System Level Impact of Non-reciprocal Interference in Adaptive MIMO-OFDM Cellular Systems

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***Abstract*— In a time-division-duple (TDD) communication system, the channel knowledge can be obtained at the transmitter due to channel reciprocity and used to increase the spectral efficiency of a multiple-input multiple-output (MIMO) communications. However, the interference structure between transmission directions does not necessarily correlate. The obtained quality of service at the receiver may differ significantly from the desired one if the transmission parameters are assigned based on the reverse link measurements only. In this paper, the performance of orthogonal frequency division multiplexing (OFDM) cellular system with adaptive MIMO transmission is studied in the presence of non-reciprocal inter-cell interference when the downlink interference structure is only known at the receiver and only limited feedback information is available at the transmitter. The results are compared to those with perfectly known interference structure per each sub-carrier. The system level impact of realistic interference non-reciprocity scenarios is studied via network simulations. Linear minimum mean squared error (MMSE) filter is applied at the receiver to suppress the impact of structured inter-cell interference together with a simple and bandwidth efficient closed-loop compensation algorithm. System level simulation results show that the proposed compensation algorithm with interference suppression at the receiver results in nearly the same performance as the ideal case. Moreover, it is demonstrated that some feedback to the transmitter is always needed in order to make cellular adaptive TDD MIMO-OFDM system to function properly.**

I. INTRODUCTION

Future broadband communication systems should provide a wide range of services at a reasonable cost and quality of service, comparable to competing wireline technologies. The key radio techniques which have to be considered in developing such systems include: multiple-input multiple-output (MIMO) antennas, multi-carrier modulation and adaptive modulation and coding. The spectral efficiency can be increased dramatically by applying eigenmode transmission combined with adaptive modulation techniques if the transmitter has the channel knowledge [1]–[3].

In a time division duplex (TDD) orthogonal frequency division multiplexing (OFDM) system with adaptive MIMO transmission, the modulation parameters in uplink (UL) and downlink (DL) can be adapted according to the channel conditions. The DL channel can be estimated accurately during the previous UL frame assuming that the frame length is shorter than the channel coherence time. However, the actual interference structure (frequency, time and space selectivity) observed by the terminal receiver in the DL can be very different from that measured by the base station receiver in UL. In such a case, the obtained quality of service at the receiver may differ significantly from the desired one if the transmission parameters are assigned based on the reverse link measurements only.

In [4], a simple and bandwidth efficient closed-loop method to compensate for the non-reciprocity between uplink and downlink interference structure was introduced for the considered adaptive MIMO–OFDM system. In [5] the idea of [4] was further improved by exploiting also the spatial correlation properties of the non-reciprocal interference at the receiver. Linear minimum mean squared error (MMSE) filter can be applied at the receiver at each sub-carrier together with the closed-loop adaptation algorithm to suppress and to compensate for the inter-eigenmode interference caused by the structured inter-cell interference. In this paper, the system level performance of the proposed system with the interference suppression and the closed-loop power offset adaptation algorithm is studied for different antenna configurations. The results are compared to those obtained in the ideal case in which the interference structure is perfectly known also at the transmitter. The ideal feedback would require the interference covariance matrix to be reported to the transmitter for each subcarrier and for each transmitted data frame. The transmitters are assumed to be unaware of the transmission parameters like beamforming weights, transmission powers, etc., used in other cells for the interfering co-channel users.

