

CO-CHANNEL INTERFERENCE CANCELLATION TECHNIQUES FOR ANTENNA ARRAY ASSISTED MULTIUSER OFDM SYSTEMS

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Abstract

Minimum Mean Square Error (MMSE)-based multiple-antenna assisted co-channel interference suppression/Space Division Multiple Access (SDMA) schemes designed for OFDM have recently drawn wide interest. In this contribution, parallel interference cancellation (PIC) is proposed and studied in comparison to maximum likelihood sequence estimation (MLSE) based schemes. The best design trade-off was constituted by the successive interference cancellation (SIC) schemes.

1 Overview

Recently a substantial amount of research has been conducted towards combining Orthogonal Frequency Division Multiplexing (OFDM) [1] with multiple-antenna reception assisted co-channel interference suppression - or more generally space-division multiple access (SDMA) techniques. One of the key issues is the separation of different users, which can be performed based on their spatial signature, assuming the knowledge of the channel parameters. A multiplicity of algorithms has been proposed for performing the task of user separation. The so-called Sample-Matrix Inversion (SMI) [2, 3] algorithm's roots can be linked back to the early years of development in this research area. Other related algorithms, which have recently drawn interest are for example the successive interference cancellation (SIC) technique proposed for OFDM by Vandenameele *et al.* in [4] as well as the parallel interference cancellation (PIC) algorithm proposed in this contribution as a further potential candidate. These schemes will be compared to the simplified soft-bit generation scheme proposed by Speth *et al.* in [5]. Following the description of our system model in Section 2 and that of the signal model in Section 3 the arrangements studied are briefly reviewed in Section 4. Simulation results are provided in Section 5 for a seven-path fading Wireless Local Area Network (WLAN) channel model having a dispersion of about $1.5\mu\text{s}$ [1], followed by our conclusions in Section 6.

2 The System Model

We assume an OFDM-SDMA uplink environment, where the number of co-channel users is upper-bounded by

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the number of antenna elements associated with the receiver array at the basestation. By contrast, each user is equipped with a single transmission antenna. In our OFDM scheme 1024 subcarriers and a 168-sample guard-interval was used, in order to avoid inter-symbol interference. Throughout the simulations conducted in Section 5 we assumed 'OFDM-symbol-invariant' fading, where the fading magnitude and phase was considered constant during each OFDM symbol. This constraint was imposed, in order to avoid inter-subcarrier interference in the system, which simplifies our comparisons. In addition, a turbo codec has been incorporated into the system, which resulted in a substantial bit error-rate (BER) reduction, although at the cost of a reduced throughput. The turbo coding parameters can be summarized as follows: the coding rate was $R = 1/2$, the constraint length was $K = 3$, the octally represented generator-polynomials of $\{7, 5\}$ were used, and 4 iterations were performed.

3 The Signal Model

The $(P \times 1)$ -dimensional vector of complex signals, $\mathbf{x}[n, k]$, received by the P -element antenna array in the k -th subcarrier of the n -th OFDM symbol is constituted by the superposition of the independently faded signals associated with the L users sharing the same space-frequency resource. The received signal was corrupted by the Gaussian noise at the array elements. The indices $[n, k]$ have been omitted for notational convenience during our forthcoming discourse, yielding:

$$\mathbf{x} = \mathbf{H}\mathbf{s} + \mathbf{n}, \quad (1)$$

where the $(P \times 1)$ -dimensional vector of received signals \mathbf{x} , the vector of transmitted signals \mathbf{s} and the array noise vector \mathbf{n} , respectively, are given by:

$$\mathbf{x} = (x_1, x_2, \dots, x_P)^T, \quad (2)$$

$$\mathbf{s} = (s^{(1)}, s^{(2)}, \dots, s^{(L)})^T, \quad (3)$$

$$\mathbf{n} = (n_1, n_2, \dots, n_P)^T. \quad (4)$$

The frequency domain channel transfer function matrix \mathbf{H} of dimension $L \times P$ is constituted by the set of channel vectors of the L users:

$$\mathbf{H} = (\mathbf{h}^{(1)}, \mathbf{h}^{(2)}, \dots, \mathbf{h}^{(L)}), \quad (5)$$

