Joint Detection and CFO Compensation in Asynchronous Multi-User MIMO OFDM Systems

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***Abstract*—It is well known that carrier frequency offsets caused by inaccuracies of local oscillators between transmitter and receiver stations destroy the orthogonality among OFDM subcarriers and induce additional intercarrier interference. In conjunction with MIMO transmission where different users interfere with each other, this effect strongly degrades the signal detection performance if it is not compensated beforehand. In the uplink of multi-user systems different CFOs between all mobile terminals occur whereas its correction is more difficult. In this paper we consider the uplink of a fully asynchronous system with different transmitter and receiver CFOs which appears in distributed MIMO systems with separate users and base stations. We describe a combined scheme for joint CFO correction and spatial layer detection in frequency domain. The detection performance is shown for some selected equaliser concepts. In this way we present interference cancellation algorithms with low and scalable complexity. Finally, numerical simulation results are provided to compare the detection performance.**

I. Introduction

It has been shown in many publications that the orthogonal frequency division multiplex (*OFDM*) modulation scheme in multiple-input multiple-output (*MIMO*) systems is a promising candidate to fulfil requirements for achieving high spectral efficiency at adequate computational effort [1]. In recent standardisation processes, OFDM has been chosen as cellular mobile communications system, namely orthogonal frequency division multiple access (*OFDMA*). In conjunction with space division multiple access (*SDMA*) techniques it is possible to assign time-frequency-space resources to users fairly flexibly. A general overview on state of the art MIMO OFDM/SDMA systems is given in [2]. In such systems, separate base stations and mobile terminals have to suffer from inaccuracies in terms of synchronisation mismatches in time and frequency [3]. Timing offsets are caused by propagation delays through transmission channels, but not considered throughout this paper. Carrier frequency offsets (*CFOs*) occur due to different unsynchronised local oscillators at each mobile terminal in the network as well as Doppler shifts in relative movements to the base station respectively. In cellular systems the different mobiles have to be frequency aligned to its serving base station such as in single frequency networks. One approach is to use downlink signals as reference to estimate and track the CFO in each terminal relatively to the base station [4]. Thus it is possible to de-rotate the phase on the uplink stream to accomplish a single frequency network. Approaches to future mobile communication systems consist of more than one cooperative base station in a so called network MIMO cluster to improve the spectral efficiency in cellular environments. In such distributed systems there are high requirements on the frequency alignment to achieve the anticipated gains. It is difficult to maintain the carrier frequency offset targets in such scenarios unless high effort is spent for base station synchronisation over satellite reference signals, like global positioning systems. Furthermore, it is problematic to ensure that the terminals in a multi cellular network have a perfect synchronisation behaviour at all. Transmitter CFOs in OFDMA systems are already treated in [4] and [5]. The main focus there was to mitigate the multi-user and multiple access interference in the SISO case. In [6] the effects of different transmitter CFOs in an OFDM SDMA system are described with a low complexity equalising approach for its compensation. In this paper, we provide a general system model where K asynchronous users transmit their data over the same channel resource and will be detected by M asynchronous base stations where $M \geq K$. We derive a convenient approximation for the intercarrier interference (*ICI*) power which is used in a new algorithm for the joint CFO compensation and user detection in such a network MIMO OFDM SDMA system where the base stations perform a cooperative signal processing at a central entity. We also consider different complexity levels for the interference cancellation and show the trade-off between the signal processing effort and the user detection performance.

The paper is organised as follows: In section II we derive our multi-user MIMO OFDM system model and the CFO impact. Afterwards in section III we present an analytical SINR analysis. The combined CFO compensation and detection algorithm is described in section IV. Performance evaluations and simulation results are given in section V, before concluding remarks are provided in section VI.

II. System Model

We consider an OFDMA system where K active users simultaneously transmit data on a subset of $\mathcal{D}$ subcarriers. The time domain representation of the base band signal for user k can be denoted as IDFT operation

$$\underline{s}_i^k[n] = \frac{1}{\sqrt{N}} \sum_{l \in \mathcal{D}} \underline{X}_i^k[l] \; e^{\frac{j2\pi l n}{N}} \;, \; 0 \leq n \leq N-1 \quad (1)$$

where $\underline{X}_i^k[l] = 0 \;\forall\; l \notin \mathcal{D}$ and N represents the DFT size. The OFDM symbol index is given by $i = 0, ..., N_S$.