II. SYSTEM MODEL

The considered adaptive MIMO-OFDM system [4]–[6] has N_C sub-carriers, N_T transmit and N_R receive antennas; it is denoted by (N_C, N_T, N_R) in the sequel. Assuming perfect frequency and sample clock synchronization between the transmitted and the received signals, the input-output relation for OFDM modulator/demodulator chain at each time instant, l, can be written as

$$\begin{aligned}\mathbf{y}_c(l) &= \mathbf{H}_c(l)\mathbf{x}_c(l) + \sum_{i=1}^{M} \sqrt{\omega}_i \mathbf{H}_{i,c}(l)\mathbf{x}_{i,c}(l) + \mathbf{n}_c(l) \\ &= \mathbf{H}_c(l)\mathbf{x}_c(l) + \mathbf{i}_c(l) + \mathbf{n}_c(l), \end{aligned} \tag{1}$$

where $c = 1, \ldots, N_C$ represents the sub-carrier index, $\mathbf{x}_c(l)$ and $\mathbf{y}_c(l)$ denote the c'th columns in $N_T \times N_C$ dimensional matrix $\mathbf{X}(l)$ and $N_R \times N_C$ dimensional matrix $\mathbf{Y}(l)$, respectively, M is then number of independent interference sources (adjacent base stations), $\mathbf{n}_c(l) \sim \mathcal{CN}(0, N_0\mathbf{I}_{N_R})$ represents the additive noise sample vector, $\mathbf{H}_c(l)$ represents the channel matrix for the desired signal at time instant l and $\mathbf{H}_{i,c}(l)$ is the channel matrix for the i'th interference source. All M interfering signals $\mathbf{x}_{i,c} \sim \mathcal{CN}(0, N_0/N_{T_i}\mathbf{I}_{N_{T_i}})$ are assumed to be unknown to the receiver. The factor ω_i represents the power ratio between the received interference from the interference source i and N_0. The factor $I_0/N_0 = \sum_{i=1}^{M} \omega_i$ models the ratio between the total average interference power density I_0 and the noise power density N_0. The frame length, L, is chosen to be smaller than the coherence time of the channel, so that the fading can be modelled constant over one frame. The time index, l, will be skipped in following in order to simplify the notation.

The transmitted vector at the c'th sub-carrier is generated as $\mathbf{x}_c = \mathbf{B}_c\mathbf{d}_c$, where $\mathbf{B}_c \in \mathbb{C}^{N_T \times r_c}$ is the pre-coding matrix, $\mathbf{d}_c = [d_{1,c}, \ldots, d_{r_c,c}]^{\mathrm{T}}$ is the vector of complex data symbols transmitted at sub-carrier c, and $r_c \leq \min(N_T, N_R)$ denotes the number of active data streams decided by the bit and power loading algorithm [6].

The received signal at sub-carrier c becomes now:

$$\mathbf{y}_c = \mathbf{T}_c\mathbf{d}_c + \mathbf{i}_c + \mathbf{n}_c \tag{2}$$

where $\mathbf{T}_c = \mathbf{H}_c\mathbf{B}_c$ represents the equivalent channel matrix of the desired signal at subcarrier c. It includes the accumulated effect of signal processing at the transmitter side and channel propagation on the transmitted data signal. Matrix $\mathbf{T}_c$ is assumed to be perfectly known at the receiver regardless of $\mathbf{B}_c$ used at the transmitter.

The receiver is equipped with linear MMSE filter and the decision variables are generated as $\widehat{\mathbf{d}}_c = \mathbf{W}_c^{\mathrm{H}}\mathbf{y}_c$. The weight matrix $\mathbf{W}_c \in \mathbb{C}^{N_R \times r_c}$ of the MMSE filter is found by a minimization $\mathbf{W}_c = \arg\min_{\mathbf{W}_c} \mathrm{E}\left[\|\mathbf{d}_c - \mathbf{W}_c^{\mathrm{H}}\mathbf{y}_c\|\right]$ and is given as

$$\begin{aligned} \mathbf{W}_c^{\mathrm{H}} &= \mathbf{T}_c^{\mathrm{H}}\left(\mathbf{T}_c\mathbf{T}_c^{\mathrm{H}} + \mathbf{R}_c\right)^{-1} \\ &= \left(\mathbf{T}_c^{\mathrm{H}}\mathbf{R}_c^{-1}\mathbf{T}_c + \mathbf{I}\right)^{-1}\mathbf{T}_c^{\mathrm{H}}\mathbf{R}_c^{-1} \end{aligned} \tag{3}$$

