

Low Complexity and Power Efficient Space-Time-Frequency Coding for OFDMA

Tobias Frank, Anja Klein, Alexander Kühne
 Darmstadt University of Technology
 Communications Engineering Lab
 Merckstr. 25, 64283 Darmstadt, Germany
 Email: t.frank@nt.tu-darmstadt.de

Elena Costa
 Siemens AG, ICM N PG NT RC FR
 Sankt-Martin Str. 76, 81541 München, Germany
 Email: elena.costa@siemens.com

Abstract—In this paper, a combination of Space-Time Block Coding (STBC) and Discrete Fourier Transform (DFT) frequency coding for OFDMA systems is proposed which provides low computational complexity, low envelope fluctuations of the transmit signals at each transmit antenna and significant performance gains for transmission over mobile radio channels due to exploitation of spatial diversity. The resulting scheme is investigated for blockwise and interleaved subcarrier allocation. The signal properties and the computational complexity are discussed and it is shown that the scheme with interleaved subcarrier allocation outperforms the scheme with block allocation for transmission over spatially correlated and uncorrelated mobile radio channels.

I. 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.

A suitable multiple access scheme providing good coded performance for transmission over mobile radio channels and high flexibility as well as low computational complexity for user separation and channel equalization even for high data rates is Orthogonal Frequency Division Multiple Access (OFDMA). However, OFDMA suffers from a high PAPR of the transmit signal and, thus, requires costly amplifiers [1]. Moreover, in order to exploit frequency diversity additional coding has to be applied.

Other promising and well known multiple access schemes result from the application of Discrete Fourier Transform (DFT) precoding to OFDMA in combination with equidistant subcarrier allocation. The resulting schemes combine the advantages of OFDMA with a low PAPR of the transmit signal due to DFT precoding and computationally efficient implementation due to equidistant subcarrier allocation [2], [3]. 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 [4] a method is introduced that provides sufficient flexibility in terms of different data rates also for systems with equidistant subcarrier allocation.

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. [5], [6] and references therein. 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 but is more robust to frequency offsets due to block allocation of the subcarriers.

A convenient way to provide reliable high data rate services and to further increase the system capacity of a mobile radio system is the exploitation of spatial diversity using multiple antenna techniques such as, e.g., Space-Time coding (STC) [8]. The application of STC to OFDMA is well known [9]. In this paper, we propose a novel space-time-frequency coded OFDMA system that combines Space-Time-Block Coding (STBC) with DFT precoding in frequency direction. The proposed combination is designed such that the resulting space-time-frequency coded OFDMA scheme maintains the advantages of DFT precoded OFDMA and, additionally, provides significant performance improvement due to exploitation of spatial diversity. Because of its signal properties and its good performance, the resulting scheme is shown to be a well suited candidate for future mobile radio applications, especially in the uplink.

The paper is organized as follows: In Section II, a system model is introduced and in Section III, the application of STBC to DFT precoded OFDMA is described. In Section IV, the properties of the resulting systems are discussed and the computational complexity is given. Finally, in Section V performance results for coded transmission over uncorrelated and over correlated spatial mobile radio channel models are discussed and the performance gains for IFDMA and SC-FDMA are compared to each other.

II. SYSTEM MODEL

In this section, a system model for OFDMA is introduced. In the following, all signals are represented by their discrete time equivalents in the complex baseband. We assume a system with

K users. With $(\cdot)^T$ denoting the transpose of a vector, let

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

designate a block of Q data symbols $d_q^{(k)}$, $q = 0, \dots, Q-1$, at symbol rate $1/T_s$ transmitted by a user with index k , $k = 0, \dots, K-1$. 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, the signal

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

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

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

Let

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

designate the vector representation of a channel with L coefficients $h_l^{(k)}$, $l = 0, \dots, L-1$, at chip rate $1/T_c$. For the time interval T required for transmission of a modulated version of 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

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

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

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

In the following, we regard an uplink scenario. Let

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

designate an Additional White Gaussian Noise (AWGN) vector with samples n_n , $n = 0, \dots, N-1$ 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 of user k of an OFDMA receiver 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 of user k is compensated by application of a Frequency Domain Equalizer (FDE) as described in [11] and [12], e.g., based on the Zero-Forcing (ZF) criterion. Let $\mathbf{E}^{(k)}$ denote the $Q \times Q$ matrix representation of the FDE. Thus, at the receiver, an estimate $\hat{\mathbf{d}}^{(k)}$ of the transmitted data block $\mathbf{d}^{(k)}$ for user k is given by