each of which describes the frequency domain channel transfer function between the single transmitter antenna associated with a particular user l and the reception array elements $p \in \{1, \dots, P\}$:

$$\mathbf{h}^{(l)} = (h_1^{(l)}, h_2^{(l)}, \dots, h_P^{(l)})^T, \quad (6)$$

with $l \in \{1, \dots, L\}$. Regarding the statistical properties of the vector components in Equations 2 and 6, we assume that the complex data signal $s^{(l)}$ transmitted by the l -th user has zero-mean and unit variance. The AWGN noise process n_p at any antenna array element p exhibits also zero-mean and a variance of σ^2 . The frequency domain channel transfer functions $h_p^{(l)}$ for different array elements $p \in \{1, \dots, P\}$ or users $l \in \{1, \dots, L\}$ are independent, stationary, complex Gaussian distributed processes with zero-mean and different variance σ_l^2 [2].

4 Co-Channel Interference Cancellation Techniques

In this section we will provide a brief overview of the most popular co-channel interference cancellation / SDMA techniques, which have been proposed for application in multiple antenna reception assisted OFDM-based communication systems. A variety of schemes has been proposed based on the SMI algorithm [2-4], which constitutes the direct solution to the well-known Wiener-equation, delivering the required antenna weights for the P -element receiver array.

The SMI-Algorithm - The SMI algorithm [2, 3] generates the optimum antenna weights in terms of the minimum mean square error (MMSE) between the associated combiner's output signal and a reference signal known to the receiver. The transmitted signal vector \mathbf{y} estimated by the combiner hosting the L different users' signals is hence given by:

$$\begin{aligned} \mathbf{y} &= (y^{(1)}, y^{(2)}, \dots, y^{(L)})^T \\ &= \mathbf{W}_{\text{MMSE}}^H \mathbf{x}, \end{aligned} \quad (7)$$

where the $(P \times L)$ -dimensional antenna weight matrix \mathbf{W}_{MMSE} is given by [2, 3]:

$$\begin{aligned} \mathbf{W}_{\text{MMSE}} &= (\mathbf{w}_{\text{MMSE}}^{(1)}, \mathbf{w}_{\text{MMSE}}^{(2)}, \dots, \mathbf{w}_{\text{MMSE}}^{(L)}) \\ &= (\mathbf{H}\mathbf{H}^H + \sigma^2\mathbf{I})^{-1} \mathbf{H}, \end{aligned} \quad (8)$$

as the direct solution of the Wiener-equation in the context of a multiuser scenario. Capitalizing on Equations 1, 7 and 8, the signal-to-noise ratio (SNR) at the combiner's output for different users is given by the following diagonal matrix [6]:

$$\begin{aligned} \text{SNR} &= \text{diag}(\text{SNR}^{(1)}, \text{SNR}^{(2)}, \dots, \text{SNR}^{(L)}) \\ &= (\mathbf{W}_{\text{MMSE}}^H \mathbf{H}\mathbf{H}^H \mathbf{W}_{\text{MMSE}}) \cdot \\ &\quad (\sigma^2 \mathbf{W}_{\text{MMSE}}^H \mathbf{W}_{\text{MMSE}})^{-1}. \end{aligned} \quad (9)$$

Equation 9 will be invoked in conjunction with turbo-coding in order to generate the soft-input information for the turbo decoder in Section 5. In the next paragraph we will briefly review the state-insertion (SI) assisted SIC technique proposed in [4], which we denote by SIC-SI.

The SIC-SI-Algorithm - The philosophy of this scheme is motivated by two observations. Firstly, in an OFDM system different users encounter different attenuations of the same subcarrier due to the uncorrelated fading processes associated with the channels between the users' transmit antennas and the receiver antenna array. This results in an increased error probability for

