11th International OFDM-Workshop (InOWo), Aug 2006, Hamburg, Germany

Channel Estimation for DFT precoded OFDMA with blockwise and interleaved subcarrier allocation

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Abstract— **In this paper, a channel estimation algorithm in frequency domain for IFDMA is introduced. Moreover, a post processing algorithm for further improvement of the channel estimation is presented. The performance of IFDMA is analyzed and compared to DFT precoded OFDMA with blockwise subcarrier allocation with and without frequency hopping in terms of overhead that occurs due to pilot assisted channel estimation and in terms of performance gains due to frequency diversity. Taking both effects into account, it is shown that IFDMA outperforms DFT precoded OFDMA with blockwise subcarrier allocation even if a frequency hopping scheme is applied.**

I. Introduction

At present, research activities for beyond 3rd generation of mobile radio systems are in progress worldwide. Several multiple access (MA) schemes are under discussion as candidates for future mobile radio systems, including Orthogonal Frequency Division Multiple Access (OFDMA) because of its favourable properties as they have been described e.g. in [1].

Other promising MA schemes result from the application of Discrete Fourier Transform (DFT) precoding to OFDMA. OFDMA with DFT precoding combines the advantages of OFDMA with a low Peak-to-Average Power Ratio (PAPR) of the transmit signal [2], [3]. In this work, the focus will be on DFT-precoded OFDMA with interleaved and blockwise subcarrier allocation. Interleaved subcarrier allocation results in the well known Interleaved Frequency Division Multiple Access (IFDMA) scheme [4],[5],[6], whereas blockwise subcarrier allocation leads to the Single Carrier Frequency Division Multiple Access (SC-FDMA) scheme [8]. SC-FDMA can also be combined with a Frequency Hopping (FH) scheme, which will be denoted as FH-SC-FDMA in the following.

For IFDMA, an efficient pilot assisted channel estimation (CE) algorithm has not been introduced so far. In [7], an iterative time domain approach of joint CE and data detection for an IFDMA system is given. However, for high data rates, increasing number of users and higher order modulation schemes, this algorithm becomes complex and, thus, a less expensive CE is required. Since the IFDMA scheme assigns mutually orthogonal subcarriers to each user that are equidistantly distributed over the total bandwidth, the application of existing pilot assisted CE techniques to IFDMA requires several adjustments. In contrast to DFT-precoded OFDMA systems with blockwise subcarrier allocation, CE for IFDMA in frequency domain cannot be realized by interpolation between scattered samples of pilot symbols. In order to perform pilot assisted CE in frequency domain, for IFDMA the pilot symbols have to be distributed over the total bandwidth and located at the positions of the allocated subcarriers. Furthermore, the advantageous low PAPR of the IFDMA time domain transmit signal should be maintained when inserting a pilot sequence and the CE algorithm should be appropriate to support the transmission of variable data rates.

In this paper, a pilot assisted CE for IFDMA in frequency domain is introduced. A post processing algorithm for further improvement of CE performance is described. Moreover, IFDMA, SC-FDMA and FH-SC-FDMA is compared concerning the loss due to CE overhead and the performance gain due to frequency diversity. An overall tradeoff comparing the three MA schemes is given.

The paper is organized as follows. In section II, the system model is described. In section III, a pilot-assisted CE algorithm for IFDMA is introduced. In section IV, the CE overhead for IFDMA, SC-FDMA and FH-SC-FDMA is calculated. In section V, the performance results for IFDMA with perfect and imperfect channel knowledge are discussed. Moreover, the performance and overhead results for IFDMA, SC-FDMA and FH-SC-FDMA as well as the overall tradeoff are given. Section VI concludes this work.

II. System Model

In this section, a system model for IFDMA and SC-FDMA will be derived from the general description of the DFT-precoded OFDMA system model.

