LOW COMPLEXITY MULTI CARRIER MULTIPLE ACCESS WITH CYCLIC DELAY DIVERSITY

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ABSTRACT

In this paper, the performance of Discrete Fourier Transform (DFT) precoded Orthogonal Frequency Division Multiple Access (OFDMA) schemes with interleaved and with blockwise subcarrier allocation is investigated and compared. Transmission with one transmit antenna is compared to transmission with multiple transmit antennas using Cyclic Delay Diversity (CDD). It is shown that the investigated DFT precoded OFDMA schemes with CDD provide low computational complexity and good power efficiency due to low envelope fluctuations of the transmit signal at each transmit antenna. Moreover, it is shown that, depending on the data rate, for one transmit antenna interleaved subcarrier allocation results in a significant performance gain compared to block allocation, whereas for multiple antennas this gain reduces. However, it is shown that even for high data rates DFT precoded OFDMA with interleaved subcarrier allocation is shown to outperform the scheme with block allocation.

1. INTRODUCTION

Presently, transmission schemes for beyond 3rd generation mobile radio systems are under research worldwide. Future mobile radio systems have to meet challenging requirements such as, on the one hand, good performance, high spectral efficiency, high flexibility and sufficient granularity in terms of different data rates. On the other hand, especially in the uplink, low cost terminals and, thus, schemes providing low computational complexity and a low Peak-to-Average Power Ratio (PAPR) of the transmit signal are desirable.

Promising and well known multiple access schemes result from the application of Discrete Fourier Transform (DFT) precoding to Orthogonal Frequency Division Multiple Access (OFDMA) in combination with equidistant subcarrier allocation [1]. The resulting schemes provide low computational complexity for signal generation, user separation and channel equalization even for high data rates and a low PAPR of the transmit signal. The advantages of the schemes are obtained essentially at the expense of flexibility, e.g., in terms of the support of different data rates. However, in [2] a method is introduced that provides sufficient flexibility in terms different data rates also for systems with equidistant subcarrier allocation [3].

Common equidistant subcarrier allocation schemes are, e.g., interleaved subcarrier allocation, where the subcarriers assigned to a specific user are equidistantly distributed over the total available bandwidth, and block allocation, where a set of adjacent subcarriers is assigned to each user. The application of DFT-precoding to OFDMA with interleaved subcarrier allocation (I-OFDMA) results in the well known Interleaved Orthogonal Frequency Division Multiple Access (IFDMA) scheme, cf. [4–6]. In addition to the already mentioned advantages of DFT precoded OFDMA, IFDMA provides high frequency diversity due to the interleaved subcarrier allocation. The application of DFT precoding to OFDMA with block allocation (B-OFDMA) results in a scheme designated as Single Carrier Frequency Division Multiple Access with localized mapping [7], in the following denoted as SC-FDMA. Compared to IFDMA, SC-FDMA suffers from lower frequency diversity, especially for moderate to low data rates, but is more robust to frequency offsets due to block allocation of the subcarriers.

An adequate method to further improve the performance of a system and, thus, to make reliable high data rate services possible is the use of multiple antennas. A simple and well known multi-antenna scheme is Cyclic Delay Diversity (CDD), cf. [8–10] and references therein. For application of CDD, only low computational complexity is required, the receiver can be left unchanged and there is no need for Multiple-Input Multiple-Output (MIMO) channel estimation [11]. Moreover, in contrast to, e.g., the Alamouti scheme [12], for CDD there is no rate loss for more then 2 transmit antennas, the number of transmit antennas is arbitrary and there is no quasi stationary channel assumption for transmission of a specific number of symbols.

In this paper, the combination of CDD with IFDMA and

SC-FDMA, respectively, is investigated. CDD converts spatial diversity into additional frequency diversity. Especially for moderate to low data rates, SC-FDMA provides only low frequency diversity and, thus, a significant gain due to additional frequency diversity provided by CDD can be expected. For IFDMA, which provides higher frequency diversity compared to SC-FDMA, the performance gain due to additional frequency diversity is expected to be lower. This tradeoff is investigated in the following. Different data rates are considered and the results obtained for CDD are compared to results obtained for application of space-time block codes. Moreover, it is shown that the investigated schemes provide low computational complexity, especially at the transmitter, and good power efficiency due to low envelope fluctuations of the signals at each transmit antenna.

