COMBINED INTERFERENCE SUPPRESSION AND FREQUENCY DOMAIN EQUALIZATION FOR SPACE-TIME BLOCK CODED TRANSMISSION

Ayman F. Naguib

Qualcomm Inc.
675 Campbell Technology Pkway, Suite 200
Campbell, CA95008
Email:anaguib@qualcomm.com, Tel:+1 408-626-0584, Fax:+1 408-557-1001

ABSTRACT

Diversity transmission at the base station is an effective technique to combat adverse effects of fading. This is suggested as an alternative to diversity at the terminal, thereby reducing the implementation complexity. In this paper, we present a transmission scheme that combines space-time block coding over frequency selective channels with single carrier frequency domain interference suppression and equalization. It is shown that this scheme will provide the diversity benefit of both the frequency selective channel and the space-time block code while completely suppressing the interference from another co-channel transmitter that occupies exactly the same channel (time & frequency) as the desired transmitter.

1. INTRODUCTION

Recently, there have been a number of proposals that use multiple antennas at the transmitter with the appropriate signal processing to jointly combat the above wireless channel impairments and provide antenna diversity for the downlink while placing most of the diversity burden on the base station. Substantial benefits can be achieved by using channel codes that are specifically designed taken into account multiple transmit antennas. The first bandwidth efficient transmit diversity scheme was proposed by Wittneben [1] and it included the transmit diversity scheme of [2] as a special case. In [3] space-time trellis codes were introduced, where a general theory for design of combined trellis coding and modulation for transmit diversity is proposed. Another approach for space-time coding, space-time block codes, was introduced in [4] and later generalized in [5]. Space-time codes have been recently adopted in third generation cellular standard (e.g. CDMA-2000 [6] and W-CDMA [7]). A scheme for combined interference suppression and space-time block decoding based on Alamouti scheme in [4] was presented in [8] and later generalized in [9]

In this paper we present a combined frequency domain equalization (FDE) and interference suppression (IS) technique for

the STBC for flat fading channels, the frequency selective channel model, and the FDE technique for single input single output channels. Next, we describe a STBC transmission scheme for frequency selective channel and FDE in Section 3. Then, in Section 4, we present a combined FDE and IS technique for STBC. In Section 5 we present a simulation example for the proposed scheme.

2. PRELIMINARIES

2.1. SPACE-TIME BLOCK CODING

In [4], Alamouti presented a transmit diversity scheme using a space-time block coding approach with two transmit antennas. This approach was designed for flat channels. In addition, it was assumed that the channel will remain constant over at least consecutive symbols. In this scheme, the original symbol sequence $x(n)$ is divided into blocks of two symbols each $x_k(n)$ and $x_{k+1}(n)$. Then, every pair of symbols is mapped according to (we will drop the index n for simplicity of notation)

$$\begin{bmatrix} x_k \\ x_{k+1} \end{bmatrix} \longrightarrow \begin{bmatrix} x_k & -x_{k+1} \\ x^*_{k+1} & x^*_k \end{bmatrix} = \boldsymbol{X} \qquad (1)$$

In the first symbol period, the first column of $\boldsymbol{X}$ is transmitted from antenna 1 and antenna 2, respectively. The corresponding received signal is

$$r_k = x_k h_1 + x^*_{k+1} h_2 + n_k \qquad (2)$$

At the next symbol period, the second column of $\boldsymbol{X}$ is transmitted from antenna 1 and antenna 2 in a similar fashion such that the corresponding received signal is

$$r_{k+1} = -x_{k+1} h_1 + x^*_k h_2 + n_{k+1} \qquad (3)$$

In (2) and (3), h_1 and h_2 represent the complex gains of the channel (which is assumed there to be flat) between the first and second transmit antennas and the receive antenna, respectively. Equations (2) and (3) can be put in a matrix form as follows

$$\boldsymbol{r}_k = \begin{bmatrix} r_k \\ r^*_{k+1} \end{bmatrix} = \begin{bmatrix} h_1 & -h_2 \\ h^*_2 & h^*_1 \end{bmatrix} \begin{bmatrix} x_k \\ x^*_{k+1} \end{bmatrix} + \begin{bmatrix} n_k \\ n^*_{k+1} \end{bmatrix} \qquad (4)$$

diversity scheme is to multiply $\mathbf{r}_k$ by $\boldsymbol{H}^*$, which turns out to be the matched filter receiver. As pointed out earlier, this scheme will achieve the maximum possible diversity order over a flat channel (which is two in this case). Despite of the simplicity of the decoding processing when used over flat channel, using this scheme in its present form over a frequency selective channel poses an extremely hard equalization problem due to the non-linear complex conjugate operation used in generating the code and in the decoding process . To enable the use of this scheme over frequency selective the channel, the above STBC scheme was first modified in [10] for use in WCDMA, in [11] for use with time domain equalization methods and in [12] for use with frequency domain equalization (FDE) [13, 14]. Below we will describe a scheme for use with FDE that combines equalization, ST decoding, and interference suppression.