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

Finally, let the user specific $Q \times Q$ matrix $\mathbf{D}^{(k)}$ designate the overall channel for OFDMA of user k including mapping according to matrix $\mathbf{M}^{(k)}$, OFDMA modulation represented by \mathbf{F}_N^H , transmission over the equivalent circulant channel $\mathbf{H}^{(k)}$, OFDMA demodulation by \mathbf{F}_N and demapping according to $\mathbf{M}^{(k)\dagger}$. Thus, the overall channel matrix of user k is given by

$$\mathbf{D}^{(k)} = \mathbf{M}^{(k)\dagger} \cdot \mathbf{F}_N \cdot \mathbf{H}^{(k)} \cdot \mathbf{F}_N^H \cdot \mathbf{M}^{(k)}. \quad (10)$$

Note that since $\mathbf{H}^{(k)}$ is circulant, $\mathbf{F}_N \cdot \mathbf{H}^{(k)} \cdot \mathbf{F}_N^H$ results in a diagonal matrix. Thus, the overall channel matrix $\mathbf{D}^{(k)}$ is diagonal, too.

The overall channel matrix $\tilde{\mathbf{D}}^{(k)}$ for DFT-precoded OFDMA can be obtained from Eq. (10) by extension with a Q -point DFT precoding matrix \mathbf{F}_Q at the transmitter side and a Q -point IDFT matrix \mathbf{F}_Q^H for decoding at the receiver side according to

$$\tilde{\mathbf{D}}^{(k)} = \mathbf{F}_Q^H \cdot \mathbf{D}^{(k)} \cdot \mathbf{F}_Q. \quad (11)$$

Since $\mathbf{D}^{(k)}$ is diagonal, the overall channel matrix $\tilde{\mathbf{D}}^{(k)}$ for DFT-precoded OFDMA is a circulant matrix.

A. IFDMA

In the following, a system model for IFDMA is derived as a special case of the previously introduced system model for OFDMA. 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

$$M_I^{(k)}(n, q) = \begin{cases} 1 & n = q \cdot K + k \\ 0 & \text{else} \end{cases}. \quad (12)$$

Moreover, a Q -point DFT precoding \mathbf{F}_Q is applied to the OFDMA transmitter described by Eq. (3) before mapping. Thus, the transmit signal for IFDMA is given by

$$\mathbf{x}_I^{(k)} = \mathbf{F}_N^H \cdot \mathbf{M}_I^{(k)} \cdot \mathbf{F}_Q \cdot \mathbf{d}^{(k)}. \quad (13)$$

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

$$\mathbf{R} = (\mathbf{I}_Q, \dots, \mathbf{I}_Q)^T. \quad (14)$$

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

$$\Phi^{(k)} = \text{diag}(\phi_0^{(k)}, \dots, \phi_{N-1}^{(k)}), \quad (15)$$

where $\text{diag}(\phi_0^{(k)}, \dots, \phi_{N-1}^{(k)})$ 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. (13) can be expressed as

$$\mathbf{x}_I^{(k)} = \mathbf{R} \cdot \mathbf{\Phi}^{(k)} \cdot \mathbf{d}^{(k)}. \quad (16)$$

From Eq. (16) 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], [13].

In the uplink, the IFDMA receiver is typically implemented in frequency domain. The receiver is given by extension of the OFDMA receiver described in Eq. (9) by a Q -point IDFT operation \mathbf{F}_Q^H for inversion of the DFT precoding. Hence, for IFDMA an estimate $\hat{\mathbf{d}}^{(k)}$ of the transmitted data block $\mathbf{d}^{(k)}$ is given by

$$\hat{\mathbf{d}}^{(k)} = \mathbf{F}_Q^H \cdot \mathbf{E}^{(k)} \cdot \mathbf{M}_I^{(k)\dagger} \cdot \mathbf{F}_N \cdot \mathbf{r}. \quad (17)$$

Note, that for IFDMA a FDE based on the Minimum Mean Square Error (MMSE) criterion provides better performance compared to a ZF-FDE [14], [12], [15].