In the following, all signals are represented by their discrete time equivalents in the complex baseband. Further on, $(\cdot)^T$ denotes the transpose and $(\cdot)^H$ the Hermitian of a vector or a matrix. Assuming a system with K users, let

$$\mathbf{d}^{(k)} = (d_0^{(k)}, \cdots, d_{Q-1}^{(k)})^T \quad (1)$$

denote a block of Q data symbols $d_q^{(k)}, q = 0, \cdots, Q-1$, at symbol rate $1/T_s$ transmitted by a user with index k, $k = 0, \cdots, K-1$. The data symbols $d_q^{(k)}$ can be taken from the alphabet of a modulation scheme like Phase Shift Keying (PSK), that is applied to coded or uncoded bits. Let $\mathbf{F}_N$ and $\mathbf{F}_N^H$ denote the matrix representation of an N-point DFT and an N-point Inverse DFT (IDFT) matrix, respectively, where $N = K \cdot Q$ is the number of available subcarriers in the system. The assignment of the data symbols $d_q^{(k)}$ to the user specific set of Q subcarriers can be described by a Q-point DFT precoding matrix $\mathbf{F}_Q$, a $N \times Q$ mapping matrix $\mathbf{M}^{(k)}$ and an N-point IDFT matrix $\mathbf{F}_N^H$ [14]. Thus, a precoded OFDMA signal at chip rate $1/T_c = K/T_S$ is given by

$$\mathbf{x}^{(k)} = \mathbf{F}_N^H \cdot \mathbf{M}^{(k)} \cdot \mathbf{F}_Q \cdot \mathbf{d}^{(k)}. \quad (2)$$

The insertion of a Cyclic Prefix, as well as the transmission over a channel and subsequent demodulation is given in [14] and will not be described in this work.

A. IFDMA

To derive a system model for IFDMA as a special case of the introduced DFT-precoded OFDMA system model, the mapping matrix has to perform an interleaved subcarrier allocation. The IFDMA mapping matrix $\mathbf{M}_I^{(k)}$ is given by its elements $M_I^{(k)}(n,q)$, with $n = 0, \cdots, N-1$, and $q = 0, \cdots, Q-1$.

$$M_I^{(k)}(n,q) = \begin{cases} 1 & n = q \cdot K + k \\ 0 & \text{else} \end{cases} . \quad (3)$$

Therefore, the IFDMA transmit signal $\mathbf{x}_I^{(k)}$ becomes

$$\mathbf{x}_I^{(k)} = \mathbf{F}_N^H \cdot \mathbf{M}_I^{(k)} \cdot \mathbf{F}_Q \cdot \mathbf{d}^{(k)}. \quad (4)$$

B. SC-FDMA

A system model for SC-FDMA can be derived as a special case of the DFT-precoded OFDMA system model with blockwise subcarrier allocation. The elements $M_B^{(k)}(n,q)$ with $n = 0, \cdots, N-1$ and $q = 0, \cdots, Q-1$ of the mapping matrix $\mathbf{M}_B^{(k)}$ are given by

$$M_B^{(k)}(n,q) = \begin{cases} 1 & n = k \cdot Q + q \\ 0 & \text{else} \end{cases} . \quad (5)$$

Therefore, the SC-FDMA transmit signal $\mathbf{x}_B^{(k)}$ becomes

$$\mathbf{x}_B^{(k)} = \mathbf{F}_N^H \cdot \mathbf{M}_B^{(k)} \cdot \mathbf{F}_Q \cdot \mathbf{d}^{(k)}. \quad (6)$$

III. Channel Estimation for IFDMA

In this section, a pilot-assisted CE algorithm for IFDMA in frequency domain is introduced. For SC-FDMA, CE is already solved, cf. e.g. [12]. Therefore, it will not be considered in this section. To apply pilot assisted CE to IFDMA, several adjustments have to be made due to its interleaved subcarrier allocation.

