EVALUATION OF AN ORTHOGONAL SFBC CHANNEL ESTIMATION SCHEME FOR MIMO OFDM WIRELESS SYSTEMS

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Abstract: This paper presents the design and evaluation of a channel estimation scheme that is efficient by means of both the mean square error (MSE) of channel estimation/tracking and its incorporation in a real MIMO system. The evaluation has been performed over the spatial channel model developed for MIMO simulations according to 802.16e case of 3GPP.25.996, taking also into account all IF and RF stages in the communication chain. Orthogonality has been applied in space-frequency dimension for both preamble and pilot symbols, as well as for the data symbols, with the application of Alamouti's scheme. In 4G multicarrier systems that use space-time-frequency coding, orthogonal design turns into a key factor for the performance of the system since the channel has to remain about constant during the transmission of one orthogonal block, something which becomes quite challenging in highly time-variant propagation channels. Furthermore, space-frequency block coding (SFBC) becomes more efficient as the number of subcarriers increases (802.16e, 802.20, etc). The modified channel estimation scheme applied to MIMO transceiver is also efficient in minimization of the processing requirements at the receiver side by estimating only those channel properties that have been changed assuming that the general channel conditions (low/high mobility) are known. The results presented refer to the normalized MSE of the channel estimator and the overall performance evaluation (BER) of the system in various propagations channels, data rates and forward error correction modes.

1 INTRODUCTION

Wireless broadband systems have to support services that demand information transmission with very high data rates over the wireless propagation medium. It has been proven (Adjoudani et al., 2003) that the use of multiple antenna elements at both ends of a wireless link offers both capacity gain and improvement of robustness and reliability. Therefore, multiple input multiple output (MIMO) architecture has been incorporated in the development of various wireless systems operating in challenging propagation environments. In addition, the various sources of diversity should be properly exploited by means of coding and transmission scheme (Tarokh et al., 1998). Temporal diversity is realized through FEC schemes (scrambling, Reed-Solomon, convolution, interleaving). Frequency diversity is exploited by orthogonal frequency division multiple (OFDM) access systems and spatial diversity is obtained by multiple antennas. Furthermore, the above diversity options are combined in space-time, space-frequency, or space-time-frequency codes where the orthogonal property and the ability to be preserved through the propagation channel is a key factor in the total system performance.

Optimum space-frequency coding schemes that maximize the diversity gain have been proposed in (Bolcskei and Paulraj, 2001), but the processing requirements at the receiver are quite high. Space-time block codes proposed by Alamouti and extended in (Tarokh et al., 1999), provide a simple transmit diversity scheme with maximum diversity in flat fading MIMO channels, which are assumed about constant during the transmission of one orthogonal block. OFDM provides flat fading channel for each subcarrier, making space-time block codes well suited for OFDM systems assuming that the channel coefficients remain constant during two or more consecutive OFDM symbols.

In propagation environments with high Doppler shift loss of orthogonality, that is assembled in space-time domain, becomes possible, whilst in space-frequency structure the orthogonal design is not distorted. Furthermore, as the number of subcarriers increases, for a given total bandwidth of transmission, the prob-

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ability of non constant affected neighbor subcarriers in severe frequency selective channels becomes quite small. Also, the greater number of subcarriers (WiMAX vs WiFi), the larger range is achieved, since larger delay spreads are tolerated (up to 10 times for WiMAX with respect to WiFi).

Channel estimation is a crucial design parameter in the performance of a real system since it has to estimate, track and compensate all channel distortions as well as the distortions caused in RF stages in transmitter and receiver units. Especially, in a MIMO-OFDM system the channel distortion is described by a complex factor per subcarrier requiring from the estimator $(N_{sym} \cdot N_c \cdot M_T \cdot M_R)$ estimations/compensations per frame (N_{sym}=number of OFDM symbols, N_c=number of subcarriers per OFDM symbol, M_T=number of transmit antennas, M_R=number of receive antennas). Such an operation can be particularly demanding in terms of computational effort (Li et al., 1999). Following the design of space-frequency orthogonality also for the preamble and pilot transmission the proposed approach causes a pilot overhead of 3.12% per OFDM symbol in which only phase estimation is used for the pilots that have been carefully placed in predefined positions.