B. SC-FDMA

In the following, a system model for SC-FDMA is derived as a special case of the previously introduced system model for 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

$$M_B^{(k)}(n, q) = \begin{cases} 1 & n = k \cdot Q + q \\ 0 & \text{else} \end{cases}. \quad (18)$$

Again we extend the OFDMA transmitter from Eq. (3) by a Q -point DFT precoding \mathbf{F}_Q before mapping. Thus, the transmit signal for SC-FDMA is given by

$$\mathbf{x}_B^{(k)} = \mathbf{F}_N^H \cdot \mathbf{M}_B^{(k)} \cdot \mathbf{F}_Q \cdot \mathbf{d}^{(k)}. \quad (19)$$

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

$$\mathbf{x}_B^{(k)} = \mathbf{P}^{(k)} \cdot \mathbf{d}^{(k)}, \quad (20)$$

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

$$\hat{\mathbf{d}}^{(k)} = \mathbf{F}_Q^H \cdot \mathbf{E}^{(k)} \cdot \mathbf{M}_B^{(k)\dagger} \cdot \mathbf{F}_N \cdot \mathbf{r}. \quad (21)$$

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

III. LOW COMPLEXITY AND POWER EFFICIENT STBC FOR IFDMA AND SC-FDMA

In Section II, OFDMA with DFT precoding is described. DFT precoding can be understood as linear non-redundant frequency coding. In this section, in addition, STBC is applied such that, on the one hand, the advantages of DFT precoded OFDMA

are maintained and, on the other hand, significant performance improvement due to exploitation of spatial diversity is obtained.

In the following, the received signal after demodulation and demapping is described utilizing the the overall channel for DFT precoded OFDMA according to Eq. (11). Application of the interleaved mapping matrix $\mathbf{M}_I^{(k)}$ or the blockwise mapping matrix $\mathbf{M}_B^{(k)}$ results in the description of the overall channel for IFDMA or SC-FDMA, respectively.

We assume a multi antenna system with n_T transmit antennas and n_R receive antennas. Let $i = 1, \dots, n_T$ and $j = 1, \dots, n_R$ denote the indices for the transmit and receive antennas, respectively. Let further $\bar{\mathbf{r}}_j^{(k)}(p)$ designate the p -th received data block at receive antenna j of user k given by

$$\bar{\mathbf{r}}_j^{(k)}(p) = \sum_{i=1}^{n_T} \mathbf{D}_{i,j}^{(k)}(p) \cdot \mathbf{d}_i^{(k)}(p), \quad (22)$$

where $\mathbf{d}_i^{(k)}(p)$ designates the p -th data block of user k transmitted from the i -th transmit antenna and $\mathbf{D}_{i,j}^{(k)}(p)$ designates the overall channel of user k between the i -th transmit and j -th receive antenna.

In [16] a method for application of STBC to Orthogonal Frequency Division Multiplex (OFDM) is introduced for transmission over time dispersive channels. In the following, a method is described that can be understood as an extension of the method described in [16] to DFT precoded OFDMA including IFDMA and SC-FDMA. A similar method is described in [17] for single carrier transmission. In order to illustrate the application of STBC to DFT-precoded OFDMA, in the following, the well known Alamouti scheme [18] is regarded as an example. However, note that the presented method can be also applied to other STBC [16].

For the Alamouti scheme, $n_T = 2$ and $n_R = 1$ is assumed. Thus, in the following, index j can be skipped. We regard two consecutive time slots, each of length T . According to [16], it is assumed that for the duration of the two consecutive time slots, i.e., for $2 \cdot T$, the channel is time invariant. Within $2 \cdot T$ a space-time-frequency coded and modulated version of the data blocks $\mathbf{d}_1^{(k)}$ and $\mathbf{d}_2^{(k)}$ is transmitted. According to [17], we assume that in the first time slot a modulated version of the double DFT of $\mathbf{d}_1^{(k)}$ is transmitted from antenna $i = 1$ and a modulated version of the double DFT of $\mathbf{d}_2^{(k)}$ is transmitted from antenna $i = 2$. In the second time interval, a modulated version of $-(\mathbf{d}_2^{(k)})^*$ is transmitted from antenna $i = 1$ and a modulated version of $(\mathbf{d}_1^{(k)})^*$ is transmitted from antenna $i = 2$. Let $\bar{\mathbf{r}}^{(k)}(1)$ and $\bar{\mathbf{r}}^{(k)}(2)$ denote the received signals in the first and the second regarded time slot, respectively. Thus, together with Eq. (22), $\bar{\mathbf{r}}^{(k)}(1)$ and $\bar{\mathbf{r}}^{(k)}(2)$ are given by