2.2. FREQUENCY SELECTIVE CHANNEL AND DATA MODEL

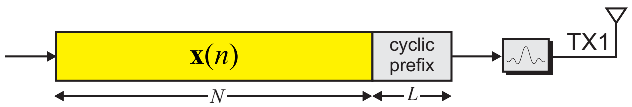

Figure 1: Block Transmission

We consider a burst of N information symbols at time k, $x_k(0), x_k(1), \cdots, x_k(N-1)$ that is transmitted over an additive white Gaussian noise (AWGN) and frequency selective channel $h(t)$ with memory L. Let

$$h(k) = \sum_{l=0}^{L} h_l \delta(k-l) \qquad (6)$$

be discrete time impulse response of the equivalent channel that includes the effects of the wireless propagation channel $h(t)$, the pulse shaping function $g(t)$, and the receive filter impulse $f(t)$. The channel complex tap gains $h_l, l = 0, 1, \cdots, L$ are modelled as Gaussian random variables with zero mean and variance σ_l^2. Without loss of generality, another important assumption we make here is that the complex tap gains are invariant within a data burst, although they may be varying from burst to burst. This assumption relaxes the necessity of time-varying channel models and simplifies the analysis. In cellular systems such as GSM and W-CDMA, the length of a data burst is about of 0.58 and 0.67 ms, respectively. Compared to the coherence time of the channel at 60 MPH mobile velocity and 1.9 GHz carrier frequency, which is approximately 5.6 ms, the burst length is small enough such that the block time-invariant channel model is valid.

We assume that each burst is appended with a cyclic prefix of length L. This is done to eliminate the inter-burst interfer-

remaining received samples as

$$\boldsymbol{y}_k = \boldsymbol{H} \cdot \boldsymbol{x}_k + \boldsymbol{n}_k \qquad (7)$$

where $\boldsymbol{x}_k = [x_k(N-1), x_k(N-2), \cdots, x_k(0)]^T$ is the input symbols vector which is assumed to be zero mean with covariance $E_s \cdot \boldsymbol{I}$ (*i.e.* the input symbols are assumed to be independent and identically distributed (i.i.d)), $\boldsymbol{y}_k = [y_k(N-1), y_k(N-2), \cdots, y_k(0)]^T$ is the received signal vector, and $\boldsymbol{n}_k = [n_k(N-1), n_k(N-2), \cdots, n_k(0)]^T$ is the additive Gaussian noise vector which is assumed to be zero mean with covariance $\boldsymbol{R}_n$. If the noise vector is also white then $\boldsymbol{R}_n = \mathcal{N}_o \cdot \boldsymbol{I}$ [1]. E_s and $\mathcal{N}_o$ are the energy per symbol and the noise power spectral density, respectively. Note that $(.)^T$ and $(.)^*$ donate the transpose and the conjugate transpose, respectively.

In the above setup, $N \times N$ channel matrix $\boldsymbol{H}$ is a circulant matrix [15-17]. A basic result from matrix theory is that a circulant matrix will have the eigenvalue decomposition

$$\boldsymbol{H} = \boldsymbol{Q}^* \boldsymbol{\Lambda}_h \boldsymbol{Q} \qquad (8)$$

where $\boldsymbol{Q}$ is the discrete Fourier transform matrix (DFT) whose (i, n) element is

$$\boldsymbol{Q}(i,n) = \frac{1}{\sqrt{N}} e^{-j2\pi in/N} \qquad 0 \le i, n \le N-1$$

and $\boldsymbol{\Lambda}_h$ is the diagonal eigenvalue matrix whose diagonal is the N point DFT of $h_0, h_1, \cdots, h_L$ [15-17].

2.3. SINGLE CARRIER FREQUENCY DOMAIN EQUALIZER (FDE)

Let us consider the DFT of the received signal vector $\boldsymbol{y}_k$

$$\begin{aligned} \boldsymbol{Y}_k &= \text{DFT}(\boldsymbol{y}_k) = \boldsymbol{Q}\boldsymbol{y}_k \\ &= \boldsymbol{\Lambda}_h \boldsymbol{Q}\boldsymbol{x}_k + \boldsymbol{Q}\boldsymbol{n}_k = \boldsymbol{\Lambda}_h \boldsymbol{X}_k + \boldsymbol{N}_k \end{aligned} \qquad (9)$$

where $\boldsymbol{X}_k$ is DFT of the input symbols vector and $\boldsymbol{N}_k$ is the DFT of the noise vector. Here we used the fact that $\boldsymbol{Q}\boldsymbol{Q}^* = \boldsymbol{I}$ since $\boldsymbol{Q}$ is orthonormal. In general, the noise vector $\boldsymbol{N}_k$ will have a covariance $\tilde{\boldsymbol{R}}_n = \boldsymbol{Q}\boldsymbol{R}_n\boldsymbol{Q}^*$ and when the noise sequence is white, the covariance matrix of the noise will be $\mathcal{N}_o \cdot \boldsymbol{I}$, i.e. the noise vector is still white. The single carrier (SC) FDE is the $N \times N$ matrix filter $\boldsymbol{W}$ that will minimize the mean-squared error (MSE)

$$e^2 = \mathbb{E}\left\{||\boldsymbol{W}^*\boldsymbol{Y}_k - \boldsymbol{X}_k||^2\right\} \qquad (10)$$

It can be shown that the MMSE SC-FDE is given by

$$\boldsymbol{W} = \left(\boldsymbol{\Lambda}_h \boldsymbol{R}_{xx} \boldsymbol{\Lambda}_h^* + \boldsymbol{R}_n\right)^{-1} \boldsymbol{\Lambda}_h \boldsymbol{R}_{xx} \qquad (11)$$

[1] the case when the noise vector is not white will be useful when considering colored co-channel interference.

