

# Channels and Parameters Acquisition in Cooperative OFDM Systems

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## 1. Introduction

Cooperative techniques are promising solutions for cellular wireless systems to improve system fairness, extend the coverage and increase the capacity. Antenna array schemes, also referred as MIMO systems, exploit the benefits from the spatial diversity to enhance the link reliability and achieve high throughput (Foschini & Gans, 1998). On the other hand, orthogonal frequency division multiplexing (OFDM) is a simple technique to mitigate the effects of inter-symbol interference in frequency selective channels (Laroia et al., 2004). The integration of multiple antenna elements is in some situation unpractical especially in the mobile terminals because of the size constraints, and the reduced spacing does not guarantee decorrelation between the channels. An effective way to overcome these limitations is generate a virtual antenna-array (VAA) in a multi-user and single antenna devices environment, this is referred as cooperative diversity. The use of dedicated terminals with relaying capabilities has been emerging as a promising key to expanded coverage, system wide power savings and better immunity against signal fading (Liu, K. et al., 2009).

A large number of cooperative techniques have been reported in the literature the potential of cooperation in scenarios with single antennas. In what concerns channel estimation, some works have discussed how the channel estimator designed to point-to-point systems impacts on the performance of the relay-assisted (RA) systems and many cooperative schemes consider that perfect channel state information (CSI) is available (Muhaidat & Uysal, 2008), (Moco et al., 2009), (Teodoro et al., 2009), (Fouillot et al., 2010). Nevertheless, to exploit the full potential of cooperative communication accurate estimates for the different links are required. Although some work has evaluated the impact of the imperfect channel estimation in cooperative schemes (Chen et al., 2009), (Fouillot et al., 2010), (Gedik & Uysal, 2009), (Hadizadeh & Muhaidat, 2010), (Han et al., 2009), (Ikki et al. 2010), (Muhaidat et al., 2009), new techniques have been derived to address the specificities of such systems.

Channel estimation for cooperative communication depends on the employed relaying protocol, e.g., decode and forward (DF) (Laneman et al., 2004) when the relay has the capability to regenerate and re-encode the whole frame; amplify and forward (AF) (Laneman et al., 2004) where only amplification takes place; and what we term equalize and forward (EF) (Moco et al., 2010), (Teodoro et al., 2009), where more sophisticated filtering operations are used.

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. In this protocol the relaying node are able to perform all the receiver's processes including channel estimation and the point-to-point estimators can be adopted in these cooperative systems. However the situation is different with AF and Equalize-and-Forward (EF) which are protocols less complex than the DF. In the former case (AF),  $B \rightarrow R \rightarrow U$  (base station-relay node-user terminal) channel is the cascaded of the  $B \rightarrow R$  and  $R \rightarrow U$  channels, which has a larger delay spread than the individual channels and additional noise introduced at the relay, this model has been addressed in (Liu M. et al., 2009), (Ma et al., 2009), (Neves et al., 2009), (Wu & Patzold, 2009), (Zhang et al., 2009), (Zhou et al., 2009).

Channel estimation process is an issue that impacts in the overall system complexity reason why it is desirable use a low complex and optimal estimator as well. This tradeoff has been achieved in (Ribeiro & Gameiro, 2008) where the MMSE in time domain (TD-MMSE) can decrease the estimator complexity comparatively to the frequency domain implementation. In (Neves et al., 2009) it is showed that under some considerations the TD-MMSE can provide the cascaded channel estimate in a cooperative system. Also regarding the receiver complexity (Wu & Patzold, 2009) proposed a criterion for the choice of the Wiener filter length, pilot spacing and power. (Zhang et al., 2009) proposed a permutation pilot matrix to eliminate inter-relay signals interference and such approach allows the use of the least square estimator in the presence of frequency off-sets. Based on the non-Gaussian dual-hop relay link nature (Zhou et al., 2009) proposed a first-order autoregressive channel model and derived an estimator based on Kalman filter. In (Liu, M. et al., 2009) the authors propose an estimator scheme to disintegrate the compound channel which implies insertion of pilots at the relay, in the same way (Ma et al., 2009) developed an approach based on a known pilot amplifying matrix sequence to improve the compound channel estimate taking into account the interim channels estimate. To separately estimate  $B \rightarrow R$  and  $R \rightarrow U$  channels (Sheu & Sheen. 2010) proposed an iterative channel estimator based on the expectation maximization. Regarding that the  $B \rightarrow R$  and  $R \rightarrow U$  links are independent and point-to-point links (Xing et al. 2010) investigated a transceiver scheme that jointly design the relay forward matrix and the destination equalizer which minimize the MSE. Concerning the two-way relay (Wang et al. 2010) proposed an estimator based on new training strategy to jointly estimate the channels and frequency offset. For MIMO relay channels (Pang et al. 2010) derived the linear mean square error estimator and optimal training sequences to minimize the MSE. However to the best of our knowledge channel estimation for EF protocol that use Alamouti coding from the base station (BS) to relay node (RN), equalizes, amplifies the signals and then forward it to the UT 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 perform the equalization. The derivation of proper channel estimator for this scenario is the objective of this chapter. We analyze the requirements in terms of channels and parameters estimation

to obtain optimal equalization. We evaluate the sensitivity of required parameters in the performance of the system and devise scheme to make these parameters available at the destination. We consider a scenario with a multiple antenna BS employing the EF protocol, and propose a time domain pilot-based scheme (Neves, et al. 2010) 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) the TD-MMSE estimator, estimates the equivalent channel from the source to destination, taking into account the Alamouti equalization performed at the RN. The estimator scheme we consider operates in time domain because of the reduced complexity when compared against its implementation in frequency domain, e.g. (Ribeiro & Gameiro, 2008).

The remainder of this chapter is organized as follows. In Section 2, we present the scenario description, the relaying protocol used in this work and the corresponding block diagram of the proposed scheme. The mathematical description involving the transmission in our scheme is presented in Section 3. In Section 4, we present the channel estimation issues such as the estimator method used in this work and the channels and parameters estimates to be assess at the RN or UT. The results in terms of BER and MSE are presented in Section 5. Finally, the conclusion is pointed out in Section 6.

## 2. System model

### 2.1 Definition

Throughout the text index  $n$  and  $k$  denote time and frequency domain variables, respectively. Complex conjugate and the Hermitian transposition are denoted by  $(\cdot)^*$  and  $(\cdot)^H$ , respectively.  $E\{\cdot\}$  and  $(*)$  correspondently denote the statistical expectation and the convolution operator.  $\mathcal{N}(m, \sigma^2)$  refers to a complex Gaussian random variable with mean  $m$  and variance  $\sigma^2$ .  $\text{diag}(\cdot)$  stands for diagonal matrix,  $|\cdot|$  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 as well. Variables, vectors or matrixes in time domain are denoted by  $(\sim)$ . All estimates are denoted by  $(\wedge)$ .