where the quasi-static interference-plus-noise covariance matrix $\mathbf{R}_c$ is

$$\mathbf{R}_c = \sum_{i=1}^{M} \omega_i\left(\mathbf{H}_{i,c}\mathbf{H}_{i,c}^{\mathrm{H}}\right) + N_0\mathbf{I}_{N_R} \tag{4}$$

and it is assumed to be perfectly known at the receiver. The matrices $\mathbf{H}_{i,c}$ are assumed to remain unchanged during the transmission of a frame.

III. Transmitter Design

This section presents the pre-coder design with different levels of channel state information available at the transmitter. We consider two cases as follows:

- optimal eigenmode transmission with full knowledge of $\mathbf{H}_c$ and $\mathbf{R}_c$ per sub-carrier at the transmitter
- sub-optimal eigenmode transmission with more practical assumptions where $\mathbf{R}_c$ is not known at (not reported to) the transmitter and only $\mathbf{H}_c$ is available at the transmitter together with a scalar feedback value

The instantaneous mutual information [bit/s/Hz] of the considered MIMO-OFDM system is [7]

$$I^{(\mathrm{inst})} = \frac{1}{N_C}\sum_{c=1}^{N_C} \log_2 \det\left(\mathbf{I} + \mathbf{R}_c^{-\frac{1}{2}}\mathbf{H}_c\mathbf{C}_c\mathbf{H}_c^{\mathrm{H}}\mathbf{R}_c^{-\frac{1}{2}}\right) \tag{5}$$

where $\mathbf{C}_c = \mathrm{E}\left[\mathbf{x}_c\mathbf{x}_c^{\mathrm{H}}\right] = \mathbf{B}_c\mathbf{B}_c^{\mathrm{H}} \in \mathbb{C}^{N_T \times N_T}$ is the covariance matrix of the signal transmitted on the c'th sub-carrier.

A. Optimal Eigenmode Transmission

With $\mathbf{H}_c$ and $\mathbf{R}_c$ known at the transmitter the optimum pre-combiner $\mathbf{B}_c$ which maximizes (5) is given by $\mathbf{B}_c = \tilde{\mathbf{V}}_c\mathbf{P}_c^{1/2}$, where $\tilde{\mathbf{V}}_c = [\mathbf{v}_{1,c}, \ldots, \mathbf{v}_{r_c,c}]$ contains the first r_c columns of unitary matrix $\mathbf{V}_c$, which is obtained by singular value decomposition (SVD) of the pre-whitened channel matrix $\mathbf{R}_c^{-\frac{1}{2}}\mathbf{H}_c = \mathbf{U}_c\mathbf{\Lambda}_c^{\frac{1}{2}}\mathbf{V}_c{}^{\mathrm{H}}$ [3], [7]. In this way, a set of r_c orthogonal spatial sub-channels are obtained at each sub-carrier and the diagonal matrix $\mathbf{P}_c = \mathrm{diag}\left(P_{1,c}, \ldots, P_{r_c,c}\right)$ controls the powers allocated for each of the r_c eigenmodes. The diagonal matrix $\tilde{\mathbf{\Lambda}}_c = \mathrm{diag}\left(\lambda_{1,c}, \ldots, \lambda_{r_c,c}\right)$ includes the first r_c eigenvalues of the hermitian matrix $\mathbf{H}_c^{\mathrm{H}}\mathbf{R_c}^{-1}\mathbf{H}_c$.