the highly attenuated subcarriers of the users - also depending on their spatial signatures. Secondly, it is well-known [4] that the SNR experienced by each user at the MMSE combiner's output also depends on the total number of users relative to the number of antenna elements of the receiver antenna array. Specifically, for a lower number of users more of the system's diversity potential is dedicated to mitigating the serious channel fades, rather than to suppressing the interfering signal sources. Hence the basic idea behind the SIC scheme proposed by Vandenameele *et al.* in [4] is to extract in each iteration the specific user's signal exhibiting the highest overall power. This can be accomplished with the aid of the MMSE combining algorithm outlined in the previous paragraph, followed by hard-decision, re-modulation and subtraction of the reconstructed received signal associated with a specific user from the residual received array output signal. This technique originates from DS-CDMA systems. Accordingly, the dimension of the $L \times P$ channel transfer function matrix \mathbf{H} given by Equation 5, has to be reduced in each iteration by removing the channel transfer function vector \mathbf{h}^l associated with the most recently detected user. Consequently, an update of the inverse channel correlation matrix of Equation 8 has to be performed. The so-called Matrix-Inversion Lemma [2] can be employed in order to substantially reduce the complexity of this operation, especially for receiver antenna array dimensions in excess of $P = 3$ elements. A further improvement of this algorithm - known as *state-insertion* - has been introduced in [4], where in the subcarriers exhibiting the lowest Signal-to-Interference Ratio (SIR) additional states are tracked after hard-decision, commencing with the first user. The final decision as to which symbol constellation vector to select - where the vector contains the estimated signals of all users - is performed upon measuring the residual signal power after subtraction of the last detected user's signal from the received interfered signal.

The PIC-Algorithm - In this paragraph we will outline our proposal, namely the MMSE-based PIC scheme. One of the key justifications for proposing SIC was that along with a decreasing number of users more antenna array diversity potential can be dedicated by the MMSE combiner to the mitigation of the serious channel transfer function fades, rather than to suppressing the interfering signal sources. Hence the highest array noise mitigation is achieved during the last SIC iteration, when after correct detection and subtraction of the co-channel users with higher power, the residual array output vector is constituted by the transmitted signal of the last user plus an array noise contribution. In this case the MMSE combiner is virtually an MMSE maximum-ratio combiner (MRC), since the combiner weights are adjusted such that the SNR at the combiner's output is maximized. The above interference cancellation principle - which was portrayed in the context of SIC - can be invoked also in form of a PIC scheme, as depicted in Figure 1. Specifically, during the first PIC iteration each user is detected separately by means of the MMSE combiner defined by Equations 7 and 8, invoking the same inverse channel correlation matrix for generating the different users' weight vectors. This is followed by demodulation,

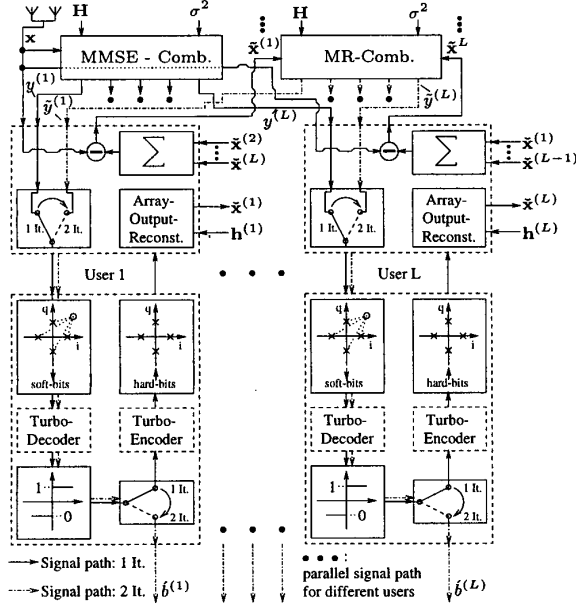


Figure 1: PIC scheme which optionally employs turbo coding to reduce error propagation during decision feedback.

optional turbo decoding, hard-decision, optional turbo re-encoding, re-modulation and reconstruction of each user's P -dimensional antenna array output signal vector $\tilde{\mathbf{x}}^{(l)}$, $l \in \{1, \dots, L\}$. The MMSE-MRC input vector $\tilde{\mathbf{x}}^{(k)}$ for the k -th user, for $k \in \{1, \dots, L\}$ is then given by the original P -dimensional array output vector \mathbf{x} , from which the vectors containing the reconstructed interfering users' signals, namely $\tilde{\mathbf{x}}^{(l)}$, $l \neq k$ have been subtracted according to:

$$\tilde{\mathbf{x}}^{(k)} = \mathbf{x} - \sum_{l=1, l \neq k}^L \tilde{\mathbf{x}}^{(l)}. \quad (10)$$

