A Time Domain Channel Estimation Scheme for Equalize-and-Forward Relay-Assisted Systems

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Abstract – **In this paper we propose a scheme with a dedicated relay node for an OFDM–based system. We consider an antenna array at the base station and a single antenna at both the user terminal and the relay node. The relay–assisted protocol considered is equalize–and–forward. The proposed scheme requires the computation of parameters which depend on unknown variables at the destination and we show that under some assumptions the scheme is suitable to provide all those requirements. Also, we show that since the pilots at the relay node are properly designed the TD–MMSE estimator can still provide the required channel estimate. The system performance is evaluated by considering a typical pedestrian scenario based on WiMAX specification. The results show that the degradation in the estimator performance is negligible.**

Keywords – OFDM systems, cooperative schemes, Pilot–aided channel estimation, equalize–and–forward/estimate–and–forward relaying protocol.

I. INTRODUCTION

Wireless communications will be a key element for the information society and current research focuses on providing high throughput pervasive coverage and fairness in access. One solution to achieve high throughput is to make use of antenna arrays in order to exploit benefits from the spatial diversity, this scheme may also be referred as MIMO systems. However, integration of multiple antenna elements is difficult especially in the terminal because of the size limitations, and the reduced spacing does not guarantee decorrelation between the channels. Cooperative diversity through the use of dedicated relays or terminals with relaying capabilities has been emerging as a promising solution which enables the single antenna devices to generate a virtual antenna–array (VAA) in a multi–user environment. Furthermore the use of relays may lead to expanded coverage, system wide power savings and better immunity against signal fading [1].

A significant number of techniques have been reported in the literature [1]–[4], that have shown the potential of cooperative diversity. Although there is some work on the impact of imperfect channel state information [5], [6] most of the work reported considers that perfect channel state information (CSI) is available. Nevertheless, to exploit the full potential of cooperative communication good estimates for the different channel impulse responses are required, and the impact needs to be analyzed as well as derivation of new techniques to address the specificities of such systems.

Channel estimation for cooperative communication depends on the employed relaying protocol. Most of the protocols studied in cooperative communications fall in three main categories: decode and forward (DF) [4] when the relay has the capability to regenerate and re–encode the whole frame; Amplify–and–Forward (AF) [4] where only amplification takes place; and what we term Equalize–and–Forward (EF) where more sophisticated filtering operations are used [7].

While in the case of DF, the effects of the $B \rightarrow R$ (base station–relay node) channel are reflected in the error rate of the decided frame and therefore the samples received at the destination only depend on the $R \rightarrow U$ (relay node–user terminal) channel the situation is different with AF and EF. In the former case AF, the $B \rightarrow R \rightarrow U$ (base station–relay node–user terminal) channel is simply the cascaded of the $B \rightarrow R$ and $R \rightarrow U$ channels, which has a larger delay spread than the individual channels [5] and additional noise introduced at the relay. This model has been addressed in [5], [8]–[10], with the authors of [5] showing that with some considerations classical estimators can be used to estimate the compound channel. In [9] the authors propose a channel estimation scheme to disintegrate the compound channel, which implies insertion of pilots at the relay, while [10] shows that although the LMMSE outperforms the LS and Lr–MMSE estimators its implementation is very complicated.

However to the best of our knowledge channel estimation for EF strategies that use Alamouti coding from the base station (BS) to relay node (RN) to provide spatial diversity has not been considered from the channel estimation point of view in the literature. Such a scenario is of practical importance in the downlink of cellular systems since the BS has less constraints than user terminals (or terminals acting as relays) in what concerns antenna integration, and therefore it is appealing to consider the use of multiple antennas at the BS improving through the diversity achieved the performance in the $B \rightarrow R$ link.

However due to the Alamouti coding–decoding operations, the channel $B \rightarrow R \rightarrow U$ is not just the cascade of the $B \rightarrow R$ and $R \rightarrow U$ channels, but a more complex channel. The channel estimator at the UT needs therefore to estimate this equivalent channel in order to do the equalization.

The derivation of proper channel estimators for this scenario is the objective of this paper. We consider a scenario with a multiple antenna BS employing the EF protocol and Alamouti coding [7], and propose a time domain pilot–based scheme to estimate the channel impulse response. The $B \rightarrow R$ channels are estimated at the RN and the information about the equivalent channel inserted in the pilot positions. At the user terminal (UT) a time domain MMSE estimator, estimates the equivalent channel from the source to destination, taking into account the Alamouti equalization

978-1-4244-3574-6/10/$25.00

performed at the RN. The estimator scheme we consider operates in the time domain because of the reduced complexity when compared against its implementation in frequency domain (e.g. [11]).