In this paper, initially (section 2) the system model is depicted giving a detailed insight of transceiver architecture, as well as the channel models used for the evaluation. In section 3, the channel estimation is described giving rise to all advantages and trade offs caused by the low computational complexity at the receiver side. Finally (section 4), evaluation results of the channel estimation (MSE) and the overall system performance (BER) are given for 2x1 and 2x2 cases evaluated in various propagation models according to 802.16e (Mobile Broadband Wireless Access, MBWA) case of 3GPP.25.996 (3GPP, 2003-2009) using various FEC codes and mapping formats following the 802.16-2004 standard.

2 SYSTEM ARCHITECTURE

2.1 Transmission Scheme

The MIMO-OFDM transmitter with two branches employing space-frequency block coding (SFBC) is shown in fig.1. A binary data block $D[k]$ of k bits is scrambled, encoded by a concatenated Reed-Solomon and Convolutional encoder, followed by a puncturer and an interleaver. The resultant bit stream is mapped using a set of predefined constellation diagrams (BPSK-1/2, QPSK-1/2, QPSK-3/4, 16QAM-1/2, 16QAM-3/4, 64QAM-2/3, and 64QAM-3/4) giving a symbol stream $S[m]$ of m symbols. The same procedure is followed as well as for the frame control header (FCH) (IEEE, 2004) with fixed QPSK mapping. These symbol streams are then frequency multiplexed with 8 pilot symbols and the output is SFB coded based on Alamouti's scheme. The output symbols are packetized in blocks of 200 symbols, zero padded and inserted in a 256-IFFT OFDM modulator. Subsequently, the outputs are time multiplexed with the OFDM output of the SFB coded preamble symbols P. The produced digital signals at the two chains are converted to analog ones and up-converted to the carrier frequency through RF stages with common oscillator. Hence, time synchronization and frequency offset compensation at the receiver are exactly the same as in the case of a SISO system.

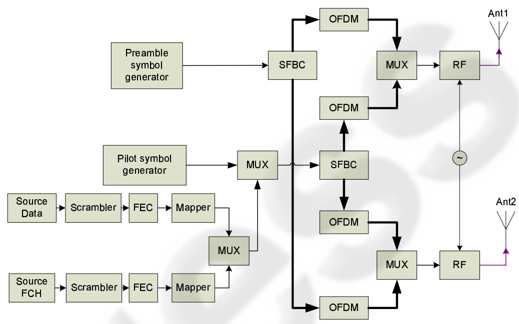

Figure 1: Transmitter Block Diagram.

2.2 Reception Scheme

At the receiver an equivalent procedure is followed. Alamouti's encoding scheme (Alamouti, 1998) (applied on a basis of 2 neighbor subcarriers) offers a simple combining scheme assuming that the channel estimates are available. Hence, extra attention has been paid in the channel estimation stage as shown in fig.2 (in which only one part of the 2×2 system has been depicted). The received signal at the frequency domain, either for data symbol stream, or for preamble symbol stream at the receiver chain is expressed as follows:

$$R_i^{(m_R)} = \sum_{j=1}^{M_T} H_i^{(m_R j)} \cdot S_i^{(j)} + N_i \qquad (1)$$

where i corresponds to the subcarrier index at the m_R-th receive antenna, j corresponds to the transmitter antenna index out of M_T transmit antennas ($M_T = 2$), N_i corresponds to additive complex Gaussian noise per subcarrier i with zero mean and variance σ_n^2. Also, $H_i^{(m_R j)}$ corresponds to the channel coefficient between the j-th transmit antenna and the m_R-th receive antenna for the i-th subcarrier (Stuber et al., 2004). The combiner outputs are fed to