The paper is organized as follows: In Section 2, a system model for DFT precoded OFDMA, IFDMA and SC-FDMA is introduced and in Section 3, the application of CDD to IFDMA and SC-FDMA and is described. In Section 4, a design rule for the cyclic delay for blockwise and interleaved subcarrier distribution is described. Finally, in Section 5 the performance results for coded IFDMA and SC-FDMA transmission with and without multiple transmit antennas over uncorrelated spatial mobile radio channels are discussed for different data rates.

2. SYSTEM MODEL

In this section, system models for IFDMA and SC-FDMA are derived as special cases of a system model for DFT precoded OFDMA. In the following, all signals are represented by their discrete time equivalents in the complex baseband. We assume a system with K users. With $(\cdot)^{\mathrm{T}}$ denoting the transpose of a vector, let

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

designate a block of Q data symbols $d_q^{(k)}$, $q = 0, \ldots, Q-1$, at symbol rate $1/T_s$ transmitted by a user with index k, $k = 0, \ldots, K-1$. The data symbols $d_q^{(k)}$ may result from application of a mapping scheme like Phase Shift Keying (PSK) or Quadrature Amplitude Modulation (QAM) to either Forward Error Control (FEC) coded or to uncoded data bits. Let further $\mathbf{F}_N$ and $\mathbf{F}_N^H$ denote the matrix representation of an N-point DFT and an N-point Inverse DFT (IDFT) matrix, respectively, where $N = K \cdot Q$ denotes the number of subcarriers available in the system and $(\cdot)^H$ denotes the Hermitian of a matrix. The assignment of the data symbols $d_q^{(k)}$ to a user specific set of Q subcarriers is assumed to be described by an $N \times Q$ mapping matrix $\mathbf{M}^{(k)}$. Thus, a precoded OFDMA signal

$$\mathbf{x}^{(k)} = (x_0^{(k)}, \ldots, x_{N-1}^{(k)})^{\mathrm{T}} \quad (2)$$

with elements $x_n^{(k)}$, $n = 0, \ldots, N-1$, at chip rate $1/T_c = K/T_s$ is given by

$$\mathbf{x}^{(k)} = \mathbf{F}_N^H \cdot \mathbf{M}^{(k)} \cdot \mathbf{F}_Q \cdot \mathbf{d}^{(k)}, \quad (3)$$

where $\mathbf{F}_Q$ denotes the Q-point DFT precoding matrix. In the following, we consider an uplink scenario. Let

$$\mathbf{h}^{(k)} = (h_0^{(k)}, \ldots, h_{L-1}^{(k)}, 0, \ldots, 0)^{\mathrm{T}} \quad (4)$$

designate an $N \times 1$ vector representation of a channel with L non-zero coefficients $h_l^{(k)}$, $l = 0, \ldots, L-1$, at chip rate $1/T_c$ and $L \leq N$. For the time interval T required for transmission of a modulated version $\mathbf{x}^{(k)}$ of the data block $\mathbf{d}^{(k)}$ the channel is assumed to be time invariant. Before transmission over the channel $\mathbf{h}^{(k)}$, a Cyclic Prefix (CP) is applied to $\mathbf{x}^{(k)}$ which is removed at the receiver before demodulation. It is well known that insertion of the CP, transmission over the channel and removal of the CP at the receiver can be described by an equivalent $N \times N$ circulant channel matrix

$$\mathbf{H}^{(k)} = \begin{pmatrix} h_0^{(k)} & \ldots & h_{L-1}^{(k)} & 0 & \ldots & \ldots & 0 \\ 0 & h_0^{(k)} & \ldots & h_{L-1}^{(k)} & 0 & \ldots & 0 \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ h_1^{(k)} & \ldots & h_{L-1}^{(k)} & 0 & \ldots & 0 & h_0^{(k)} \end{pmatrix}^{\mathrm{T}} \quad (5)$$