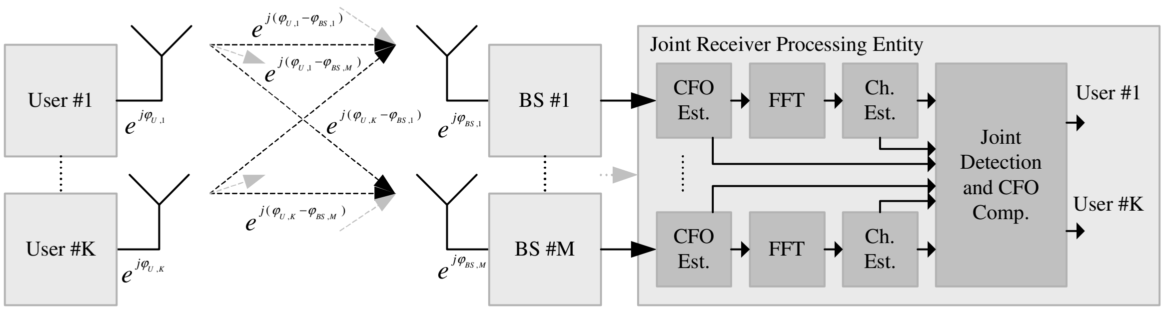

Fig. 1. System Data Flow

To avoid intersymbol interference among OFDM symbols, a cyclic prefix (*CP*) is prepended by copying the last N_{CP} time domain samples of an OFDM symbol before the beginning. After the transmission over the channel with the impulse response vector $\underline{h}$ which consists of N_{CIR} discrete channel taps, the signal at the mth receiver branch can be obtained by

$$\underline{r}_i^m[n] = \left(\sum_{k=1}^{K} \left(\sum_{\lambda=1}^{N_{CIR}} \underline{h}^{m,k}[\lambda] \underline{s}_i^k[n-\lambda] \right) e^{j\Delta\varphi_i^{m,k}[n]} \right) + \underline{w}_i^m[n] \tag{2}$$

where w incorporates additive white Gaussian noise. $\Delta\varphi^{m,k} = \varphi^k - \varphi^m$ implies the phase rotation error between the up and the down conversion process on the link between the kth user and the mth base station. φ is in general referred to as phase noise, but in our model we only investigate the CFO impact such that $\Delta\varphi_i^{m,k}[n]$ can be defined as a linear phase process for each link:

$$\Delta\varphi_i^{m,k}[n] = 2\pi\ \Delta f^{m,k}\ n\ T_S + \varphi_0^{m,k} \tag{3}$$

For a general system model the CFO $\Delta f^{m,k}$ is normalised to subcarrier spacing B_{SC} which leads with $nT_S = \frac{n}{NB_{SC}}$ to

$$\Delta\varphi_i^{m,k}[n] = 2\pi\ \Delta\epsilon^{m,k}\ \frac{n}{N} + \varphi_0^{m,k} \tag{4}$$

with $\Delta\epsilon^{m,k} = \Delta f^{m,k}/B_{SC}$. If OFDM symbols are transmitted consecutively φ_0 is given as $\varphi_0 = \Delta\varphi_{i-1}^{m,k}[N-1]$. The channel impulse response $\underline{h}^{m,k}$ is modelled as zero-mean complex Gaussian variable with variance $\sigma_h^2 = \mathcal{E}\left\{|h_\lambda^{m,k}|^2\right\}$. For the sake of simplicity the OFDM symbol index i is omitted for the rest of this section. With $l \in \mathcal{D}$ the received signal at the lth subcarrier in frequency domain is obtained by a DFT operation

$$\underline{Y}^m[l] = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} \underline{r}^m[n]\ e^{\frac{-j2\pi ln}{N}} \tag{5}$$

