Frequency Offset Estimation for IFDMA Uplink Systems

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***Abstract*— This paper proposes two frequency offset estimation algorithms for the uplink of an Interleaved Frequency-Division Multiple-Access (IFDMA) system. One algorithm performs estimation in the frequency domain and the other in the time domain. Both algorithms are based on the maximum likelihood estimation (MLE) principle and use knowledge about pilot symbols. IFDMA utilizes a block-interleaved frequency allocation scheme to exploit the frequency diversity of the channel. In the presence of frequency offsets between users, multiple-access interference (MAI) appears, which has a negative impact on existing frequency offset estimation algorithms. The proposed algorithms are robust, since a special construction of pilot symbols allows to exclude a large amount of MAI in the presence of frequency offsets between users. As a result, the proposed time domain frequency estimation algorithm outperforms the frequency domain algorithm and all other known schemes.**

I. Introduction

Orthogonal frequency-division multiplexing (OFDM) has been accepted for several wireless multiple-access systems, e.g., IEEE 802.16 [1] and the wireless LAN 802.11 family [2]. Combining OFDM with FDMA [3] results in orthogonal frequency-division multiple access (OFDMA) [3], which is a promising approach for mobile radio uplink transmission. An alternative to OFDMA is IFDMA which has been introduced in [4] and described in detail in [5]. Recently, IFDMA attained significant importance and is considered as an air interface candidate for 4G mobile communications [6]. IFDMA is equivalent to an OFDMA-CDM [7] scheme with block interleaved frequency allocation and Fourier spreading [8]. Compared to conventional OFDMA, IFDMA seems more attractive due to its lower complexity, since IFDMA requires no discrete Fourier transform (DFT) at the transmitter. Moreover, IFDMA shows a significantly lower peak-to-average power ratio (PAPR) than conventional OFDMA.

Like OFDMA, IFDMA is very sensitive to timing and carrier frequency offsets [9]. In IFDMA and OFDMA uplink systems, frequency offsets between mobile users and the base station cause MAI. Several studies have been accomplished to estimate the frequency offset in OFDMA uplink systems. Recently, Morelli presented a joint timing and frequency offset estimator for OFDMA uplink systems, which uses a block-interleaved subcarrier allocation [10]. This scheme relies on the repetition of fixed pilot symbols and the frequency estimator is derived from ad hoc reasoning.

In this contribution, we propose a frequency domain and a time domain algorithms for frequency offset estimation of a new user entering the system, as in [10]. The algorithms use pilot symbols and exploit the signal structure of IFDMA for maximum likelihood frequency offset estimation. As in [10], we assume that the other users have been acquired and aligned to the base station reference. The obtained frequency estimation is returned to the mobile terminal via downlink control channel and is used to adjust the mobile terminal carrier frequency.

We use the following notation: variables denoting signals in time domain are written in small letters, variables denoting signals in frequency domain are written in capital letters. Bold variables designate matrices or vectors.

The remainder of the paper is organized as follows. Section II introduces the IFDMA scheme. In section III time and frequency domain algorithms for frequency offset estimation are presented. Simulation results are discussed in section IV and, finally, section V concludes the paper.

II. Interleaved Frequency-Division Multiple Access (IFDMA) Scheme

In the following, IFDMA is described as a special case of OFDMA-CDM. For additional descriptions of the IFDMA scheme refer to [5]. The standard OFDMA-CDM [7] uplink transmission system is able to support up to K simultaneously active users. The number of active users is defined as $N_u \leq K$. Each OFDMA-CDM user $i = 1, \ldots, N_u$, performs a blocked transmission of Q complex-valued symbols $d_q^{(i)}, q = 1, \ldots, Q$, which are assumed to be equally likely and are taken from the underlying modulation alphabet. The duration of one data symbol $d_q^{(i)}$ is T_s. The Q symbols are arranged to build the symbol vector $\mathbf{d}^{(i)}$ of user i

$$\mathbf{d}^{(i)} = (d_1^{(i)}, d_2^{(i)}, \ldots, d_Q^{(i)}). \quad (1)$$

The symbol vector $\mathbf{d}^{(i)}$ is first serial-to-parallel converted. Then, a code-division multiplexing (CDM) component is introduced by spreading each symbol with a different spreading code $\mathbf{c}_q, q = 1, \ldots, Q$, of length $L \geq Q$ and adding together the resulting Q chips. After summation, the resulting vector $\mathbf{s}^{(i)}$ can be represented by