The proposed CE for IFDMA is performed by replacing one block $\mathbf{x}_I^{(k)}$ of the IFDMA time-domain signal by an appropriate pilot sequence

$$\mathbf{p}_I^{(k)} = (p_0^{(k)}, \cdots, p_{N-1}^{(k)})^T. \quad (7)$$

While inserting $\mathbf{p}_I^{(k)}$ in the IFDMA signal, the orthogonality to the signals of other users has to be maintained. The pilots have to be located at the same subcarriers that are allocated to a certain user. Thus, the pilot sequence $\mathbf{p}_I^{(k)}$ can be built in time domain in analogy to a DFT precoded OFDMA signal with interleaved subcarrier allocation, as given in (4). Thus, the transmitted pilot sequence $\mathbf{p}_I^{(k)}$ is given by

$$\mathbf{p}_I^{(k)} = \mathbf{F}_N^H \cdot \mathbf{M}_I^{(k)} \cdot \mathbf{F}_Q \cdot \tilde{\mathbf{p}}^{(k)}, \quad (8)$$

where

$$\tilde{\mathbf{p}}^{(k)} = (\tilde{p}_0^{(k)}, \cdots, \tilde{p}_{Q-1}^{(k)})^T \quad (9)$$

denotes the pilot sequence before IFDMA-modulation. The unmodulated pilot symbols $\tilde{p}_q^{(k)}, q = 0, \cdots, Q-1$, are complex values. In this work, a Constant Amplitude Zero Autocorrelation (CAZAC) sequence [10], that is built of complex values with equal amplitudes, is used as pilot sequence. The CAZAC sequences exhibit a constant amplitude at the transmitter in time domain, which is desirable for the insertion into an IFDMA signal, and a constant Power Density Spectrum (PDS), which is preferable for CE in frequency domain.

The frequency domain representation of the modulated pilot sequence $\mathbf{p}_I^{(k)}$ of user k is given by its N-point DFT

$$\mathbf{P}_I^{(k)} = \mathbf{F}_N \cdot \mathbf{p}_I^{(k)} = (P_0^{(k)}, \cdots, P_{N-1}^{(k)})^T \quad (10)$$

with the non-zero elements $P_{(n\, modulo\, Q) \cdot K + k}^{(k)}$, $n = 0, \cdots, N-1$, of $\mathbf{P}_I^{(k)}$. $K = N/Q$ denotes the total number of users. Due to the IFDMA-modulation, the spectral components of the pilot sequence $\mathbf{p}_I^{(k)}$ are allocated to the same set of evenly distributed subcarriers as the spectral components of the IFDMA transmit signal $\mathbf{x}_I^{(k)}$ in (4) and the orthogonality of the K users' signals is maintained. In the following, only one user will be considered and the index k will be omitted for simplicity. Let

$$\mathbf{h} = (h_0, \cdots, h_{N-1})^T \quad (11)$$

denote the vector representation of a channel with N coefficients h_i, $i = 0, \cdots, N-1$, at chip rate $1/T_c$. The channel is assumed to be time invariant for the transmission of the pilot sequence $\mathbf{p}_I$. The values V_n with $n = 0, \cdots, N-1$, received on each subcarrier after transmission over the channel $\mathbf{h}$ can be described by one complex channel coefficient H_n due to flat fading on each subcarrier in frequency domain and are given by

$$V_n = H_n \cdot P_n + \check{N}_n. \quad (12)$$

The values H_n denote the complex coefficients of the Channel Transfer Function (CTF)

$$\mathbf{H} = \mathbf{F}_N \cdot \mathbf{h} = (H_0, \cdots, H_{N-1})^T \quad (13)$$

and

$$\check{\mathbf{N}} = \mathbf{F}_N \cdot \mathbf{n} = (\check{N}_0, \cdots, \check{N}_{N-1})^T \quad (14)$$

denotes the Additive White Gaussian Noise (AWGN) on each subcarrier in frequency domain. At the non-zero samples $P^{(k)}_{(n\, modulo\, Q)\cdot K+k}$ of $\mathbf{P}_I$ with $n = 0; \cdots, N-1$, the channel transfer coefficients H_n can be estimated by