Figure 2: FFT Based Single Carrier MMSE-FDE

Where $\boldsymbol{R}_{xx} = \mathbb{E}\left\{\boldsymbol{X}_k \boldsymbol{X}_k^*\right\}$ and $\boldsymbol{R}_n = \mathbb{E}\left\{\boldsymbol{N}_k \boldsymbol{N}_k^*\right\}$. Using the assumptions that the input symbols vector and the noise vector are white, the MMSE SC-FDE matrix filter $\boldsymbol{W}$ reduces to:

$$\boldsymbol{W} = \left(\boldsymbol{\Lambda}_h \boldsymbol{\Lambda}_h^* + \frac{1}{\rho} \cdot \boldsymbol{I}\right)^{-1} \boldsymbol{\Lambda}_h \qquad (12)$$

where $\rho = \frac{E_s}{\mathcal{N}_o}$ is the signal to noise ratio (SNR). We immediately notice that, in this case, the matrix filter $\boldsymbol{W}$ is a diagonal matrix whose (i, i) element is given by

$$\boldsymbol{W}(i, i) = \frac{\boldsymbol{\Lambda}_h(i, i)}{|\boldsymbol{\Lambda}_h(i, i)|^2 + \frac{1}{\rho}} \qquad (13)$$

The output of MMSE SC-FDE defined as $\boldsymbol{Z}_k = \boldsymbol{W}^* \boldsymbol{Y}_k$ is transformed back to the time domain (by applying inverse DFT matrix $\boldsymbol{Q}^*$ to yield the *soft decisions* $\hat{\boldsymbol{x}}_k$ for the input symbols vector $\boldsymbol{x}_k$

$$\begin{aligned} \hat{\boldsymbol{x}}_k &= \boldsymbol{Q}^* \boldsymbol{Z}_k \\ &= \boldsymbol{Q}^* \boldsymbol{\Lambda}_h^* \left(\boldsymbol{\Lambda}_h \boldsymbol{\Lambda}_h^* + \frac{1}{\rho} \cdot \boldsymbol{I}\right)^{-1} \boldsymbol{\Lambda}_h \boldsymbol{Q} \boldsymbol{x}_k + \boldsymbol{v}_k \qquad (14) \end{aligned}$$

where $\boldsymbol{v}_k = \boldsymbol{Q} \boldsymbol{W}^* \boldsymbol{N}_k$ is the output noise vector. Final decision can be made on the transmitted symbols by feeding the soft decisions $\hat{\boldsymbol{x}}_k$ to a slicer in the case of uncoded transmission, otherwise they are fed into channel decoder in case of coded transmission. Note that when the number of symbols in each block N is a power of 2, the DFT can be efficiently implemented using the Fast Fourier Transform (FFT). Figure 2 shows and FFT based implementation for the FDE.

3. SPACE-TIME BLOCK CODING FOR FREQUENCY SELECTIVE CHANNELS

We assume that the transmitter is equipped with two transmit antennas, and consider two consecutive blocks $\boldsymbol{x}_k$ and $\boldsymbol{x}_{k+1}$ of N symbols each that need to be transmitted . Let $\boldsymbol{x}_k(n) = x_k(n), n = 0, 1, \cdots, N-1$, the n-th symbol of the k-th block. Given $\boldsymbol{x}_k$ and $\boldsymbol{x}_{k+1}$, we define two complementary sequences $\tilde{\boldsymbol{x}}_k$ and $\tilde{\boldsymbol{x}}_{k+1}$ as

$$\begin{aligned} \tilde{\boldsymbol{x}}_k(n) &= \bar{\boldsymbol{x}}_k(N-n) \qquad 0 \le n \le N-1 \qquad (15) \\ \tilde{\boldsymbol{x}}_{k+1}(n) &= \bar{\boldsymbol{x}}_{k+1}(N-n) \qquad 0 \le n \le N-1 \qquad (16) \end{aligned}$$

where $\bar{(.)}$ denotes the complex conjugate operation for scalars and element by element complex conjugate for vectors and matrices.