In order to avoid the matrix inversion associated with generating the weight vector for the MMSE-MRC similarly to Equation 8, it can be shown that one could employ a standard maximum-ratio combiner at the cost of a slight performance penalty at low SNRs. Specifically, the associated MRC scheme is described by the following weight matrix:

$$\mathbf{W}_{\text{MRC}} = \left(\frac{1}{\|\mathbf{h}^{(1)}\|_2^2} \mathbf{h}^{(1)}, \dots, \frac{1}{\|\mathbf{h}^{(L)}\|_2^2} \mathbf{h}^{(L)} \right)^T. \quad (11)$$

The vector of final signal estimates of the users, $\tilde{\mathbf{y}}$, is then given with the aid of Equation 7 upon replacing \mathbf{W}_{MMSE} by \mathbf{W}_{MRC} :

$$\tilde{\mathbf{y}} = \mathbf{W}_{\text{MRC}}^H \tilde{\mathbf{x}}, \quad (12)$$

where the entries of vector $\tilde{\mathbf{y}}$ are defined by Equation 12:

$$\tilde{\mathbf{y}} = (\tilde{y}^{(1)}, \tilde{y}^{(2)}, \dots, \tilde{y}^{(L)})^T, \quad (13)$$

and are also portrayed in Figure 1. The symbols at the output of the MRC of Figure 1 have to be demodulated and decoded once more to yield the transmitted bit vectors $\tilde{\mathbf{b}} = (\tilde{b}^{(1)}, \tilde{b}^{(2)}, \dots, \tilde{b}^{(L)})^T$. A problem associated with the feedback structure of the scheme of Figure 1

- which is also reflected by our simulation results - is the error leakage between different users resulting from erroneous symbol decisions prior to the signal subtraction of Equation 10. In order to alleviate this problem, turbo coding can be effectively employed, which will be underlined by the results in Section 5. In this context an advantage of the system is that the total number of affordable turbo iterations can be split between the turbo decoding processes in the first- and second PIC iteration.

The MLSE-Algorithm - The last algorithm studied is the MLSE soft-bit generation algorithm proposed by Speth *et al.* in [5], which can be employed in both coded- and uncoded systems after applying hard-decisions to the so-called log-likelihood ratios (LLR) of the bits. This approach is based on the idea of detecting one out of L users at a time, so as to avoid the complexity imposed by the joint detection of all users in a Viterbi decoder. In order to avoid the explicit evaluation of the exponential terms which constitute the LLR-values, Speth *et al.* [5] proposed to transform the received signal vector $\mathbf{x}[n, k]$ of the P -element array into a so-called trial-set using the maximum likelihood (ML) estimate $\hat{\mathbf{s}}$ of the L -dimensional transmitted signal vectors \mathbf{s} , which is given similarly to Equation 7 by:

$$\hat{\mathbf{s}} = \mathbf{W}_{\text{ML}}^H \mathbf{x}, \quad \text{with} \quad (14)$$

$$\mathbf{W}_{\text{ML}} = (\mathbf{H}\mathbf{H}^H)^{-1} \mathbf{H}, \quad (15)$$

once again, invoking the signal model outlined in Section 3. It should be noted that the ML transform of Equation 14 maximizes the SIR at the combiner's output. This is in contrast to the MMSE combiner defined by Equations 7 and 8, which rather seeks to maximize the Signal-to-Interference plus Noise Ratio (SINR). Capitalizing on the so-called maximum approximation of the exponential LLR terms [5], the authors' elaborations result in the following log-likelihood value for the first user [5]:

$$L_{1,k} \approx (\hat{\mathbf{s}} - \tilde{\mathbf{s}}_k^{0x})^H \mathbf{R}_\epsilon^{-1} (\hat{\mathbf{s}} - \tilde{\mathbf{s}}_k^{0x}) - (\hat{\mathbf{s}} - \tilde{\mathbf{s}}_k^{1x})^H \mathbf{R}_\epsilon^{-1} (\hat{\mathbf{s}} - \tilde{\mathbf{s}}_k^{1x}), \quad (16)$$