The remaining of this paper is organized as follows. In Section II we present the system description with the designed relay–assisted scheme. Section III brings an overview of the TD–MMSE estimator and we present the analysis regarding the relay–assisted scheme and channel estimation, and evaluate the impact of simplications that can be made. The system performance is evaluated in terms of BER and MSE in Section IV. Finally, Section V concludes this work.

II. SYSTEM DESCRIPTION

Throughout the text index n and k denote time and frequency domain variables, respectively. Complex conjugate and the statistical expectation operator are correspondently denoted by $(\cdot)^*$ and $\mathrm{E}\{\cdot\}$. $\mathcal{N}(m,\sigma^2)$ denotes a complex Gaussian random variable with mean m and variance σ^2. $|\cdot|$ denotes absolute value and $\mathbf{I}_Q$ denotes the identity matrix of size Q. Regular small letters denote variables in frequency domain while boldface small and capital letters denote matrices and vectors, respectively in frequency domain (FD) as well. Variables, vectors or matrices in time domain (TD) are denoted by $(\sim)$. All estimates are denoted by $(\wedge)$.

A. Scenario Description

The studied scenario, depicted in Figure 1, corresponds to the proposed relay–assisted scheme for downlink OFDM–based system. The BS is equipped with two antennas and both the UT and RN are single antenna devices.

The channels are assumed to exhibit Rayleigh fading, and since the RN and UT are mobile the Doppler's effect is considered in all channels.

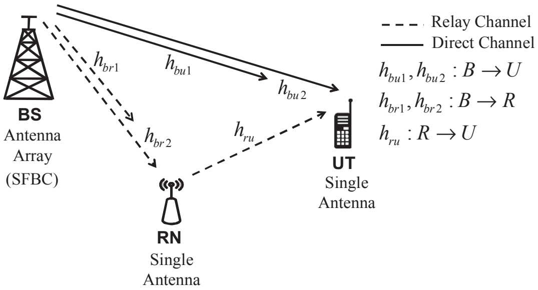

Figure 1 – MISO cooperative scenario.

B. The Equalize–and–Forward (EF)Relaying Protocol

The half–duplex EF protocol considered requires two phases:

1) *Phase I:* the BS broadcast the data to the RN and UT. The signals follow the direct and relay paths simultaneously and independently.

2) *Phase II:* the BS is idle and the RN retransmits to the UT the equalized signal which was received at the RN from the BS in phase I.

C. The Cooperative Scheme

Figure 2 shows the block diagram of the studied scenario with indication of the signals at the different points. The superscripts $(\cdot)^{(1)}$ and $(\cdot)^{(2)}$ denote the first and the second phase of the EF protocol, respectively. In the different variables used, the subscripts u, r and b mean that these variables are related to the UT, RN and BS, respectively. The soft–decision block corresponds to the SFBC–decode of the data d which per subcarrier k is coded at the BS, according to the Table 1, assuming the transmitted power as unitary.

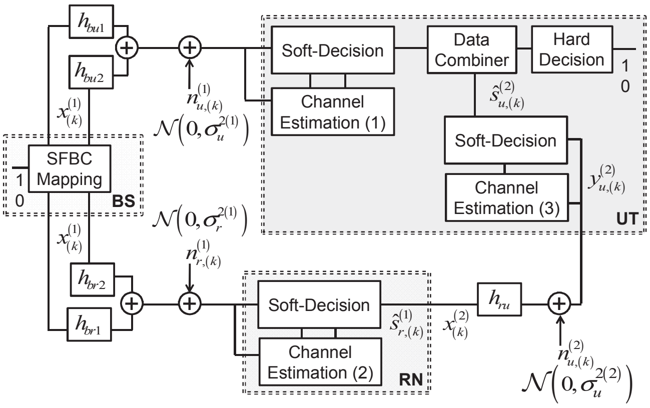

Figure 2 – The block diagram of the relay–assisted scenario.