$$s_n^{(i)} = \sum_{q=1}^{Q} c_{n,q} \cdot d_q^{(i)}, n = 0, \ldots, L-1, \quad (2)$$

where $c_{n,q}$ is the n-th component of the spreading code $\mathbf{c}_q$, $q = 1, \ldots, Q$. Usually, Walsh-Hadamard (WH) codes are utilized for the spreading. Finally, an inverse DFT (IDFT) with user-specific frequency mapping is performed for the row vector $\mathbf{s}^{(i)}$. Block interleaving achieves optimal frequency diversity, since it distributes the L chips of the Q symbols of user i equally over the whole transmission bandwidth. Note, OFDMA-CDM exclusively assigns to each of the K users a set of L frequencies out of all possible $N_c = KL$ subcarriers. If all subcarriers assigned to user i are uniformly spaced over the whole available transmission bandwidth at a distance N_c/L, the time domain transmit signal $x_l^{(i)}, l = 0, \ldots N_c-1$, obtained after the IDFT operation in the transmitter, can be represented by

$$x_l^{(i)} = e^{j\frac{2\pi}{N_c}il} \frac{1}{\sqrt{N_c}} \sum_{n=0}^{L-1} s_n^{(i)} e^{j\frac{2\pi}{L}nl}. \tag{3}$$

As in standard OFDMA, a guard interval larger than the maximum channel delay is added as a cyclic prefix in OFDMA-CDM in order to avoid interference from preceding OFDMA-CDM transmission symbols.

IFDMA as a special case of OFDMA-CDM can be realized if a fully loaded system, i.e. $L = Q$, is considered and Fourier codes (instead of WH codes) with components $c_{n,q} = \{e^{-j\frac{2\pi}{Q}nq}\}, n = 0, \ldots, L-1, q = 1, \ldots, Q$, are used for spreading. In this case, the Fourier codes for spreading and the rotation factors of the Fourier transform cancel out and the transmit signal given in (3) reduces to

$$x_l^{(i)} = \underbrace{e^{\frac{j2\pi}{N_c}il}}_{\text{user specific phase factor}} \cdot \underbrace{\frac{1}{\sqrt{K}} d_{l \bmod L}^{(i)}}_{\text{transmitted chips}}, l = 0, \ldots, N_c - 1. \tag{4}$$

Comparing (4) with [4] [5] it follows that IFDMA is equivalent to OFDMA-CDM with block interleaved frequency allocation and can be completely generated in the time domain without using the Fourier transform. The time domain generation of IFDMA uses compression and repetition as described in detail in [5]. Note, the multiplication with the user specific phase factor in (4) ensures the user discrimination by assigning to each user i a set of subcarriers orthogonal to all other users' subcarrier sets. The duration of the resulting chips $x_l^{(i)}$ is T_s/K.

In the following, we assume transmission over a time-dispersive mobile radio channel described by its impulse response vector $\mathbf{h}^{(i)}$ of dimension M with components $h_m, m = 1, \ldots, M$. By inserting a guard interval with a length no less than M at the transmitter, we can convert the linear convolution of the transmitted symbol vector $\mathbf{x}^{(i)}$ and the channel impulse response $\mathbf{h}^{(i)}$ into a circular convolution. Therefore, the orthogonality between the subcarrier sets of different users is maintained. The received signal vector $\mathbf{y}$ after guard interval removal can be represented by

$$\mathbf{y} = \mathbf{f}^{(i)}\hat{\mathbf{h}}^{(i)}\mathbf{x}^{(i)} + \sum_{\substack{k=1 \\ k\neq i}}^{K} \mathbf{f}^{(k)}\hat{\mathbf{h}}^{(k)}\mathbf{x}^{(k)} + \mathbf{n}, \tag{5}$$

where $\mathbf{n}$ is a vector of white Gaussian noise samples and $\mathbf{f}^{(i)}$ denotes the diagonal matrix of user frequency offsets

$$\mathbf{f}^{(i)} = \text{diag}\{e^{2\pi\epsilon^{(i)}/N_c}, \ldots, e^{2\pi\epsilon^{(i)}}\}. \tag{6}$$