[13]. Hence, the signal of user k after transmission over the equivalent circulant channel is given by

$$\mathbf{y}^{(k)} = \mathbf{H}^{(k)} \cdot \mathbf{x}^{(k)}. \quad (6)$$

Let

$$\mathbf{n} = (n_0, \ldots, n_{N-1})^{\mathrm{T}} \quad (7)$$

designate an Additional White Gaussian Noise (AWGN) vector with samples n_n, $n = 0, \ldots, N-1$ at chip rate $1/T_c$. The received signal after removal of the cyclic prefix is given by the superposition of the signals of the K users transmitted over the user specific equivalent circulant channels $\mathbf{H}^{(k)}$ and distorted by AWGN according to

$$\mathbf{r} = \sum_{k=0}^{K-1} \mathbf{y}^{(k)} + \mathbf{n} = \sum_{k=0}^{K-1} \mathbf{H}^{(k)} \cdot \mathbf{x}^{(k)} + \mathbf{n}. \quad (8)$$

In the following, we regard the user specific part for user k of a receiver for DFT precoded OFDMA in the uplink. Firstly, for OFDMA demodulation, an N-point DFT is applied to the received signal $\mathbf{r}$. Secondly, the demodulated signal is user specifically demapped. The demapping matrix is given by the $Q \times N$ matrix ${\mathbf{M}^{(k)}}^{\dagger}$, where $(\cdot)^{\dagger}$ denotes the pseudo-inverse of a matrix. Furthermore, we assume that after demapping, the impact of the channel for user k is compensated by application of a Frequency Domain Equalizer (FDE) as described in [14], [15] e.g., based on the Minimum Mean Square Error

(MMSE) criterion. Let $\mathbf{E}^{(k)}$ denote the $Q \times Q$ matrix representation of the FDE. Finally, the DFT precoding has to be compensated by a Q-point IDFT $\mathbf{F}_Q^H$. Thus, at the receiver, an estimate $\hat{\mathbf{d}}^{(k)}$ of the transmitted data block $\mathbf{d}^{(k)}$ for user k is given by

$$\hat{\mathbf{d}}^{(k)} = \mathbf{F}_Q \cdot \mathbf{E}^{(k)} \cdot \mathbf{M}^{(k)\dagger} \cdot \mathbf{F}_N \cdot \mathbf{r}. \tag{9}$$

2.1. IFDMA

In the following, a system model for IFDMA is derived as a special case of the system model for DFT precoded OFDMA introduced above. As already described, for IFDMA we assume interleaved subcarrier mapping, i.e., the elements $M_I^{(k)}(n,q)$ in the n-th row, $n = 0, \dots, N-1$, and q-th column, $q = 0, \dots, Q-1$, of the mapping matrix $\mathbf{M}_I^{(k)}$ are given by

$$M_I^{(k)}(n,q) = \begin{cases} 1 & n = q \cdot K + k \\ 0 & \text{else} \end{cases}. \tag{10}$$

Thus, the transmit signal $\tilde{\mathbf{x}}_I^{(k)}$ for IFDMA is given by

$$\mathbf{x}_I^{(k)} = \mathbf{F}_N^H \cdot \mathbf{M}_I^{(k)} \cdot \mathbf{F}_Q \cdot \mathbf{d}^{(k)}. \tag{11}$$

Let $\mathbf{I}_Q$ designate a $Q \times Q$ identity matrix. We define an $N \times Q$ repetition matrix $\mathbf{R}$ by stacking K identity matrices according to

$$\mathbf{R} = (\mathbf{I}_Q, \dots, \mathbf{I}_Q)^{\mathrm{T}}. \tag{12}$$

Let us further define a user dependent $N \times N$ rotation matrix

$$\mathbf{\Phi}^{(k)} = \mathrm{diag}(\phi_0^{(k)}, \dots, \phi_{N-1}^{(k)}), \tag{13}$$

where $\mathrm{diag}(\phi_0^{(k)}, \dots, \phi_{N-1}^{(k)})$ denotes a diagonal matrix with diagonal elements $\phi_n^{(k)} = 1/\sqrt{K} \cdot e^{j\frac{2\pi}{N}\cdot k \cdot n}$; $n = 0, \dots, N-1$. It is shown in [6] that the signal from Eq. (11) can be expressed as

$$\mathbf{x}_I^{(k)} = \mathbf{R} \cdot \mathbf{\Phi}^{(k)} \cdot \mathbf{d}^{(k)}. \tag{14}$$