TABLE 1 – TWO TRANSMIT ANTENNA SFBC MAPPING

Subcarrier	Antenna 1	Antenna 2
k	$d_{(k)}/\sqrt{2}$	$-d^*_{(k+1)}/\sqrt{2}$
$k+1$	$d_{(k+1)}/\sqrt{2}$	$d^*_{(k)}/\sqrt{2}$

In phase I of the relay protocol, the received signals in FD at the UT and RN, in (k) and $(k+1)$ subcarriers are given by respectively (1) and (2):

$$\begin{cases} y^{(1)}_{u,(k)} = \dfrac{1}{\sqrt{2}}\left(h_{bu1,(k)}d_{(k)} - h_{bu2,(k+1)}d^*_{(k+1)}\right) + n^{(1)}_{u,(k)} \\ y^{(1)}_{u,(k+1)} = \dfrac{1}{\sqrt{2}}\left(h_{bu2,(k)}d^*_{(k)} + h_{bu1,(k+1)}d_{(k+1)}\right) + n^{(1)}_{u,(k+1)} \end{cases}, \tag{1}$$

$$\begin{cases} y^{(1)}_{r,(k)} = \dfrac{1}{\sqrt{2}}\left(h_{br1,(k)}d_{(k)} - h_{br2,(k+1)}d^*_{(k+1)}\right) + n^{(1)}_{r,(k)} \\ y^{(1)}_{r,(k+1)} = \dfrac{1}{\sqrt{2}}\left(h_{br2,(k)}d^*_{(k)} + h_{br1,(k+1)}d_{(k+1)}\right) + n^{(1)}_{r,(k+1)} \end{cases}, \tag{2}$$

where $n^{(1)}_{u,(k)}$ and $n^{(1)}_{r,(k)}$ are the additive white Gaussian noise with zero mean and have unitary variance $\sigma_u^{2(1)}$ and $\sigma_r^{2(1)}$, respectively.

The RN performs Alamouti equalization resulting in the soft–samples:

$$\begin{cases} \hat{s}^{(1)}_{r,(k)} = \hat{\Gamma}_{(k)}d_{(k)} + \dfrac{\hat{h}^*_{br1,(k)}}{\sqrt{2}}n^{(1)}_{r,(k)} + \dfrac{\hat{h}_{br2,(k)}}{\sqrt{2}}n^{(1)*}_{r,(k)} \\ \hat{s}^{(1)}_{r,(k+1)} = \hat{\Gamma}_{(k+1)}d_{(k+1)} - \dfrac{\hat{h}_{br1,(k+1)}}{\sqrt{2}}n^{(1)*}_{r,(k+1)} + \dfrac{\hat{h}^*_{br2,(k+1)}}{\sqrt{2}}n^{(1)}_{r,(k+1)} \end{cases}, \tag{3}$$

where $\Gamma = \frac{1}{2}\left(|h_{br1}|^2 + |h_{br2}|^2\right)^\dagger$, $(\cdot)^\dagger$ denotes that the indices (k) and $(k+1)$ are dropped for simplicity.

In order to transmit an unitary power signal the RN normalizes (3) by considering the normalization factor $\hat{\alpha}_{(k)}$:

$$\hat{\alpha}_{(k)} = \frac{1}{\sqrt{\hat{\Gamma}^2_{(k)} + \hat{\Gamma}_{(k)}\sigma_r^{2(1)}}}. \tag{4}$$

During phase II the information is sent via $R \rightarrow U$. The received signal at the UT is given by:

$$\begin{aligned} y_{u,(k)}^{(2)} &= \hat{\alpha}_{(k)}\hat{\Gamma}_{(k)}h_{ru,(k)}d_{(k)} + \alpha_{(k)}\frac{\hat{h}_{br1,(k)}^{*}}{\sqrt{2}}n_{r,(k)}^{(1)}h_{ru,(k)} \\ &+ \hat{\alpha}_{(k)}\frac{\hat{h}_{br2,(k)}}{\sqrt{2}}n_{r,(k)}^{(1)*}h_{ru,(k)} + n_{u,(k)}^{(2)}, \end{aligned} \tag{5}$$

where $n_{u,(k)}^{(2)}$ is the additive white Gaussian noise which is zero mean and has unitary variance $\sigma_u^{2(2)}$.