In (6), $\epsilon^{(i)}$, refers to the frequency offset of user i normalized to the subcarrier spacing, such that $|\epsilon^{(i)}| \leq 1$. The channel matrix $\hat{\mathbf{h}}^{(i)}$

$$\hat{\mathbf{h}}^{(i)} = \begin{pmatrix} h_1^{(i)} & \cdots & h_M^{(i)} & 0 & \cdots & 0 \\ 0 & h_1^{(i)} & \cdots & h_M^{(i)} & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots & \vdots & \vdots \\ 0 & \ldots & h_1^{(i)} & \ldots & h_M^{(i)} & 0 \\ 0 & 0 & \ldots & h_1^{(i)} & \ldots & h_M^{(i)} \end{pmatrix} \tag{7}$$

is the $N_c \times N_c$ right circular matrix whose rows are cyclically shifted versions of the vector $\tilde{\mathbf{h}}^{(i)}$. The vector $\tilde{\mathbf{h}}^{(i)}$ is obtained by appending $N_c - M$ zeros to $\mathbf{h}^{(i)}$.

Defining $\mathbf{W}$ as an $N_c \times N_c$ DFT matrix, the frequency domain representation of the received signal vector $\mathbf{y}$ can be written as $\mathbf{Y} = \mathbf{W}\mathbf{y}$. The L elements $Y_{nK+i}, n = 0, \ldots, L-1$, of vector $\mathbf{Y}$ determine the data symbols of user i on the L subcarriers

$$Y_{nK+i} = H_{nK+i}^{(i)} s_n^{(i)} I_0^{(i)} + \sum_{\nu=0}^{L-1} \sum_{\substack{k=1 \\ k\neq i}}^{K} H_{\nu K+k}^{(k)} s_\nu^{(k)} I_{\nu K+k-nK-i}, \tag{8}$$

where the fading coefficients $H_{nK+i}^{(i)}, n = 0, \ldots, L-1$, are obtained by DFT transformation of matrix $\hat{\mathbf{h}}^{(i)}$. The coefficients $I_u^{(i)}$ depend on the frequency offset $\epsilon^{(i)}$ and are given by [11]

$$I_u^{(i)} = \frac{\sin \pi(u + \epsilon^{(i)})}{N_c \sin \frac{\pi}{N_c}(u + \epsilon^{(i)})} e^{j\pi(1-\frac{1}{N_c})(u+\epsilon^{(i)})}. \tag{9}$$

The first term in (8) represents the useful data of user i and the second term can be considered as MAI caused by the frequency offsets.

III. Maximum-likelihood Algorithms for Frequency Offset Estimation

In this section, we propose two novel algorithms for frequency offset estimation which are based on the maximum-likelihood principle. The algorithms can be applied to both IFDMA and OFDMA. One of the algorithms performs estimation in the frequency domain whereas the second performs estimation in the time domain. We assume a transmission scheme where several transmission symbols are grouped into a transmission frame and where several pilot symbols at the beginning of each frame are available for frequency offset estimation. Frequency offset estimation is performed in two steps. In the first step, the MAI within the pilot symbols caused by the frequency offsets is reduced. In the second step, the actual maximum-likelihood frequency estimation is performed. Note, reducing the MAI within the pilot symbols prior to performing the frequency estimation improves the performance of the considered algorithms and makes them robust against MAI.

Reduction of MAI

The reduction of MAI within the pilot symbols is achieved by applying additional spreading in time direction. For this purpose, we divide the K possible users i, $i = 1, \ldots, N_u$, into groups of P users and assign to each user i a unique orthogonal spreading code $\mathbf{b}^{(\upsilon)}$, $\upsilon = i \bmod P$, of length P, e.g. a Walsh-Hadamard code if P is a power of two. Starting with a known transmit signal vector $\mathbf{x}^{(i)}$, altogether P pilot symbols for user i are constructed by spreading $\mathbf{x}^{(i)}$ with $\mathbf{b}^{(\upsilon)}$, $\upsilon = i \bmod P$, in time direction. As a result, the μ-th received pilot symbol $\mathbf{y}_\mu, \mu = 1, \ldots, P$, can be written according to (5) as

$$\mathbf{y}_\mu = \mathbf{f}^{(i)}\hat{\mathbf{h}}^{(i)}\mathbf{x}^{(i)}b_\mu^{(i \bmod P)} + \sum_{\substack{k=1\\k\neq i}}^{K} \mathbf{f}^{(k)}\hat{\mathbf{h}}^{(k)}\mathbf{x}^{(k)}b_\mu^{(k \bmod P)} + \mathbf{n}, \tag{10}$$