From Eq. (14) follows that the IFDMA time domain signal can be obtained by compression in time and subsequent K-fold repetition of the data block $\mathbf{d}^{(k)}$ of user k and subsequent user specific phase rotation according to matrix $\mathbf{\Phi}^{(k)}$, cf. [6], [5].

In the uplink, the IFDMA receiver is typically implemented in frequency domain. Hence, for IFDMA an estimate $\hat{\mathbf{d}}^{(k)}$ of the transmitted data block $\mathbf{d}^{(k)}$ is given by

$$\hat{\mathbf{d}}^{(k)} = \mathbf{F}_Q^H \cdot \mathbf{E}^{(k)} \cdot \mathbf{M}_I^{(k)\dagger} \cdot \mathbf{F}_N \cdot \mathbf{r}. \tag{15}$$

Note that for IFDMA a FDE based on MMSE criterion provides better performance compared to a FDE based of the zero forcing (ZF) criterion [16], [15], [17].

2.2. SC-FDMA

In the following, a system model for SC-FDMA is derived as a special case of the previously introduced system model for DFT precoded OFDMA. For that purpose, we assume blockwise subcarrier allocation, i.e., the elements $M_B^{(k)}(n,q)$, $n = 0, \dots, N-1$; $q = 0, \dots, Q-1$, of the mapping matrix $\mathbf{M}_B^{(k)}$ are given by

$$M_B^{(k)}(n,q) = \begin{cases} 1 & n = k \cdot Q + q \\ 0 & \text{else} \end{cases}. \tag{16}$$

Thus, the transmit signal for SC-FDMA is given by

$$\mathbf{x}_B^{(k)} = \mathbf{F}_N^H \cdot \mathbf{M}_B^{(k)} \cdot \mathbf{F}_Q \cdot \mathbf{d}^{(k)}. \tag{17}$$

Also for SC-FDMA, the combination of OFDMA modulation, blockwise mapping and DFT precoding can be simplified and signal generation in time domain is given by

$$\mathbf{x}_B^{(k)} = \mathbf{P}^{(k)} \cdot \mathbf{d}^{(k)}, \tag{18}$$

where $\mathbf{P}^{(k)} = \mathbf{F}_N^H \cdot \mathbf{M}_B^{(k)} \cdot \mathbf{F}_Q$ designates a user specific $N \times Q$ matrix which can be calculated offline.

In the uplink, also for SC-FDMA the receiver is implemented in frequency domain. Hence, for SC-FDMA an estimate $\hat{\mathbf{d}}^{(k)}$ of the transmitted data block $\mathbf{d}^{(k)}$ is given by

$$\hat{\mathbf{d}}^{(k)} = \mathbf{F}_Q^H \cdot \mathbf{E}^{(k)} \cdot \mathbf{M}_B^{(k)\dagger} \cdot \mathbf{F}_N \cdot \mathbf{r}. \tag{19}$$

Similar to IFDMA, also for SC-FDMA an MMSE-FDE provides better performance compared to a ZF-FDE.

3. IFDMA AND SC-FDMA WITH CDD

In this section, it is shown how IFDMA and SC-FDMA can be combined with CDD and the signal properties of the resulting schemes are discussed.