$$\hat{H}_n = \frac{V_n}{P_n} = H_n + \frac{\check{N}_n}{P_n} = H_n + E_n, \quad (15)$$

where $\hat{H}_n$ denotes the estimate of the channel transfer coefficient H_n and E_n the estimation error. In order to apply frequency domain equalization to the IFDMA signal, only the coefficients H_n have to be estimated that correspond to the actually allocated subcarriers of a certain user. Thus, the total number of estimates $\hat{H}_n$ is Q and is dependent on the data rate provided to the user. The lower the data rate, the smaller is Q and the less coefficients need to be estimated per user.

The estimation error E_n occurring on every estimated coefficient $\hat{H}_n$ increases the mean square error (MSE) of the estimated data symbols at the receiver. The increasing MSE can be described by a degradation d_{CE} of the Signal-to-Noise-Ratio (SNR) [11]

$$d_{CE} = 10 \cdot \log_{10}\left(\frac{\tilde{\text{MSE}}}{\text{MSE}}\right), \quad (16)$$

where $\tilde{\text{MSE}}$ is the data estimation error for the imperfect channel estimation and MSE is the data estimation error for perfect channel knowledge. In [11], it has been shown for a Code Division Multiple Access (CDMA) system that the SNR-degradation d_{CE} approaches a value of $d_{CE_b} = 3$ dB for decreasing bit error rate. This result also applies for the proposed CE algorithm.

In the following, the channel $\mathbf{h}$ in (11) is assumed to consist of $L < N$ non-zero coefficients h_i. To improve the CE performance and, thus, decrease the SNR-degradation d_{CE_b}, a Post Processing (PP) algorithm can be applied to the time domain representation $\hat{\mathbf{h}} = \mathbf{F}_N^H \cdot \hat{\mathbf{H}} = (\hat{h}_0, \cdots, \hat{h}_{N-1})^T$ of the estimated CTF

$$\hat{\mathbf{H}} = (\hat{H}_0, \cdots, \hat{H}_{N-1})^T \quad (17)$$

with the non-zero elements $\hat{H}_{(n\, modulo\, Q)\cdot K+k}$, $n = 0, \cdots, N-1$. $\hat{\mathbf{h}}$ is denoted as estimated Channel Impulse Response (CIR) in the following. For PP a threshold-test [11] is introduced that is used to generate the post processed CIR given by

$$\tilde{\mathbf{h}} = (\tilde{h}_0, \cdots, \tilde{h}_{N-1})^T. \quad (18)$$

The elements $\tilde{h}_n$, $n = 0, \cdots, N-1$, of the post processed CIR are formed by the following rule:

$$\tilde{h}_n = \begin{cases} \hat{h}_n & \text{if} \quad |\hat{h}_n| \geq \lambda \\ 0 & \text{else} \end{cases} . \quad (19)$$

The threshold λ can be deduced from [11] and is given by

$$\lambda = M \cdot \sqrt{\frac{\sigma_n^2}{E_P}} = 2 \cdot \frac{Q}{N} \cdot \sqrt{\frac{\sigma_n^2}{E_P}}, \quad (20)$$

where σ_n^2 denotes the variance of the White Gaussian Noise $\mathbf{n}$ and E_P is given by

$$E_P = \frac{1}{N}\sum_{n=0}^{N-1} |P_n|^2 \quad (21)$$

The factor M in (20) is adapted to the particular value of Q, because the estimated CTF $\hat{\mathbf{H}}$ of length N consists of Q non-zero elements. Therefore, the energy of the estimated CIR $\hat{\mathbf{h}}$ is reduced by factor $\frac{Q}{N}$ with decreasing Q and the threshold λ has to be also reduced with decreasing Q. The additional factor 2 in (20) has been derivated heuristically.