The signal given by (5) is equalized using the coefficients $g_{ru} = \left(\hat{\alpha}\hat{\Gamma}\hat{h}_{ru}/\sigma_t^2\right)^{\dagger}$ which after some mathematical manipulation give:

$$\begin{aligned} \hat{s}_{u,(k)}^{(2)} &= \hat{\alpha}_{(k)}^2\hat{\Gamma}_{(k)}^2\frac{\left|\hat{h}_{ru,(k)}\right|^2}{\sigma_t^2}d_{(k)} + \hat{\alpha}_{(k)}\hat{\Gamma}_{(k)}\frac{\hat{h}_{ru,(k)}^{*}}{\sigma_t^2} + n_{u,(k)}^{(2)} \\ &+ \hat{\alpha}_{(k)}^2\hat{\Gamma}_{(k)}\left|\hat{h}_{ru,(k)}\right|^2\frac{1}{\sqrt{2}\sigma_t^2}\left(\hat{h}_{br1,(k)}^{*}n_{r,(k)}^{(1)} + \hat{h}_{br2,(k)}n_{r,(k)}^{*(1)}\right). \end{aligned} \tag{6}$$

The equalization coefficient g_{ru} is a function dependent on the channel estimate $\hat{h}_{ru,(k)}$ and the variance of the total noise. Moreover, the statistics of the total noise is conditioned to the channel realization $B \rightarrow R$. Therefore the variance of the total noise can be computed as conditioned to these channel realizations or averaged over all the channel realizations. We denote by $\sigma_{t(h_{ru},(k))}^2$ the variance of the total noise conditioned to the specific channel realization at k subcarrier and by σ_t^2 the averaged noise variance.

To obtain the soft variables used for the data regeneration, the UT combines the signals received in the two phases i.e. (1) and (6) using the MRC. The equalization coefficients for (1) are $g_{bu1,2} = \left(\hat{h}_{bu1,2}/\sqrt{2}\sigma_u^2\right)^{\dagger}$. The result is the soft–decision variable to be hard–decoded

III. CHANNEL ESTIMATION

A. *Pilot pattern*

The system considered is pilot–aided which at the BS are inserted according to the pilot pattern of Figure 2. Without loss of generality pilots are assumed to have unitary value. Two consecutives pilots are separated by a spacing of N_f in frequency and by N_t in time.

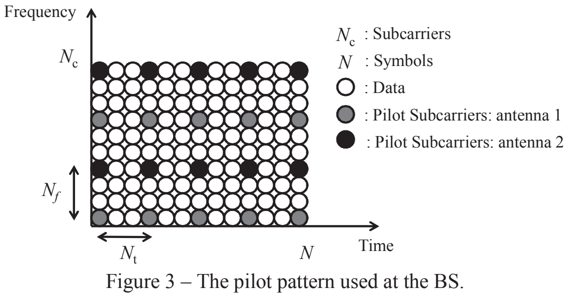

Figure 3 – The pilot pattern used at the BS.

The pilot pattern used at the RN to convey the information about the equivalent $B \rightarrow R$ channel is identical to the one of Fig. 3, but as both RN and UT are single antenna, all pilot subcarriers are allocated to estimate a single channel. However in order to keep the same pilot density in the BS–RN and RN–UT frames, the numerical results reported in Section IV consider only one half of the pilot subcarriers, i.e. one half of the pilot assume zero as value.

B. *TD– Channel Estimation*

To introduce the TD–MMSE channel estimator let us consider the classical input–output through a discrete additive Gaussian noise channel:

$$\tilde{y}_{(n)} = \sum_{k=0}^{N_c-1}\tilde{h}_{(k)}\tilde{x}_{(n-k)} + \tilde{n}_{(n)}, \tag{7}$$

where $\tilde{n}_{(n)}$ is the additive white Gaussian noise samples.

The key idea in the TD channel estimation comes from the fact that if the pilots are equal amplitude and equispaced in the frequency domain, they will correspond in the TD to a periodic train of pulses:

$$\tilde{p}_{(n)} = \frac{1}{N_f}\sum_{m=0}^{N_f-1}\delta_{\left(n-mN_c/N_f\right)}. \tag{8}$$

Therefore considering that $\tilde{x}_{(n-k)}$ is the sum of the data (d) plus pilots, (7) can be written as:

$$\tilde{y}_{(n)} = \sum_{k=0}^{N_c-1}\tilde{h}_{(k)}d_{(n-k)} + \frac{1}{N_f}\sum_{m=0}^{N_f-1}\tilde{h}_{\left(n-mN_c/N_f\right)} + \tilde{n}_{(n)}. \tag{9}$$