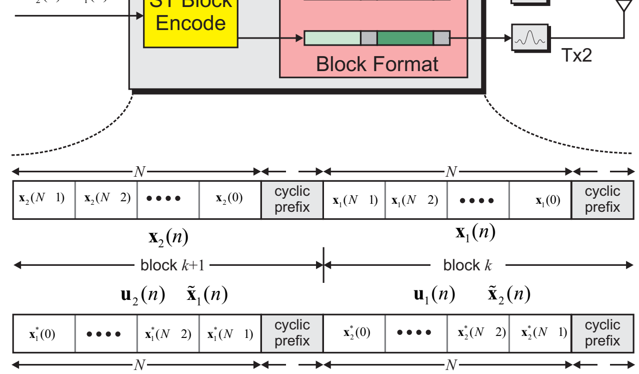

Figure 3: Block Space-Time Coding for frequency selective Channels:Transmitter

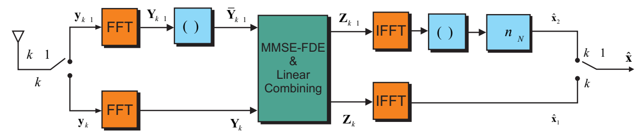

Figure 4: Block Space-Time Coding for frequency selective Channels:Receiver

time k, $\boldsymbol{x}_k$ is appended with the cyclic prefix and transmitted from antenna 1 and $-\tilde{\boldsymbol{x}}_{k+1}$ is appended with the cyclic prefix and transmitted from antenna 2. At time $k+1$, $\boldsymbol{x}_{k+1}$ is appended with the cyclic prefix and transmitted from antenna 1 and $\tilde{\boldsymbol{x}}_k$ is appended with the cyclic prefix and transmitted from antenna 2. Without loss of generality, let us assume that the receiver has one receive antenna and let $\boldsymbol{H}_1$ and $\boldsymbol{H}_2$ be the channel matrices as defined in (**??**) from transmit antenna 1 and transmit antenna 2 to the receive antenna, respectively. Note that both $\boldsymbol{H}_1$ and $\boldsymbol{H}_2$ are also cyclic and hence they will have eigenvalue decomposition similar to that in (8). This transmission scheme is shown in Figure 3. Then, we can write the received signal vectors corresponding to the block transmissions at times k and $k+1$ as

$$\begin{aligned} \boldsymbol{y}_k &= \boldsymbol{H}_1 \boldsymbol{x}_k - \boldsymbol{H}_2 \tilde{\boldsymbol{x}}_{k+1} + \boldsymbol{n}_k \qquad (17) \\ \boldsymbol{y}_{k+1} &= \boldsymbol{H}_1 \boldsymbol{x}_{k+1} + \boldsymbol{H}_2 \tilde{\boldsymbol{x}}_k + \boldsymbol{n}_{k+1} \qquad (18) \end{aligned}$$

Next, we consider the DFT of the received signal vectors $\boldsymbol{y}_k$ and $\boldsymbol{y}_{k+1}$

$$\begin{aligned} \boldsymbol{Y}_k &= \boldsymbol{Q} \boldsymbol{y}_k = \boldsymbol{\Lambda}_1 \boldsymbol{X}_k - \boldsymbol{\Lambda}_2 \bar{\boldsymbol{X}}_{k+1} + \boldsymbol{N}_k \qquad (19) \\ \boldsymbol{Y}_{k+1} &= \boldsymbol{Q} \boldsymbol{y}_{k+1} = \boldsymbol{\Lambda}_1 \boldsymbol{X}_{k+1} + \boldsymbol{\Lambda}_2 \bar{\boldsymbol{X}}_k + \boldsymbol{N}_{k+1} \qquad (20) \end{aligned}$$

where $\boldsymbol{\Lambda}_1$ and $\boldsymbol{\Lambda}_2$ are the DFT of channel impulse response from transmit antenna 1 and 2 to the receive antenna, respectively, $\boldsymbol{X}_k = \mathrm{DFT}(\boldsymbol{x}_k)$, and $\boldsymbol{N}_k = \mathrm{DFT}(\boldsymbol{n}_k)$. Here we used the DFT

$$\boldsymbol{S} \triangleq \begin{bmatrix} \boldsymbol{Y}_k \\ \bar{\boldsymbol{Y}}_{k+1} \end{bmatrix} = \begin{bmatrix} \boldsymbol{\Lambda}_1 & -\boldsymbol{\Lambda}_2 \\ \boldsymbol{\Lambda}_2^* & \boldsymbol{\Lambda}_1^* \end{bmatrix} \begin{bmatrix} \boldsymbol{X}_k \\ \bar{\boldsymbol{X}}_{k+1} \end{bmatrix} + \begin{bmatrix} \boldsymbol{N}_k \\ \bar{\boldsymbol{N}}_{k+1} \end{bmatrix}$$

$$\triangleq \boldsymbol{\Gamma} \cdot \boldsymbol{\mathcal{X}} + \boldsymbol{\mathcal{N}} \quad (21)$$

As before, using the assumptions that the input symbols vector and the noise vector are weight, we can easily show that the combined MMSE SC-FDE / ST Decoder matrix filter $\boldsymbol{\mathcal{W}}$ in this case is

$$\boldsymbol{\mathcal{W}} = \left(\boldsymbol{\Gamma}\boldsymbol{\Gamma}^* + \frac{1}{\rho} \cdot \boldsymbol{I} \right)^{-1} \boldsymbol{\Gamma} \quad (22)$$