Subsequently we use vector matrix notation to simplify the transmission model. With the Fourier transform matrix $\mathbf{F}$ and the link CFO matrix $\mathbf{\Phi}^{m,k} = \mathrm{diag}(e^{j\Delta\varphi^{m,k}[n]})$ where $\mathbf{\Phi}^{m,k} \in \mathbb{C}^{N\times N}$, (5) can be rewritten as

$$\underline{Y}^m = \mathbf{F} \left(\sum_{k=1}^{K} \left(\mathbf{\Phi}^{m,k} \underline{h}^{m,k} \otimes \underline{s}^k \right) + \underline{w}^m \right) \tag{6}$$

$$\underline{Y}^m = \sum_{k=1}^{K} \underbrace{\mathbf{F}\mathbf{\Phi}^{m,k}\mathbf{F}^H}_{\mathbf{E}^{m,k}}\ \underbrace{\mathbf{F}\underline{h}^{m,k}\mathbf{F}^H}_{\mathbf{H}^{m,k}}\ \underline{X}^k + \mathbf{F}\underline{w}^m \tag{7}$$

and in stacked form

$$\underline{Y}^m = \left[\mathbf{E}^{m,1}\ \mathbf{H}^{m,1} \ldots \mathbf{E}^{m,K}\ \mathbf{H}^{m,K}\right]\ \underline{X}' + \underline{W}^m \tag{8}$$

$$\underline{Y}' = \mathbf{E}'\ \mathbf{H}'\ \underline{X}' + \underline{W}' \tag{9}$$

with $\mathbf{E}', \mathbf{H}' \in \mathbb{C}^{(NM)\times(NK)}$, the user symbol vector $\underline{X}' = [\underline{X}^1 \ldots \underline{X}^K]^T$ and the receiver vector $\underline{Y}' = [\underline{Y}^1 \ldots \underline{Y}^M]^T$. With $\mathbf{F}\mathbf{F}^H = I_N$, the unitary similarity transformation of $\mathbf{\Phi}^{m,k}$ is used and the elements in $\mathbf{E}^{m,k}$ are stated as

$$\mathbf{E}^{m,k}[l,p] = \frac{1}{N} \sum_{n=0}^{N-1} e^{j\Delta\varphi^{m,k}[n]}\ e^{\frac{j2\pi n}{N}(p-l)} \tag{10}$$

or equivalently as Dirichlet kernel

$$\mathbf{E}^{m,k}[l,p] = \begin{cases} 1 & \text{for } \kappa = 0 \\ \frac{1}{N}\ e^{j\pi\kappa\frac{N-1}{N}}\ \frac{\sin(\pi\kappa)}{\sin\left(\frac{\pi\kappa}{N}\right)} & otherwise \end{cases} \tag{11}$$

with $\kappa = \Delta\epsilon^{m,k} + p - l\ \forall\ l, p = 1...N$. Based on this model, the frequency domain transmission of (6) can be rewritten for the lth subcarrier to

$$\begin{aligned} \underline{Y}^m[l] &= \sum_{k=1}^{K} \left(\sum_{p\in\mathcal{D}} \underbrace{\mathbf{E}^{m,k}[l,p]\mathbf{H}^{m,k}[p,p]}_{\widetilde{\mathbf{H}}^{m,k}[l,p]}\ \underline{X}^k[p] \right) + \underline{W}[l] \\ \underline{Y}^m[l] &= \sum_{k=1}^{K} \left(\widetilde{\mathbf{H}}^{m,k}[l,l]\ \underline{X}^k[l] \right) \\ &\quad + \underbrace{\sum_{k=1}^{K} \left(\sum_{p\in\mathcal{D}, p\neq l} \widetilde{\mathbf{H}}^{m,k}[l,p]\ \underline{X}^k[p] \right)}_{Y_{ICI}} + \underline{W}[l] \end{aligned} \tag{12}$$

in which $\widetilde{\mathbf{H}}^{m,k}[l,p]$ represents the effective transmission channel in frequency domain. The first term of Eq. (12) represents the ICI free spatial multiplexing signal transmission with a user common phase error (*CPE*) term $\mathbf{E}^{m,k}[l,l]$ which causes an additional phase increment of the channel of one OFDM symbol. The second term Y_{ICI} includes self and multi-user ICI which destroys the orthogonality of the subcarrier symbols $\underline{X}^k[l]$ and leads to amplitude and phase errors of the received symbols.