If we now incorporate the term $\mathbf{T}_c^{\mathrm{H}}\mathbf{R}_c^{-1}\mathbf{T}_c = \tilde{\mathbf{\Lambda}}_c\mathbf{P}_c$ into (3), the MMSE filter gets the form $\mathbf{W}_c^{\mathrm{H}} = \left(\tilde{\mathbf{\Lambda}}_c\mathbf{P}_c + \mathbf{I}\right)^{-1}\mathbf{T}_c^{\mathrm{H}}\mathbf{R}_c^{-1}$ and the decision variables are given by:

$$\hat{\mathbf{d}}_c = \left(\tilde{\mathbf{\Lambda}}_c\mathbf{P}_c + \mathbf{I}\right)^{-1}\left(\tilde{\mathbf{\Lambda}}_c\mathbf{P}_c\mathbf{d}_c + \mathbf{T}_c^{\mathrm{H}}\mathbf{R}_c^{-1}\left(\mathbf{i}_c + \mathbf{n}_c\right)\right). \tag{6}$$

The signal-to-interference-plus-noise ratio (SINR) for each spatial sub-channel k, denoted as $\gamma_{k,c}$, becomes

$$\gamma_{k,c} = \frac{\left(\lambda_{k,c}P_{k,c}\right)^2}{\left[\mathbf{T}_c^{\mathrm{H}}\mathbf{R}_c^{-1}\mathbf{T}_c\right]_{k,k}} = \lambda_{k,c}P_{k,c}. \tag{7}$$

The SINR values $\gamma_{k,c}$ can be independently controlled for each sub-channel k by the bit and power loading algorithm [6].

B. Interference Structure Known Only at Receiver

Let us now consider the more practical case where the interference covariance matrix $\mathbf{R}_c$ is known only at the receiver. The channel matrix $\mathbf{H}_c$ is assumed still to be known at the the transmitter. Without $\mathbf{R}_c$ at the transmitter the optimum combiner $\mathbf{B}_c$ which maximizes the mutual information (5) cannot be computed. Therefore, we propose a sub-optimal but still efficient design of the pre-coder. Specifically, we found a lower bound of $I^{(\mathrm{inst})}$ which does not depend on the statistical interference structure in frequency or space and then design a pre-coder which maximizes this bound. It is easy to observe that

$$\hat{\lambda}_{\max}\mathbf{I} \succeq \mathbf{R}_c \tag{8}$$

where $\hat{\lambda}_{\max} = \max_{k,c} \hat{\lambda}_{k,c}$, $\hat{\lambda}_{k,c}$ are eigenvalues of $\mathbf{R}_c$ and the notation $\mathbf{A} \succeq \mathbf{B}$ means that $\mathbf{A} - \mathbf{B}$ is positive semidefinite.

By combining (8) with (5) the instantaneous mutual information of the considered system can be now bounded as

$$I^{(\mathrm{inst})} \geq \frac{1}{N_C}\sum_{c=1}^{N_C} \log_2 \det\left(\mathbf{I} + \frac{1}{\hat{\lambda}_{\max}}\mathbf{H}_c\mathbf{C}_c\mathbf{H}_c^{\mathrm{H}}\right) \tag{9}$$

where we have utilized the fact that $\log$ is strictly increasing function, and, the following implications are valid for $\forall \mathbf{A}, \mathbf{B}, \mathbf{C} \succeq 0 : \mathbf{B} \succeq \mathbf{C} \Rightarrow \mathbf{AB} \succeq \mathbf{AC}$ and $\mathbf{A} \succeq \mathbf{B} \Rightarrow \det \mathbf{A} \geq \det \mathbf{B}$.

The lower bound (9) can be maximized by using the SVD of the channel matrix $\mathbf{H}_c = \mathbf{U'}_c \mathbf{\Lambda'}_c^{\frac{1}{2}} \mathbf{V'}_c^{\mathrm{H}}$. Thus, the transmit covariance matrix is given by $\mathbf{C}_c = \mathbf{V'}_c \mathbf{P'}_c \mathbf{V'}_c^{\mathrm{H}}$. The transmit power optimization for the lower bound, can be performed as

$$P'_{k,c} = \left(\mu' - \hat{\lambda}_{\max}/\lambda'_{k,c}\right)^{+}. \qquad (10)$$

The mutual information of such a system can be then calculated as in (5). Both the optimum signalling and the suboptimal signalling provide the same number of eigenmodes, and, hence, both have the same slope of increase with increasing SINR. The difference to the optimum signalling capacity case decreases as the SINR increases or the interference becomes white.