Let $\boldsymbol{D} = \boldsymbol{\Lambda}_1\boldsymbol{\Lambda}_1^* + \boldsymbol{\Lambda}_2\boldsymbol{\Lambda}_2^*$. It is straightforward to see that $\boldsymbol{D}$ is an $N \times N$ diagonal matrix whose (i, i) element d_{ii} is given by $|\boldsymbol{\Lambda}_1(i,i)|^2 + |\boldsymbol{\Lambda}_2(i,i)|^2$. Also, let and $\tilde{\boldsymbol{D}} = \boldsymbol{D} + \frac{1}{\rho} \cdot \boldsymbol{I}$. We can easily verify that $\tilde{\boldsymbol{D}}^{-1}\boldsymbol{\Lambda}_j = \boldsymbol{\Lambda}_j\tilde{\boldsymbol{D}}^{-1}$ and $\tilde{\boldsymbol{D}}^{-1}\boldsymbol{\Lambda}_j^* = \boldsymbol{\Lambda}_j^*\tilde{\boldsymbol{D}}^{-1}$. Hence, we can rewrite $\boldsymbol{\mathcal{W}}$ as

$$\boldsymbol{\mathcal{W}} = \begin{bmatrix} \tilde{\boldsymbol{D}}^{-1} & 0 \\ 0 & \tilde{\boldsymbol{D}}^{-1} \end{bmatrix} \begin{bmatrix} \boldsymbol{\Lambda}_1 & -\boldsymbol{\Lambda}_2 \\ \boldsymbol{\Lambda}_2^* & \boldsymbol{\Lambda}_1^* \end{bmatrix} \quad (23)$$

$$= \begin{bmatrix} \boldsymbol{\Lambda}_1 & -\boldsymbol{\Lambda}_2 \\ \boldsymbol{\Lambda}_2^* & \boldsymbol{\Lambda}_1^* \end{bmatrix} \begin{bmatrix} \tilde{\boldsymbol{D}}^{-1} & 0 \\ 0 & \tilde{\boldsymbol{D}}^{-1} \end{bmatrix} \quad (24)$$

$$= \boldsymbol{\mathcal{W}}_c \cdot \boldsymbol{\mathcal{W}}_e \quad (25)$$

Let us consider the output of the matrix filter

$$\boldsymbol{Z} \triangleq \begin{bmatrix} \boldsymbol{Z}_k \\ \bar{\boldsymbol{Z}}_{k+1} \end{bmatrix} = \boldsymbol{\mathcal{W}}^* \cdot \boldsymbol{S}$$

$$= \begin{bmatrix} \tilde{\boldsymbol{D}}^{-1}\boldsymbol{D} & 0 \\ 0 & \tilde{\boldsymbol{D}}^{-1}\boldsymbol{D} \end{bmatrix} \begin{bmatrix} \boldsymbol{X}_k \\ \bar{\boldsymbol{X}}_{k+1} \end{bmatrix} + \begin{bmatrix} \boldsymbol{V}_k \\ \bar{\boldsymbol{V}}_{k+1} \end{bmatrix} \quad (26)$$

The output vector $\boldsymbol{Z}$ is transformed back to the time domain via inverse DFT to yield

$$\begin{bmatrix} \hat{\boldsymbol{x}}_k \\ \hat{\tilde{\boldsymbol{x}}}_{k+1} \end{bmatrix} \triangleq \begin{bmatrix} \boldsymbol{Q}^*\boldsymbol{Z}_k \\ \boldsymbol{Q}^*\bar{\boldsymbol{Z}}_{k+1} \end{bmatrix}$$

$$= \begin{bmatrix} \boldsymbol{Q}^*\tilde{\boldsymbol{D}}^{-1}\boldsymbol{D}\,\boldsymbol{Q} \\ \boldsymbol{Q}^*\tilde{\boldsymbol{D}}^{-1}\boldsymbol{D}\,\boldsymbol{Q} \end{bmatrix} \begin{bmatrix} \boldsymbol{x}_k \\ \tilde{\boldsymbol{x}}_{k+1} \end{bmatrix} + \begin{bmatrix} \tilde{\boldsymbol{n}}_k \\ \tilde{\boldsymbol{n}}_{k+1} \end{bmatrix} \quad (27)$$

In the time domain, $\hat{\boldsymbol{x}}_k$ is the soft-decision for $\boldsymbol{x}_k$ and $\hat{\tilde{\boldsymbol{x}}}_{k+1}$ is time reversed an complex conjugated to give the soft decision for $\boldsymbol{x}_{k+1}$. These soft decisions can be either applied to a slicer to get final (hard) decisions or to the channel decoder in case of coded transmission. ***We can easily verify that a two fold diversity benefit is provided by this scheme***. Figure 4 shows a block diagram of the receiver in this case. We make the following observations. By considering the matrix filter in (25), we notice that it is split into two parts. The first part $\boldsymbol{\mathcal{W}}_c$ represents the decoding operation of the space-time block code, similar to that of the original Alamouti scheme [4]. The second part $\boldsymbol{\mathcal{W}}_e$ represents the MMSE FDE part similar to that in [12]. We make another

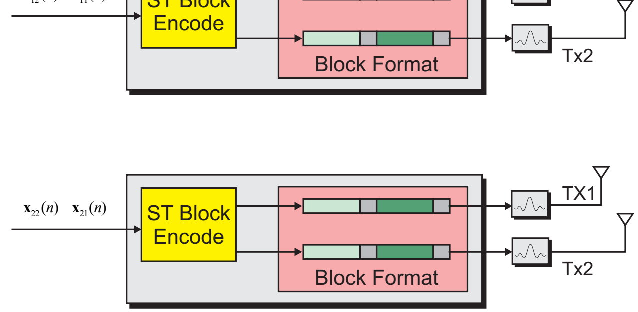

Figure 5: Block Space-Time Coding for frequency selective Channels with Parallel Transmission