MAI reduction can be performed in frequency or time domain. For the frequency domain estimation, the DFT operation must be applied first. After DFT operation, the obtained frequency domain vector $\mathbf{Y}_\mu = \mathbf{W}\mathbf{y}_\mu, \mu = 1, \ldots, P$, is placed in a $P \times N_c$ - dimensional matrix as the μth column. Despreading is performed by multiplication of the obtained matrix with the spreading code $\mathbf{b}^{(i \bmod P)}$ and results in a new vector $\mathbf{Y}_\mathrm{r}$ which can be represented as

$$\mathbf{Y}_\mathrm{r} = \{\mathbf{Y}_1, \ldots, \mathbf{Y}_\mu, \ldots, \mathbf{Y}_P\}\mathbf{b}^{(i \bmod P)}. \tag{11}$$

Assuming that the channel coefficients remain constant at least for the P pilot symbols $\mathbf{Y}_\mu, \mu = 1, \ldots, P$, allows us to simplify (11). As a result, we obtain the L elements $Y_{\mathrm{r},nK+i}, n = 0, \ldots, L-1$ of vector $\mathbf{Y}_\mathrm{r}$, which determine the data symbols of user i as

$$\begin{aligned} Y_{\mathrm{r},nK+i} &= P^2\Big[H_{nK+i}^{(i)}s_n^{(i)}I_0^{(i)} \\ &+ \sum_{\nu=0}^{L-1}\sum_{\substack{k\leq K\\(k \bmod P)=(i \bmod P)\\k\neq i}} H_{\nu K+k}^{(k)}s_\nu^{(k)}I_{\nu K+k-nK-i}\Big]. \end{aligned} \tag{12}$$

A comparison of (12) with (8) reveals that in (12) the amount of MAI is significantly reduced. Especially, the interference from the $P-1$ neighboring subcarriers is completely removed in (12). Moreover, an analysis shows that the coefficients in (9) degrade rapidly with increasing u, which means that the significant interference contribution due to the frequency offset stems from the neighboring subcarriers.

MAI reduction can also be performed in time domain without DFT operation. The time domain signal $\mathbf{y}_\mathrm{r}$ can be written as

$$\mathbf{y}_\mathrm{r} = \{\mathbf{y}_1, \ldots, \mathbf{y}_\mu, \ldots, \mathbf{y}_P\}\mathbf{b}^{(i \bmod P)}. \tag{13}$$

and is used, in the following, for frequency offset estimation in time domain.

Thus, the MAI-free signal of user i is obtained either in (11) or in (13), where the influence caused by the frequency offset $\epsilon^{(i)}$ is reconstructed on the N_c subcarriers.

For the sake of simplicity, but without loss of generality, only the frequency estimation for user #1 ($i = 1$) is considered in the following. Thus, we omit the subscript i, in the rest of the paper.

Frequency offset estimation in the frequency domain

For frequency offset estimation we consider those subcarriers which are used for transmitting the chips $\mathbf{s}$ of user #1. Additionally, we consider one adjacent subcarrier for each transmitted chip. The vector $\bar{\mathbf{Y}}_\mathrm{r}$, which consists of the received values for the considered subcarriers, has a length of $2L$ and is given by

$$\bar{\mathbf{Y}}_\mathrm{r} = \{Y_{\mathrm{r},1}, Y_{\mathrm{r},2}, \ldots, Y_{\mathrm{r},(L-1)K+1}, Y_{\mathrm{r},(L-1)K+2}\}^T. \tag{14}$$

Moreover, we introduce the vector of fading coefficients $\mathbf{H}$ for subcarrriers, where chips $\mathbf{s}$ have been transmitted. Note that the vector $\mathbf{H}$ is unknown at the receiver. The equation which describes the relation between vector $\mathbf{H}$ and $\bar{\mathbf{Y}}_\mathrm{r}$ can be written as

$$\bar{\mathbf{Y}}_\mathrm{r} = \boldsymbol{\theta}(\epsilon)\mathbf{H} + \mathbf{n}, \tag{15}$$

where $\boldsymbol{\theta}(\epsilon)$ is a $2L \times L$-dimensional matrix, which is determined by the frequency offset coefficients given in (9)