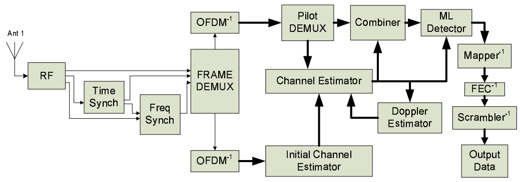

Figure 2: Receiver Block Diagram.

the maximum likelihood (ML) detector which estimates the most probable symbol stream according to the equation:

$$J = \arg\min_{S_k \varepsilon C} = \sum_{k=1}^{N_c-1} \left\| R_k - \hat{H} \cdot S_k \right\|^2 \qquad (2)$$

where $C = [S_0, S_1, , S_{N-1}]$. Time and frequency synchronization are performed based on the time-correlation properties of the relative preamble (fig.3). The correction factor is fed back to the oscillator causing a delay. During this session, the analog automatic gain control (AGC) is adapted and remains constant during the subsequent frame period.

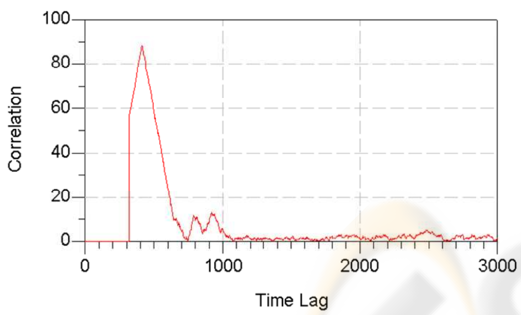

Figure 3: Correlation of time synchronization preamble at the receiver.

2.3 Channel Models

The evaluation of SFBC MIMO-OFDM scheme has been performed over realistic conditions taking into account not only the channel propagation characteristics, like time variability (Doppler shift) and multipath propagation (frequency selectivity), but also the correlations between the antennas at the transmitter and the receiver (described by Tx and Rx correlation matrices). The physical parameters used for link level modelling have been based on pedestrian level of mobility with line of sight (Rice factor K=6dB) according to the relative standard. Also, the proposed correlation values have been taken into account for an inter-element spacing of $\lambda/2$, where λ denotes the wavelength.

3 CHANNEL ESTIMATION

Channel state information (CSI) is acquired by the receiver on a two-step procedure (fig.2) whereas no CSI is fed back to the transmitter, establishing an open-loop system with equal transmission power on the antennas. The first step in the channel estimation procedure employs the OFDM preamble symbols which are orthogonal on a SFBC subcarrier basis. The estimation has been implemented using a MMSE approach. In the second step, the pilot symbols are used only for the phase estimation compensating the Doppler distortion. Then, using interpolation the correction factor for each subcarrier is taken into account in the preamble based estimation. The final channel estimates are used for both channel compensation and soft decision stages. Furthermore, Doppler estimation gives a figure of merit of the channel's time variation which can be potentially used to increase the number of pilot data in time dimension or for adapting a higher order interpolation filter.
The MIMO channel estimation problem can be decomposed into several MISO channel estimations in parallel (Stuber et al., 2004). The initial channel estimation is based on the preamble OFDM symbols that have been transmitted from the 2 antennas in an orthogonal space-frequency format. Taking into account only the adjacent subcarriers i and $i+1$ that convey pilot information in an orthogonal format it will be:

$$\begin{aligned} R_i^{(1)} &= H_i^{(11)} \cdot S_i^{(1)} + H_i^{(12)} \cdot S_i^{(2)} \\ R_{i+1}^{(1)} &= H_{i+1}^{(11)} \cdot S_{i+1}^{(1)} + H_{i+1}^{(12)} \cdot S_{i+1}^{(2)} \\ R_i^{(2)} &= H_i^{(21)} \cdot S_i^{(1)} + H_i^{(22)} \cdot S_i^{(2)} \\ R_{i+1}^{(2)} &= H_{i+1}^{(21)} \cdot S_{i+1}^{(1)} + H_{i+1}^{(22)} \cdot S_{i+1}^{(2)} \end{aligned} \qquad (3)$$