where \mathbf{R}_ϵ denotes the covariance matrix of the estimation error $\epsilon = \hat{\mathbf{s}} - \mathbf{s}$, which is given by [5]:

$$\mathbf{R}_\epsilon = \sigma^2 (\mathbf{H}^H \mathbf{H})^{-1}, \quad (17)$$

and $\tilde{\mathbf{s}} = \tilde{\mathbf{s}}_k^{bx}$ represents the optimum trial-vector defined by [5]:

$$\arg \min_{\tilde{\mathbf{s}} \in \mathcal{M}_k^{bx}} |(\hat{\mathbf{s}} - \tilde{\mathbf{s}})^H \mathbf{R}_\epsilon^{-1} (\hat{\mathbf{s}} - \tilde{\mathbf{s}})|^2, \quad (18)$$

where the minimization is performed over all trial-vectors of the form $\tilde{\mathbf{s}} = (\tilde{s}_1, \tilde{s}_2)^T$ representative of the two-user scenario and the trial-vectors are contained in the set:

$$\mathcal{M}_k^{bx} = \left\{ \begin{pmatrix} \tilde{s}_1 \\ \tilde{s}_2 \end{pmatrix} \mid \tilde{s}_1 \in \mathcal{M}_k^b, \tilde{s}_2 \in \mathcal{M} \right\}. \quad (19)$$

In Equation 19, \mathcal{M} denotes the set of constellation points associated with a particular modulation scheme. For example, we have $M = 4$ elements in \mathcal{M} for 4-QAM, and \mathcal{M}_k^{bx} is the subset of the M^L possible symbol combinations of the L users, comprising only those constellation points, which have an associated bit combination that carries a binary value b at bit position k . We have now determined the soft-bit value $L_{1,k}$ of user 1 at bit-position k . In order to obtain the soft-bit values of user

'Frame-Invariant' Fading WLAN/ A2, U2, SIR=0dB/ 4-QAM

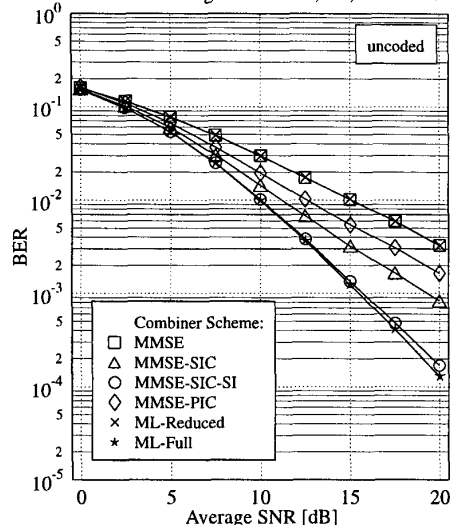


Figure 2: BER performance of the MMSE, MMSE-SIC, MMSE-SIC-SI, MMSE-PIC, MLSE-Reduced and MLSE-Full algorithms in conjunction with the $P = 2$ antenna, $L = 2$ user scenario (SIR=0dB) and 4-QAM - without channel coding - in the 'frame-invariant' fading WLAN channel environment of [1], assuming frequency domain channel knowledge

2, the trial set given by Equation 19 has to be modified correspondingly, by exchanging the appropriate indices. The computational complexity associated with evaluating Equation 18 strongly depends on the size M^L of the 'search-set'. Since the associated complexity is unacceptably high, particularly for QAM schemes of higher order, Speth *et al.* [5] conceived a further simplification, which - instead of Equation 16 - leads to the following expression for the soft-bit values of the first user [5]:

$$L_{1,k} = \frac{|\hat{s}_1 - \hat{s}_{1,k}^0|^2 - |\hat{s}_1 - \hat{s}_{1,k}^1|^2}{R_{e11}}, \quad (20)$$

where

$$\hat{s}_{1,k}^b = \arg \min_{\tilde{s} \in \mathcal{M}_k^b} |\hat{s}_1 - \tilde{s}|^2. \quad (21)$$

Hence a significant complexity reduction is achieved by confining the original trial-set defined by Equation 18 required for generating the soft-bit values for each user to one 'dimension', as seen in Equation 21. This is in contrast to the approach of Equation 16, where each additional user increased the size of the trial-set by one 'dimension'. In our forthcoming performance comparisons, the full-search will be denoted by 'ML-Full', and the simplified or reduced search by 'ML-Reduced'.