III. SINR Analysis

For the SINR analysis of the received signal the focus of our interests is on the Y_{ICI} term in Eq. (12) which represents the impact of the intercarrier interference. As it is already stated in [7] and [8] in the case of randomized transmit symbols the first term in (12) is uncorrelated to Y_{ICI}. Thus, the SINR on the lth subcarrier at the mth receiver is given by

$$\mathrm{SINR}_l^m = \frac{\sum_{k=1}^{K} \mathcal{E}\left\{\left|\mathbf{E}^{m,k}[l,l]\right|^2\right\} \sigma_H^2 \sigma_X^2}{\sum_{k=1}^{K} \underbrace{\sum_{p\in\mathcal{D},p\neq l} \mathcal{E}\left\{\left|\mathbf{E}^{m,k}[l,p]\right|^2\right\} \sigma_H^2 \sigma_X^2}_{\gamma_l^{m,k}} + \sigma_W^2} \tag{13}$$

In the numerator the CPE term in Eq. (13) is fully compensated by the receiver algorithms. For the ICI power estimation of $\gamma_l^{m,k}$ we use results from [9] with the adaptation to the CFO context. $\gamma_l^{m,k}$ can be expressed as $\mathcal{E}\left\{Y_{ICI}\ Y_{ICI}^H\right\}$ as

$$\begin{aligned}\gamma_l^{m,k} &= \sum_{p\in\mathcal{D},p\neq l} \mathcal{E}\left\{\left|\mathbf{E}^{m,k}[l,p]\right|^2\right\} \sigma_H^2 \sigma_X^2 \\ &= \sigma_H^2 \sigma_X^2 \left(1 - \mathcal{E}\left\{\left|\mathbf{E}^{m,k}[l,l]\right|^2\right\}\right)\end{aligned} \tag{14}$$

With (11) Eq. (14) can be written as

$$\gamma_l^{m,k} = \sigma_H^2 \sigma_X^2 \left(1 - \frac{1}{N^2} \frac{\sin^2(\pi\Delta\varepsilon^{m,k})}{\sin^2(\pi/N\Delta\varepsilon^{m,k})}\right) \tag{15}$$

With $\sin(\pi/N\Delta\varepsilon^{m,k}) \approx \pi/N\Delta\varepsilon^{m,k}$ for large N the ICI power can be approximated with

$$\begin{aligned}\gamma_l^{m,k} &= \sigma_H^2 \sigma_X^2 \cdot \left[1 - \mathrm{sinc}^2(\pi\Delta\varepsilon^{m,k})^2\right] \\ \mathrm{sinc}^2(\pi\Delta\varepsilon^{m,k}) &= \Big(\sum_{r=0}^{\infty} (-1)^r \frac{(\pi\Delta\varepsilon^{m,k})^{2r}}{(2k+1)!}\Big)^2 \\ &\approx 1 - 2 \cdot \frac{(\pi\Delta\varepsilon^{m,k})^2}{3!} \\ \gamma_l^{m,k} &\approx \sigma_H^2 \sigma_X^2 \cdot (\pi\Delta\varepsilon^{m,k})^2/3.\end{aligned} \tag{16}$$

with $\Delta\varepsilon^{m,k}$ as an approximately quadratic factor. Equation (16) gives an convenient term to be evaluated in CFO compensation algorithms. From Eq. (13) we derive that in the proposed multi-user MIMO system the interference power of each link superimpose each other and the resulting approximated ICI power at one receiver branch can be obtained by

$$\sigma_{ICI,m}^2 = \sum_{k=1}^{K} \gamma_l^{m,k} \tag{17}$$

A closer look at Eq. (13) yields that the joint noise and interference term at each subcarrier consists of $\sigma_{Y,m}^2 = \sigma_{ICI,m}^2 + \sigma_W^2$. From Eq. (16) we know that $\sigma_{ICI,m}^2$ can vary according to the $\Delta\varepsilon^{m,k}$ distribution. This means that for each receiver station m there is another effective noise term $\sigma_{Y,m}^2$.

IV. Receiver Processing

The steps to be done at the receiver processing entity are depicted in Fig. 1. In this section, we will concentrate on the joint CFO compensation and detection process. A short overview over CFO estimation is given beforehand.