$$\boldsymbol{\theta}(\epsilon) = \begin{pmatrix} I_0 & I_K & \cdots & I_{(n-1)K} \\ I_{-1} & I_{K-1} & \cdots & I_{(n-1)K-1} \\ I_{-K} & I_0 & \cdots & I_{(n-2)K} \\ \vdots & \vdots & \vdots & \vdots \\ I_{(1-n)K-1} & \cdots & I_{-K-1} & I_{-1} \end{pmatrix}\begin{pmatrix} s_1 \\ s_2 \\ \vdots \\ \vdots \\ s_L \end{pmatrix} \tag{16}$$

Our task is to estimate the frequency offset $\hat{\epsilon}$ exploiting the vector $\bar{\mathbf{Y}}_\mathrm{r}$ and equation (15). An ML approach can be utilized for this purpose. We build a trial function and variate $\tilde{\epsilon}$. Initially, we keep $\tilde{\epsilon}$ constant and try to find the estimation $\hat{\mathbf{H}}$ for the unknown vector $\mathbf{H}$. The ML solution is given by [12]

$$\hat{\mathbf{H}} = [\boldsymbol{\theta}^H(\tilde{\epsilon})\boldsymbol{\theta}(\tilde{\epsilon})]^{-1}\boldsymbol{\theta}^H(\tilde{\epsilon})\bar{\mathbf{Y}}_\mathrm{r}. \tag{17}$$

Substituting (17) into (15) and maximizing with respect to $\tilde{\epsilon}$ produces the trial function of ML estimate $\hat{\epsilon}$ of the frequency offset ϵ

$$\hat{\epsilon} = \arg\max_{\tilde{\epsilon}} \bar{\mathbf{Y}}_\mathrm{r}^H\boldsymbol{\theta}(\tilde{\epsilon})\Big[\boldsymbol{\theta}^H(\tilde{\epsilon})\boldsymbol{\theta}(\tilde{\epsilon})\Big]^{-1}\boldsymbol{\theta}^H(\tilde{\epsilon})\bar{\mathbf{Y}}_\mathrm{r}. \tag{18}$$

This algorithm is referred to as the FD in the following. The FD algorithm requires an one-dimensional search over the possible range for ϵ. Moreover, the inversion of an $L \times L$-dimensional matrix is needed. The next step is to develop a practical and simple algorithm for the estimate $\hat{\epsilon}$ of the frequency offsets, ϵ, in the time domain without any exhaustive search.

Frequency offset estimation in the time domain

After successful reduction of the MAI, the time domain received signal $\mathbf{y}_\mathrm{r}$, obtained in (13), is used for frequency offset estimation. For this purpose, the signal $\mathbf{y}_\mathrm{r}$ is divided into K subgroups $\mathbf{r}_k, k = 1, \ldots, K$. The L elements $r_{k,n}, n = 0, \ldots, L-1$ of each subgroup k can be defined as

$$r_{k,n} = y_{\mathrm{r},kL+n}e^{-j2\pi(kL+n)i/N_c}, \tag{19}$$

where $y_{\mathrm{r},kL+n}, k = 1, \ldots, K, n = 0, \ldots, L-1$, are the elements of the vector $\mathbf{y}_\mathrm{r}$. The rotating phase $e^{-j2\pi(kL+n)i/N_c}$ is added in (19) in order to eliminate the influence of the user-specific phase vector. Note, for the considered user #1, we have $i = 1$ in (19).

A closer analysis reveals that due to the repeating structure of the IFDMA time domain signal in (4) and the property of the right-circulant matrix in (7), the obtained groups in (19) experience exactly the same influence of the transmission channel.

If K is even, the obtained subgroups can be arranged in two one-row matrices $\boldsymbol{\zeta}_1$ and $\boldsymbol{\zeta}_2$ such that

$$\begin{aligned} \boldsymbol{\zeta}_1 &= \{\mathbf{r}_1^T, \mathbf{r}_2^T, \ldots, \mathbf{r}_{K/2}^T\}, \\ \boldsymbol{\zeta}_2 &= \{\mathbf{r}_{K/2+1}^T, \mathbf{r}_{K/2+1}^T, \ldots, \mathbf{r}_K^T\}, \end{aligned} \tag{20}$$

and the following expression is valid

$$\boldsymbol{\zeta}_2 = e^{j\pi\epsilon}\boldsymbol{\zeta}_1. \tag{21}$$

The obtained vectors have length $LK/2$ and differ only by a constant value, which is determined by the frequency offset. Therefore, if ϵ is determined using the observation in (21), it is possible to obtain an accurate estimation even if the offset is too large for successful data demodulation.