### 2.2 Channel model

The OFDM symbol  $\mathbf{x} = \mathbf{d} + \mathbf{p}$  where  $\mathbf{p}$  corresponds to the pilots which are multiplexed with data  $\mathbf{d}$  subcarriers. The element  $x_{(k)}$  of the OFDM symbol vector is transmitted over a channel which the discrete impulse response is given by:

$$\tilde{h}_{(n)} = \sum_{g=0}^{G-1} \beta_g \delta(n - \tau_g), \quad (1.1)$$

where  $G$  is total number of paths,  $\beta_g$  and  $\tau_g$  are the complex amplitude and delay of the  $g$ th path, respectively.  $\beta_g$  is modelled as wide-sense stationary uncorrelated scattering (WSSUS) process. The  $g$ th path has a variance  $\sigma_g^2$  which is determined by the power delay profile and satisfies  $\sum_{g=0}^{G-1} \sigma_g^2 = 1$ . Although the channel is time-variant we assume it is constant during one OFDM symbol interval and its time dependence is not present in notation for simplicity.

### 2.3 Scenario description

The studied scenario, depicted in Fig. 1, corresponds to the proposed RA schemes for downlink OFDM-based system. The BS and the RN are equipped with  $M$  and  $L$  antennas, respectively. The BS is a double antenna array and the UT is equipped with a single antenna. Throughout this chapter we analyze two RA schemes: the RN as a single antenna or an array terminal. These scenarios are referred as  $M \times L \times 1$  schemes.

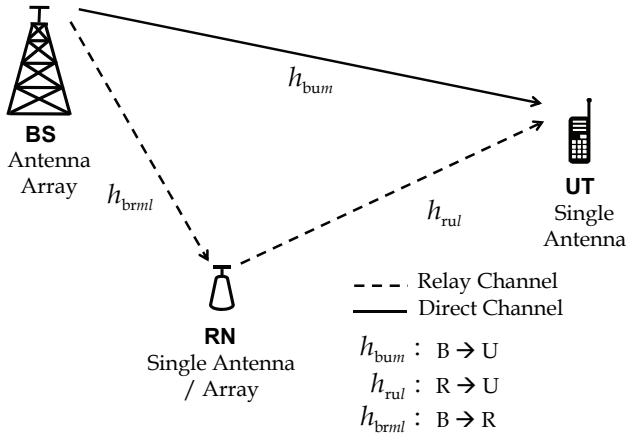


Fig. 1. Proposed RA scenario

The following channels per  $k$  subcarrier are involved in this scheme:

- $M \times 1$  MISO channel between the BS and UT ( $B \rightarrow U$ ):  $h_{bum,(k)}$ ,  $m = 1, 2$
- $L \times 1$  MISO channel between the RN and UT ( $R \rightarrow U$ ):  $h_{rul,(k)}$ ,  $l = 1, 2$
- $M \times L$  channel between the BS and RN ( $B \rightarrow R$ ):  $h_{bvm,(k)}$ ,  $m = 1, 2$  and  $l = 1, 2$

All 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 and the power transmitted by the BS is equally allocated between the two antennas.

### 2.4 The Equalize-and-Forward (EF) relaying protocol

For the single antenna relay scenario, the amplify-and-forward protocol studied in (Moco et al., 2010) is equivalent to the RA EF protocol considered here. However, if the signal at the relay is collected by two antennas, doing just a simple amplify-and-forward is not the best strategy. We need to perform some kind of equalization at the RN to combine the received signals before re-transmission. Since we assume the relay is half-duplex, the communication cycle for the aforementioned cooperative scheme requires two phases:

**Phase I:** the BS broadcasts its own data to the UT and RN, which does not transmit data during this stage.

**Phase II:** while the BS is idle, the RN retransmits to the UT the equalized signal which was received from the BS in phase I. The UT terminal receives the signal from the RN and after reception is complete, combines it with the signal received in phase I from the BS, and provides estimates of the information symbols.

### 2.5 The cooperative system

Fig. 2 shows the corresponding block diagram of the scenario depicted in Fig. 1, with indication of the signals at the different points. The superscripts (1) and (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.

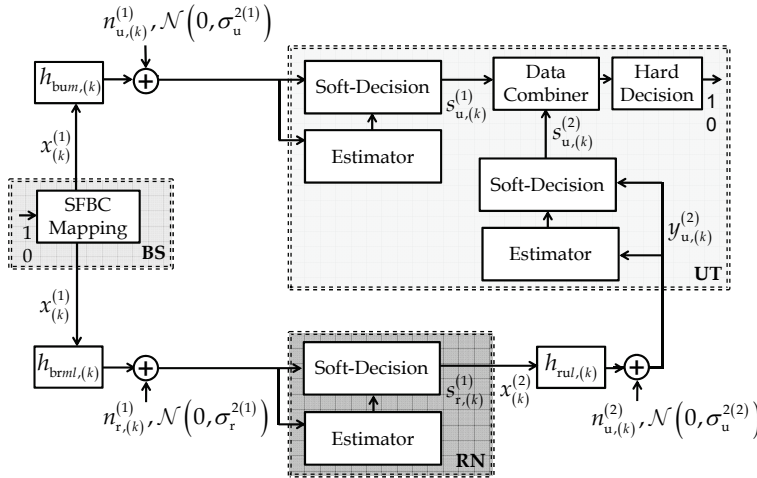


Fig. 2. The corresponding block diagram of the  $M \times L \times 1$  RA scenario

Let  $\mathbf{d} = (d_0 d_1 \dots d_{N_d} - 1)^T$  be the symbol sequence to be transmitted where  $N_d$  is the number of data symbols, then for  $k$  even the SFBC (Teodoro et al., 2009) mapping rule is defined in Table 1. The symbols  $d_{(k)}$  are assumed to have unit average energy, i.e.  $|d_{(k)}|^2 = 1, \forall k$ , and therefore the factor  $1/\sqrt{2}$  used in the mapping, is to ensure that the total energy transmitted by the two antennas per subcarrier is normalized to 1.