A. Frequency Error Estimation

For the upcoming layer detection, the frequency errors have to be known at the receiving entity. The multi-user CFO estimation is treated in [3] and the results can be reused here. [5] also introduces a multi-user CPE estimation scheme for asynchronous OFDMA systems. According to Eq. (10), the CPE can be interpreted as a mean phase rotation during one OFDM symbol. With Eq. (4) and (10) the CPE $\Delta\theta$ of the ith OFDM symbol with respect to the CP on each link can be denoted as

$$\begin{aligned}\Delta\theta_i^{m,k} &= \angle\ \mathbf{E}^{m,k}[l,l] = \angle\left\{\frac{1}{N}\sum_{n=0}^{N-1} e^{j\Delta\varphi_i^{m,k}[n]}\right\} = \\ &\angle\left\{\frac{1}{N}\ e^{j\pi\Delta\epsilon^{m,k}\frac{N-1}{N}}\ \frac{\sin(\pi\Delta\epsilon^{m,k})}{\sin\left(\frac{\pi\Delta\epsilon^{m,k}}{N}\right)}\ \underbrace{e^{j2\pi\Delta\epsilon^{m,k}i\left(1+\frac{N_{CP}}{N}\right)}}_{\Delta\varphi_i^{m,k}[0]=\Delta\varphi_{i-1}^{m,k}[N-1]}\right\}\end{aligned} \tag{18}$$

As a result for every link on different subcarriers an equal but over more than one OFDM symbol a changing CPE occurs. Normally, this is an additional phase offset on each OFDM symbol which will be inherently estimated by the channel estimation process and will also be taken into account during the equalisation process by estimating the spatial decorrelation filter $\mathbf{G}$. Alternatively, the CPE impact on the channel can directly be obtained from the multi-user CFO estimates and can be developed into the time series as it is done in (18). For the rest of this letter we assume that each CFO and CPE value is perfectly known for each link.

B. Multi-User Spatial Layer Detection

Based on Eq. (12), we describe two possible user layer detection strategies in the sequel. Unless otherwise stated, the following derivations are always denoted for the lth subcarrier and for K users and M receiver stations.

1) Layer Detection with CPE Correction: As we have identified in section III, the intercarrier interference can be interpreted as additional noise with $\sigma_{ICI,m}^2$. It is clear that in the case of low ICI the thermal noise floor is much stronger than the part caused by the interference. In cellular systems we suppose that the CFO is partially precompensated at terminal side, such that the low ICI assumption holds. However, the CPE phase de-rotation cannot be neglected because the phase error leads to performance degradations. With (18) we could form a matrix which contains all link CPEs in the ith OFDM symbol derived from the measured CFOs:

$$\Delta\Theta_i = \begin{bmatrix} e^{j\Delta\theta_i^{1,1}} & \cdots & e^{j\Delta\theta_i^{1,K}} \\ \vdots & \ddots & \vdots \\ e^{j\Delta\theta_i^{M,1}} & \cdots & e^{j\Delta\theta_i^{M,K}} \end{bmatrix} \tag{19}$$

$\Delta\Theta_i$ is inherently superimposed leading to the effective channel $\widetilde{\mathbf{H}}$. Due to the nature of the phase difference matrix, the time variant channel can be decomposed in a static part and a dynamic phase rotation part. Then we can simplify the channel inversion such that a continuously tracking because of the static part is not necessary. With the following equations

$$
\begin{aligned}
\widetilde{\mathbf{H}}_i &= \Theta_{i,R}\ \mathbf{H}\ \Theta_{i,T} = \mathbf{H}_l.\Delta\Theta_i && (20)\\
\widetilde{\mathbf{H}}_i^{-1} &= \Theta_{i,T}^{-1}\ \mathbf{H}^{-1}\ \Theta_{i,R}^{-1} = \mathbf{H}^{-1}.\Delta\Theta_i^H && (21)\\
&= \mathbf{G}.\Delta\Theta_i^H && (22)
\end{aligned}
$$

we are able to invert the effective channel in a low complexity manner. Note that (.) means the element wise matrix product. Hence, we can apply known multi-user detection filters with the restriction to the coupled channel model shown in the following transmission equation

$$\underline{\widehat{X}}_i = \mathbf{G}.\Delta\Theta_i^H\ \underline{Y}_i = \mathbf{G}.\Delta\Theta_i^H\ \left(\Delta\Theta_i.\mathbf{H}\ \underline{X}_i + \underline{\widetilde{W}}_i\right) \quad (23)$$

where $\underline{\widetilde{W}}$ is the effective noise vector. It is known that the Maximum-Likelihood detector yields the optimum results by minimizing the error $\underline{Y} - \mathbf{H}\underline{S}$. Therefore we get the optimum detector by using the following equation:

$$\underline{\widehat{X}}_i = \underline{X}_i + \underline{W}_i = \arg\min_{X_i} \left\{ \|\underline{Y}_i - \mathbf{H}.\Delta\Theta_i\ \underline{Y}_i\|^2 \right\} \quad (24)$$