As shown in [13], the MLE can be written as

$$\hat{\epsilon} = \frac{1}{\pi}\tan^{-1}\left\{\frac{\Im\{\boldsymbol{\zeta}_2\boldsymbol{\zeta}_1^H\}}{\Re\{\boldsymbol{\zeta}_2\boldsymbol{\zeta}_1^H\}}\right\}, \tag{22}$$

where operators $\Im(x)$ and $\Re(x)$ denote real and imaginary part of x, accordingly.

The following remarks are of interest:

- Function $\tan^{-1}$ can be approximated by its argument if ϵ is small enough.
- The solution presented in (22) does not require knowledge of the channel state information.
- The limits of the estimation given by (22) are $|\epsilon| \leq 0.5$ which is half of the subcarrier spacing. If $\epsilon > 0.5$, the expression (21) is no longer valid for each subgroup k. When this happens, the estimate given by (22) becomes useless.
- The standard deviation of the estimate $\hat{\epsilon}$ can be addressed as follows

  $$\mathrm{var}(\hat{\epsilon}) = \left(\frac{1}{\pi}\right)^2 \frac{1}{\gamma LP}, \tag{23}$$

  where γ represents the signal-to-noise ratio (SNR).

IV. Simulation Results

The performance of the proposed frequency estimation algorithms is investigated by computer simulation in a frequency-selective fading channel. The following assumptions are made.

System Parameters

The transmission system under investigation has a bandwidth of 2 MHz and the carrier frequency is located at 5 GHz. The sampling interval is $T_c = 0.5\mu s$. The total number of subcarriers is $N_c = 256$. The resulting subcarrier spacing is 7.81 kHz. The spreading length is set to $L = 32$ and the total number of users is $K = 8$. All users are assumed to be perfectly aligned in the time direction. This can be considered as a realistic scenario, since for each user, the start of the transmission is determined by the instructions transmitted from the base station via the control channel.

Channel Model

For the simulations, a multipath channel model is considered, which consists of a tapped delay line with $M = 11$ statistically independent Rayleigh fading taps. The average power for each channel component $h_m, m = 1, \ldots, M$, is modelled as

$$E\{|h_m|^2\} = -m + 1 \text{ (dB)}. \tag{24}$$

Additionally, the average channel attenuation is normalized to unity in our simulations. For each simulation run, a channel snapshot is generated which is kept constant for the P observed pilot symbols. The maximum frequency offset $\epsilon_{\max}$ is 0.15, which corresponds to a mobile speed greater than 200 km/h. In particular, frequency offsets are modelled as a random variable with uniform distribution over $[-\epsilon_{\max}, \epsilon_{\max}]$.

System Performance

We evaluate the performance of both the TD and the FD algorithms in terms of the mean square error (MSE) of the frequency estimates and the standard deviation of such an estimate.

In Fig. 1, the standard deviation of the estimate is simulated for different values of P. Only user #1 is active and $N_u = 1$. The theoretical curves are plotted according to (23). At high SNR values, simulation results perfectly match the theoretical curves, validating our approach.

In Fig. 2, a performance comparison between the TD and the FD frequency estimation algorithms is presented. As a result, the TD algorithm outperforms the FD algorithm. The reason for that is that in the FD algorithm the trial function (18) has a flat dependence on $\tilde{\epsilon}$ and is easily affected by noise. The performance of the FD estimation algorithm degrades rapidly since flatness in (18) increases with decreasing $\epsilon_{\max}$.