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

Table 1. Two transmit antenna SFBC mapping

The pilot symbols are multiplexed with data and the BS broadcasts the information  $x^{(1)}(k)$  (data and pilot) to the RN and UT. This processing corresponds to the phase I of the EF relay protocol. At the UT, the direct channels are estimated and the data are SFBC de-mapped and equalized. These two operations are referred as soft-decision which the result is the soft-decision variable, in this case,  $s_{u,(k)}^{(1)}$ . At the RN, pilots and data are separated; based on pilots, the channels  $B \rightarrow R$  are estimated and the soft-decision is performed. The result is the soft-decision variable  $s_{r,(k)}^{(1)}$ . Then, the new pilot symbols are multiplexed in  $s_{r,(k)}^{(1)}$  and the

information  $x_{(k)}^{(2)}$  is transmitted / forwarded to the UT via R→U channel. This second transmission corresponds to the phase II of the EF protocol. At the final destination, the required channel is estimated and the soft-decision is performed in order to obtain the soft-decision variable  $s_{u,(k)}^{(2)}$ . After the phase II the UT has the soft-variable provided by both the BS and RN. These variables are combined and hard-decoded.

### 3. Mathematical description of the proposed cooperative scheme

The mathematical description for transmit and receive processing is described in this section. As this work is focused on channel estimation, this scheme is designed in order to be capable to provide all the channels and parameters that the equalization requires in both phases of the relaying protocol.

#### 3.1 Phase I

During the first phase the information is broadcasted by the BS. The frequency domain (FD) signals received at the UT in data-subcarriers  $k$  and  $k+1$  are given by

$$\begin{cases} y_{u,(k)}^{(1)} = \frac{1}{\sqrt{2}}(h_{bu1,(k)}d_{(k)} - h_{bu2,(k+1)}d_{(k+1)}^*) + n_{u,(k)}^{(1)} \\ y_{u,(k+1)}^{(1)} = \frac{1}{\sqrt{2}}(h_{bu2,(k)}d_{(k)}^* + h_{bu1,(k+1)}d_{(k+1)}) + n_{u,(k+1)}^{(1)} \end{cases}, \quad (3.1)$$

where  $n_{u,(k)}^{(1)}$  is the additive white Gaussian noise with zero mean unit variance  $\sigma_u^{2(1)}$  and for  $m=1,2$ ,  $h_{bum,(k)}$  represent the channels between the BS and the UT terminal. The FD signals received at the RN in data-subcarriers  $k$  and  $k+1$  are expressed by:

$$\begin{cases} y_{rl,(k)}^{(1)} = \frac{1}{\sqrt{2}}(h_{br1l,(k)}d_{(k)} - h_{br2l,(k+1)}d_{(k+1)}^*) + n_{r,(k)}^{(1)} \\ y_{rl,(k+1)}^{(1)} = \frac{1}{\sqrt{2}}(h_{br2l,(k)}d_{(k)}^* + h_{br1l,(k+1)}d_{(k+1)}) + n_{r,(k+1)}^{(1)} \end{cases}, l=1,2 \quad (3.2)$$

where  $h_{brml,(k)}$  represent the channels between the antenna  $m$  of the BS and antenna  $l$  of the RN terminal and  $n_{r,(k)}^{(1)}$  is the additive white Gaussian noise with zero mean unit variance  $\sigma_r^{2(1)}$ .

Since the data are SFBC mapped at the BS the SFBC de-mapping at the terminals RN and UT also includes the MRC (maximum ration combining) equalization which coefficients are functions dependent on the channels estimates. It is widely known that in the OFDM systems the subcarrier separation is significantly lower than the coherence bandwidth of the channel. Accordingly, the fading in two adjacent subcarriers can be considered flat and without loss of generality we can assume for generic channel  $h_{(k)} = h_{(k+1)}$ . Thus, in phase I the soft-decision variables at the UT follow the expression.

$$\begin{cases} s_{u,(k)}^{(1)} = g_{bu1,(k)}^* y_{u,(k)}^{(1)} + g_{bu2,(k)} y_{u,(k+1)}^{*(1)} \\ s_{u,(k+1)}^{(1)} = -g_{bu2,(k)} y_{u,(k)}^{*(1)} + g_{bu1,(k)}^* y_{u,(k+1)}^{(1)} \end{cases}, \quad (3.3)$$

where the equalization coefficients for  $m=1,2$  are given by  $g_{\text{bum},(k)} = \hat{h}_{\text{bu},m,(k)} / \sqrt{2}\sigma_u^2$ . After some mathematical manipulation, these soft-decision variables may be expressed as:

$$\begin{cases} s_{\text{u},(k)}^{(1)} = \Gamma_{\text{bu},(k)} d_{(k)} + \frac{\hat{h}_{\text{bu}1,(k)}^*}{\sqrt{2}} n_{\text{u},(k)}^{(1)} + \frac{\hat{h}_{\text{bu}2,(k)}}{\sqrt{2}} n_{\text{u},(k)}^{(1)*} \\ s_{\text{u},(k+1)}^{(1)} = \Gamma_{\text{bu},(k+1)} d_{(k+1)} + \frac{\hat{h}_{\text{bu}1,(k+1)}^*}{\sqrt{2}} n_{\text{u},(k+1)}^{(1)} - \frac{\hat{h}_{\text{bu}2,(k+1)}}{\sqrt{2}} n_{\text{u},(k+1)}^{(1)*} \end{cases}, \quad (3.4)$$

$$\text{where } \Gamma_{\text{bu}} = \frac{1}{2} \sum_{m=1}^2 |\hat{h}_{\text{bu},m}|^2.$$

The soft-decision variables should be kept in a buffer, waiting for the information to be provided by the RN in the second phase of the protocol. The mathematical formulation of the next phase varies according to the number of antennas at the RN, *i.e.*  $L$ , and these cases are separately presented in the next sub-sections.

## 3.2 Phase II

### 3.2.1 RN equipped with a single antenna

The FD soft-decision variables at the RN in data-subcarriers  $k$  and  $k+1$  are expressed by:

$$\begin{cases} s_{\text{r},(k)}^{(1)} = g_{\text{br}11}^* y_{\text{r}1,(k)}^{(1)} + g_{\text{br}21} y_{\text{r}1,(k+1)}^{*(1)} \\ s_{\text{r},(k+1)}^{(1)} = -g_{\text{br}21} y_{\text{r}1,(k)}^{*(1)} + g_{\text{br}11}^* y_{\text{r}1,(k+1)}^{(1)} \end{cases}, \quad (3.5)$$

where the equalization coefficient are given by  $g_{\text{br}ml,(k)} = (\hat{h}_{\text{br}ml,(k)} / \sqrt{2})$ , for  $m=1,2$  and  $l=1$ . After some mathematical manipulation, these soft-decision variables are given by:

$$\begin{cases} s_{\text{r},(k)}^{(1)} = \Gamma_{\text{br},(k)} d_{(k)} + \frac{\hat{h}_{\text{br}11,(k)}^*}{\sqrt{2}} n_{\text{r},(k)}^{(1)} + \frac{\hat{h}_{\text{br}21,(k)}}{\sqrt{2}} n_{\text{r},(k)}^{(1)*} \\ s_{\text{r},(k+1)}^{(1)} = \Gamma_{\text{br},(k+1)} d_{(k+1)} + \frac{\hat{h}_{\text{br}11,(k+1)}^*}{\sqrt{2}} n_{\text{r},(k+1)}^{(1)} - \frac{\hat{h}_{\text{br}21,(k+1)}}{\sqrt{2}} n_{\text{r},(k+1)}^{(1)*} \end{cases}, \quad (3.6)$$

$$\text{where } \Gamma_{\text{br},(k)} = \frac{1}{2} \sum_{m=1}^2 |\hat{h}_{\text{br}ml}|^2.$$