The data component in (10) is eliminated convolving with the transmitted pilots (8). The resulting signal is dependent only on the CIR, and is made–up of N_f replicas of the CIR separated by N_c/N_f.

$$\tilde{y}_{(n)} = \sum_{k=0}^{N_c-1}\tilde{h}_{(k)}d_{(n-k)} + \frac{1}{N_f}\sum_{m=0}^{N_f-1}\tilde{h}_{\left(n-mN_c/N_f\right)} + \tilde{n}_{(n)}. \tag{10}$$

The CIR estimate is improved by considering the MMSE filter [11] to reduce the noise variance of the estimate. If the channel taps are properly separated by the sampling interval the MMSE filter is implemented by the matrix is:

$$\mathbf{W}_{\text{MMSE},\,h} = \mathbf{R}_{h\hat{h}}\mathbf{R}_{\hat{h}\hat{h}}^{-1}, \tag{11}$$

where $\mathbf{R}_{\hat{h}\hat{h}}$ is the filter input correlation matrix, which is given by $\mathbf{R}_{hh} + \sigma_n^2\mathbf{I}_{N_c/N_f}$, and $\mathbf{R}_{h\hat{h}}$ is the filter input–output cross–correlation matrix.

The matrices in (11) are sparse with non–null elements whose number is equal to the number of taps L occurring only in the diagonal. This leads to a considerable simplicity reduction when compared to a FD implementation of the MMSE estimator.

C. *Estimation of the noise variance*

The channels to be estimated at the UT are the direct channels $B \rightarrow R$ and the equivalent channel $B \rightarrow R \rightarrow U$. Furthermore there is need to estimate the noise variances, which is done resorting to the TD–MMSE. [11]. Since we know that the CIR energy is limited to the number of taps L, the noise variance estimate $\hat{\sigma}_n^2$ can be calculated by take into account the samples out of the number of taps, i.e. $n \notin \{L\}$ and by averaging the number of symbols N. Thus $\hat{\sigma}_n^2$ is given by:

$$\hat{\sigma}_n^2 = \frac{N_c/N_f}{\left[\left(N_c/N_f\right) - L\right]N}\sum_{i=1}^{N}\sum_{n\notin\{L\}}\left|\hat{\tilde{h}}_{(n)}\right|^2. \tag{12}$$

The channels $B \rightarrow U$ and the noise variance $\sigma_u^{2(1)}$ are estimated at the UT during the phase I of the protocol, using the point–to–point TD–MMSE channel estimator according to (11) and (12).

D. Estimation of the equivalent channel $B \rightarrow R \rightarrow U$

During phase I, the channels $B \rightarrow R$ and the noise variance $\sigma_r^{2(1)}$ are estimated at the RN. As this corresponds to a point–to–point link the conventional TD–MMSE is used. In phase II of the protocol, the equivalent channel $B \rightarrow R \rightarrow U$ $\hat{\alpha}_{(k)}\hat{\Gamma}_{(k)}h_{ru,(k)}$ and the noise variance $\sigma_u^{2(2)}$ need to be estimated at the UT. The equivalent channel can be estimated since $\hat{\alpha}_{(k)}\hat{\Gamma}_{(k)}$ is inserted in the pilot subcarriers at the RN.

There are two issues that arise from the specifities of the $B \rightarrow R \rightarrow U$ link that need to be considered in the channel estimator design:

1. The fact that the pilots inserted at the RN do not have a constant value.
2. The fact that the statistics of the total noise at the UT are for a specific frame conditioned to the realization of the $B \rightarrow R$ channel.

1) The equivalent channel $\hat{\alpha}_{(k)}\hat{\Gamma}_{(k)}$ amplitudes including Alamouti is estimated at the RN and the estimated value inserted in subcarriers reserved for pilots. However, the new pilots are no longer constant and we will not get in the TD a periodic sequence of pulses resulting in overlapping of the CIRs.

Developing (4) it is easy to verify that the product $\hat{\alpha}_{(k)}\hat{\Gamma}_{(k)}$ tends to one for a high SNR value as illustrated in Figure 4. Even for low values of SNR the factor is quite close to 1, and simulations have shown that the degradation in the operation of the TD–MMSE is negligible.