IV. Channel Estimation Overhead

In this section, the overhead that occurs due to CE will be derived for DFT-precoded OFDMA with interleaved and blockwise subcarrier allocation.

The pilot arrangement in a frame of N_S precoded OFDMA symbols x_i, $i = 0, \cdots, N_S - 1$, of one user in time domain and N subcarriers in frequency domain is given by applying the two-dimensional sampling theorem [10]. With τ_{max} the maximum delay of the CIR, $f_{D,max}$ the maximum Doppler-shift, T the symbol duration of one precoded OFDMA symbol and B_Δ the subcarrier spacing, the pilot distance in frequency direction is given by [12],[13]

$$D_f < \frac{1}{\tau_{max} \cdot B_\Delta} \quad (22)$$

and in time direction by

$$D_t < \frac{1}{2 \cdot f_{D,max} \cdot T} \quad (23)$$

Practical values for D_f and D_t are selected by an over-sampling with factor $P_f = 5$ in frequency and with factor $P_t = 5$ in time domain. With the coherence bandwidth $B_k = 1/\tau_{max}$ and the coherence time $T_k = 1/(2f_{D,max})$, the pilot distances are given by

$$D_f = \frac{B_k}{P_f \cdot B_\Delta} = \frac{B_k}{5 \cdot B_\Delta} \quad (24)$$

and

$$D_t = \frac{T_k}{P_t \cdot T} = \frac{T_k}{5 \cdot T}. \quad (25)$$

The overall CE overhead Λ is given by [10]

$$\Lambda = 10 \cdot \log_{10}\left(\frac{N \cdot N_S}{N \cdot N_S - N_t \cdot N_f}\right), \quad (26)$$

with N_f the number of required pilots in frequency direction and N_t the number of required pilots in time direction of N overall symbols in frequency direction and N_S overall symbols in time direction, respectively. Due to pilot assisted CE, the energy which has to be spent per data bit is increased relatively by the value of Λ for achieving a certain performance at the receiver. Therefore, the CE overhead Λ is equivalent to an SNR-degradation. For IFDMA, interpolation between the pilot positions in frequency domain is not possible as long as $K \cdot B_\Delta \geq B_k/5$. For low data rates, a pilot has to be transmitted on

TABLE I

SIMULATION PARAMETERS

Carrier Frequency	5 GHz
Bandwidth	20 MHz
No. of Subcarriers	512
Modulation	QPSK
Guard Interval	7 μs
Channel	WINNER SCM, Urban Macro
Velocity	70 km/h

each subcarrier. Dependent on the data rate, the number N_{fI} of pilots in frequency domain for IFDMA is given by

$$N_{fI} = \begin{cases} Q & \text{if} \quad Q \leq \frac{N}{D_f} \\ \lceil \frac{N}{D_f} \rceil & \text{else} \end{cases} \tag{27}$$

where $\lceil x \rceil$ denotes the smallest integer greater than or equal to x.
The number N_{fS} of pilots in frequency domain for SC-FDMA can be calculated by

$$N_{fB} = \lceil \frac{Q}{D_f} \rceil, \tag{28}$$

because interpolation is possible for all data rates.
Regarding the number N_t of pilots in time direction, interpolation is possible for both subcarrier allocations and given by

$$N_t = \lceil \frac{N_S}{D_t} \rceil \tag{29}$$

In contrast to SC-FDMA, IFDMA suffers from increased CE overhead as a consequence of missing interpolation in frequency domain. However, a well known advantage of IFDMA compared to SC-FDMA is its higher frequency diversity leading to a better BER performance. There is a tradeoff between increasing SNR-degradation due to CE overhead on the one hand and decreasing SNR-degradation due to frequency diversity on the other hand. The frequency diversity of SC-FDMA can be improved by applying frequency hopping. However, for FH-SC-FDMA, the CE overhead in time direction increases, since interpolation in time direction is impossible as long as the number of possible hops, which is dependent on Q, is higher than the number of concerned OFDMA symbols. In frequency direction, FH-SC-FDMA shows the same possibility of interpolation as SC-FDMA.