In order to transmit a unit power signal the RN normalizes the expression in (3.6) by considering the normalization factor  $\alpha_{(k)}$  which is given by:

$$\alpha_{(k)} = \frac{1}{\sqrt{\Gamma_{\text{br},(k)}^2 + \Gamma_{\text{br},(k)} \sigma_r^2}} \quad (3.7)$$

During the phase II, the normalized soft-decision variable is sent via the second hop of the relay channel R→U. The FD signal received at the UT per  $k$  subcarrier is expressed according to:

$$y_{u,(k)}^{(2)} = \alpha_{(k)} s_{r,(k)}^{(1)} h_{ru,(k)} + n_{u,(k)}^{(2)}, \quad (3.8)$$

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

The signal in (3.8) is equalized using the coefficients  $g_{ru,l,(k)} = \left( \alpha_{(k)} \Gamma_{br,(k)} \hat{h}_{ru,(k)} / \sigma_t^2 \right)$  which after some mathematical manipulation is given by:

$$\begin{aligned} s_{u,(k)}^{(2)} &= \alpha_{(k)}^2 \Gamma_{br,(k)}^2 \frac{|\hat{h}_{ru1,(k)}|^2}{\sigma_t^2} d_{(k)} + \alpha_{(k)} \Gamma_{(k)} \frac{\hat{h}_{ru1,(k)}^*}{\sigma_t^2} n_{u,(k)}^{(2)} \\ &+ \alpha_{(k)}^2 \Gamma_{br,(k)} |\hat{h}_{ru,l,(k)}|^2 \frac{1}{\sqrt{2}\sigma_t^2} \left( \hat{h}_{br11,(k)}^* n_{r,(k)}^{(1)} + \hat{h}_{br21,(k)} n_{r,(k)}^{*(1)} \right). \end{aligned} \quad (3.9)$$

The equalization coefficient  $g_{ru1,(k)}$  is a function dependent on the channel estimate  $\hat{h}_{ru1,(k)}$  and the variance of the total noise. Moreover, the statistics of the total noise is conditioned to the channel realization  $\hat{h}_{brml,(k)}$ . 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  $\sigma_{t(h_{ru1,(k)})}^2$  the noise variance conditioned to the specific channel realization per  $k$  subcarrier and by  $\sigma_t^2$  the unconditioned noise variance. The noise variance of the total noise conditioned to channel realizations is found to be:

$$\sigma_{t(h_{ru,(k)})}^2 = \alpha_{(k)}^2 \Gamma_{br,(k)} |h_{ru1,(k)}|^2 \sigma_u^{2(2)} + \sigma_u^{2(1)}. \quad (3.10)$$

### 3.2.2 RN equipped with a double antenna array

When the array is equipped with two antennas the soft-decision variables are expressed by:

$$\begin{cases} s_{r,(k)}^{(1)} = \sum_{l=1}^2 \left( g_{br1l,(k)}^* y_{rl,(k)}^{(1)} + g_{br2l,(k)} y_{rl,(k+1)}^{(1)*} \right) \\ s_{r,(k+1)}^{(1)} = \sum_{l=1}^2 \left( -g_{br2l,(k)} y_{rl,(k)}^{(1)*} + g_{br1l,(k)} y_{rl,(k+1)}^{(1)} \right) \end{cases}, \quad l = 1, 2 \quad (3.11)$$

where the equalization coefficients are expressed by  $g_{brml,(k)} = \left( \hat{h}_{brml,(k)} / \sqrt{2} \right)$ .

The RN transmits to the UT a unit power signal following the normalization factor in (3.7).

However, in this scenario  $\Gamma_{br,(k)}$  is expressed by  $\Gamma_{br,(k)} = \frac{1}{2} \sum_{m=1}^2 \sum_{l=1}^2 |\hat{h}_{brml,(k)}|^2$ . The FD signals received at the UT are given by:

$$\begin{cases} y_{u,(k)}^{(2)} = \frac{1}{\sqrt{2}} \left( h_{ru1,(k)} \alpha_{(k)} s_{r,(k)}^{(1)} - h_{ru2,(k+1)} \alpha_{(k+1)}^* s_{r,(k+1)}^{(1)*} \right) + n_{u,(k)}^{(2)} \\ y_{u,(k+1)}^{(2)} = \frac{1}{\sqrt{2}} \left( h_{ru2,(k)} \alpha_{(k)}^* s_{r,(k)}^{(1)*} + h_{ru1,(k+1)} \alpha_{(k+1)} s_{r,(k+1)}^{(1)} \right) + n_{u,(k+1)}^{(2)} \end{cases}, \quad (3.12)$$



where  $\hat{h}_{ru,l,(k)}$  represent the channels between the RN and the UT terminal. The soft-decision variables found at the UT in phase II are expressed by:

$$\begin{cases} s_{u,(k)}^{(2)} = g_{ru1}^* y_{u,(k)}^{(2)} + g_{ru2} y_{u,(k+1)}^{*(2)} \\ s_{u,(k+1)}^{(2)} = -g_{ru2} y_{u,(k)}^{*(2)} + g_{ru1}^* y_{u,(k+1)}^{(2)} \end{cases}, \quad (3.13)$$

where the equalization coefficients are given by  $g_{ru,l,(k)} = (\alpha_{(k)} \Gamma_{br,(k)} \hat{h}_{ru,l,(k)} / \sqrt{2\sigma_i^2})$ , for  $l = 1, 2$ . In this scenario the soft-decision variables also depend on the variance of the total noise  $\sigma_i^2$  which conditioned to the channel specific realization can be calculated from (3.12) and is expressed by,

$$\sigma_{t,\hat{h}_{ru,l,(k)}}^2 = \alpha_{(k)}^2 \Gamma_{br,(k)} \Gamma_{ru,(k)} \sigma_r^{2(1)} + \sigma_u^{2(2)} \quad (3.14)$$

where  $\Gamma_{ru,(k)} = \frac{1}{2} \sum_{l=1}^2 |\hat{h}_{ru,l,(k)}|^2$ .

