PER-TONE CHIP EQUALIZATION FOR SPACE-TIME BLOCK CODED SINGLE-CARRIER BLOCK TRANSMISSION DS-CDMA DOWNLINK

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ABSTRACT

In this paper, we extend space-time block coding techniques, originally proposed for point-to-point communication links, to point-to-multipoint communication links. In specific, we combine space-time block coding with single-carrier block transmission DS-CDMA for downlink multi-user MIMO communications. Moreover, we propose a pilot-trained space-time chip equalizer that acts on a per-tone basis in the frequency-domain. With M_t transmit antennas at the base station, M_r receive antennas at the mobile station and L the order of the multi-path channel, it comes close to extracting the full diversity of order $M_t \cdot M_r \cdot (L+1)$ in reduced as well as full load settings.

1. INTRODUCTION

Space-time coding techniques, that introduce both temporal and spatial correlation between the transmitted signals, are capable of supporting reliable high-data-rate communications without sacrificing precious bandwidth resources [1]. However, up till now the focus was mainly on point-to-point communication links, thereby neglecting the multiple access technique in the design of the transmission scheme. In this paper, we consider point-to-multipoint Multiple Input Multiple Output (MIMO) communication links and add the multi-user dimension to the problem. In specific, we focus on the downlink and combine Space-Time Block Coding (STBC) with Single-Carrier Block Transmission (SCBT) DS-CDMA [2] because the resulting transmission technique inherits the advantages of CDMA, STBC and SCBT. On the one hand, CDMA allows universal frequency reuse in a cellular network and therefore provides higher capacity and easier network planning over conventional access techniques like TDMA and FDMA. On the other hand, space-time block coded SCBT with zero padding (ZP) achieves maximum diversity gains over frequency-selective point-to-point communication channels [3]. Moreover SCBT with ZP enables simple frequency-domain equalization at the mobile station but avoids the Peak-to-Average-Power-Ratio (PAPR) problem of classical OFDM systems. In this perspective, we also design a pilot-trained space-time chip-level equalizer that acts on a per-tone basis in the frequency-domain. With M_t transmit antennas at the base-station, M_r receive antennas at the mobile station and L the order of the multi-path channel, it comes close to extracting the full diversity of order $M_t.M_r.(L+1)$ independently of the system load.

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2. SCBT-DS-CDMA DOWNLINK SYSTEM MODEL

Let us consider the downlink of a single-cell space-time block coded SCBT DS-CDMA system with U active mobile stations. The base-station is equipped with M_t transmit antennas whereas the mobile station of interest is equipped with M_r receive antennas.

2.1. Transmitter model for the base station

For simplicity reasons, we will assume in the following that the base station has only $M_t = 2$ transmit antennas. Note however that the proposed techniques can be extended to the more general case of $M_t > 2$ transmit antennas when resorting to the generalized orthogonal designs of [1]. As shown in Figure 1, each user's data symbol sequence $s^u[i]$ (similar for the pilot symbol sequence $s^p[i]$) is demultiplexed into M_t parallel lower rate sequences $\{s^u_{m_t}[i] := s^u[iM_t + m_t - 1]\}_{m_t=1}^{M_t}$, where M_t is the number of transmit antennas. Each of the u-th user's symbol sequences $\{s^u_{m_t}[i]\}_{m_t=1}^{M_t}$ is serial-to-parallel converted into blocks of B symbols, leading to the symbol block sequences $\left\{\mathbf{s}^u_{m_t}[i] := \begin{bmatrix} s^u_{m_t}[iB] & \dots & s^u_{m_t}[(i+1)B-1] \end{bmatrix}^T\right\}_{m_t=1}^{M_t}$ that are subsequently spread by a factor N with the same user composite code sequence $c_u[n]$ which is the multiplication of the user specific orthogonal Walsh-Hadamard spreading code and the base station specific scrambling code. For each of the M_t parallel streams, the different user chip block sequences are added up together with the pilot chip block sequence, resulting into the m_t-th multi-user chip block sequence :

$$\mathbf{x}_{m_t}[n] = \sum_{u=1}^{U} \mathbf{s}^u_{m_t}[i]c_u[n] + \mathbf{s}^p_{m_t}[i]c_p[n] \tag{1}$$