where $R_i^{(m_R)}$ is the received signal at m_R-th receive antenna in i-th subcarrier, $H_i^{(m_R m_T)}$ is the channel coefficient from the m_T-th transmit antenna to m_R-th receive antenna in i-th subcarrier, and $S_i^{(m_T)}$ is the transmitted symbol from m_T-th antenna in i-th subcarrier. Since, the Alamouti scheme has been adapted in space-frequency dimension the transmitted symbols in i-th and $(i+1)$-th subcarriers will be: $P_a = S_i^{(1)}$, ${P_a}^* = {S_{i+1}}^{(1)}$, $P_b = {S_i}^{(2)}$, $-{P_b}^* = {S_{i+1}}^{(2)}$, where $(\cdot)^*$ denotes the complex conjugate operation. In addition, the channel is assumed constant for the subcarriers i and $i+1$ giving:

$$\begin{aligned} H_i^{(11)} = H_{i+1}^{(11)} = H^{(11)}, &\quad H_i^{(21)} = H_{i+1}^{(21)} = H^{(21)} \\ H_i^{(12)} = H_{i+1}^{(12)} = H^{(12)}, &\quad H_i^{(22)} = H_{i+1}^{(22)} = H^{(2)} \end{aligned} \qquad (4)$$

Hence, eq.3 becomes:

$$\left.\begin{array}{r} R_i^{(1)} = H^{(11)} \cdot P_a + H^{(12)} \cdot P_b \\ R_{i+1}^{(1)} = H^{(11)} \cdot P_a^* - H^{(12)} \cdot P_b^* \\ R_i^{(2)} = H^{(21)} \cdot P_a + H^{(22)} \cdot P_b \\ R_{i+1}^{(2)} = H^{(21)} \cdot P_a^* - H^{(22)} \cdot P_b^* \end{array}\right\} \Rightarrow$$

$$\begin{array}{r} R_i^{(1)} = H^{(11)} \cdot P_a + H^{(12)} \cdot P_b \\ R_{i+1}^{(1)^*} = H^{(11)^*} \cdot P_a - H^{(12)^*} \cdot P_b \\ R_i^{(2)} = H^{(21)} \cdot P_a + H^{(22)} \cdot P_b \\ R_{i+1}^{(2)^*} = H^{(21)^*} \cdot P_a - H^{(22)^*} \cdot P_b \end{array} \tag{5}$$

Expressing the above formula in matrix notation, it will be:

$$\begin{bmatrix} R_i^{(1)} \\ R_{i+1}^{(1)^*} \\ R_i^{(2)} \\ R_{i+1}^{(2)^*} \end{bmatrix} = \begin{bmatrix} H^{(11)} & H^{(12)} \\ H^{(11)^*} & -H^{(12)^*} \\ H^{(21)} & H^{(22)} \\ H^{(21)^*} & -H^{(22)^*} \end{bmatrix} \cdot \begin{bmatrix} P_a \\ P_b \end{bmatrix}$$

$$\Leftrightarrow \mathbf{R}_p = \mathbf{H}_p \cdot \mathbf{S} \tag{6}$$

where the index p denotes the processed nature of the relative receive vector and the channel matrix. The matrix $\mathbf{H}_p$ has unitary properties, i.e.

$$\begin{array}{c} \mathbf{H}_p^H \cdot \mathbf{H}_p = \\ \left(\left|H^{(11)}\right|^2 + \left|H^{(12)}\right|^2 + \left|H^{(21)}\right|^2 + \left|H^{(22)}\right|^2\right) \cdot \mathbf{I}_2 \\ = \mu \cdot \mathbf{I}_2 \end{array} \tag{7}$$

where $\mathbf{I}_2$ is the identity matrix of dimension 2, and $(\cdot)^H$ denotes the conjugate transpose matrix operation. For the case of perfect channel knowledge at the receiver, the output of the combiner representing the decision statistics (soft decisions) are given as follows:

$$\begin{array}{c} \tilde{\mathbf{S}} = \frac{1}{\mu}\mathbf{H}_p^H \cdot \mathbf{R}_p = \frac{1}{\mu}\mathbf{H}_p^H \cdot (\mathbf{H}_P \cdot \mathbf{S} + \mathbf{N}) \Leftrightarrow \\ \tilde{\mathbf{S}} = \mathbf{S} + \mathbf{N}_m \end{array} \tag{8}$$

which indicates that except the noise term the symbols have been recovered at the combiner's output. In a real system the MIMO channel has been estimated at the receiver non perfectly giving the following soft decision metric:

$$\begin{array}{c} \tilde{\mathbf{S}} = \frac{1}{\mu}\hat{\mathbf{H}}_p^H \cdot \mathbf{R}_p = \frac{1}{\mu}\hat{\mathbf{H}}_p^H \cdot (\mathbf{H}_p \cdot \mathbf{S} + \mathbf{N}) \Leftrightarrow \\ \tilde{\mathbf{S}} = \mathbf{H}_{res} \cdot \mathbf{S} + \mathbf{N}_m \end{array} \tag{9}$$

where the subscript res indicates the residual channel effect that have to be compensated by the ML decoder. In case of significant temporal channel variation during the transmission of an orthogonal scheme, the orthogonality is lost causing intersymbol interference. Hence, in order to preserve the orthogonality, the sampling theorem has to be applied determining the relative distances in time and frequency dimension that the pilots have to be placed. In space-time block codes the distance is proportional to the coherence time, while in space-frequency block coding is proportional to the coherence bandwidth. The channel tracking is performed through the phase estimation at the pilot positions based on the ML criterion according to the equation:

$$\hat{\theta}_c = \arg\min_{\angle\hat{H}} \sum_{i=1}^{8} \hat{H}_i \cdot \hat{H}_i^{(pre)*} \tag{10}$$

where $\hat{H}_i$ and $\hat{H}_i^{(pre)}$ are the current channel estimates at pilot positions and the estimates at the same subcarrier i during the preamble OFDM symbol respectively. The estimated phase difference updates the preamble based estimation. The final channel estimates at the pilot positions are interpolated (linearly in our case) in order to obtain the estimates in all subcarriers. A sample of the channel compensated symbols, just before the detector, is given in fig.4, where the blue dots are the transmitted ones before the SFBC encoder. The specific snapshot corresponds to a 2×2 MIMO-OFDM system with 16QAM and total coding (RS-CC) 3/4 in a channel type A (802.16e, 3GPP standard) operating in $E_b/N_o = 15$dB. In addition, QPSK modulation is observed due to the FCH symbols. Quantitatively, the maximum achievable diversity order is a product of the number of transmit antennas, the number of receive antennas, and the number of resolvable paths (Bolcskei and Paulraj, 2001).

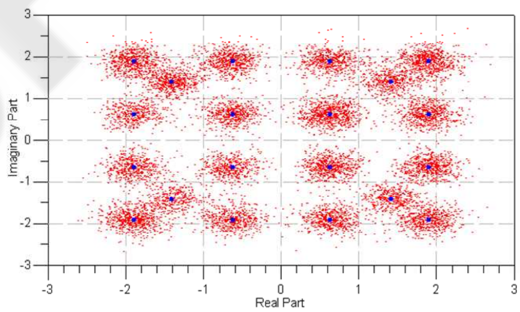

Figure 4: Constellation map of the channel compensated symbols (red) with respect to the transmitted ones (blue).

4 EVALUATION RESULTS

To study the impact of realistic channel estimation architecture on MIMO-OFDM performance, a 2×1 and 2×2 MIMO-OFDM system with orthogonal space frequency design has been designed, modelled and simulated, taking into account all stages in RF, IF and baseband level, as well as their relative requirements.