The UT combines the signals received from the RN and the BS. By performing this combining the diversity of the relay path is exploited. This processing is conducted by taking into account  $s_{u,(k)}^{(1)} + s_{u,(k)}^{(2)}$  with ensure and MRC combining. The result corresponds to the variable to be hard-decoded.

## 4. Channel estimation

### 4.1 Time Domain Minimum Mean Square Error (TD-MMSE) estimator

The TD-MMSE (Ribeiro & Gameiro, 2008) corresponds to the version of the MMSE estimator which was originally implemented in FD. This estimator comprises the least square (LS) estimation and the MMSE filtering, both processed in time domain (TD). The TD-MMSE is a pilot-aided estimator, *i.e.* the channel estimation is not performed blindly. It is based on pilot symbols which are transmitted by the source and are known at destination.

The pilot subcarriers convey these symbols that are multiplexed with data subcarriers according to a pattern, Fig. 3, where  $N_f$  and  $N_t$  correspond to distance between two consecutive pilots in frequency and in time, respectively.  $N$  is the number of OFDM symbols and  $N_c$  is the number of subcarriers. The pattern presented in Fig. 3 is adopted during the transmission stages of the envisioned cooperative scheme.

It is usual the pilot symbols assume a unitary value and be constant during an OFDM symbol transmission. Thus, at  $k$  subcarrier the element  $p$  of the vector  $\mathbf{p}$  may be expressed by a pulse train equispaced by  $N_f$  with unitary amplitude. The corresponding expression in TD is also given by a pulse train with elements in the instants  $(n - mN_c/N_f)$ , for  $m \in \{0; N_f - 1\}$ , according to the following expression.

$$p^{(k)} = \sum_{m=0}^{N_f} \delta_{(k-mN_f)} \xleftrightarrow{F^{-1}} \tilde{p}^{(n)} = \frac{1}{N_f} \sum_{m=0}^{N_f-1} \delta_{(n-mN_c/N_f)}. \quad (4.1)$$

The transmitted signal is made-up of data and pilot components. Consequently, at the receiver side the component of the received signal in TD is given by the expression in (4.2).

$$\tilde{y}_{(n)} = \sum_{k=0}^{N_c-1} \tilde{h}_{(k)} d_{(n-k)} + \frac{1}{N_f} \sum_{m=0}^{N_f-1} \tilde{h}_{(n-mN_c/N_f)} + \tilde{n}_{(n)}, \quad (4.1)$$

where  $n_{(n)}$  is the complex white Gaussian noise.

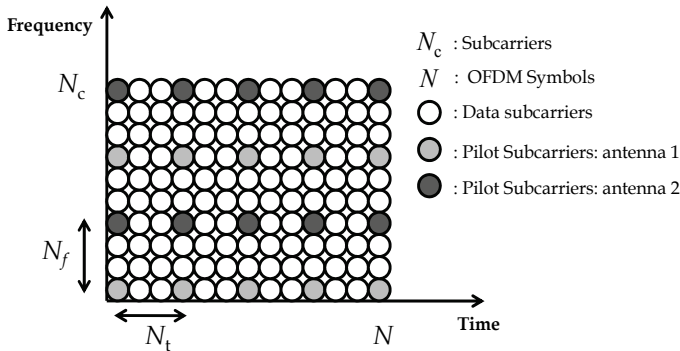


Fig. 3. Pilot pattern

In order to perform the LS estimation in TD, *i.e.* TD-LS, the received signal is convolved with the TD pilot symbols  $\tilde{p}_{(n)}$ . This convolution corresponds to multiply by 1 the subcarriers at frequencies  $N_f$  as by design these are the positions reserved to the pilots thus the data component in the received signal vanishes. The resulting CIR estimate  $\hat{h}_{LS}$  is made-up of  $N_f$  replicas of the CIR separated by  $N_c/N_f$ .

$$\hat{h}_{LS} = \underbrace{\sum_{m=0}^{N_f-1} \tilde{h}_{(n-mN_c/N_f)}}_{\text{CIR}} + \underbrace{\sum_{m=0}^{N_f-1} \tilde{n}_{(n-mN_c/N_f)}}_{\text{Noise}}, \quad n = 0, 1, \dots, \frac{N_c}{N_f} - 1. \quad (4.3)$$

Besides to estimate the CIR the TD-MMSE in (Ribeiro & Gameiro, 2008) can estimate the noise variance as well. It corresponds to an essential requirement when the UT has no knowledge of this parameter. Since we know that the CIR energy is limited to the number of taps, or the set of the taps  $\{G\}$ , 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.*  $n \notin \{G\}$  and by averaging the number of OFDM symbols  $N$ . Thus  $\hat{\sigma}_n^2$  is given by:

$$\hat{\sigma}_n^2 = \frac{N_c/N_f}{\left[ \left( \frac{N_c}{N_f} \right) - G \right] N} \sum_{i=1}^N \sum_{n \notin \{G\}} \left| \hat{h}_{(n)} \right|^2. \quad (4.4)$$

The MMSE filter can improve the LS estimates by reducing its noise variance. The TD-MMSE filter corresponds to a diagonal matrix with non-zero elements according to the number of taps, *i.e.*  $G$ , thus it can be implemented simultaneously with the TD-LS estimator and this operation simplifies the estimator implementation. The MMSE filter implemented by the  $(N_c/N_f) \times (N_c/N_f)$  matrix and for a generic channel  $h$  it is expressed by:

$$\mathbf{W}_{MMSE,\hat{h}} = \mathbf{R}_{\hat{h}\hat{h}}^{-1} \mathbf{R}_{\hat{h}\hat{h}}^{-1}, \quad (4.5)$$

where  $\mathbf{R}_{\hat{h}\hat{h}}$  is the  $(N_c/N_f) \times (N_c/N_f)$  filter input correlation matrix,  $E\{\hat{\mathbf{h}}\hat{\mathbf{h}}^H\}$ , which is given by  $\mathbf{R}_{\hat{h}\hat{h}} + \sigma_n^2 \mathbf{I}_{N_c/N_f}$ , and  $\mathbf{R}_{\hat{h}\hat{h}}$  is the  $(N_c/N_f) \times (N_c/N_f)$  filter input-output cross-correlation matrix,  $E\{\hat{\mathbf{h}}\hat{\mathbf{h}}^H\}$ , which is given by  $\mathbf{R}_{\hat{h}\hat{h}} = \text{diag}\{\sigma_0^2, \sigma_1^2, \dots, \sigma_{G-1}^2, 0, \dots, 0\}$ .