$$\begin{aligned} \bar{\mathbf{r}}^{(k)}(1) &= \tilde{\mathbf{D}}_1^{(k)} \mathbf{F}_Q \mathbf{F}_Q \mathbf{d}_1^{(k)} + \tilde{\mathbf{D}}_2^{(k)} \mathbf{F}_Q \mathbf{F}_Q \mathbf{d}_2^{(k)} \\ \bar{\mathbf{r}}^{(k)}(2) &= -\tilde{\mathbf{D}}_1^{(k)} (\mathbf{d}_2^{(k)})^* + \tilde{\mathbf{D}}_2^{(k)} (\mathbf{d}_1^{(k)})^*, \end{aligned} \quad (23)$$

where $\tilde{\mathbf{D}}_i^{(k)}$, $i = 1, 2$, denotes the overall channel from the first and the second transmit antenna, respectively, to the receive antenna. The eigenvalue decomposition of $\tilde{\mathbf{D}}_i^{(k)}$ is given by

$$\tilde{\mathbf{D}}_i^{(k)} = \mathbf{F}_Q \cdot \mathbf{\Lambda}_i^{(k)} \cdot \mathbf{F}_Q^H, \quad (24)$$

where $\mathbf{\Lambda}_i^{(k)}$ is a diagonal matrix with the eigenvalues of $\tilde{\mathbf{D}}_i^{(k)}$ on its diagonal. Insertion of Eq. (24) into Eq. (23) and multiplication

with \mathbf{F}_Q^H from the left leads to

$$\begin{aligned}\mathbf{F}_Q^H \bar{\mathbf{r}}^{(k)}(1) &= \mathbf{\Lambda}_1^{(k)} \mathbf{F}_Q \mathbf{d}_1^{(k)} + \mathbf{\Lambda}_2^{(k)} \mathbf{F}_Q \mathbf{d}_2^{(k)} \\ \mathbf{F}_Q^H \bar{\mathbf{r}}^{(k)}(2) &= -\mathbf{\Lambda}_1^{(k)} \mathbf{F}_Q^H (\mathbf{d}_2^{(k)})^* + \mathbf{\Lambda}_2^{(k)} \mathbf{F}_Q^H (\mathbf{d}_1^{(k)})^*.\end{aligned}\quad (25)$$

Eq. (25) can be rewritten as

$$\begin{pmatrix} \mathbf{F}_Q^H \cdot \bar{\mathbf{r}}^{(k)}(1) \\ (\mathbf{F}_Q^H \cdot \bar{\mathbf{r}}^{(k)}(2))^* \end{pmatrix} = \underline{\mathbf{\Lambda}}^{(k)} \cdot \begin{pmatrix} \mathbf{F}_Q \mathbf{d}_1^{(k)} \\ \mathbf{F}_Q \mathbf{d}_2^{(k)} \end{pmatrix} \quad (26)$$

with

$$\underline{\mathbf{\Lambda}}^{(k)} = \begin{pmatrix} \mathbf{\Lambda}_1^{(k)} & \mathbf{\Lambda}_2^{(k)} \\ (\mathbf{\Lambda}_2^{(k)})^* & -(\mathbf{\Lambda}_1^{(k)})^* \end{pmatrix}. \quad (27)$$

At the receiver, $(\underline{\mathbf{\Lambda}}^{(k)})^H$ is applied according to

$$(\underline{\mathbf{\Lambda}}^{(k)})^H \begin{pmatrix} \mathbf{F}_Q^H \cdot \bar{\mathbf{r}}^{(k)}(1) \\ (\mathbf{F}_Q^H \cdot \bar{\mathbf{r}}^{(k)}(2))^* \end{pmatrix} = (\underline{\mathbf{\Lambda}}^{(k)})^H \underline{\mathbf{\Lambda}}^{(k)} \begin{pmatrix} \mathbf{F}_Q \mathbf{d}_1^{(k)} \\ \mathbf{F}_Q \mathbf{d}_2^{(k)} \end{pmatrix}. \quad (28)$$

Matrix $(\underline{\mathbf{\Lambda}}^{(k)})^H \underline{\mathbf{\Lambda}}^{(k)}$ is a diagonal matrix which can be easily compensated at the receiver by a FDE based on the MMSE criterion as described in [11].

IV. SIGNAL PROPERTIES AND COMPLEXITY

In the following, the properties of the signals at each transmit antenna and the computational complexity of the system are discussed.