We assume a system with n_T transmit antennas with index $i = 1, \dots, n_T$. Let

$$\tilde{\mathbf{x}}^{(k,i)} = (\tilde{x}_0^{(k,i)}, \dots, \tilde{x}_{N-1}^{(k,i)})^{\mathrm{T}}. \tag{20}$$

designate the transmit signal of user k at the i-th transmit antenna for DFT precoded OFDMA according to Eq. (3). According to [8, 9], $\tilde{\mathbf{x}}^{(k,i)}$ can be obtained from the transmit signal of DFT precoded OFDMA from Eq. (3) by application of an antenna specific Cyclic Delay (CD) $\delta_i^{(k)}$. Thus, the elements $\tilde{x}_n^{(k,i)}$, $n = 0, \dots, N-1$; of $\tilde{\mathbf{x}}^{(k,i)}$ are cyclically delayed versions of the elements $x_n^{(k)}$ of the DFT precoded OFDMA signal from Eq. (2) according to

$$\tilde{x}_n^{(k,i)} = x_{(n-\delta_i^{(k)}) \bmod N}^{(k)}, \tag{21}$$

where mod denotes the modulo operation. As described in Section 2.1 and Section 2.2 the corresponding IFDMA and

SC-FDMA signals, respectively, can be obtained by application of either $\mathbf{M}_I^{(k)}$ or $\mathbf{M}_B^{(k)}$ as mapping matrix. Let in extension to Eq. (4)

$$\mathbf{h}^{(k,i)} = (h_0^{(k,i)}, \ldots, h_{L-1}^{(k,i)}, 0, \ldots, 0)^{\mathrm{T}} \tag{22}$$

designate the $N \times 1$ vector representation of the channel from the i-th transmit antenna to the receive antenna. Note that, in general, the number of non-zero channel coefficients of the channels $\mathbf{h}^{(k,i)}$ can be different. According to [8, 9], application of a CD according to Eq. (21) leads to an equivalent Single-Input Single-Output (SISO) channel

$$\tilde{\mathbf{h}}^{(k)} = (\tilde{h}_0^{(k)}, \ldots, \tilde{h}_{(N-1)}^{(k)})^{\mathrm{T}} \tag{23}$$

with elements $\tilde{h}_n^{(k)}$, $n = 0, \ldots, N-1$ given by

$$\tilde{h}_n^{(k)} = \sum_{i=1}^{n_T} h^{(k,i)}_{(n-\delta_i^{(k)}) \bmod N}. \tag{24}$$

Thus, the received signal is given by

$$\mathbf{r} = \sum_{k=0}^{K-1} \tilde{\mathbf{H}}^{(k)} \cdot \mathbf{x}^{(k)} + \mathbf{n}, \tag{25}$$

where $\tilde{\mathbf{H}}^{(k)}$ denotes a circulant channel matrix as defined in Eq. (5) with elements taken from to the equivalent SISO channel defined in Eq. (23).

In the following the signal properties and the computational complexity of the resulting schemes are discussed. Since the CDD scheme can be applied by a simple permutation of the elements of the IFDMA or SC-FDMA transmit signal according to Eq. (21), the transmit signals for IFDMA and SC-FDMA with CDD at each transmit antenna provide the same advantages as the transmit signals without CDD. In particular, according to Eq. (14) and Eq. (18), respectively, both schemes provide low complexity signal generation and a low PAPR due to signal generation in time domain. At the receiver, using the equivalent SISO channel model from Eq. (23), compared to transmission without CDD no modification is required. Moreover, no MIMO channel estimation is required since the equivalent SISO channel can be estimated directly. Thus, as for IFDMA and SC-FDMA transmission in the SISO case, user separation as well as channel equalization can be obtained by application of a Frequency Domain Equalizer (FDE) at the receiver, cf. [8, 9], [15], [17].