Based on the reasoning above, we choose to perform suboptimal eigenmode transmission which relies on the channel knowledge only. The pre-combining matrix at subcarrier c is then derived as $\mathbf{B}_c = \tilde{\mathbf{V}}'_c \mathbf{P'}_c^{\frac{1}{2}}$. The matrices $\tilde{\mathbf{V}}'_c$ and $\mathbf{P'}_c$ are defined similarly to the optimum case shown in the previous section. Due to the non-reciprocal interference at the receiver, the orthogonality between the eigenmodes is destroyed. The linear MMSE filter (3) is used at the receiver at each subcarrier to maximize the SINR for each r_c eigenmodes. The SINR for each sub-channel k can be derived as follows:

$$\begin{aligned} \gamma_{k,c} &= \frac{\mathrm{E}\left[\left|\mathbf{w}_{k,c}^{\mathrm{H}} \mathbf{t}_{k,c} d_{k,c}\right|^2\right]}{\mathrm{E}\left[\left|\mathbf{w}_{k,c}^{\mathrm{H}} \left(\bar{\mathbf{T}}_{k,c} \bar{\mathbf{d}}_{k,c} + \mathbf{i}_c + \mathbf{n}_c\right)\right|^2\right]} \qquad (11) \\ &= P'_{k,c} \lambda'_{k,c} \frac{\mathbf{w}_{k,c}^{\mathrm{H}} \mathbf{u'}_{k,c} \mathbf{u'}_{k,c}^{\mathrm{H}} \mathbf{w}_{k,c}}{\mathbf{w}_{k,c}^{\mathrm{H}} \left(\mathbf{R}_c + \bar{\mathbf{T}}_{k,c} \bar{\mathbf{T}}_{k,c}^{\mathrm{H}}\right) \mathbf{w}_{k,c}} \end{aligned}$$

where $\mathbf{w}_{k,c}$, $\mathbf{t}_{k,c}$ and $\mathbf{u}'_{k,c}$ are the kth column vectors of matrices $\mathbf{W}_c$, $\mathbf{T}_c$ and $\mathbf{U'}_c$, respectively. Matrices $\bar{\mathbf{T}}_{k,c} = \bar{\mathbf{U}}'_{k,c} \bar{\mathbf{\Lambda}}'_{k,c} \bar{\mathbf{P}}'_{k,c}$ and $\bar{\mathbf{d}}_{k,c}$ are defined as $\bar{\mathbf{T}}_{k,c} = [\mathbf{t}_{1,c}, \ldots, \mathbf{t}_{k-1,c}, \mathbf{t}_{k+1,c}, \ldots, \mathbf{t}_{r_c,c}]$ and $\bar{\mathbf{d}}_{k,c} = [d_{1,c}, \ldots, d_{k-1,c}, d_{k+1,c}, \ldots, d_{r_c,c}]$, respectively. The SINR values derived in (7) and (11) are used as input parameters to the decoding process at the receiver.

It can be seen from (11) that with the presence of nonreciprocal interference, the SINR per sub-channel can no longer be controlled at the transmitter as in (7) but it is affected by the structure of $\mathbf{R}_c$. If the SINR values are set to some fixed value at the transmitter by the bit and power loading algorithm to achieve certain frame error rate (FER) at the receiver [6], these targets cannot necessarily be met. Therefore, some additional mechanisms are needed to maintain the quality of service at the receiver. In the special case, where the interference term $\mathbf{i}_c$ is white, the SINR of sub-channel k at sub-carrier c is simplified to $\gamma_{k,c} = \lambda'_{k,c} P'_{k,c} / \left(I_0 + N_0\right)$.