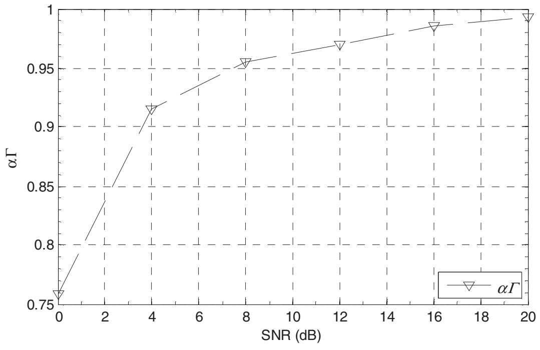

Figure 4 – $\hat{\alpha}_{(k)}\hat{\Gamma}_{(k)}$ vs. SNR.

2) Insertion of the equivalent channel $\hat{\alpha}_{(k)}\hat{\Gamma}_{(k)}$ amplitudes in the pilot positions at the RN allows the estimator at the UT estimate the equivalent channel $\hat{\alpha}_{(k)}\hat{\Gamma}_{(k)}h_{ru,(k)}$ but does not enable estimation $\hat{\alpha}^2_{(k)}\hat{\Gamma}_{(k)}\left|\hat{h}_{ru,(k)}\right|^2$ which is a required parameter in the variance of the total noise when conditioned to a specific channel realization, according to (6).

Considering $E\left\{|h|^2\right\}=1$ the averaged (over all channel realizations), σ_t^2 is given by:

$$\sigma_t^2 = \frac{1}{E\left\{\Gamma^2_{(k)}\right\}+\sigma_r^{2(1)}} E\{\Gamma\}\sigma_u^{2(2)} + \sigma_u^{2(1)}. \qquad (13)$$

Furthermore considering $\sigma_u^{2(1)} = \sigma_r^{2(1)} = \sigma_u^{2(2)}$ and the premise of high SNR in the $B \rightarrow R$ link, we have $\sigma_u^{2(2)} << 1.5$ and after some mathematical (13) may be express as $\sigma_t^2 \cong 5/3\,\sigma_u^{2(2)}$.

To assess the validity of using the averaged noise variance instead of the conditioned one we plot in Figure 5, the BER versus E_b/N0 performance assuming perfect channel estimation is available at the receiver but considering the cases where the noise variance used is the conditioned one and the averaged ones. The results refer to a channel as referred in Section IV but similar results were obtained with other models.

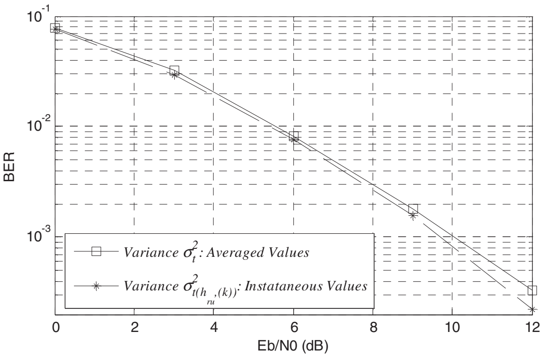

Figure 5 – The system performance.

The performance penalty by using the averaged noise variance is less than 0.8dB which is a tolerable penalty to pay in order to obtain the variance of the total noise regarding the low complexity implementation. Therefore we consider the use of σ_t^2 in our scheme.

IV. SIMULATION RESULTS

In the simulations we used the ITU pedestrian channel model B [12] at speed $v = 10\,\text{km/h}$. The taps were modified according to the sampling frequency defined on WiMAX standard, $\Delta t = 89.3e^{-9}$. We considered that 1024 subcarriers were QPSK modulated and the carrier frequency was $f_c = 2e^9$. The transmitted OFDM symbol carried pilot and data with a pilot separation $N_f = 4$ and $N_t = 1$.

We focus our analysis on the scenario presented in Figure 1 and the simulations were performed assuming that the channels are uncorrelated, the receiver is perfectly synchronized and the insertion of a long enough cyclic prefix in the transmitter ensures that the orthogonality of the subcarriers in maintained after transmission. In Figure 2 we have three estimators' blocks and in our simulations we use the TD–MMSE in all of them. We use the TD–MMSE to estimate all the noise variances as well.

The results are presented in terms of BER and the normalized MSE, both as function of E_b/N0.