The computational complexity in terms of complex multiplications and divisions required for IFDMA and SC-FDMA combined with CDD is given as follows: For IFDMA, at the transmitter N complex multiplications resulting from user specific frequency shift according to Eq. (13) are required per transmit antenna. SC-FDMA signal generation according to Eq. (18) requires $N \cdot Q$ complex multiplications per transmit antenna. It is assumed that for the permutation according to Eq. (21) no computational effort is necessary. At the receiver, in a first step, for both schemes an N-point DFT is required which can be implemented using the Fast Fourier Transform (FFT) algorithm if N is a power of 2. In this case, the computational effort for the DFT is given by $1/2 \cdot N \cdot \mathrm{ld}(N)$ complex multiplications. For determination of the FDE coefficients, for each user Q complex divisions are necessary, cf. [6]. The equalization of the received signal requires Q complex multiplications per user. Finally, for DFT decoding a Q-point IDFT is required. If Q is chosen as a power of 2, the Inverse FFT algorithm (IFFT) can be utilized and for DFT decoding $1/2 \cdot Q \cdot \mathrm{ld}(Q)$ complex multiplications are required per user. The computational effort is summarized in Table 1.

Table 1: Complexity of IFDMA and SC-FDMA

		Multiplications	Divisions
Transmitter	IFDMA	$n_T \cdot N$	-
	SC-FDMA	$n_T \cdot Q \cdot N$	-
Receiver	IFDMA	$1/2\ N\ \mathrm{ld}(N)+$	Q
	SC-FDMA	$Q+1/2Q\ \mathrm{ld}(Q)$	

4. DESIGN OF THE CYCLIC DELAY

In this section, two different design rules for the CD are described and compared to each other. On the one hand, the CD can be designed such that always maximum spatial diversity is exploited, cf. [18]. On the other hand, time-varying CD [19] can be assumed which increases the time diversity of the equivalent SISO channel.

A design rule exploiting maximum spatial diversity for OFDMA systems has been proposed in [18]. In this case, the correlation between channel coefficients related to the subcarriers assigned to a specific user has to be as low as possible. If it is assumed that at the transmitter the number of non-zero channel coefficients of the channel between the i-th transmit antenna and the receive antenna is unknown, for blockwise subcarrier allocation the cyclic delay of the i-th transmit antenna should be designed according to

$$\delta_i^{(k)} = \frac{N}{n_T} + \delta_{i-1}, \quad 1 < i \leq n_t, \tag{26}$$

cf. [18]. A convenient choice for antenna $i = 1$ is $\delta_1 = 0$. For interleaved subcarriers, a design of the cyclic delay according to

$$\delta_i^{(k)} = \frac{N \cdot (i-1)}{n_T \cdot K} \tag{27}$$

is proposed [18]. The design rules from Eq. (26) and Eq. (27) are also valid for SC-FDMA and IFDMA, respectively.

However, in a multi-user system, even for a choice of the CD according to Eq. (26) or Eq. (27) the conventional

Table 2: Simulation parameters

Carrier frequency	5 GHz
Bandwidth	20 MHz
No. of subcarriers	512
Modulation	QPSK
Code	Conv. Code
Code rate	$1/2$
Constraint length	6
Decoder	MaxLogMAP
Equalizer	MMSE FDE
Interleaving	Random
Interl. depth	0.5 ms
Guard interval	5 μs
Channel	WINNER SCM, Urban Macro
Velocity	70 km/h
No. of Tx antennas	1 or 2
No. of Rx antennas	1

CDD scheme is expected not to provide full diversity because some users are expected to suffer from deep fades whereas other users are expected to benefit from good transmission conditions. In this case, the diversity provided by the equivalent MIMO channel can be further improved by time-varying CDD [19]. For that purpose, for each transmitted data block $\tilde{\mathbf{x}}^{(k,i)}$ different CDs are assumed. The resulting time diversity can be exploited by application of a FEC code combined with a bit interleaving over more than one transmitted data block $\tilde{\mathbf{x}}^{(k,i)}$.

5. SIMULATION RESULTS

In this section, the performance of IFDMA and SC-FDMA combined with CDD is investigated and compared for transmission over time dispersive uncorrelated spatial mobile radio channels. The simulation parameters are given in Table 2. In the following, we assume that for IFDMA [SC-FDMA] transmission using the conventional CDD scheme a CD according to Eq. (27) [Eq. (26)] is applied. Alternatively, for both schemes time-varying CDD with random CD is assumed. Transmission with one transmit and one receive antenna (SISO) is compared to transmission with $n_T = 2$ transmit antennas and one receive antenna (MISO).