with $i = \lfloor \frac{n}{N} \rfloor$. Let us also define the u-th user's total symbol block sequence $\mathbf{s}^u[i] := \begin{bmatrix} \mathbf{s}^u_1[i]^T & \mathbf{s}^u_2[i]^T \end{bmatrix}^T$ and the total multi-user chip block sequence $\mathbf{x}[n] := \begin{bmatrix} \mathbf{x}^T_1[n] & \mathbf{x}^T_2[n] \end{bmatrix}^T$. Following a similar approach as in [3] for point-to-point SCBT with ZP, the block Space-Time (ST) encoder operates in the time-domain at the chip-level rather than at the symbol-level and takes the two multi-user chip blocks $\{\mathbf{x}_{m_t}[n]\}_{m_t=1}^{2}$ to output the following $2B \times 2$ matrix of ST coded multi-user chip blocks :

$$\begin{bmatrix} \bar{\mathbf{x}}_1[2n] & \bar{\mathbf{x}}_1[2n+1] \\ \bar{\mathbf{x}}_2[2n] & \bar{\mathbf{x}}_2[2n+1] \end{bmatrix} = \begin{bmatrix} \mathbf{x}_1[n] & -\mathbf{P}^{(0)}_B \cdot \mathbf{x}^*_2[n] \\ \mathbf{x}_2[n] & \mathbf{P}^{(0)}_B \cdot \mathbf{x}^*_1[n] \end{bmatrix} \tag{2}$$

where $\mathbf{P}^{(j)}_J$ is a $J \times J$ permutation matrix implementing a reversed cyclic shift over j positions. At each time interval n, the ST

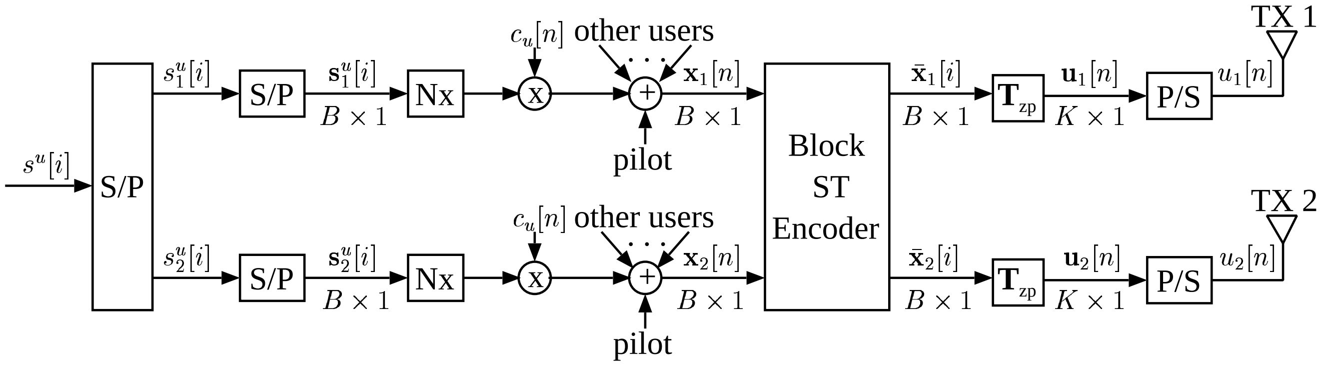

Fig. 1. Transmitter model for STBC SCBT-DS-CDMA

coded multi-user chip blocks $\bar{\mathbf{x}}_1[n]$ and $\bar{\mathbf{x}}_2[n]$ are forwarded to the first respectively the second transmit antenna. From Equation 2, we can easily verify that the transmitted multi-user chip block at time instant $2n+1$ from one antenna is the time-reversed conjugate of the transmitted multi-user chip block at time instant $2n$ from the other antenna (with a possible sign change). The $K \times B$ transmit matrix $\mathbf{T}_{zp}$, with $K = B + \mu$, pads a zero postfix of length μ to each block of the ST coded multi-user chip block sequence $\bar{\mathbf{x}}_{m_t}[n]$ leading to the transmitted multi-user chip block sequence $\mathbf{u}_{m_t}[n] = \mathbf{T}_{zp} \cdot \bar{\mathbf{x}}_{m_t}[n]$. Finally, the transmitted multi-user chip block sequence $\mathbf{u}_{m_t}[n]$ is parallel-to-serial converted into K chips, obtaining the transmitted multi-user chip sequence $\begin{bmatrix} u_{m_t}[nK] & \dots & u_{m_t}[(n+1)K-1] \end{bmatrix}^T := \mathbf{u}_{m_t}[n]$.

