{T,T} \end{bmatrix} \begin{bmatrix} \tilde{x}_1 \\ \vdots \\ \tilde{x}_T \end{bmatrix} + \begin{bmatrix} \tilde{v}_1 \\ \vdots \\ \tilde{v}_T \end{bmatrix}.$$  \(3\)

B. System Properties

The properties of the model (3) are determined by the effective channel $\tilde{h}$ and the effective noise $\tilde{v}$. $\tilde{h}_t, \tilde{v}_t \in \mathbb{C}^{M \times M}$ is a circular convolution matrix which is composed of the time domain channel impulse response $\tilde{h}_{t_1,t_2}$. Its elements are computed from the M-IDFT of the equalized frequency domain channel $\tilde{h}[m] = W[m] H[m]$ as follows $[\tilde{h}_{t_1,t_2}[1], \ldots, \tilde{h}_{t_1,t_2}[M]]^T = \frac{1}{M} F_M H_{t_1,t_2}[1], \ldots, H_{t_1,t_2}[M]^T$.

The time domain receiver noise $\tilde{v}_t$, including noise, is zero-mean complex Gaussian distributed with covariance matrix

$$\Phi_{\tilde{v}} = \mathbb{E} [\tilde{v}_t \tilde{v}_t^H] = \frac{\sigma_v^2}{N} \begin{bmatrix} \Phi_{\tilde{v}_1,\tilde{v}_1} & \cdots & \Phi_{\tilde{v}_1,\tilde{v}_T} \\ \vdots & \ddots & \vdots \\ \Phi_{\tilde{v}_T,\tilde{v}_1} & \cdots & \Phi_{\tilde{v}_T,\tilde{v}_T} \end{bmatrix}.$$  \(4\)

$\Phi_{\tilde{v}}$ is composed of circulant matrices $\Phi_{\tilde{v}_r,\tilde{v}_t} \in \mathbb{C}^{M \times M}$ which can be derived as follows. Let $W[m] = W[m] W^H[m]$. Collect the element $w_{t_1,t_2}$ of $W[m]$ at all $M$ subcarriers in a column vector. The circulant matrix $\Phi_{\tilde{v}_r,\tilde{v}_t}$ is composed of the M-IDFT of this vector, shown below.

$$[\tilde{w}_{t_1,t_2}[1], \ldots, \tilde{w}_{t_1,t_2}[M]]^T = F_M^H W_{t_1,t_2}[M], \ldots, W_{t_1,t_2}[1]^T.$$  \(5\)

A numerical example illustrating channel and noise properties, is shown in Fig. 2. Clearly, the ISI of the channel after FDE is significantly reduced. However, many small residual ISI contributions remain. From the noise covariance plot it can be seen that the noise is spatially and temporally correlated after FDE. That is, FDE reduces ISI but at the same time moves the channel memory into the receiver noise.

III. RECEIVER TYPES

A. Optimal Receiver

From the (3) the transition probability density function (pdf) is a Gaussian distribution $p(\tilde{y} | \tilde{x}_1, \ldots, \tilde{x}_T) = \mathcal{N}_C(\tilde{h} x, \Phi_{\tilde{v}})$. The optimal receiver jointly selects the $T$ input sequences which maximize this pdf according to

$$\{\tilde{x}_1, \ldots, \tilde{x}_T\} = \arg \max_{\tilde{x}_1, \ldots, \tilde{x}_T} \left\{ p(\tilde{y} | \tilde{x}_1, \ldots, \tilde{x}_T) \right\}.$$  \(6\)

Only the core OFDM symbol of length $N$ without cyclic prefix.
Maximization is carried out over the set of valid code words and all code words are assumed to be equally likely. The sum rate per channel use $I'$ achievable by the optimal detector is given by the mutual information\(^2\) [7]

$$I' = \frac{1}{M} I \left( \tilde{y}; x_1, \ldots, x_T \right).$$

Implementing (4) in terms of the Viterbi algorithm yields a complexity, growing exponentially in the channel length. To reduce complexity, we focus on simplified receiver designs building on detection metrics which assume the channel to be memoryless. (5) will serve as an upper-bound for these receivers.

\(^2\) In theory, the mutual information (5) is an achievable rate if a codeword extends over infinitely many SC-FDMA symbols which are mutually independent due to the cyclic prefix [4].

Fig. 2: CIR realization without ($\tilde{h}$, left) and with ($\tilde{h}_{1}, \tilde{h}_{2}$ middle) MMSE-FDE. Magnitude of the noise covariance $\Phi_{\tilde{v}\tilde{v}}$ with MMSE FDE (right). ($N = 128$, $M = 24$, $2 \times 2$, $SNR = 15 dB$, uniform power delay profile with $L = 9$ taps, $\rho_s = 0.9$).