$Y_{k+1,n}$ (the n-th DFT point for the N point DFT Of $\boldsymbol{Y}_k$ and $\boldsymbol{Y}_{k+1}$, respectively). These can be written as $\boldsymbol{S}^{(n)} = [\boldsymbol{Y}_k \boldsymbol{Y}_{k+1}]^T$

$$\boldsymbol{S}^{(n)} = \begin{bmatrix} \Lambda_1(n) & -\Lambda_2(n) \\ \Lambda_2^*(n) & \Lambda_1^*(n) \end{bmatrix} \begin{bmatrix} X_{k,n} \\ \bar{X}_{k+1,n} \end{bmatrix} + \begin{bmatrix} N_{k,n} \\ \bar{N}_{k+1,n} \end{bmatrix} \quad (28)$$

This looks exactly like the original Alamouti scheme in [4] except that the spaced-time block coding is now done in the frequency domain. We now, extend these results to the case when co-channel interference exist.

4. COMBINED FD EQUALIZATION AND INTERFERENCE SUPPRESSION

We now consider the scenario where two co-channel users or transmitters simultaneously transmit signals using the transmission scheme described in the previous section. Figure 5 shows a block diagram for this case. Without loss of generality, let $\boldsymbol{x}_{1,k}$, $\boldsymbol{x}_{2,k}$ denote the input information sequence to the first and second transmitters, respectively. Each information sequence, is block encoded and formatted in the same manner as described in the previous section. We also assume that the receiver is equipped with two receive antennas (extending this to the case where more than one receive antenna is straight forward). Then, we can write the DFT of the received signal at antenna j, $j = 1, 2$ due to the block transmissions from transmitter 1 and 2 at times k and $k+1$ as

$$\boldsymbol{S}_j \triangleq \begin{bmatrix} \boldsymbol{Y}_{j,k} \\ \bar{\boldsymbol{Y}}_{j,k+1} \end{bmatrix} = \begin{bmatrix} \boldsymbol{\Lambda}_{1j} & -\boldsymbol{\Lambda}_{2j} \\ \boldsymbol{\Lambda}_{2j}^* & \boldsymbol{\Lambda}_{1j}^* \end{bmatrix} \begin{bmatrix} \boldsymbol{X}_{1,k} \\ \bar{\boldsymbol{X}}_{1,k+1} \end{bmatrix} +$$

$$\begin{bmatrix} \boldsymbol{L}_{1j} & -\boldsymbol{L}_{2j} \\ \boldsymbol{L}_{2j}^* & \boldsymbol{L}_{1j}^* \end{bmatrix} \begin{bmatrix} \boldsymbol{X}_{2,k} \\ \bar{\boldsymbol{X}}_{2,k+1} \end{bmatrix} + \begin{bmatrix} \boldsymbol{N}_{1,k} \\ \bar{\boldsymbol{N}}_{2,k+1} \end{bmatrix} \quad (29)$$

$$\triangleq \boldsymbol{\Gamma}_j \cdot \boldsymbol{\mathcal{X}}_1 + \boldsymbol{L}_j \cdot \boldsymbol{\mathcal{X}}_2 + \boldsymbol{\mathcal{N}}_j \quad j = 1, 2 \quad (30)$$

where, as before, $\boldsymbol{\Lambda}_{i,j}$ is a diagonal matrix whose diagonal elements are the DFT of the channel impulse response from **first** ments are the DFT of the channel impulse response from **second user** transmit antenna i, $i = 1, 2$ to receive antenna j, $j = 1, 2$. These two equations can be put in a matrix form

$$\begin{bmatrix} \boldsymbol{S}_1 \\ \boldsymbol{S}_2 \end{bmatrix} = \begin{bmatrix} \boldsymbol{\Gamma}_1 & \boldsymbol{L}_1 \\ \boldsymbol{\Gamma}_2 & \boldsymbol{L}_2 \end{bmatrix} \begin{bmatrix} \boldsymbol{\mathcal{X}}_1 \\ \boldsymbol{\mathcal{X}}_2 \end{bmatrix} + \begin{bmatrix} \boldsymbol{\mathcal{N}}_1 \\ \boldsymbol{\mathcal{N}}_2 \end{bmatrix} \quad (31)$$

Now, we make the following observations will be useful in deriving the combined FD equalizer/interference suppression filter:

- The matrix $\boldsymbol{\Gamma}_j$ is orthogonal, i.e.

$$\boldsymbol{C}_j = \boldsymbol{\Gamma}_j^* \boldsymbol{\Gamma}_j = \begin{bmatrix} \boldsymbol{D}_j & \boldsymbol{0} \\ \boldsymbol{0} & \boldsymbol{D}_j \end{bmatrix} \quad (32)$$

  where $\boldsymbol{D}_j = \boldsymbol{\Lambda}_{1j}^* \boldsymbol{\Lambda}_{1j} + \boldsymbol{\Lambda}_{2j}^* \boldsymbol{\Lambda}_{2j}$ is and $N \times N$ diagonal matrix whose i-th element is $|\boldsymbol{\Lambda}_{1j}(i,i)|^2 + |\boldsymbol{\Lambda}_{2j}(i,i)|^2$.