2.2. Receiver model for the mobile station

We assume that the mobile station of interest is equipped with M_r receive antennas and has acquired perfect synchronisation. At each receive antenna in Figure 2, the time-domain received chip sequence $v_{m_r}[n]$ is serial-to-parallel converted into blocks of K chips, resulting into the time-domain received chip block sequence $\mathbf{v}_{m_r}[n] := \begin{bmatrix} v_{m_r}[nK] & \dots & v_{m_r}[(n+1)K-1] \end{bmatrix}^T$. The $K \times K$ receive matrix $\mathbf{R} := \mathbf{I}_K$ completely preserves each block of the time-domain received chip block sequence $\mathbf{v}_{m_r}[n]$ leading to the time-domain received ST coded chip block sequence $\bar{\mathbf{y}}_{m_r}[n] = \mathbf{R} \cdot \mathbf{v}_{m_r}[n]$. Assuming a sufficiently long zero postfix $\mu \geq L$, we obtain a simple input/ouput relationship in the time-domain :

$$\bar{\mathbf{y}}_{m_r}[n] = \sum_{m_t=1}^{M_t} \dot{\mathbf{H}}_{m_r,m_t} \cdot \mathbf{T}_{zp} \cdot \bar{\mathbf{x}}_{m_t}[n] + \bar{\mathbf{e}}_{m_r}[n] \tag{3}$$

where $\bar{\mathbf{e}}_{m_r}[n]$ is the time-domain received noise block sequence and $\dot{\mathbf{H}}_{m_r,m_t}$ is a $K \times K$ circulant channel matrix. By exploiting the ST code structure of Equation 2 in a similar way as in [3], we can write for two consecutive chip blocks $\mathbf{y}_{m_r,1}[n] := \bar{\mathbf{y}}_{m_r}[2n]$ and $\mathbf{y}_{m_r,2}[n] := \mathbf{P}_K^{(B)} \cdot \bar{\mathbf{y}}_{m_r}^*[2n+1]$ the time-domain input/output relationship of Equation 3. Transforming the resulting ST decoded chip blocks $\mathbf{y}_{m_r,1}[n]$ and $\mathbf{y}_{m_r,2}[n]$ to the frequency-domain employing the $K \times K$ FFT matrix $\mathbf{F}_K$, leads to the input/output relationship in Equation 4 where $\tilde{\mathbf{H}}_{m_r,m_t} = \mathbf{F}_K \cdot \dot{\mathbf{H}}_{m_r,m_t} \cdot \mathbf{F}_K^H$ is the $K \times K$ diagonal frequency-domain channel matrix having the FD channel response $\tilde{\mathbf{h}}_{m_r,m_t}$ as its main diagonal. Note from Equation 4 that $\tilde{\mathbf{x}}[n] = \mathcal{F}_K \cdot \mathcal{T}_{zp} \cdot \mathbf{x}[n]$ where both the compound FFT matrix $\mathcal{F}_K := diag\{\mathbf{F}_K, \mathbf{F}_K\}$ and the compound transmit matrix $\mathcal{T}_{zp} := diag\{\mathbf{T}_{zp}, \mathbf{T}_{zp}\}$ are block diagonal. Stacking the contributions of the M_r receive antennas $\tilde{\mathbf{y}}[n] = \begin{bmatrix} \tilde{\mathbf{y}}_1^T[n] & \dots & \tilde{\mathbf{y}}_{M_r}^T[n] \end{bmatrix}^T$, we obtain the following per receive antenna frequency-domain data model :

$$\tilde{\mathbf{y}}[n] = \tilde{\mathbf{H}} \cdot \tilde{\mathbf{x}}[n] + \tilde{\mathbf{e}}[n] \tag{5}$$

where the per receive antenna channel matrix $\tilde{\mathbf{H}}$ and the per receive antenna noise block $\tilde{\mathbf{e}}[n]$ are similarly defined as the per receive antenna output block $\tilde{\mathbf{y}}[n]$. Defining the receive permutation matrix $\mathbf{P}_r$ respectively the transmit permutation matrix $\mathbf{P}_t$ as follows :