Furthermore we derive the well known MMSE filter with the ICI extension

$$\mathbf{G} = \mathbf{H}^H \left(\mathbf{H}\mathbf{H}^H + \frac{1}{\sigma_s^2} \mathrm{diag}\left(\underline{\sigma}_{Y,m}^2\right) \right)^{-1} \quad (25)$$

where $\underline{\sigma}_{Y,m}^2$ represents an effective noise vector with elements for each receiver station. With (23) after extraction of the kth row of $\mathbf{G}$ and the kth column of Θ_i^H, it is possible to equalise the transmit symbols of the kth user $\widehat{X}_i^k$ with

$$\widehat{X}_i^k = \mathbf{G}^T[k,:]\ \Delta\Theta_i^H[:,k]\ \underline{Y}_i \quad (26)$$

The last equation directly leads to the successive interference cancellation (*SIC*) approach where already detected symbols X_i^k are removed from the received signals iteratively.

$$\underline{Y}_i = \underline{Y}_i - \left(\mathbf{H}[:,k]\ \Delta\Theta_i[:,k]\ \ \widetilde{X}_i^k\right) \quad (27)$$

$\widetilde{X} = \mathcal{Q}(\widehat{X})$ realises the symbol decision of the current user data to avoid error propagation at the interference reduction stage. This can be accomplished by decoding the information bits and remodulate the estimated symbols. In this context robust decoding techniques can be exploited to increase the system performance [10]. As a reference the GENIE approach is used here where always the perfect known symbols are cancelled. Furthermore, depending on the largest layer SINR, the order of the detection process can be modified which results in an additional permutation of the user layers.

2) Layer Detection with ICI Correction: Based on the first signal detection step it is further possible to reduce the remaining interference. Already detected symbols will be remodulated and feed back in order to estimate the intercarrier interference Y_{ICI} which will be subtracted before the channel equalization is applied again. In a system with limited feedback processing resources only adjacent, but strongest interfering, subcarriers are included depending on the ratio

$$\alpha^{m,k} = \left\lfloor \frac{\gamma_l^{m,k}}{\sigma_{ICI,m}^2} \cdot \alpha_{max} \right\rfloor \quad (28)$$

where $\alpha^{m,k} = 0, 1, ..., N/2-1$ and $\alpha_{max} \in [\,0:1:(N/2-1)\cdot KM\,]$. Considering Eq. (12) and (23), the iterative scalable decision feedback equalization (*DFE*) scheme can be written as

$$\underline{\widehat{X}}_i^t = \mathbf{G}^t.\Delta\Theta_i^H \left(\underline{Y}_i - \underline{\widehat{Y}}_{ICI,i}^{t-1}\right) \quad (29)$$

with

$$\underline{\widehat{Y}}_{ICI,i}^{m,t-1} = \sum_{k=1}^{K} \sum_{\substack{p=l-\alpha^{m,k} \\ p\in\mathcal{D}, p\neq l}}^{p=l+\alpha^{m,k}} \widetilde{\mathbf{H}}^{m,k}[l,p]\ \underline{\widetilde{X}}_i^{k,t-1}[p] \quad (30)$$

as the ICI estimate on the lth subcarrier and t as iteration numbering. For further simplification the processing entity can evaluate $\gamma_l^{m,k}$ and $\sigma_{ICI,m}^2$ fairly easily if the linear approximations derived to section III are used. Note that the noise variance in $\mathbf{G}^t$ changes to $\frac{\sigma_w^2}{\sigma_s^2}\mathbf{I}_M$ in the interference reduction steps.