In Fig. 3, the MSE of the proposed TD frequency estimation algorithm is compared to the reduced complexity frequency estimator (RCFE) proposed in [10]. The MSE is simulated as a function of $\epsilon_{\max}$. It is stated in [10], that RCFE is equal to the frequency estimator proposed in [13] for the case of $P = 2$. We have compared the performance of RCFE and the TD frequency estimation algorithm for the realistic scenario, where all users are active i.e., $N_u = 8$. The SNR is 14 dB. As a result, the proposed algorithm outperforms RCFE. The TD frequency estimation algorithm performs equal to RCFE, if $\epsilon_{\max}$ is low. With increasing $\epsilon_{\max}$, MAI also increases and the performance of RCFE degrades

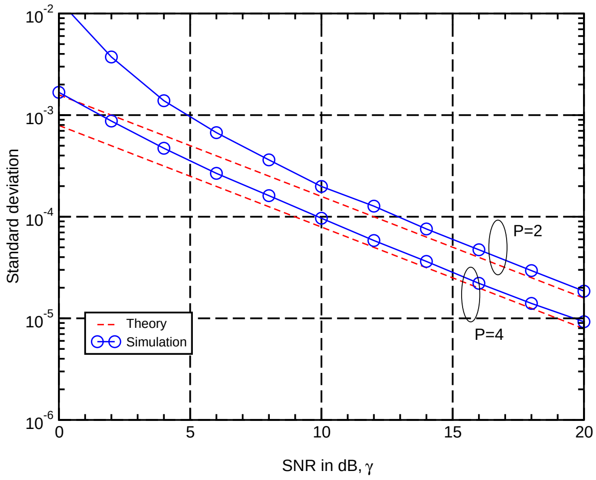

Fig. 1. Standard deviation of the frequency estimates versus SNR; $N_u = 1$

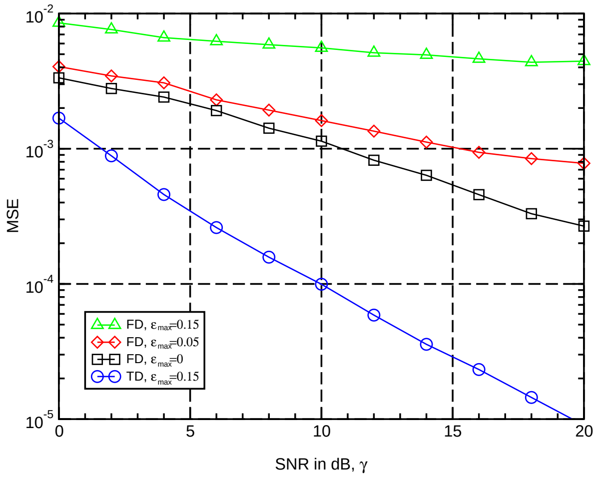

Fig. 2. MSE of the frequency estimates versus SNR; $P = 4$; $N_u = 1$

rapidly. In contrast, the performance of the TD frequency estimation algorithm remains nearly constant. The main part of MAI comes from neighboring users, which influence can be completely eliminated by the proposed additional spreading in time direction.

V. Conclusions

Two novel frequency offset estimation algorithms for uplink IFDMA systems have been proposed. Both algorithms exploit the knowledge of pilot symbols and allow maximum likelihood estimation of the frequency offset. The TD frequency estimation algorithm does not require exhaustive search and provides higher precision than the FD frequency estimation algorithm. Due to the special construction of the pilot symbols, the proposed TD algorithm is robust to MAI caused by frequency misaligned users. It has been shown that the proposed TD algorithm outperforms the RCFE algorithm proposed in [10]. The TD algorithm can also be applied for OFDMA uplink scheme with block-interleaved frequency allocation.

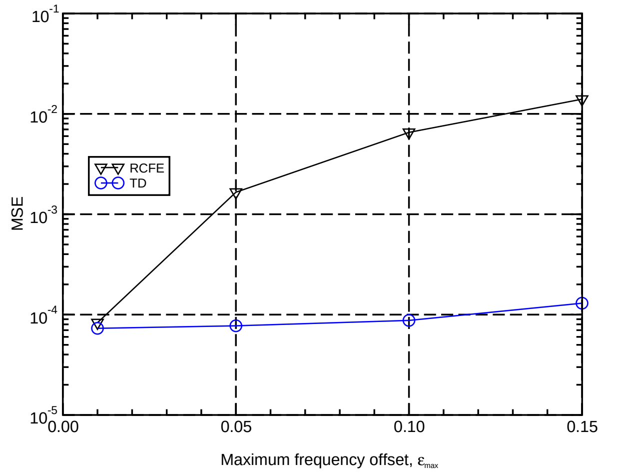

Fig. 3. MSE of the frequency estimates versus maximum frequency offset $\epsilon_{\max}$; SNR=14 dB; $P = 2$; $N_u = 8$

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