$$\acute{\mathbf{y}}[n] := \mathbf{P}_r \cdot \tilde{\mathbf{y}}[n] \qquad \tilde{\mathbf{x}}[n] := \mathbf{P}_t \cdot \acute{\mathbf{x}}[n] \tag{6}$$

where $\mathbf{P}_r$ permutes a per receive antenna ordering into a per-tone ordering and where $\mathbf{P}_t$ conversely permutes a per-tone ordering into a per transmit antenna ordering, we obtain the following per-tone data model :

$$\acute{\mathbf{y}}[n] = \acute{\mathbf{H}} \cdot \acute{\mathbf{x}}[n] + \acute{\mathbf{e}}[n] \tag{7}$$

In this Equation, $\acute{\mathbf{y}}[n] = \begin{bmatrix} \acute{\mathbf{y}}_1^T[n] & \dots & \acute{\mathbf{y}}_K^T[n] \end{bmatrix}^T$ is the per-tone output block, $\acute{\mathbf{x}}[n]$ the per-tone input block and $\acute{\mathbf{e}}[n]$ the per-tone noise block both similarly defined as $\acute{\mathbf{y}}[n]$. The per-tone channel matrix $\acute{\mathbf{H}}$ is a block diagonal matrix, given by :

$$\acute{\mathbf{H}} := \mathbf{P}_r \cdot \tilde{\mathbf{H}} \cdot \mathbf{P}_t = diag\left\{\acute{\mathbf{H}}_1, \dots, \acute{\mathbf{H}}_K\right\} \tag{8}$$

2.3. Data model for burst processing

Assuming a burst length of $M_t \cdot B \cdot I$ symbols for each user, we can stack $I \cdot N$ consecutive chip blocks $\tilde{\mathbf{y}}[n]$, defined in Equation 5, into $\tilde{\mathbf{Y}} := \begin{bmatrix} \tilde{\mathbf{y}}[0] & \dots & \tilde{\mathbf{y}}[IN-1] \end{bmatrix}$, leading to the following per receive antenna data model for burst processing :

$$\tilde{\mathbf{Y}} = \tilde{\mathbf{H}} \cdot \tilde{\mathbf{X}} + \tilde{\mathbf{E}} \tag{9}$$

where the input matrix $\tilde{\mathbf{X}}$ and the noise matrix $\tilde{\mathbf{E}}$ are similarly defined as the output matrix $\tilde{\mathbf{Y}}$. Note that

$$\tilde{\mathbf{X}} = \mathcal{F}_K \cdot \mathcal{T}_{zp} \cdot \mathbf{X} \tag{10}$$