According to Eq. (23), in the first time slot DFT precoded and OFDMA modulated versions of the double DFTs of $\mathbf{d}_i^{(k)}$, $i = 1, 2$ are transmitted from the i -th antenna. In the second time slot a DFT precoded and OFDMA modulated version of $\mathbf{d}_i^{(k)}$ is transmitted from the i -th antenna. It is well known that the double DFT of a vector $\mathbf{v} = (v_0, v_1, \dots, v_{Q-1})^T$ is given by

$$\mathbf{F}_Q \mathbf{F}_Q \mathbf{v} = (v_0, v_{Q-1}, \dots, v_1)^T, \quad (29)$$

cf. [17], i.e., application of STBC to DFT precoded OFDMA at the transmitter is given by a permutation of the elements of the transmit signals at each antenna and, hence, the signal properties of these signals are the same as for a DFT precoded OFDMA signal without STBC. Thus, applying the mapping matrices $\mathbf{M}_I^{(k)}$ or $\mathbf{M}_B^{(k)}$, respectively, the signal properties can be concluded directly from Eq. (16) or Eq. (20), respectively.

Compared to OFDMA without DFT precoding, IFDMA signals and SC-FDMA signals provide significantly lower envelope fluctuations due to the signal generation in time domain. Thus, a power efficient use of the amplifier and utilization of low cost amplifiers is possible. The envelope fluctuations of IFDMA signals are even lower compared to the envelope fluctuations of SC-FDMA signals because of the simple compression and repetition of the input data symbols.

In the following, the computational effort in terms of complex multiplications and divisions required for IFDMA and SC-FDMA with STBC is determined. The effort for channel estimation is omitted and it is assumed that for the permutation according to Eq. (29) no computational effort is necessary.

From Eq. (16) follows that IFDMA signal generation requires compression, repetition and user dependent phase rotation. Thus, at the transmitter, per transmit antenna N complex multiplications resulting from matrix multiplication with the matrix given

TABLE I: Complexity of IFDMA and SC-FDMA

	Receiver		
	Transmitter	Receiver	
	multiplications	multiplications	divisions
IFDMA	$2 \cdot N$	$1/2 N \text{ld}(N) + 2Q$	Q
SC-FDMA	$2 \cdot Q \cdot N$	$+Q + 1/2 Q \text{ld}(Q)$	

in Eq. (15) are required. According to Eq. (20), SC-FDMA signal generation requires $N \cdot Q$ complex multiplications per transmit antenna resulting from matrix multiplication with $\mathbf{P}^{(k)}$, cf. Eq. (20). Matrix $\mathbf{P}^{(k)}$ can be calculated offline. Thus, for both schemes only low computational effort at the transmitter is required which, together with the low envelope fluctuations, makes low cost and power efficient terminals possible.

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 $1/2 \cdot N \cdot \text{ld}(N)$ complex multiplications. Additionally, according to Eq. (28), a matrix multiplication with $(\underline{\mathbf{\Lambda}}^{(k)})^H$ is required. $(\underline{\mathbf{\Lambda}}^{(k)})^H$ is block diagonal, hence, the matrix multiplication with $(\underline{\mathbf{\Lambda}}^{(k)})^H$ requires $2 \cdot Q$ complex scalar 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 $1/2 \cdot Q \cdot \text{ld}(Q)$ complex multiplications are required per user. Thus, also at the receiver, for both schemes low computational effort is required. Due to the low complexity, in particular at the transmitter side, and the low PAPR of the transmit signals at each transmit antenna, both schemes are well suited especially for the uplink. The total computational effort for a system with 2 transmit antennas and 1 receive antenna using Alamout STBC is summarized in Table I.

V. SIMULATION RESULTS

In Figure 1 performance results for transmission over correlated and uncorrelated spatial mobile radio channels with simulation parameters according to Table II are given. Regarding IFDMA at a Bit Error Rate (BER) of 10^{-3} , for a net bit rate of about 5.1 Mbit/s, i.e., a number of 128 subcarriers assigned to each user, a gain due to additional spatial diversity of about 0.9 dB is obtained for transmission over correlated channels. For uncorrelated channels, the gain is about 1.7 dB. For SC-FDMA the gain is similar. For a net bit rate of about 1.25 Mbit/s, i.e. $Q = 32$ subcarriers, at a BER of 10^{-3} SC-FDMA provides a gain of about 1.6 dB for transmission over correlated channels and 2 dB for transmission over uncorrelated channels. For IFDMA, the gain for transmission over correlated channels is 0.7 dB and for transmission over uncorrelated channels 1.5 dB.