The simulated system achieves information data rates of 6.9Mbps at BPSK-1/2, of 13.8Mbps at QPSK-1/2, of 20.7Mbps at QPSK-3/4, of 27.7Mbps at 16QAM-1/2, of 41.5Mbps at 16QAM-3/4, of 55.3Mbps at 64QAM-2/3, and of 62.2Mbps at 64QAM-3/4 in a frequency bandwidth of 20MHz at a center frequency of 5.2GHz. The relative frequency spectrum at a transmit antenna and a receive antenna is given in fig.5 for the 16QAM-3/4 case in a propagation channel of type A (802.16e) and for $E_b/N_o = 10$dB. The frequency selectivity is obvious, as well as the noise distortion is becoming quite severe.
The channel adaptation is based on 8 pilot symbols per OFDM symbol placed in blocks of 2 adjacent subcarriers. OFDM stages are based on a 256-point FFT/IFFT with cyclic prefix of 1/4. Each frame carries 2400 information bits and the evaluation is performed on the basis of achieving BER estimation relative variance of 0.0001 with an upper limit of 1000 frames. The performance of the channel estimator

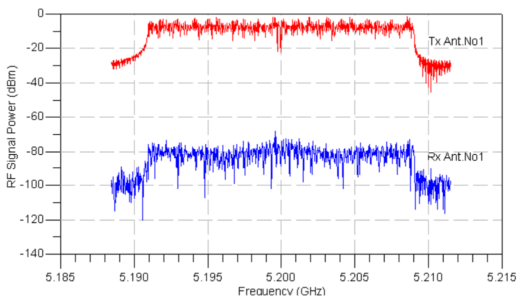

Figure 5: Frequency spectrum at the Tx antenna (red) and the corresponding Rx one (blue) for 2×2 OFDM system.

with respect to the actual channel propagation conditions is based on normalized mean square error (NMSE), according to the formula (11) and the results are given in fig.6 for the 2×2 MIMO-OFDM case and for various propagation channels and data rates. Based on these results the channel estimator is characterized by an irreducible error floor at 3dB, achieving the limit at $E_b/N_o = 13$dB for all schemes and channel conditions tested.

$$NMSE = \frac{E\left|H_i^{(m_R m_T)} - \hat{H}_i^{(m_R m_T)}\right|^2}{E\left|H_i^{(m_R m_T)}\right|^2} \quad (11)$$

The total system performance of 2×1 and 2×2 MIMO-OFDM system has been evaluated based on the information bit error rate giving an insight at the sensitivity of the channel estimation errors in the efficiency of the system. For the probability of error P_e measurement to be statistically significant, the relative variance R_{var} of P_e is taken into account for N_t transmitted bits indicating the confidence interval of P_e. In fig.7 the probability of error has been produced

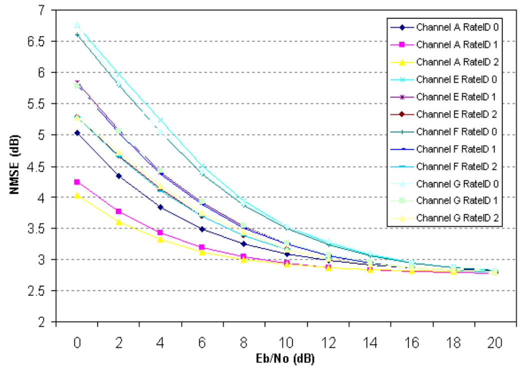

Figure 6: Normalized MSE of the overall channel estimation in various propagation channels and data rates for 2×2 MIMO-OFDM case.

for 2×1 case, in channel type A of 802.16e for all the modulation formats and coding modes. It is worth noticing that for the information data rate of 6.9Mbps a value of $E_b/N_o = 4.5$dB is enough to achieve $Pe = 10^{-4}$, while the same value for 62.2Mbps requires almost 10dB increase in E_b/N_o. In fig.8 the