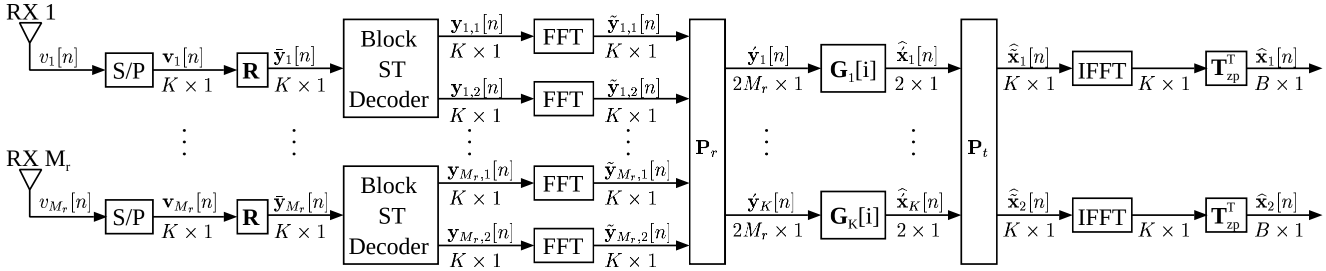

Fig. 2. Receiver model for STBC SCBT-DS-CDMA

$$\underbrace{\begin{bmatrix} \mathbf{F}_K \cdot \bar{\mathbf{y}}_{m_r}[2n] \\ \mathbf{F}_K \cdot \mathbf{P}_K^{(B)} \cdot \bar{\mathbf{y}}_{m_r}^*[2n+1] \end{bmatrix}}_{\tilde{\mathbf{y}}_{m_r}[n]} = \underbrace{\begin{bmatrix} \tilde{\mathbf{H}}_{m_r,1} & \tilde{\mathbf{H}}_{m_r,2} \\ \tilde{\mathbf{H}}_{m_r,2}^* & -\tilde{\mathbf{H}}_{m_r,1}^* \end{bmatrix}}_{\tilde{\mathbf{H}}_{m_r}} \cdot \underbrace{\begin{bmatrix} \mathbf{F}_K \cdot \mathbf{T}_{zp} \cdot \mathbf{x}_1[n] \\ \mathbf{F}_K \cdot \mathbf{T}_{zp} \cdot \mathbf{x}_2[n] \end{bmatrix}}_{\tilde{\mathbf{x}}[n]} + \underbrace{\begin{bmatrix} \mathbf{F}_K \cdot \bar{\mathbf{e}}_{m_r}[2n] \\ \mathbf{F}_K \cdot \mathbf{P}_K^{(B)} \cdot \bar{\mathbf{e}}_{m_r}^*[2n+1] \end{bmatrix}}_{\tilde{\mathbf{e}}_{m_r}[n]} \tag{4}$$

where $\mathbf{X}$ stacks $I \cdot N$ consecutive total multi-user chip blocks $\mathbf{x}[n]$. Moreover, by inspecting Equation 1, we can write $\mathbf{X}$ as follows :

$$\mathbf{X} = \mathbf{S}_d \cdot \mathbf{C}_d + \mathbf{S}_p \cdot \mathbf{C}_p \tag{11}$$

where the multi-user total data symbol matrix $\mathbf{S}_d := \begin{bmatrix} \mathbf{S}_1 & \dots & \mathbf{S}_U \end{bmatrix}$ stacks the total data symbol matrices of the different active users and the u-th user's total data symbol matrix $\mathbf{S}_u := \begin{bmatrix} \mathbf{s}^u[0] & \dots & \mathbf{s}^u[I-1] \end{bmatrix}$ stacks I consecutive total symbol blocks for the u-th user. The total pilot symbol matrix $\mathbf{S}_p$ is similarly defined as $\mathbf{S}_u$. The multi-user code matrix $\mathbf{C}_d := \begin{bmatrix} \mathbf{C}_1^T & \dots & \mathbf{C}_U^T \end{bmatrix}^T$ stacks the code matrices of the different active users. The u-th user's code matrix stacks the u-th user's composite code vectors at I consecutive symbol instants :

$$\mathbf{C}_u := diag\{\mathbf{c}_u[0], \dots, \mathbf{c}_u[I-1]\} \tag{12}$$

where $\mathbf{c}_u[i] = \begin{bmatrix} c_u[iN] & \dots & c_u[(i+1)N-1] \end{bmatrix}$ is the u-th user's composite code vector used to spread the total symbol block $\mathbf{s}^u[i]$. The pilot code matrix $\mathbf{C}_p$ is similarly defined as $\mathbf{C}_u$.

Similarly to the per receive antenna data model for burst processing in Equation 9, we can stack $I \cdot N$ consecutive chip blocks $\acute{\mathbf{y}}[n]$ leading to the following per-tone data model for burst processing :

$$\acute{\mathbf{Y}} = \acute{\mathbf{H}} \cdot \acute{\mathbf{X}} + \acute{\mathbf{E}} \tag{13}$$

Using Equation 6, 10 and 11, we can express $\acute{\mathbf{X}}$ as follows :

$$\acute{\mathbf{X}} = \acute{\mathbf{S}}_d \cdot \mathbf{C}_d + \acute{\mathbf{S}}_p \cdot \mathbf{C}_p \tag{14}$$

where $\acute{\mathbf{S}}_d := \mathbf{P}_t^T \cdot \tilde{\mathbf{S}}_d$ and $\acute{\mathbf{S}}_p := \mathbf{P}_t^T \cdot \tilde{\mathbf{S}}_p$ are the per-tone permuted versions of $\tilde{\mathbf{S}}_d := \mathcal{F}_K \cdot \mathcal{T}_{zp} \cdot \mathbf{S}_d$ respectively $\tilde{\mathbf{S}}_p := \mathcal{F}_K \cdot \mathcal{T}_{zp} \cdot \mathbf{S}_p$.