As already mentioned in Section 1, CDD transforms spatial diversity into additional frequency diversity. It is expected that the benefit of SC-FDMA due to additional frequency diversity provided by CDD is higher than for IFDMA. The reason for this is that, in the SISO case, SC-FDMA provides less frequency diversity compared to IFDMA. Using CDD, additional frequency diversity is provided which can be exploited since utilization of a CD according to Eq. (26) results in weakly correlated channel coefficients of the equivalent channel for adjacent subcarriers [18].

However, for high data rates, not only IFDMA but also SC-FDMA is expected to provide considerable frequency diversity. For that reason, for high data rates in the MISO case the gain of IFDMA compared to SC-FDMA is expected to be rather low. In the following, simulation results for a number of $Q = 128$ subcarriers per user, which corresponds to a net bit rate of 5.1 Mbit/s, are discussed. The following results can be concluded from Fig. 1:

- As expected, at a Bit Error Rate (BER) of 10^{-3} for SC-FDMA combined with conventional CDD a considerable gain of about 1.2 dB compared to the SISO case is obtained.
- The results for SC-FDMA combined with conventional CDD are very similar to the results for SC-FDMA combined with time varying CDD. The reason for this is that for an interleaving depth of 0.5 ms the channel already provides sufficient time diversity so that time varying CDD cannot provide additional gains.
- At a BER of 10^{-3} for IFDMA with conventional CDD a gain of about 0.4 dB compared to the SISO case is obtained. As expected, compared to SC-FDMA, the gain of IFDMA due to exploitation of spatial diversity is lower, because already for the SISO case IFDMA provides significantly higher frequency diversity than SC-FDMA.
- Also for IFDMA, the results for conventional CDD and time varying CDD are very similar.
- Comparison of the overall performance of SC-FDMA and IFDMA shows that for both, conventional CDD and time varying CDD, IFDMA outperforms SC-FDMA by about 0.5 dB because of its better exploitation of frequency diversity. As expected, the gain of IFDMA compared to SC-FDMA for $Q = 128$ is rather small.

In Fig. 2, the performance of IFDMA and SC-FDMA combined with time varying CDD is compared to orthogonal Space-Time Block-Coding (STBC) using the Alamouti scheme [12], [20]:

- At a BER of 10^{-3} the Alamouti scheme outperforms SC-FDMA and IFDMA, both combined with time varying CDD, by about 0.7 dB. The reason for this is that in case of orthogonal STBC an optimum combination of the channel coefficients of the channels between both transmit antennas and the receive antenna is provided, whereas for CDD this combination is not necessarily optimum, cf. [18].

For lower data rates, compared to IFDMA, SC-FDMA provides significantly less frequency diversity. Thus, it is expected that for the MISO case the performance gain of IFDMA compared to SC-FDMA increases. In the following, simulation results for a number of $Q = 32$ subcarriers per user, which corresponds to a net bit rate of 1.2 Mbit/s, are discussed. Similar to the results concluded from Fig. 1, the following results can be concluded from Fig. 3:

- At a BER of 10^{-3} SC-FDMA combined with conventional CDD provides a gain of about 1.3 dB compared to the SISO case.
- SC-FDMA can benefit from the additional time diversity which is provided by time-varying CDD resulting in an additional gain of about 0.2 dB. The reason for that is that for SC-FDMA in case of $Q = 32$ subcarriers the frequency diversity which can exploited in the SISO case is very low. Thus, any additional diversity resultnoticeableticable performance improvement.
- For IFDMA, the gain due to spatial diversity is about 0.3 dB because even for moderate to low data rates the frequency diversity which can be exploited by IFDMA is much higher compared to SC-FDMA.
- Again, for IFDMA time varying CDD does not lead to a noticeable performance improvement.
- In the MISO case, the overall gain of IFDMA compared to SC-OFDMA is about 1.2 dB because of the good exploitation of frequency diversity of IFDMA. As expected, this gain is considerably higher than for $Q = 128$. It is expected that for lower data rates the performance gain of IFDMA compared to SC-FDMA further increases.