Fig. 3: Simplified receiver designs based on time-domain detection. IC denotes interference cancelation.

B. Lower-Complexity Receivers Based on the Memoryless Channel Assumption

Instead of the optimal detection metric used in (4) we will subsequently focus on a suboptimal metric $q(\tilde{y}, x)$ which is implemented by the different receivers shown in Fig. 3 and will be detailed in the later part of this section. Based on [5], [7], [8] we derived the achievable rate, corresponding to $q(\tilde{y}, x)$ in terms of generalized mutual information (GMI) as follows\(^3\) [4]

$$I_G(s) = \mathbb{E}_{\tilde{x}, x} \left[ \log \left( \frac{q(\tilde{y}, x)^s}{\sum_{x'} \prod_{t=1}^{T} \Pr(x_t) q(\tilde{y}, x')^s} \right) \right].$$

\(^3\) The parameter $s$ is due to bounding the probability of block error. $s = 1$ maximizes (6) for matched receivers [7] but also for the suboptimal receivers in this work [4].
In the following, we focus on detection metrics that decompose into a product over the channel uses \( q(\hat{y}, x) = \prod_{m=1}^{M} q(\hat{y}[m], x[m]) \). That is, it is assumed that the \( m \)-th output symbol vector \( \hat{y}[m] \) depends only on the \( m \)-th input symbol vector \( x[m] \). A receiver designed based on this assumption will not be able to exploit the channel memory, captured mostly in the noise \( \xi \) after equalization. However, its complexity becomes independent of the channel length.

The subsequently presented receivers are based on detection metrics, relating to \( p(\hat{y}[m] | x[m]) \). Therefore, collect the \( m \)-th tap of all effective CIRs \( \hat{h}_{t_1t_2} \) in a matrix \( \hat{h}[m] \in \mathbb{C}^{T \times T} \). The input-output relation for the \( m \)-th symbol vector then reads

\[
\hat{y}[m] = \hat{h}[1]|x[m]| + \sum_{\tau = 0}^{M-1} \hat{h}((\tau + m - 1) \mod M) + 1|x[\tau + 1] + \nu[m]
\]

(7)

In general, \( p(\hat{y}[m] | x[m]) \) derives from computing the marginal of \( p(\hat{y}[m] | x) \) with regard to all, except the \( m \)-th input vector. However, by noting that \( z[m] \) contains the superposition of many, typically small discrete noise components and is hence approximately Gaussian distributed, we can state the respective transition pdf as \( p(\hat{y}[m] | x[m]) \approx \mathcal{N}(\hat{h}[1]|x[m]|, \Phi_{zz}) \). The approximation is accurate if Gaussian distributed input signals are used. The noise covariance matrix is given by \( \Phi_{zz} = \sigma_z^2 I \sum_{m=2}^{M} |h[m]|^2 H + \sigma_z^2 / N \Phi_{1} \). The matrix \( \Phi_{1} \in \mathbb{C}^{T \times T} \) contains the elements \( \nu_{t_1t_2} \) at position \( t_1, t_2 \).

1) **Joint Detection**: The receiver jointly searches for the \( T \) input sequences which maximize the metric \( q(\hat{y}, x) = \prod_{m=1}^{M} p(\hat{y}[m] | x[m]) \). Solving (7) yields the achievable sum rate

\[
I' = I(\hat{y}; x) = \sum_{i=1}^{T} I(\hat{y}; x_1, x_2, \ldots, x_{t-1}).
\]

(8)

The RHS of (8) is due to the mutual information chain rule and suggests an implementation in terms of successive cancellation detection as plotted in Fig. 3a). This becomes feasible if rate feedback is available and the rates per codeword are adjusted according to the mutual information chain rule. (8) can be solved in closed form in case of Gaussian input signals [4] but only numerically in case of discrete input alphabets, e.g. M-QAM (see e.g. [9]).

2) **Parallel Detection**: Successive cancellation detection entails a detection delay and is prone to error propagation in case of imperfect rate adaptation. Detecting the \( T \) input sequences in parallel as shown in Fig. 3b) can help mitigating this issue. When detecting the \( t \)-th input sequence, the parallel detector treats all other input sequences as noise. However, the detector accounts for the discrete structure of this noise if, e.g., the input symbols are derived from an M-QAM signal alphabet. More specifically, to implement parallel detection the receiver assumes the channel input symbols to be independent, given knowledge of the received signal. This is equivalent to jointly searching for the \( T \) input sequences which maximize the metric \( \prod_{m=1}^{M} p(\hat{y}[m] | x[m]) \) which breaks down to \( T \) independent detectors. That is, the \( t \)-th detector maximizes the metric \( \prod_{m=1}^{M} p(\hat{y}[m] | x[m]) \) for the input sequence \( x_t \).