3. PER-TONE BURST CHIP EQUALIZATION

Inspired by our related work for the DS-CDMA downlink [4], we can now deal with the design of a burst chip equalizer that processes a burst of $M_t \cdot B \cdot I$ data symbols at once. Starting from Equation 13 and assuming that the channel matrix $\acute{\mathbf{H}}$ has full column rank and the input matrix $\acute{\mathbf{X}}$ has full row rank, it is possible to find a Zero-Forcing (ZF) chip equalizer matrix $\mathbf{G}$, for which :

$$\mathbf{G} \cdot \acute{\mathbf{Y}} - \acute{\mathbf{X}} = \mathbf{0} \tag{15}$$

provided there is no noise present in the output matrix $\acute{\mathbf{Y}}$. Since the channel matrix $\acute{\mathbf{H}}$ has a block diagonal structure, as shown in Equation 8, the equalizer matrix $\mathbf{G}$ suffices to have a block diagonal structure as well :

$$\mathbf{G} := diag\{\mathbf{G}_1, \dots, \mathbf{G}_K\} \tag{16}$$

acting on a per-tone basis (see also Figure 2). For this reason, the ZF problem of Equation 15 decouples into K parallel and independent ZF problems, one for each tone. Using Equation 14, we can rewrite the original ZF problem of Equation 15 as follows :

$$\mathbf{G} \cdot \acute{\mathbf{Y}} - \acute{\mathbf{S}}_d \cdot \mathbf{C}_d - \acute{\mathbf{S}}_p \cdot \mathbf{C}_p = \mathbf{0} \tag{17}$$

which is a ZF problem in both the equalizer matrix $\mathbf{G}$ and the multi-user total data symbol matrix $\acute{\mathbf{S}}_d$.

3.1. Pilot-trained burst chip equalizer

The pilot-trained burst chip equalizer determines its equalizer coefficients from the per-tone output matrix $\acute{\mathbf{Y}}$ based on the knowledge of the pilot code matrix $\mathbf{C}_p$ and the total pilot symbol matrix $\acute{\mathbf{S}}_p$. By despreading Equation 17 with the pilot code matrix $\mathbf{C}_p$, we obtain :

$$\mathbf{G} \cdot \acute{\mathbf{Y}} \cdot \mathbf{C}_p^H - \acute{\mathbf{S}}_p = \mathbf{0} \tag{18}$$

because of the orthogonality between the multi-user code matrix $\mathbf{C}_d$ and the pilot code matrix $\mathbf{C}_p$. In case noise is present in the output matrix $\acute{\mathbf{Y}}$, we have to solve the corresponding Least Squares (LS) minimisation problem :

$$\widehat{\mathbf{G}} = \arg\min_{\mathbf{G}} \left\| \mathbf{G} \cdot \acute{\mathbf{Y}} \cdot \mathbf{C}_p^H - \acute{\mathbf{S}}_p \right\|_F^2 \tag{19}$$

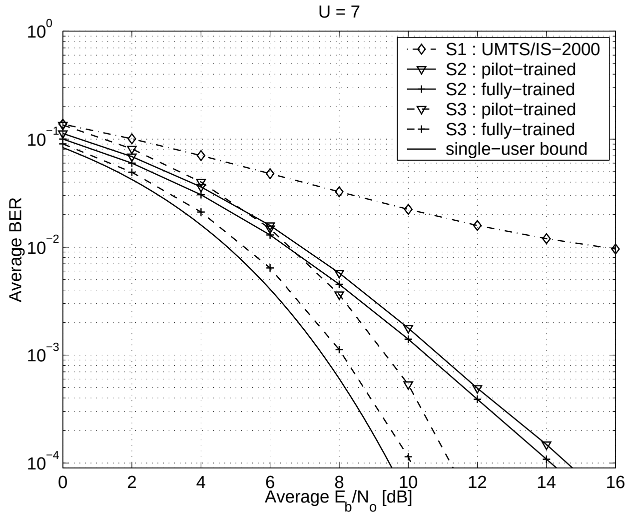

Fig. 3. Comparison of different scenarios for U=7

3.2. User-specific detection

As shown in Figure 2, the obtained per-tone pilot-trained chip equalizer matrix $\widehat{\mathbf{G}}$ may subsequently be used to extract the desired user's total data symbol matrix :