Comparison of IFDMA and SC-FDMA shows that for a net bit rate of about 5.1 Mbit/s, at a BER of 10^{-3} , IFDMA gains about 0.9 dB for the single antenna case and about 0.8 dB for combination with STBC due to its higher frequency diversity. For a net bit rate of about 1.25 Mbit/s, i.e. $Q = 32$ subcarriers per user, the effect is further reinforced. At a BER of 10^{-3} the gain of IFDMA compared to SC-FDMA is about 2.1 dB for the single antenna case and 1.9 dB for combination with STBC. For both data rates, the gain for transmission over correlated and

TABLE II: Simulation parameters

Carrier frequency	5 GHz	Decoder	MaxLogMAP
Bandwidth	20 MHz	Equalizer	MMSE FDE
No. of subcarriers	512	Interleaving	Random
Modulation	QPSK	Interl. depth	0.5 ms
Code	Conv. Code	STBC	Alamouti
Code rate	1/2	Guard interval	5 μ s
Constraint length	6	Channel	WINNER SCM [19], Urban, 70 km/h

uncorrelated channels is similar.

VI. CONCLUSIONS

In this paper, a combination of STBC and DFT frequency coding for OFDMA systems has been proposed. The resulting space-time-frequency coded OFDMA scheme provides low computational complexity for channel equalization, user separation and space-time-frequency decoding. Especially for application to DFT precoded OFDMA systems with equidistant subcarrier allocation, i.e., IFDMA and SC-FDMA, also low computational complexity for signal generation is obtained. Additionally, both, IFDMA with STBC and SC-FDMA with STBC, provide low envelope fluctuations of the signals at each transmit antenna and, thus, provide a power efficient use of the amplifiers and make low cost terminals possible. Finally, significant performance gains due to additional spatial diversity is exploited and both, SC-FDMA and IFDMA provide good coded performance for transmission over typical mobile radio channels. Due to the interleaved subcarrier allocation, IFDMA can exploit more frequency diversity and thus, outperforms SC-FDMA, especially for moderate to low data rates. However, for application of further diversity techniques such as, e.g., frequency hopping, also the performance of SC-FDMA can be further improved. Thus, due to their good performance, their signal properties and their low complexity also for high data rates, in particular at the transmitter, both schemes are well suited candidates for the uplink of future mobile radio systems.

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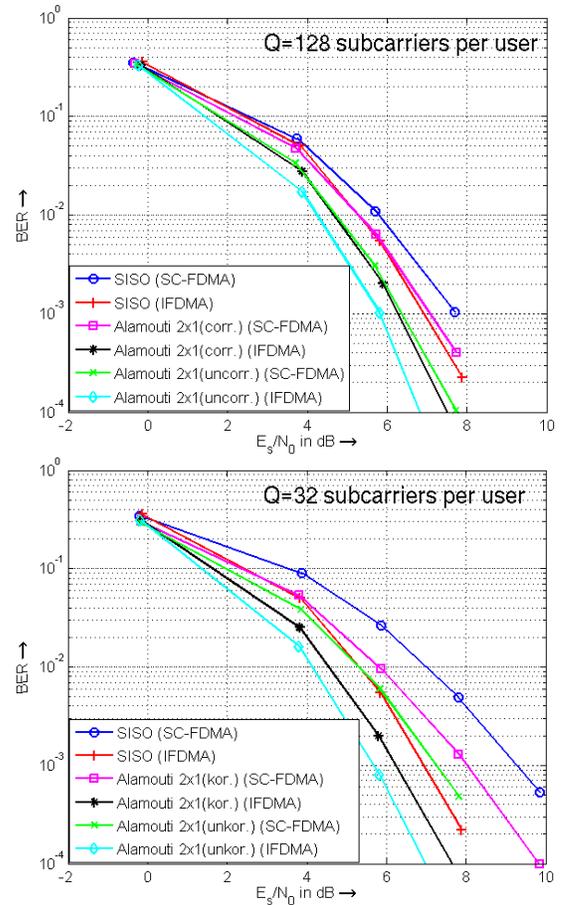


Fig. 1: Coded performance for IFDMA and SC-FDMA

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