V. PERFORMANCE ANALYSIS

In this section, the performance of the CE algorithm proposed in section III for IFDMA will be given. Moreover, the tradeoff between loss due to CE overhead and frequency diversity gain will be analyzed for IFDMA, SC-FDMA and FH-SC-FDMA. The simulation parameters are given in Table I.

A. Channel Estimation for IFDMA

In the following, the uncoded BER performance when using the CE algorithm proposed in section III is discussed for $Q = 256$ subcarriers per user, cf. Fig. 1. $Q = 256$ corresponds to a net bit rate of 20 Mbit/s. Equalization is performed in frequency domain with a

TABLE II

SNR-DEGRADATION AFTER APPLICATION OF PP

No. Q of Subcarriers per User	512	256	128	64
SNR-degradation d_{CE} in dB	1.1	1.7	2.5	3

Frequency Domain Equalizer (FDE) based on the zero-forcing criterion. For an increasing E_S/N_0, i.e. symbol energy over noise power, the SNR-degradation between Perfect Channel Knowledge (PCK) and CE without PP decreases to 3 dB as it has been described in section III and in [11]. The SNR-degradation d_{CE} after application of PP is given in Table II for different values of Q for $BER = 10^{-2}$. The lower the value of Q, i.e. the lower the data rate, the smaller the improvement by the use of the proposed PP. The reason for this is that the estimated CIR $\hat{\mathbf{h}}$ results from an N-point IDFT of Q sampling points in frequency domain spread over N subcarriers. The time domain representation shows an $\frac{N}{Q}$-fold repetition of the sampled estimate. The lower Q, the more repetitions occur and the less CIR coefficients can be discarded by applying PP. Similar SNR-degradations as for IFDMA appear for SC-FDMA with and without FH when pilot assisted CE is applied.

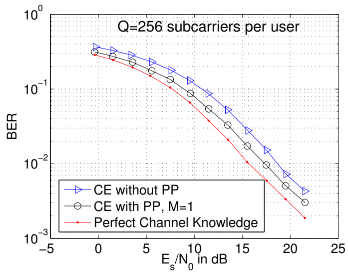

Fig. 1. Uncoded BER performance for CE with and without PP

B. Channel Estimation Overhead vs. Frequency Diversity

In the following, the results for the overhead calculations as described in section IV are given for IFDMA, SC-FDMA and FH-SC-FDMA. Further on, the tradeoff between loss due to CE overhead and frequency diversity gain is discussed for single transmit and single receive antennas. In Fig. 2 the CE overhead Λ is given in dependency of the number Q of allocated subcarriers per user. For overhead calculations, a maximum channel delay of $\tau_{max} = 2.56$ μs and a maximum Doppler frequency of $f_{D,max} = 290$ Hz is assumed. The oversampling is chosen as described in (24) and (25), so that one pilot symbol is transmitted per 24 symbols in time direction and one pilot per 2 adjacent subcarriers in frequency direction. Regarding the loss due to CE overhead, it becomes obvious that IFDMA loses 0.095 dB compared to SC-FDMA for net bit rates between 156 kbit/s and 20 Mbit/s. For Q=512, interpolation can be applied to IFDMA and both schemes show the same loss due to CE overhead. For FH-SC-FDMA, the number of possible hops increases with decreasing values of Q. FH-SC-FDMA loses about 0.6 dB compared to IFDMA for a net bit rate of 1.25 Mbit/s. For a net bit rate of 5 Mbit/s, both schemes show the same loss due to CE overhead. In Fig. 4 and Fig. 5, the coded BER is given as a