$$\widehat{\mathbf{S}}_u = \mathcal{T}_{zp}^T \cdot \mathcal{F}_K^H \cdot \mathbf{P}_t \cdot \widehat{\mathbf{G}} \cdot \acute{\mathbf{Y}} \cdot \mathbf{C}_u^H \tag{20}$$

where the transmit permutation matrix $\mathbf{P}_t$ permutes the per-tone ordering of the equalized output matrix $\widehat{\mathbf{G}} \cdot \acute{\mathbf{Y}}$ into a per transmit antenna ordering. Next, the compound IFFT matrix $\mathcal{F}_K^H$ transforms the permuted version of the equalized output matrix back to the time-domain whereas $\mathcal{T}_{zp}^T$ removes the zero postfix. Finally, the desired user's code matrix $\mathbf{C}_u^H$ despreads the resulting matrix to obtain the desired user's soft estimates.

4. SIMULATION RESULTS

We consider the downlink of a ST coded CDMA system with $M_t = 2$ transmit antennas at the base station, $M_r = 2$ receive antennas at the mobile station of interest and real orthogonal Walsh-Hadamard spreading codes of length $N = 8$ along with a random overlay code for scrambling. The QPSK modulated data symbols are transmitted in bursts of 520 symbols. We assume that each channel $h_{m_r,m_t}[l]$ is FIR with order $L = 3$ and has Rayleigh distributed channel taps of equal average power. We compare three different scenarios :

- S1. A pilot-trained space-time RAKE receiver is applied to the space-time coded downlink DS-CDMA transmission scheme that was proposed for the UMTS and the IS-2000 WCDMA standards, also known as Space-Time Spreading (STS) [5]. The pilot-trained space-time RAKE receiver is similar to the time-only RAKE receiver discussed in [5], but instead of using a time-only maximum ratio combiner based on exact channel knowledge, we use a space-time combiner that is trained with the pilot.
- S2. The pilot-trained and the ideal fully-trained time-domain space-time chip equalizer methods of [4] are applied to the space-time block coded downlink DS-CDMA transmission scheme proposed in [6].
- S3. The proposed pilot-trained and the ideal fully-trained per-tone space-time chip equalizer methods are applied to the proposed space-time block coded downlink single-carrier block transmission (SCBT) DS-CDMA transmission scheme. The burst length of 520 symbols is split into $M_t \cdot I = 40$ symbol blocks of $B = 13$ symbols each. Taking $\mu = L = 3$ and correspondingly $K = B + \mu = 16$, zero-padding results in an acceptable decrease in information rate, more specifically, a decrease with a factor $B/K \approx 0.81$.

Figure 3 compares the average BER versus the average SNR per bit of the different scenarios for a fully loaded system with $U = 7$ users. Also shown in the figure is the theoretical BER-curve for QPSK with $M_t \cdot M_r \cdot (L + 1) = 16$-fold diversity in Rayleigh fading channels (single-user bound). Firstly, we observe that S1 performs poorly compared to S2 and S3 : e.g. at a BER of 10^{-2} S2 and S3 achieve a 9 dB gain compared to S1. In contrast to S2 and S3, that completely suppress the Inter Chip Interference (ICI) and the MUI at infinite SNR, S1 does not and therefore suffers from a BER saturation level. Secondly, we observe that the ideal fully-trained (FT) version of S3 closely approaches the optimal single-user bound whereas the corresponding version of S2 does not. At a BER of 10^{-4} e.g., FT/S3 achieves a 4 dB gain compared to FT/S2. Thirdly, the pilot-trained (PT) version of S3 incurs a larger loss with respect to its ideal FT/S3 (1.2 dB at 10^{-4}) than the PT version of S2 does with respect to its ideal FT/S2 (0.2 dB at 10^{-4}). However, at a BER of 10^{-4}, PT/S3 still achieves a 3 dB gain compared to PT/S2.

5. CONCLUSION

Due to its definite advantages, we can conclude that space-time block coded SCBT-DS-CDMA is an interesting transmission technique for future 4G broadband wireless communication systems.

6. REFERENCES

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