A low complexity bit and power loading algorithm requiring a low signalling overhead is used to achieve maximum throughput at certain target frame error rate (FER) [6]. The closed-loop non-reciprocity compensation algorithm was introduced in [4]. The basic idea of the method is to apply a single power offset value at the transmitter for all eigenmodes which compensates for the frequency, time and space selectivity between TDD DL and UL. The receiver estimates signal quality metric (FER or BER) from the transmitted signal, compares the estimated metric to the target metric, and transmits the updated offset value to the transmitter. The offset value can be used either to adjust the transmission rate or the transmitted power per sub-channel. The way to use the feedback information depends on the set of the modulation and coding schemes used and the status of available resources, e.g., power and bandwidth, at the transmitter.

IV. Simulation Results

The simulated system was based on HIPERLAN/2 [8] and IEEE 802.11a assumptions, where the number of subcarriers in the OFDM air interface is N_C=64 and the cyclic prefix length is 16 samples. In all cases, the channel was assumed to be static during one coded OFDM frame. One coded OFDM frame consists of 16 OFDM symbols. Modulation schemes used in the simulations were QPSK, 16QAM and 64QAM. Half rate turbo code was used for channel coding. The number of TX antennas was varied between 2 and 8, while the number of RX antennas was fixed to 2. Power offset step used in the simulation was 0.25 dB.

System level evaluation assesses the impact of a realistic multi-cell environment (realistic interference power profiles, spatial properties, etc.) on the system performance with and without non-reciprocal interference compensation. The gain from both interference suppression and the fast adaptation observed from the link level simulation results [5] can be reduced in situations where the number of interference sources is high (either more transmit antennas in the dominant interferer or more independent interfering adjacent cells with similar average received power) and/or the I_0/N_0 ratio decreases, i.e., the interference becomes more white.

A 57-cell scenario (21 3-sector sites) was used for the multicell system level evaluation. Fig. 1 illustrates the simulation scenario. Okumura-Hata propagation model is used for modelling the path loss. The effect of antenna patterns is included in the path loss calculations [9]. In addition to the path loss, the signal suffers from the shadowing caused by large obstacles usually around the mobile. The shadowing factor is a lognormal random variable with mean of 0 dB and the variance of 6 dB. The channel coefficients for each MIMO antenna transmitter-receiver pair are generated by a standardized geometric stochastic channel model, 3GPP/3GPP2 SCM model [9] with urban micro scenario.

The simulation run consist of D user drops where K users are randomly uniform distributed within the geographic area of the system in the beginning of each drop [9]. In order to speed up the simulation time and to avoid border effects, only the center site users' data is recorded for the statistics. L frames are simulated for each center cite user during each

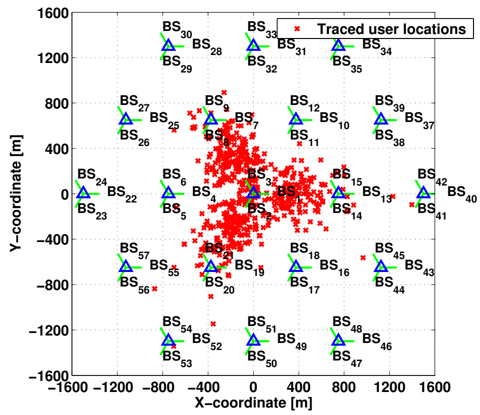

Fig. 1. System level simulation scenario

drop. The parameters D, K and L are selected such that sufficient statistical accuracy is reached.

Universal frequency reuse one is assumed. The multiple access scheme is time division multiple access, where each user is assigned a fixed length transmission slot within downlink TDD frame. Parameter K is adjusted such that 40% average downlink load (time slot occupation) is achieved. Such a loading scenario was selected in order to create sufficient amount of interference while avoiding the system to become too interference limited. A simple random transmission slot allocation scheme is selected for the system level evaluation. In each cell/sector, the users are randomly allocated in the transmission slots within the frame and the user allocation tables are maintained fixed during one simulation drop. The base station transmission power is fixed to 33 dBm. Only those center site users with average SINR more than 0 dB are traced for the statistics while the others are declared to be in the outage. The user locations and allocation tables (interference profiles) are identical between the simulation cases in order to make the results comparable.