#### 4.2 Channels and parameters estimates

According to the scenario previously presented, there are channels which correspond to point-to-point links:  $h_{\text{bwm},(k)}$  and  $h_{\text{brml},(k)}$ . Therefore, these channels can be estimated by using conventional estimators. However, for the RN  $\rightarrow$  UT links, and since the EF protocol is used, it is necessary to estimate a version of  $h_{\text{rul},(k)}$ , which depends on  $\alpha_{(k)}$  and  $\Gamma_{(k)}$ , the equivalent channel  $h_{\text{eq}} = \alpha_{(k)} \Gamma_{(k)} h_{\text{rul},(k)}$ . Note that the UT has no knowledge of  $\alpha_{(k)}$  and  $\Gamma_{(k)}$ . These factors are dependent on  $h_{\text{brml},(k)}$  which the UT has no knowledge as well. However, the channels  $h_{\text{brml},(k)}$  are estimated at the RN, and based on that,  $\alpha_{(k)} \Gamma_{(k)}$  is calculated. Therefore, we propose to transmit the factor  $\alpha_{(k)} \Gamma_{(k)}$  at the pilot subcarriers as pilots. Consequently, the new pilots are no longer constant and that may compromise the conventional TD-MMSE performance, since this estimator was designed in time domain assuming the pilots are unitary with constant values at the destination. Although our approach enables the destination had knowledge of the non-constant pilot ( $\alpha_{(k)} \Gamma_{(k)}$ ), the result of the convolution between the received signal and these pilots, would result in the overlapped replicas of CIR, according to the Fig. 4.

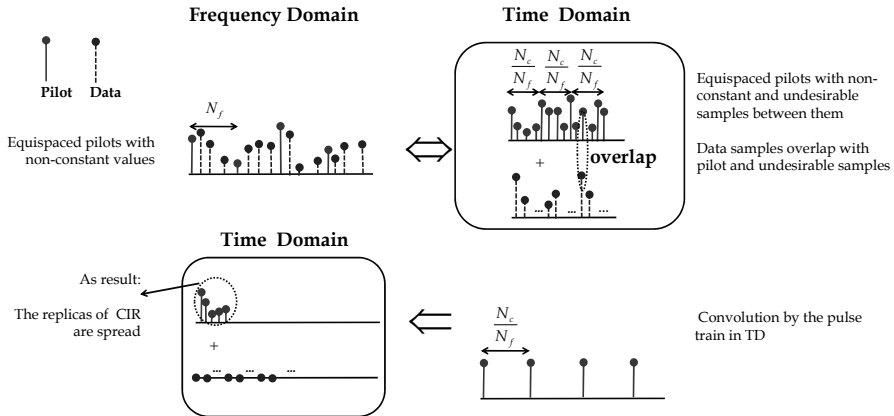


Fig. 4. Pilots with non-constant values result in the overlapped replicas of the CIR

However, the  $\alpha_{(k)}$  expressions depend on the noise variance  $\sigma_r^{2(1)}$  and the product  $\alpha_{(k)} \Gamma_{(k)}$  tends to one for a high SNR value, according to (4.6). The same equation also suggests that the factor  $\alpha_{(k)} \Gamma_{(k)}$  varies exponentially according to the SNR, as depicted in Fig. 5 for  $L = 1$ .

$$\alpha_{(k)} \Gamma_{(k)} = \left( \frac{1}{\sqrt{\Gamma_{\text{br},(k)}^2 + \Gamma_{\text{br},(k)} \sigma_r^{2(1)}}} \right) \Gamma_{\text{br},(k)} = \left( \frac{1}{\sqrt{\Gamma_{\text{br},(k)}^2 + 0}} \right) \Gamma_{\text{br},(k)} \cong 1. \quad (4.6)$$

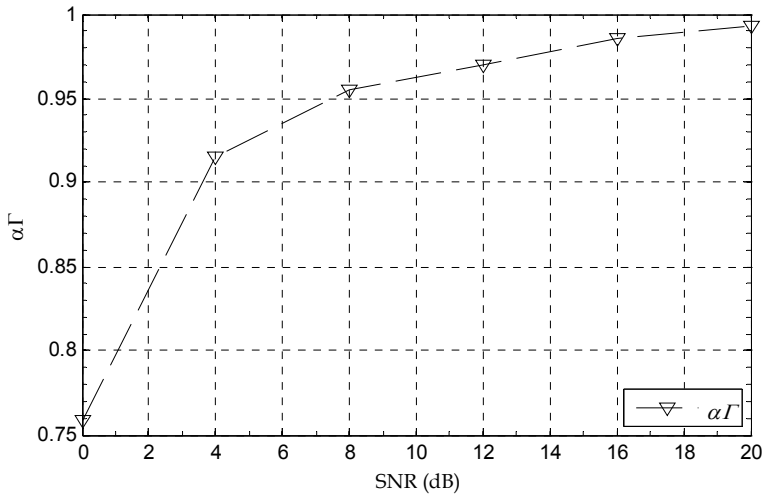


Fig. 5.  $\alpha_{(k)}\Gamma_{(k)}$  vs. SNR

Other behaviour of the factor  $\alpha_{(k)}\Gamma_{(k)}$  can be demonstrated in terms of SNR and subcarriers, as presented in Fig. 6. These plots show that in the first case, *i.e.* SNR = 20dB, the  $\alpha_{(k)}\Gamma_{(k)}$  factor presents the amplitude close to 1 with some negligible fluctuation. However, in the second case, SNR = 2dB, the result is likely different to the previous one: the  $\alpha_{(k)}\Gamma_{(k)}$  factor presents an amplitude also close to 1 but the fluctuation is not negligible.

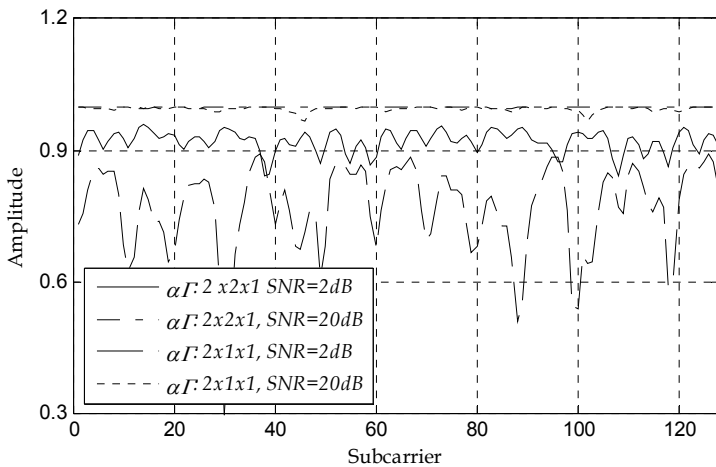


Fig. 6.  $\alpha_{(k)}\Gamma_{(k)}$  per subcarriers

The results in Fig. 6 emphasize that transmit the factor  $\alpha_{(k)}\Gamma_{(k)}$  in the pilot subcarriers may degrade the estimator performance and the causes are:

1. Pilots with some fluctuation in amplitude:

- As the amplitude of the pilots at the destination are constant and equal to one, the result of the estimation is a spread of the replicas of the CIR.
2. *Decreasing the amplitude of the pilots:*
    - The SNR of the pilots is decreased as well.
  3. *The MMSE filter depends on the statistics of the channel  $B \rightarrow R$*

Despite we are considering the TD-MMSE estimator in our analysis, the causes presented previously degrade the performance of any other estimator scheme as well. In order to quantify how these effects can degrade the estimator performance we evaluated the impact of both of them, separately, in a SISO system, since the  $B \rightarrow R$  and  $R \rightarrow U$  channels correspond to point-to-point links.