A comparison of performance results for CDD and STBC is shown in Fig. 4:

- At a BER of 10^{-3} for SC-FDMA and IFDMA combined with STBC an additional gain of about 0.8 dB compared to combination with time varying CDD can be obtained.

6. REFERENCES

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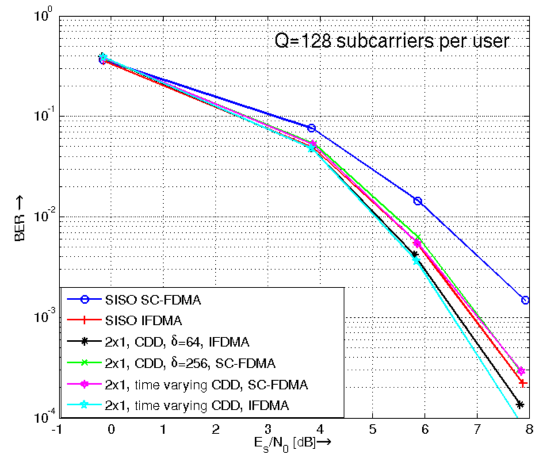

Fig. 1: Coded performance for IFDMA and SC-FDMA, Q=128

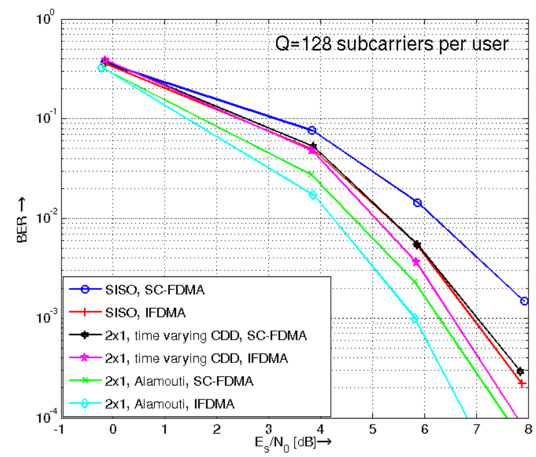

Fig. 2: Coded performance for IFDMA and SC-FDMA, comparison with Alamouti, Q=128

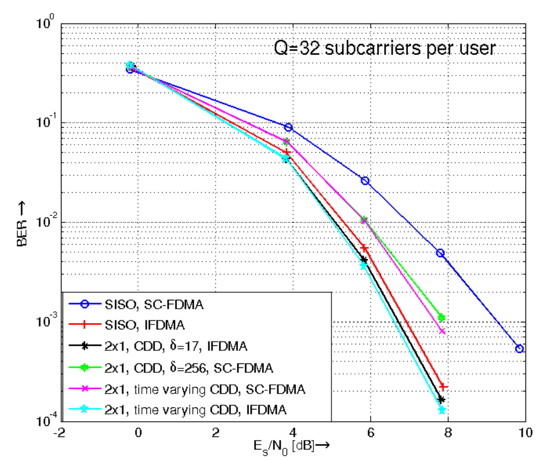

Fig. 3: Coded performance for IFDMA and SC-FDMA, Q=32

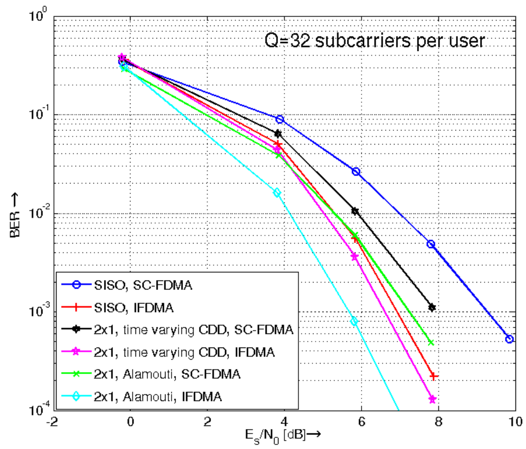

Fig. 4: Coded performance for IFDMA and SC-FDMA, comparison with Alamouti, Q=32

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