V. Performance Evaluation

In this section, we evaluate the system performance under various CFO impairments. Therefore, we consider a set-up with three mobile stations ($K = 3$) and three base stations ($M = 3$), where we continuously increase the CFO at each station. The simulation parameters are listed in table I. Within this simulation set-up we compare the introduced linear MMSE and the SIC detection schemes only with CPE correction and further the decision feedback equaliser with an increasing number of adjacent subcarriers. Our lower bound is the case where no CFO correction is applied. In Fig. 2, the uncoded bit error rate degradation for a Rayleigh flat channel and in Fig. 3 the uncoded bit error rate degradation for a Rayleigh channel weighted with the denoted power delay profile is pictured, whereas the normalised CFO ϵ is continuously increased in a range from $10^{-3}, ..., 0.2$. The ϵ distribution at the users and base stations is applied as it is stated in table I. We can see in both figures that without any correction of the CFO impact the BER strongly degrades after a few percent's of subcarrier spacing. The initial CPE correction with linear detection yields good results. This shows us that we can significantly increase the system performance if we apply this first correction step. As it was mentioned before in cellular mobile networks the mobile terminals can pre-compensate its coarse CFOs such that only residual frequency offsets occur at the receiver sides. In such scenarios the CPE correction often achieves sufficient

TABLE I
SIMULATION SETTINGS

Parameter	Value
User CFOs	$[\epsilon/2 \;\; 0 \;\; -\epsilon/2]$
Base Station CFOs	$[\epsilon \;\; 0 \;\; -\epsilon]$
CFO Range	$\epsilon = 10^{-3}, ..., 0.2$
SNR	20 dB
Channel Type	IID Rayleigh
Power Delay Profile	Pedestrian A [11]
Modulation Order	16-QAM
Subcarrier Number	$N = 512$
User Subcarrier Allocation	$\mathcal{D} = [106...226]$
Number of OFDM Symbols	$N_S = 7$
CP Length	$N_{CP} = 40$

results with the advantage of a low complexity effort. In the other cases as expected the system performance can be improved by applying the successive interference cancellation schemes. Furthermore, the performance of both detection techniques can be increased by reducing further the ICI at each detection step. There we can see that if we include only a small number of adjacent subcarriers into the ICI reduction process the BER performance can be improved.

VI. CONCLUSION

In this paper we presented how to cope with asynchronous multi-user MIMO systems. We have shown the complete system model for such systems which we used for further evaluation of signal detection and CFO compensation processes. The proposed multi-user detection techniques are evaluated in a simulation environment and performance evaluations show good improvements for our approaches. One advantage of the described algorithm is the channel decomposition into a static and a dynamic part. At this, with the knowledge of the estimated CFOs it is not necessary to track the time invariant channel in the case that time variations are only caused by these CFOs. All parameters for the CFO compensation process can be derived from the CFO estimates. With the ICI power approximation we are also able to correct the intercarrier interference terms. The proposed scalable detection algorithm compensates both low and high ICI with proportional processing effort. The given results can be used as an overview for the synchronisation policies which have to be taken into account at cellular network MIMO planning.

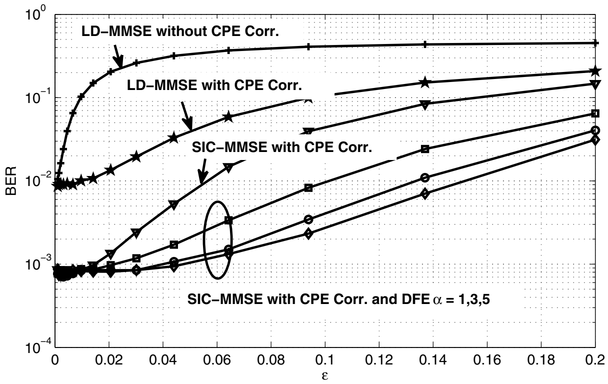

Fig. 2. BER degradation for flat Rayleigh channels

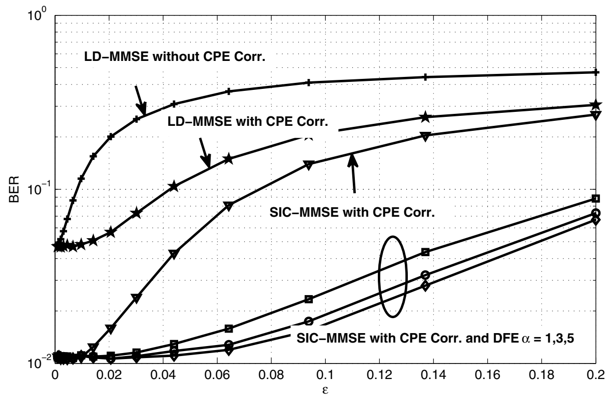

Fig. 3. BER degradation for Rayleigh channels weighted with PDP

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