- The matrix $\boldsymbol{L}_j$ is orthogonal, i.e.

$$\boldsymbol{E}_j = \boldsymbol{L}_j^* \boldsymbol{L}_j = \begin{bmatrix} \boldsymbol{F}_j & \boldsymbol{0} \\ \boldsymbol{0} & \boldsymbol{F}_j \end{bmatrix} \quad (33)$$

  where $\boldsymbol{F}_j = \boldsymbol{L}_{1j}^* \boldsymbol{L}_{1j} + \boldsymbol{L}_{2j}^* \boldsymbol{L}_{2j}$ is and $N \times N$ diagonal matrix whose i-th element is $|\boldsymbol{L}_{1j}(i,i)|^2 + |\boldsymbol{L}_{2j}(i,i)|^2$.

- The matrices $\boldsymbol{A} = \boldsymbol{\Gamma}_1 \boldsymbol{\Gamma}_2^*$ and $\boldsymbol{B} = \boldsymbol{L}_1 \boldsymbol{L}_2^*$ are orthogonal. That is

$$\boldsymbol{A}\boldsymbol{A}^* = \boldsymbol{\Gamma}_1 \boldsymbol{\Gamma}_2^* \boldsymbol{\Gamma}_2 \boldsymbol{\Gamma}_1^* = \begin{bmatrix} \boldsymbol{D}_\Gamma & \boldsymbol{0} \\ \boldsymbol{0} & \boldsymbol{D}_\Gamma \end{bmatrix} \quad (34)$$

$$\boldsymbol{B}\boldsymbol{B}^* = \boldsymbol{L}_1 \boldsymbol{L}_2^* \boldsymbol{L}_2 \boldsymbol{L}_1^* = \begin{bmatrix} \boldsymbol{F}_L & \boldsymbol{0} \\ \boldsymbol{0} & \boldsymbol{F}_L \end{bmatrix} \quad (35)$$

$$(36)$$

  where $\boldsymbol{D}_\Gamma$ is an $N \times N$ diagonal matrix whose (i,i) entry is $(|\boldsymbol{\Lambda}_{12}(i,i)|^2 + |\boldsymbol{\Lambda}_{22}(i,i)|^2)(|\boldsymbol{\Lambda}_{11}(i,i)|^2 + |\boldsymbol{\Lambda}_{21}(i,i)|^2)$ and $\boldsymbol{F}_L$ is an $N \times N$ diagonal matrix whose (i,i) entry is $(|\boldsymbol{L}_{12}(i,i)|^2 + |\boldsymbol{L}_{22}(i,i)|^2)(|\boldsymbol{L}_{11}(i,i)|^2 + |\boldsymbol{L}_{21}(i,i)|^2)$

- Moreover, $\boldsymbol{A}$ and $\boldsymbol{B}$ have the same structure, and hence $\boldsymbol{A} + \boldsymbol{B}$ will also have the same structure (we will omit the proof for this claim here for lack of space, although it is straightforward proof)

Now define

$$\boldsymbol{R}_j = \boldsymbol{C}_j + \boldsymbol{E}_j \quad j = 1, 2 \quad \text{and} \quad \boldsymbol{\Psi} = \boldsymbol{A} + \boldsymbol{B}$$

4.1. MMSE SC-FD EQUALIZER/INTERFERENCE SUPPRESSION FILTER

As before, here we want to find two $4N \times 4N$ matrix filter $\boldsymbol{\mathcal{W}}_1$ and $\boldsymbol{\mathcal{W}}_2$ such that the mean squared errors (MMSE)

$$e_1^2 = \mathbb{E}\left\{ \left| \boldsymbol{\mathcal{W}}_1^* \cdot \begin{bmatrix} \boldsymbol{S}_1 \\ \boldsymbol{S}_2 \end{bmatrix} - \boldsymbol{\mathcal{X}}_1 \right|^2 \right\} \quad (37)$$

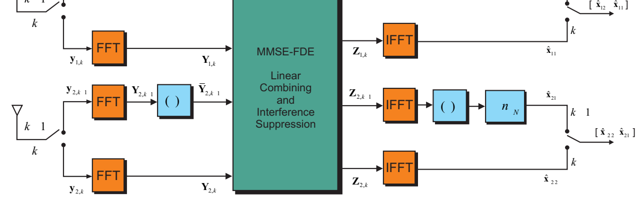

Figure 6: Block Space-Time Coding for frequency selective Channels:Receiver for Parallel Transmission

are minimized. Again, it is easy to show that the matrix filters $\boldsymbol{\mathcal{W}}_1$ and $\boldsymbol{\mathcal{W}}_2$ are given by

$$\boldsymbol{\mathcal{W}}_1 = \begin{bmatrix} \boldsymbol{R}_1 + \frac{1}{\rho}\boldsymbol{I} & \boldsymbol{\Psi} \\ \boldsymbol{\Psi}^* & \boldsymbol{R}_2 + \frac{1}{\rho}\boldsymbol{I} \end{bmatrix}^{-1} \begin{bmatrix} \boldsymbol{\Gamma}_1 \\ \boldsymbol{\Gamma}_2 \end{bmatrix} \quad (39)$$

$$\boldsymbol{\mathcal{W}}_2 = \begin{bmatrix} \boldsymbol{R}_1 + \frac{1}{\rho}\boldsymbol{I} & \boldsymbol{\Psi} \\ \boldsymbol{\Psi}^* & \boldsymbol{R}_2 + \frac{1}{\rho}\boldsymbol{I} \end{bmatrix}^{-1} \begin{bmatrix} \boldsymbol{L}_1 \\ \boldsymbol{L}_2 \end{bmatrix} \quad (40)$$