First, we evaluate the case when the pilots have some fluctuation in amplitude. We consider that the pilots (originally with unit amplitude) had their amplitude disturbed by a noise with zero mean and variance equal to  $\sigma_{\alpha r}^2 = E\left\{\left|1 - \alpha_{(k)}\Gamma_{(k)}\right|^2\right\}$ ,  $\sigma_{\alpha r}^2$  quantifies how far the factor  $\alpha_{(k)}\Gamma_{(k)}$  would be from the pilots with unitary amplitude. We can express  $\sigma_{\alpha r}^2$  as:

$$\sigma_{\alpha r}^2 = 1 + E\left\{\left(\alpha_{(k)}\Gamma_{(k)}\right)^2\right\} - 2E\left\{\alpha_{(k)}\Gamma_{(k)}\right\}. \quad (4.7)$$

Therefore, the pilots are no longer constant and unitary. They have some fluctuation in amplitude which depend on  $\alpha_{(k)}\Gamma_{(k)}$  and they are equal to  $p_{\sigma_{\alpha r}^2} = 1 + z$ , where  $z$  has a normal distribution with zero mean and power  $\sigma_{\alpha r}^2$ . The performance of a SISO system which the pilots correspond to  $p_{\sigma_{\alpha r}^2}$  is shown in Fig. 7 (dash line). For reference, we also include the SISO performance for unit pilots,  $p_1$ . Since we are focus on the degradation of the estimator performance, the results are presented in terms of the normalized mean square error (MSE) and  $E_b/N_0$ , where  $E_b$  corresponds to the energy per bit received at UT and  $N_0$  is the power spectrum density of the total noise which affects the information conveying signals. The normalized MSE for a generic channel  $h$  is given by:

$$\text{MSE}_h = \frac{E\left\{\left|\hat{h} - h\right|^2\right\}}{E\left\{|h|^2\right\}}. \quad (4.8)$$

For low values of SNR,  $[0 - 4]$ , the major difference in performance between the two results is approximately 0.5dB, which is not a noticeable degradation and the estimator performance is not compromised.

The second effect to be evaluated is the decreasing of the amplitude of the transmitted pilot. In order to evaluate this effect we also consider a SISO system, for which the transmitted pilots assume different constant values, and we consider different values of the factor  $\alpha_{(k)}\Gamma_{(k)}$  as pilot. The normalized MSE is given by (4.9). The result is shown in Fig. 8 and for reference we include the SISO performance for unit pilots,  $p_1$ , as well.

$$\text{MSE}_{\text{eq}} = \frac{E\left\{\left|\hat{h}_{\text{eq}} - h_{\text{eq}}\right|^2\right\}}{E\left\{|h_{\text{eq}}|^2\right\}}. \quad (4.9)$$

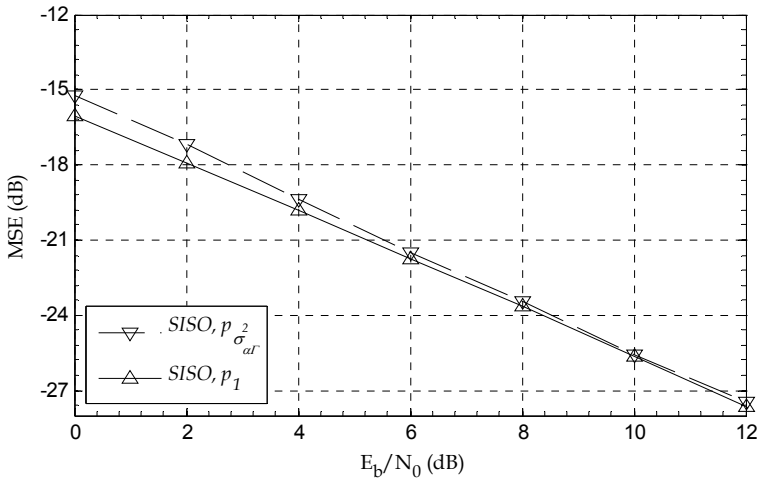


Fig. 7. Channel estimation MSE performance

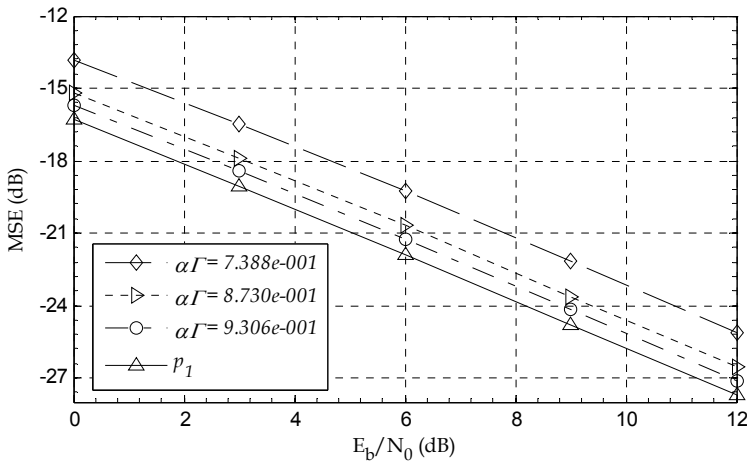


Fig. 8. Channel estimation MSE performance

The results show a constant shift in the MSE when the amplitude of the pilots is not unitary. The shift presents in all results is not a real degradation. It is caused by the normalization present in the MSE in (4.9). In fact, assuming a MSE without normalization the results are all the same.

Transmit the factor  $\alpha_{(k)}\Gamma_{(k)}$  as pilot does not bring any noticeable degradation in the TD-MMSE performance comparing to transmitting unitary pilots. The major degradation occurs only when the pilots have some fluctuation in amplitude, as shown in Fig. 7, and solely for low SNR values.