The performance of the sub-optimal receiver with interference suppression ('MMSE') was compared to that without interference suppression at the receiver (labelled as 'MF'), and to the optimal case where $\mathbf{R_c}$ is known both at the transmitter and at the receiver (labelled as 'R known at TX/RX'). In the case without interference suppression, the receiver assumed the interference to be white Gaussian ($\mathbf{R}_c = \mathbf{I}$), and thus, the receiver (3) is reduced to the matched filter (MF). The closed-loop interference non-reciprocity compensation algorithm able to follow the time-continuous changes in the interference structure [4], is used in the system level simulations. The two sub-optimal cases were simulated with fast power offset adaptation, denoted as 'tuning' in the figures. The power offset value feedback is used to adjust the reference level in the bit and power loading algorithm, i.e., the lower the received power offset value the higher the transmission rate and vice versa.

Figs. 2 and 3 illustrate the cumulative density function (CDF) of the average user FER for all the simulated cases for 4×2 and 8×2 antenna structures, respectively. It can be

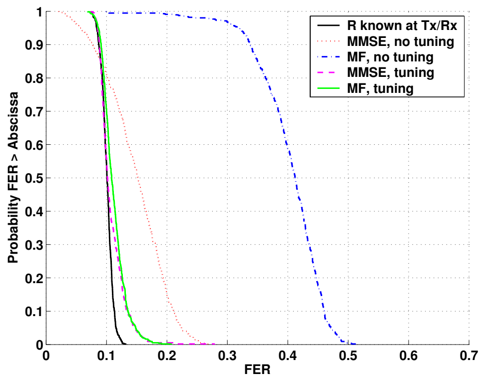

Fig. 2. CDF of average user FER, 4×2 scenario.

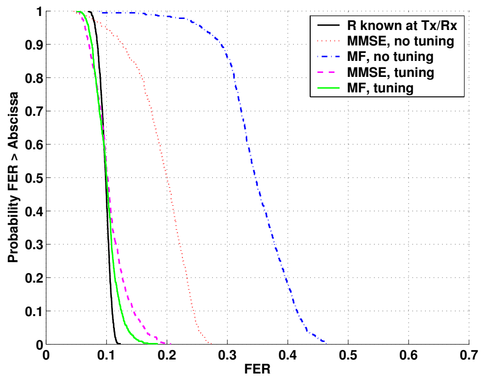

Fig. 3. CDF of average user FER, 8×2 scenario.

seen from the results that the performance without interference suppression or power offset tuning is poor, and the target FER cannot be achieved. For example, the median of the FER is more than 40% while the target was set at 10% in Fig. 2. The case with interference suppression but without power offset tuning results in significantly better performance than the case without interference suppression. However, the resulting FER cannot be controlled at the transmitter due to varying SINR per sub-carrier as shown in (11), and thus, the FER variance is rather large. In the cases with power offset tuning, the average user FER is maintained close to the target. The variation is almost as small as in the optimum case where the interference structure is perfectly known at the transmitter.

Figs. 4, 5 and 6 show the cumulative density function of the average user spectral efficiency [bits/s/Hz] at the receiver for different antenna configurations. Due to high FER, the spectral efficiency of the case without interference suppression or tuning is considerably lower than in the other cases. In general, the cases labelled as 'MMSE, no tuning' and 'MF, tuning' have very similar performance with all antenna configurations. However, the users with high spectral efficiency (high SINR, full eigenmode transmission) gain more from the fast power offset adaptation especially if the number of transmit antennas of the interference source is high, as seen from Fig. 6.