Let

$$\boldsymbol{Z}_1 = \boldsymbol{\mathcal{W}}_1^* \begin{bmatrix} \boldsymbol{S}_1 \\ \boldsymbol{S}_2 \end{bmatrix} = \begin{bmatrix} \boldsymbol{Z}_{1,k} \\ \boldsymbol{Z}_{1,k+1} \end{bmatrix} \quad (41)$$

The vector $\boldsymbol{Z}_{1,k}$ is transformed back to the time domain via inverse DFT to yield the soft decisions $\hat{\boldsymbol{x}}_{1,k} = \boldsymbol{Q}^* \boldsymbol{Z}_{1,k}$. Similarly, The vector $\boldsymbol{Z}_{1,k+1}$ is transformed back to the time domain via inverse DFT to yield the soft decisions $\tilde{\hat{\boldsymbol{x}}}_{1,k+1} = \boldsymbol{Q}^* \boldsymbol{Z}_{1,k+1}$ (the time reversed and complex conjugated soft decisions for $\boldsymbol{x}_{1,k+1}$). The soft decisions for $\boldsymbol{x}_{2,k}$ and $\boldsymbol{x}_{2,k+1}$ are obtained in a similar fashion. Figure 6 shows a block diagram for the overall receiver in this case.

We will state without proof that the columns matrix filters $\boldsymbol{\mathcal{W}}_1$ and $\boldsymbol{\mathcal{W}}_2$ are orthogonal, that is

$$\boldsymbol{\mathcal{W}}_1^* \boldsymbol{\mathcal{W}}_1 = \Delta_{\mathcal{W}_1} \cdot \boldsymbol{I} \quad \text{and} \quad \boldsymbol{\mathcal{W}}_2^* \boldsymbol{\mathcal{W}}_2 = \Delta_{\mathcal{W}_2} \cdot \boldsymbol{I} \quad (42)$$

The structure of the matrices $\boldsymbol{C}_j, \boldsymbol{E}_j, \boldsymbol{A},$ and $\boldsymbol{B}$ can be used to efficiently compute the $4N \times 4N$ matrix inverse in the above solutions ((39) and (40)) for $\boldsymbol{\mathcal{W}}_1$ and $\boldsymbol{\mathcal{W}}_2$. Nevertheless, for moderate values for the block size $N = 64$ and 128 this would be still a prohibitive complexity (at least from the point of view of the memory requirement if not the number of operations required). The next approach, provides a more efficient way to compute the matrix filters $\boldsymbol{\mathcal{W}}_1$ and $\boldsymbol{\mathcal{W}}_2$.

5. SIMULATION RESULTS

In this section, we present a simulation example (more simulation results will be included in the final submission) for the combined FDE and interference suppression scheme for space-time block

Gaussian pulse shape with 2 samples per baud (the Gaussian pulse was truncated to ± 2 symbols for pulse shaping and ideal low pass filtering was assumed at the receiver. The overall digital FIR channel for a typical urban channel at 3 km/h (TU3) and the linearized Gaussian pulse shaping function will have, in general, 5 taps. However, the 5th tap will have an average energy of -50 dB compared to the tap at delay 0. In our simulations, we ignored this tap and assumed that the channel has 4 taps only as it will have very little effect on the performance. We used a basic slot structure that consisted of 12 training symbols followed by two block each has 64 data symbols (thus the DFT will be a 64 point DFT) and 4 cyclic prefix symbols.

Figure 7 shows the bit error probability as a function of the symbol energy to noise ratio ρ. We show the performance for the case of only one transmitter and one receive antennas and two transmitter and two receive antennas with the combined equalization and interference suppression technique described above. We can easily see that the scheme described above is effective in eliminating the interference. In fact the performance is slightly better due to the MMSE nature of the approach.

6. SUMMARY

In this paper we presented a combined frequency domain equalization and interference suppression technique for space-time bl[illegible] Re-
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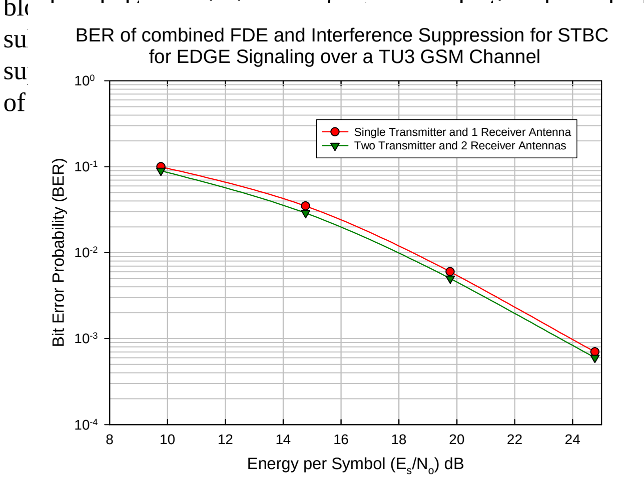

Figure 7: BER of combined FDE and Interference Suppression for STBC for EDGE Signaling over a TU3 GSM Channel

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