5 Simulation Results

Simulations have been conducted for the schemes introduced in Section 4, namely for the MMSE, MMSE-SIC, MMSE-SIC-SI, MMSE-PIC, MLSE-Reduced and for the MLSE-Full arrangements over the 'frame-invariant' fading WLAN channel model of [1]. The MMSE-SIC-SI scheme was configured to track two symbol states after each decision. Hence in this scheme in conjunction with

'Frame-Invariant' Fading WLAN/ A2, U2, SIR=0dB/ 4-QAM

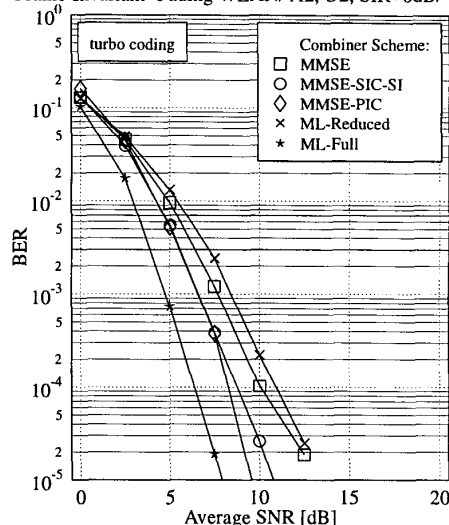


Figure 3: BER performance of the MMSE, MMSE-SIC-SI, MMSE-PIC, MLSE-Reduced and MLSE-Full algorithms in conjunction with the $P = 2$ antenna, $L = 2$ user scenario (SIR=0dB) and 4-QAM - further improved by turbo coding - in the 'frame-invariant' fading WLAN channel environment of [1], assuming perfect frequency domain channel knowledge

an $L = 2$ -user scenario - after subtraction of the last user's signal - a decision has to be performed between 4 possible received symbol vectors, where the vectors contain the estimates of the transmitted symbols. In our initial simulations we assumed an SDMA scenario with two simultaneously transmitting users and a base station (BS) equipped with a receiver antenna array of $P = 2$ elements. A 4-QAM scheme was invoked for signal modulation. Furthermore, we stipulated perfect frequency domain channel transfer function knowledge, which is not available in practice. A reliable channel transfer function estimate can be inferred by capitalizing on frequency domain pilot tones, as described in [3]. The associated bit error-rate (BER) results are depicted in Figure 2. As we expected, the MMSE-SIC - and in particular the MMSE-SIC-SI scheme substantially outperformed the conventional MMSE combiner. This is, because the users encountering a high channel attenuation and hence exhibiting a low effective SNR at the receiver, profit more from the receiver array's diversity potential as a result of the iterative subtraction of already detected users. The ML-Reduced approach using sliced soft-bits exhibited virtually the same performance, as the MMSE combiner. By contrast, the full-search ML scheme performed as well as the MMSE-SIC-SI scheme. The PIC scheme proposed in this contribution on the other hand performed better, than the standard MMSE combiner, although worse than the basic MMSE-SIC scheme. An explanation of this was offered already in Section 4 - erroneous symbol decisions associated with users, who encounter a relatively high channel attenuation may be propagated to users, who produced potentially error-free symbol decisions after the first PIC iteration.

Further simulations were conducted in conjunction with turbo coding, employing a total of four turbo decoding iterations and once again invoking the SDMA scenario.