By solving (6) for this detection metric, the achievable sum rate turns out to be given by the mutual information

\[
I' = \sum_{i=1}^{T} I(\hat{y}; x_t).
\]

(9)

The rate of the \( t \)-th input codeword needs to be adjusted according to \( I(\hat{y}; x_t) \)

3) **Linear Detection**: The linear receiver, shown in Fig. 3c) assumes the \( T \) channel input - equalizer output channels \( x_1[m] \rightarrow \hat{y}_1[m] \) to be independent. That is, the metric \( q(\hat{y}, x) = \prod_{m=1}^{M} \prod_{t=1}^{T} p(\hat{y}[m] | x_t[m]) \) is used and the joint detection of \( T \) inputs signals decomposes into detecting them in parallel. The \( t \)-th detector maximizes \( \prod_{m=1}^{M} p(\hat{y}[m] | x_t[m]) \) with respect to the \( t \)-th input sequence \( x_t \).

By solving (6) for this detection metric, the achievable sum rate is

\[
I' = \sum_{i=1}^{T} I(\hat{y}_t; x_t).
\]

(10)

The rate of the \( t \)-th input codeword needs to be adjusted according to the mutual information \( I(\hat{y}_t; x_t) \). Note that for Gaussian distributed input signals, the MMSE equalizer output \( \hat{y}[m] \) is a sufficient statistic with regard to \( x[m] \). Therefore, the rates of the parallel receiver (9) and the linear receiver (10) will coincide in this special case.

**Example**: A first receiver comparison under severe ISI and severe spatial correlation is shown in Fig. 4. The joint receiver (Section III-B1) with Gaussian input signals experiences a loss compared to the optimal receiver as a consequence of the memoryless channel assumption. Note that the mutual information of a memoryless Gaussian vector channel in (8) is maximized by Gaussian input signals. Statically using 64-QAM at both channel inputs almost approaches this mutual information. Also, there is no gain of adaptively choosing another combination of M-QAM modulation schemes.
parallel receiver with Gaussian input signals experiences an additional loss due to assuming the channel input symbols to be independent given the channel output which would only be valid for orthogonal MIMO channel matrices. For this receiver, however, the rate can be significantly increased by choosing a proper combination of input signal alphabets (adaptive modulation) such that it almost approaches the joint receiver, which serves as an upper bound for the parallel receiver performance. For the linear receiver whose performance is determined by the post equalization signal-to-interference-and-noise-ratio, the parallel receiver with Gaussian input signals is an upper bound which is almost achieved by 64-QAM input signals. Adaptive modulation cannot further improve the rate. Knowing that the linear receiver with M-QAM input signals will almost approach the parallel receiver with Gaussian inputs and furthermore knowing that the parallel receiver with M-QAM inputs will operate between the joint receiver and the parallel receiver with Gaussian inputs, we subsequently only analyze Gaussian input signals.

4) Frequency Domain Interference Cancelation: In order to overcome the loss caused by the memoryless channel assumption recall that the spatio-temporal correlation of the noise \( \tilde{\mathbf{x}} \) directly depends on the FDE filter matrices \( \mathbf{W}[m] \).

To improve the post equalization noise properties, the interference of a decoded codeword could be perfectly removed in frequency domain, followed by computing updated FDE filter matrices, equalizing the remaining (reduced) channel matrix, followed by detecting the next codeword in time domain whose interference could then also be removed in frequency domain and so forth. This scheme is plotted in Fig. 3d).

Denote by \( \mathbf{H}^t[m] \) the remaining channel columns \( \mathbf{H}[m]_{:,t:T} \) in frequency domain after canceling the contributions of the channel inputs \( t, \ldots, T - 1 \). The FDE filter matrix \( \mathbf{W}^t[m] \) is computed according to (2), replacing \( \mathbf{H}[m] \) by \( \mathbf{H}^t[m] \). The time domain signal \( \tilde{\mathbf{y}}^t \), containing only the contributions of the channel inputs \( t, \ldots, T \), is obtained from an M-IDFT of all equalizer outputs. Similar to (7), the input-output relation for the \( m \)-th vector symbol reads

\[
\tilde{\mathbf{y}}^t[m] = \mathbf{H}^t[1]|\mathbf{x}^t[m] + \mathbf{z}^t[m],
\]

where \( \mathbf{x}^t[m] = [x_t, \ldots, x_T]^T \). The elements of \( \mathbf{H}^t[m] \) derive from the M-IDFT of the equalized channel transfer function as described in Section II-B. The noise covariance matrix of the respective transition pdf \( p(\mathbf{y}^t[m]|\mathbf{x}^t[m]) \approx \mathcal{N}_c(\mathbf{H}^t[1]|\mathbf{x}^t[m], \mathbf{\Phi}^t_{\tilde{x}^t}) \), which is accurate in case of Gaussian input signals, derives as discussed in III-B. In the \( t \)-th detection step, the detector maximizes \( \prod_{m=1}^M p(\tilde{\mathbf{y}}^t|m|\mathbf{x}^t[m]) \) with respect to the input sequence \( \mathbf{x}^t \), similar to the parallel receiver. From (6) the achievable sum rate is

\[
I' = \sum_{t=1}^T I(\tilde{\mathbf{y}}^t; \mathbf{x}^t) \tag{12}
\]

The rate at the \( t \)-th channel input has to be adjusted according to the mutual information \( I(\tilde{\mathbf{y}}^t; \mathbf{x}^t) \).