The system performance is further improved if both the interference suppression and the fast power offset tuning are

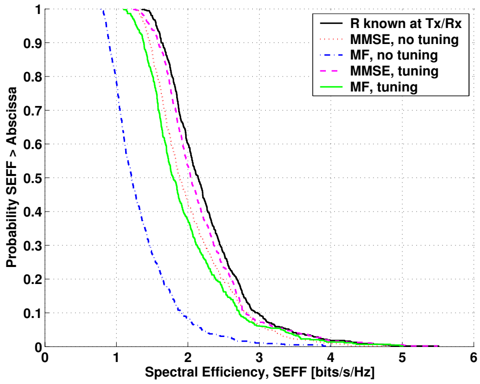

Fig. 4. CDF of average spectral efficiency per user, 2×2 scenario.

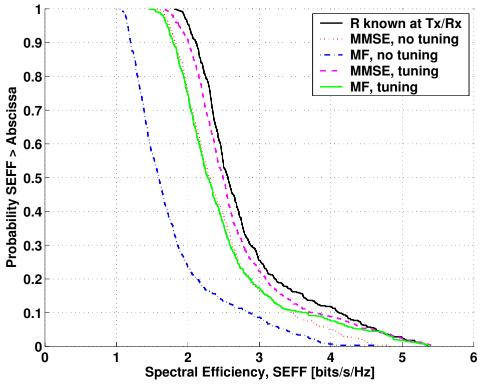

Fig. 5. CDF of average spectral efficiency per user, 4×2 scenario.

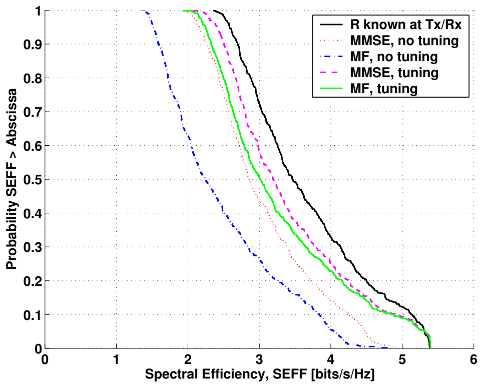

Fig. 6. CDF of average spectral efficiency per user, 8×2 scenario.

enabled for all the users. The gain as compared to the cases labelled as 'MMSE, no tuning' and 'MF, tuning' is about 15% at a 10% outage probability for 4×2 case depicted in Fig. 5, for example. Moreover, the 2×2 and the 4×2 scenarios show almost as good performance as the optimal case where $\mathbf{R_c}$ is perfectly known at the transmitter, labelled as 'R known at TX/RX'. In the 8×2 case, the spatial selectivity of the interference is reduced and at the same time the majority of the users operate in the full eigenmode excitation mode. Therefore, the benefits from MMSE receiver are reduced and the gap to the ideal feedback case is somewhat increased.

V. Conclusion

The performance of an adaptive MIMO-OFDM system was studied in the presence of non-reciprocal inter-cell interference when the downlink interference structure is known only at the receiver. Linear MMSE filter is applied at the receiver to suppress the effect of structured interference together with a simple and bandwidth efficient closed-loop compensation algorithm. The results were compared to those in the ideal case where the interference structure per sub-carrier is also known at the transmitter. The proposed closed-loop compensation algorithm with a simple scalar power offset feedback combined with interference suppression at the receiver results in nearly the same performance as the ideal case where the interference structure per sub-carrier is perfectly known at the transmitter in system level studies. The performance of the ideal case was found to be still better, especially at high SINR region with full eigenmode excitation. However, the required signalling feedback would make this approach unpractical in most applications. The results also demonstrate that in the presence of non-reciprocal inter-cell interference the quality of service at the receiver cannot be controlled if the transmission parameters are defined based on the reverse link measurements only. Therefore, some feedback to the transmitter is always needed in order to make cellular adaptive TDD MIMO-OFDM system to function properly.

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