The MSE performance of the cooperative channel is evaluated by averaging the MSE's of the direct and the relaying channel [8]. Thus the resulting MSE, i.e. the MSE of the cooperative channel, is given by:

$$\mathrm{MSE}=\frac{1}{2}\left(\frac{1}{2}\left(\mathrm{MSE}_{h_{bu1,(k)}}+\mathrm{MSE}_{h_{bu2,(k)}}\right)+\mathrm{MSE}_{\hat{\alpha}_{(k)}\hat{\Gamma}_{(k)}h_{ru,(k)}}\right). \quad (14)$$

A. Performance Evaluation

In order to validate the proposed scheme, channel estimation simulations were performed as well. We consider that the RN was employing the proposed pilots and the variance of the total noise was given by σ_t^2. Also, we assume that the links $B \to U$, $B \to R$ and $R \to U$ have the same statistics.

Figure 6 depicts the BER attained with perfect CSI and the TD–MMSE estimator and the difference of performance is minimal in most of the cases and in the worst case is 0.5dB.

Figure 7 depicts the normalized MSE's performance. These results show that the proposed pilot allocation, at the RN, according to Section III.D, allows the TD–MMSE satisfactory estimate the required channel. When comparing the channel estimator for the link with relay against the one of the direct link, there is some penalty which accounts for the additional noise added at the relay. The relative penalty decreases as $\mathrm{E}_b/\mathrm{N0}$ increases and can be verified to converge to 2.2dB which is the factor of $5/3$ that relates the total and individual noises in the asymptotic case of high SNR.

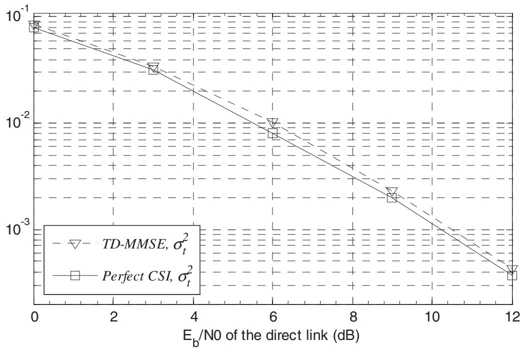

Figure 6 – System BER performance.

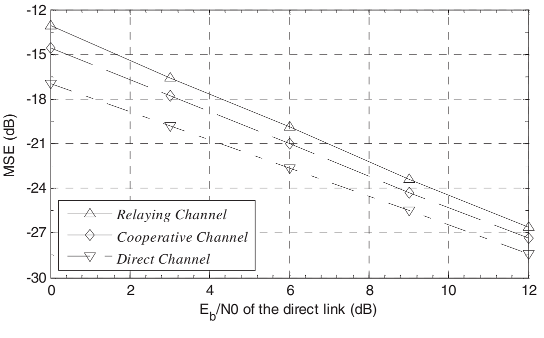

Figure 7 – Channel estimation MSE performance.

V. CONCLUSION

In this paper we considered two problems of channel estimation in a scenario where spatial diversity provided by Alamouti is complemented with the use of a half–duplex relay node using the EF protocol. The estimator scheme was based on the TD–MMSE which leads to a significant complexity reduction when compared to its FD counterpart. We proposed a scheme where the estimates of the $B \to R$ link are inserted in the pilot positions in the $R \to U$ transmission. For the channel estimation at the destination of the $B \to R \to U$ equivalent channel we analyzed several simplifying options enabling the operation of channel estimation namely the use of averaged statistics for the overall noise and the impact of the fluctuations in the amplitude of the $B \to R$ equivalent channel. It is shown that in the asymptotic case of high SNR, and equal noise statistics at the relay and destination the penalty in the estimation $B \to R \to U$ equivalent channel if 2.2dB relatively to the case of a direct link using the same pilot density. The resulting estimation was assessed in terms of the BER of the overall link through simulation with channel representative of a real scenario and the results have shown its effectiveness despite a moderate complexity.

ACKNOWLEDMENTS

The authors wish to acknowledge the support of the European project Enhanced Wireless Communication Systems Employing Cooperative Diversity – CODIV, FP7/ICT/2007/215477, Portuguese Cooperative and Antenna Diversity for Broadband Wireless Networks – CADWIN, PTDC/EEA–TEL/099241/2008 and Portuguese Foundation for Science and Technology (FCT) grant for the first author.

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