IV. Receiver Comparison

The performance of the optimal receiver and the four suboptimal receivers is assessed in this section. As an example, we investigate a \( 4 \times 4 \) MIMO system comprising the smallest 3GPP-LTE bandwidth of \( N = 128 \) subcarriers. The channel is randomly drawn according to a uniform power delay profile with length \( L = 9 \) which equals the cyclic prefix length and may be viewed as a worst case from the channel memory perspective. The length \( M \) of an SC-FDMA symbol is chosen as a multiple of the 3GPP-LTE physical resource block (PRB) size which contains 12 subcarriers. Note that for \( L = 9 \), four PRB cover 5 coherence bandwidth, i.e., the channel is strongly frequency selective. Throughout the following, Gaussian input signals are analyzed.

A. Impact of the Occupied Bandwidth

Fig. 5 shows the mutual information for transmitting over 1 or 10 PRB, respectively. The average performance of the optimal receiver does not depend on the occupied bandwidth. For small bandwidth, the rate of the joint receiver is close to optimal. From the large gap between the parallel and the joint receiver, the possible gain of a parallel receiver with adaptive M-QAM modulation, as compared to a linear receiver, becomes obvious (see the example in Section III-B3). At large bandwidth, the joint and parallel receiver performance severely
...degrades and both receivers almost coincide. This result is due to the dominating effect of channel memory, rather than spatial interference. Joint, parallel, and linear receivers will achieve almost the same performance in that case.

The frequency domain interference cancelation receiver performance, however, is almost not influenced by increasing the bandwidth. To explain this result, we analyze the properties of the time-domain noise covariance matrix $\Phi_{xx}$ (see Fig. 2, right) before detection of the first channel input sequence and before detection of the $T$-th channel input sequence, i.e., after canceling the interference of the first three input sequences. More specifically, we investigate the ratio of the average sum of diagonal elements (magnitude) of each submatrix $\Phi_{xx}$, divided by the average sum of off-diagonal elements (magnitude). That is, the ratio of spatial noise correlation and spatio-temporal noise correlation. The smaller this ratio is, the higher is the impact of channel memory. The results in Fig. 6 illustrate that the higher the bandwidth is, the larger becomes the temporal noise correlation in the first iteration. However, this effect is mitigated (dashed curve) after canceling three channel input sequences. From a different perspective, the FDE-MMSE equalizer reduces to a maximum ratio combiner for the last remaining channel input, i.e., the scalar term $W_{[m]}$ comprises the sum of squared channel magnitudes in that case. This sum becomes less frequency selective, the more receive antennas are comprised in the system which, in turn, reduces the temporal noise correlation.

**B. Impact of Subcarrier Mapping**

Results for non-contiguous subcarrier mapping, shown in Fig. 7, expose a similar behavior as compared to increasing the bandwidth. The performance of the joint receiver and the parallel receiver almost coincides at higher $Q$, while the frequency domain interference cancelation receiver performance remains almost constant.

**C. Impact of Spatial Correlation**

Fig. 8 illustrates the impact of spatial correlation. Most obvious, spatial correlation degrades the performance of all receivers. The parallel receiver shows the largest degradations.

On the one hand, this result highlights the susceptibility of linear receivers to spatial correlation, but, on the other hand, it also highlights the possible gains of a parallel receiver with adaptive modulation (M-QAM) whose performance may get close to the joint receiver. The frequency domain interference cancelation receiver is robust with regard to spatial correlation.

**V. CONCLUSIONS AND FUTURE WORK**

In this work, we derived four different low-complexity MIMO SC-FDMA receiver types based on assuming a channel with memory to be memoryless, and compared their achievable rate to an optimal receiver. It turns out that a low-complexity linear receiver may be a good choice at small bandwidth, short channel impulse responses, consecutive subcarrier mappings and low spatial correlation. The parallel receiver, in addition, may cope with spatial correlation if adaptive modulation is applied. The frequency domain interference cancelation receiver shows almost no performance degradation as compared to an optimal receiver and is hence suited for the widest range of operation conditions. Future work will compare these information theoretic results to realistic coded transmission.

**REFERENCES**