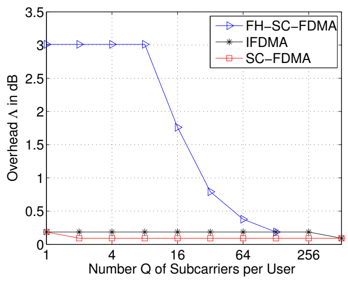

Fig. 2. CE Overhead Λ

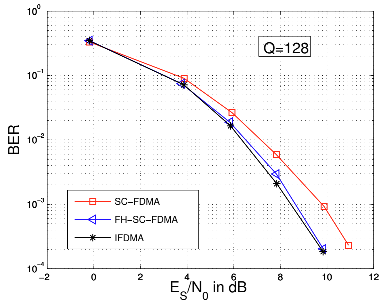

Fig.4. Performance Results for 5 Mbit/s

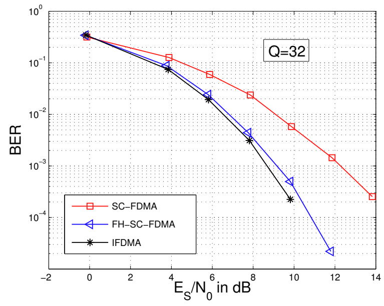

Fig.5. Performance Results for 1.25 Mbit/s

TABLE III
COMPARISON IFDMA VS. SC-FDMA/FH-SC-FDMA

Net Bit Rate of 1.25 Mbit/s		
MA scheme	IFDMA/SC-FDMA	IFDMA/FH-SC-FDMA
CE Overhead Λ in dB	-0.095 dB	$+0.6$ dB
Frequency Diversity Gain in dB	$+3.4$ dB	$+0.4$ dB
Overall Gain in dB	$+3.305$ dB	$+1.0$ dB

TABLE IV
COMPARISON IFDMA VS. SC-FDMA/FH-SC-FDMA

Net Bit Rate of 5 Mbit/s		
MA scheme	IFDMA/SC-FDMA	IFDMA/FH-SC-FDMA
CE Overhead Λ in dB	-0.095 dB	$+0.0$ dB
Frequency Diversity Gain in dB	$+1.5$ dB	$+0.2$ dB
Overall Gain in dB	$+1.405$ dB	$+0.2$ dB

function of E_S/N_0. The results show performance gains due to frequency diversity and are valid for PCK as the frequency diversity gain is independent of CE algorithms. A convolutional code with code rate $1/2$ is used and the equalization is performed with an FDE corresponding to the Minimum Mean Square Error criterion. In Table III and IV, the pilot overhead Λ, the frequency diversity gain and the overall gain is given for IFDMA related to the loss due to CE overhead and frequency diversity gain for SC-FDMA and FH-SC-FDMA. Regarding the overall gain, it can be seen that IFDMA shows an overall gain compared to SC-FDMA and FH-SC-FDMA for both bit rates. The highest gain of 3.3 dB occurs for a net bit rate of 1.25 Mbit/s in comparison with SC-FDMA, because IFDMA benefits from its spread subcarriers that lead to higher frequency diversity. The lower the bit rate, the higher the benefit due to frequency diversity of IFDMA compared to SC-FDMA. Thus, the overall gain of IFDMA compared to SC-FDMA increases with decreasing Q.

VI. CONCLUSIONS

In this paper, a pilot assisted CE for IFDMA in frequency domain has been introduced. It has been shown that pilot transmission is solely necessary at the Q subcarriers that are allocated to a certain user and that the performance of the proposed CE algorithm can be improved by applying time domain PP with a threshold-test. Moreover, the CE overhead for IFDMA has been compared to the CE overhead for SC-FDMA and FH-SC-FDMA. Finally, regarding CE overhead and frequency diversity, IFDMA shows an overall performance gain compared to SC-FDMA and FH-SC-FDMA. The loss due to higher CE overhead for IFDMA is overcompensated by the performance gain due to frequency diversity. Thus, IFDMA can be regarded as a promising candidate for future mobile radio systems.

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