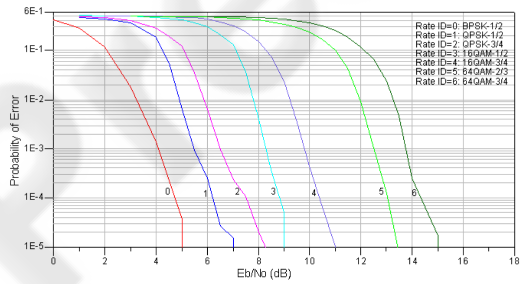

Figure 7: Probability of error for a 2×1 OFDM system in propagation channel type A.

performance of a 2×1 OFDM system has been evaluated for various channel types of 802.16e standards. The system performs better for channel type A, but for more demanding channels (E,F,G), with higher mobility or frequency selectivity, the system fails to support increased data rates in relatively small values E_b/N_o. In fig.9 the performance of 2×2 MIMO OFDM system has been depicted for various propagation channels and data rates indicating that the proposed scheme is quite efficient in propagation channels that follow the channel models of type A, E, or G. Furthermore, the proposed scheme is characterized by bit error rate achievable floors as E_b/N_o increases in channel models with increased frequency selectivity.

Finally, for comparison reasons, the system has been also implemented with space-time block coding (STBC) and for the 2×1 OFDM case the results are given in fig.10. The performance gain for the SFBC

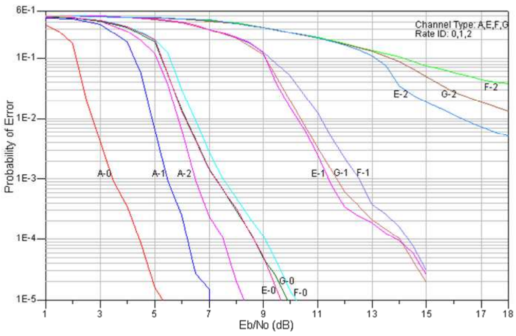

Figure 8: Perfromance of a 2×1 OFDM system in various propagation channels and data rates.

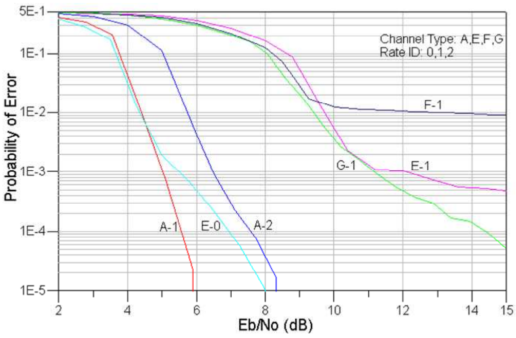

Figure 9: Performance evaluation of a 2×2 OFDM system.

scheme at a probability of error $P_e = 10^{-3}$ is about 2dB for a data rate of 20.7Mbps in a channel of type A, whilst for 13.8Mbps the gain is about 1.5dB in channels of types G and F.

5 CONCLUSIONS

In this paper a MIMO multicarrier system has been designed and evaluated giving rise to space-frequency orthogonality. In addition, all the RF stages at the transmitter and the receiver were taken into account approaching a real architecture as close as possible. An efficient channel estimation scheme was incorporated in the system achieving not only a good efficiency, but also low computational requirements since the processing is performed in one OFDM symbol and the channel estimation during the frame is limited only in the varying propagation characteristics. The overall system performance of SFBC MISO/MIMO OFDM was evaluated for various propagation channels resulting in very good performances for propagation conditions that are characterized by low frequency selectivity, since the pilot overhead is only 3.12%.

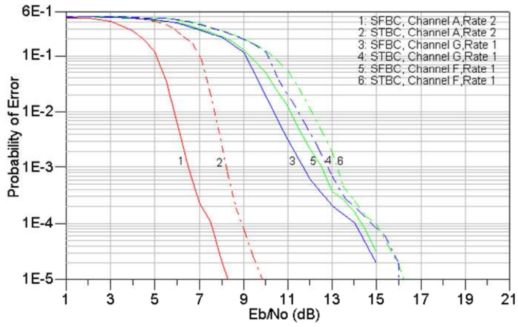

Figure 10: Comparison between SFBC (solid) and STBC (dash) schemes for 2×1 case.

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