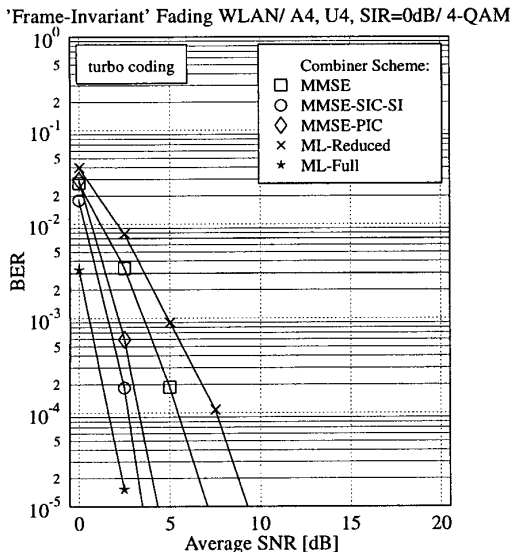


Figure 4: BER performance of the MMSE, MMSE-SIC-SI, MMSE-PIC, MLSE-Reduced and MLSE-Full algorithms in conjunction with the $P = 4$ antenna, $L = 4$ user scenario ($SIR=0dB$) and 4-QAM - further improved by turbo coding - in the 'frame-invariant' fading WLAN channel environment of [1], assuming perfect frequency domain channel knowledge

The corresponding simulation results are illustrated in Figure 3. When employing turbo coding, the full-search MLSE scheme substantially outperformed all other candidates. The full performance potential of this scheme did not emerge in the uncoded scenario of Figure 2, since due to the slicing or hard-decision operation information was lost. Furthermore, we observed that the turbo coded PIC scheme performed equally well, as the SIC-SI scheme. This was achieved by assigning two of the four turbo decoder iterations to the decoding process during the first PIC iteration, and two to the turbo decoding after the second PIC iteration, as portrayed in Figure 1. In this scenario the ML-Reduced approach performed slightly worse, than the MMSE approach. This phenomenon can be explained with the lower accuracy of the absolute values of the soft-bits imposed by the assumptions stipulated in [5]. Further simulations were then conducted for a SDMA scenario of higher complexity. Specifically, $L = 4$ simultaneous users of equal signal power were supported and a $P = 4$ -element receiver antenna array was used at the base station - once again, invoking turbo coding. The corresponding results are portrayed in Figure 4. In comparison to the more simple SDMA scheme described in our previous experiment, the performance of all investigated algorithms was substantially improved - mainly due to the increased diversity potential provided by the $P = 4$ -element receiver antenna array. Using the complex full-search MLSE algorithm, the BER was as low as $1.5 \cdot 10^{-5}$ at a SNR of only 2.5 dB. In this scenario the SIC-SI algorithm performed slightly better, than the PIC algorithm, which is mainly due to the increased probability of error propagation after the first PIC iteration.

6 Discussion and Conclusion

In this paper we have comparatively studied a variety of co-channel interference suppression / SDMA schemes. In terms of performance the full-search MLSE scheme proposed in [5] constitutes the best choice. However, in practice, particularly in conjunction with higher-order modulation schemes, such as for example 16QAM, the computational complexity is prohibitive. For scenarios of relatively low numbers of antennas and users, the turbo coded MMSE-SIC-SI scheme and the coded MMSE-PIC scheme perform equally well. By contrast, in scenarios of higher complexity the coded MMSE-SIC-SI scheme exhibits a slight performance advantage over the MMSE-PIC scheme. The performance of the coded MMSE-PIC scheme may be further improved for each user by switching between the MMSE-combiner output of the first PIC iteration in Figure 1 and the MMSE-MRC output of the second PIC iteration based on the soft-bit information generated after the first iteration, thus mitigating error propagation between different users. In terms of complexity the SIC and PIC schemes have different advantages and disadvantages. The PIC scheme is less complex, than the SIC scheme in terms of the number of operations required for the antenna weight calculation, since the correlation matrix inversion of Equation 8 for the MMSE combiner has to be performed only once. By contrast, the SIC combiner requires continuous updating of the correlation matrix, although the PIC combiner requires the generation of a second set of weights, namely the MRC weights. On the other hand, the demodulation process in the PIC combiner of Figure 1 is potentially of higher complexity - since it is performed twice - unless state insertion [4] is employed in the SIC combiner scheme. An attractive scheme which - in conjunction with PIC - is likely to outperform SIC is subcarrier-based transmitter power control. This will constitute our future research in this area.

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