The conventional MMSE filter in (4.5) is implemented to improve a channel estimate  $\hat{h}$  when the required channel corresponds to  $h$ . However, according to our cooperative

scheme, we need estimate an equivalent channel  $h_{\text{eq}} = \alpha_{(k)} \Gamma_{(k)} h_{\text{ru},(k)}$ . Since the factor  $\widetilde{\alpha\Gamma}$  does not depend on  $\tilde{h}_{\text{ru}}$ , the MMSE filter input correlation matrix for the channel  $\tilde{h}_{\text{eq}}$ , referred  $\mathbf{R}_{\tilde{h}_{\text{eq}}\tilde{h}_{\text{eq}}}$ , is expressed by  $\mathbb{E}\left\{\hat{\mathbf{h}}_{\text{eq}}\hat{\mathbf{h}}_{\text{eq}}^H\right\} = \mathbf{R}_{\left\{\widehat{\alpha\Gamma}\widehat{\alpha\Gamma}\right\}\tilde{h}_{\text{ru}}\tilde{h}_{\text{ru}}} + \sigma_n^2 \mathbf{I}_{N_c/N_f}$  while the filter input-output cross-correlation, referred as  $\mathbf{R}_{\tilde{h}_{\text{eq}}\hat{\mathbf{h}}_{\text{eq}}}$ , is given by  $\mathbb{E}\left\{\tilde{\mathbf{h}}_{\text{eq}}\hat{\mathbf{h}}_{\text{eq}}^H\right\} = \mathbf{R}_{\left\{\widehat{\alpha\Gamma}\widehat{\alpha\Gamma}\right\}\tilde{h}_{\text{ru}}\tilde{h}_{\text{ru}}}$ , both  $\mathbf{R}_{\tilde{h}_{\text{eq}}\tilde{h}_{\text{eq}}}$  and  $\mathbf{R}_{\tilde{h}_{\text{eq}}\hat{\mathbf{h}}_{\text{eq}}}$  are  $(N_c/N_f) \times (N_c/N_f)$  matrices. Thus the MMSE filter, when  $\tilde{h}_{\text{eq}}$  is required, may be express as:

$$\mathbf{W}_{\text{MMSE}, \tilde{h}_{\text{eq}}} = \mathbf{R}_{\left\{\widehat{\alpha\Gamma}\widehat{\alpha\Gamma}\right\}\tilde{h}_{\text{ru}}\tilde{h}_{\text{ru}}} \left( \mathbf{R}_{\left\{\widehat{\alpha\Gamma}\widehat{\alpha\Gamma}\right\}\tilde{h}_{\text{ru}}\tilde{h}_{\text{ru}}} + \sigma_n^2 \mathbf{I}_{N_c/N_f} \right)^{-1}. \quad (4.10)$$

As we shown previously in (4.6) the factor  $\alpha_{(k)} \Gamma_{(k)}$  tends to one for high values of SNR and examining (4.10), which depends on  $\widehat{\alpha\Gamma}$ , it is clear that (4.10) tends to (4.5) for high values of SNR as well. In order to show that several simulation were performed by taking into account  $\mathbf{R}_{\tilde{h}_{\text{eq}}\tilde{h}_{\text{eq}}}$  and the noise variance  $\sigma_r^{2(1)}$ . According to Fig. 9 the results show that the maximum value in the  $\mathbf{R}_{\tilde{h}_{\text{eq}}\tilde{h}_{\text{eq}}}$  matrix is close to  $-40\text{dB}$  for high values of the noise variance.

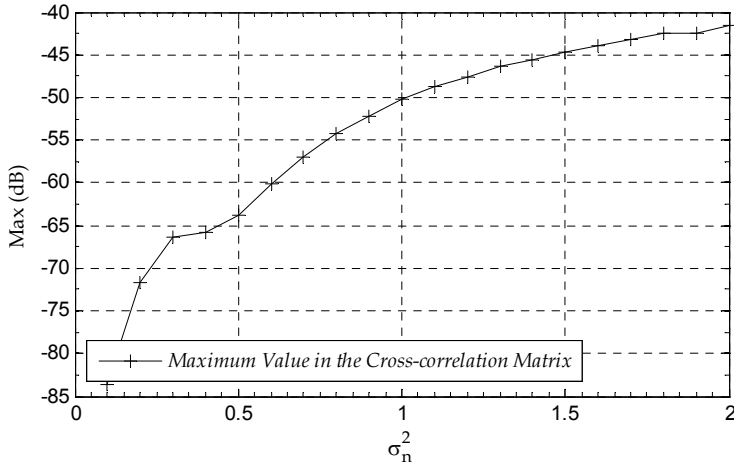


Fig. 9. Maximum value in the correlation matrix vs. noise variance

According to the results in Fig. 7 and Fig. 8 in terms of MSE, transmitting the factor  $\widehat{\alpha\Gamma}$  brings, in the worst case, 0.5dB of degradation reason why there is no need to increase the system complexity by implementing the filter in (4.10). We have shown that our cooperative scheme allows the use of the TD-MMSE estimator without compromising its estimate. According to (4.6) the behaviour of the factor  $\alpha_{(k)} \Gamma_{(k)}$  does not depend on number of taps of the channel, it depends only on the noise variance  $\sigma_r^{2(1)}$ . Therefore, this analysis is applied to

any other channel without loss of generality. Besides to estimate the equivalent channel it is necessary estimate others factors  $\alpha_{(k)}^2 \Gamma_{(k)} |h_{ru,(k)}|^2$  and  $\alpha_{(k)}^2 \Gamma_{br,(k)} \Gamma_{ru,(k)}$  for  $L=1$  and  $L=2$ , respectively. These factors are required parameters in the variance of the total noise  $\sigma_{t(h_{ru,(k)})}^2$ , previously presented in (3.10) and (3.14).

Although the UT does not have individual knowledge of  $h_{brml,(k)}$ ,  $\alpha_{(k)}$  and  $\Gamma_{(k)}$  it has knowledge of the second moment of the expected value of the channels, *i.e.* for all channels  $E\{|h|^2\} = 1$ . Thus, we propose the use of the noise variance unconditioned to the channel realization,  $\sigma_t^2$ , instead of its instantaneous value  $\sigma_{t(h_{ru,(k)})}^2$ . Therefore,  $\sigma_t^2$  is referred as the expectation value of the variance of the total noise. Also we consider that the channels have identical statistics, *i.e.*  $\sigma_u^{2(1)} = \sigma_r^{2(1)} = \sigma_u^{2(2)}$ , thus  $\sigma_t^2$  can be expressed numerically by (4.11) and (4.12) for  $L=1$  and  $L=2$ , respectively.

$$\sigma_t^2 \cong \frac{1}{1.5 + \sigma_u^{2(2)}} \sigma_u^{2(2)} + \sigma_u^{2(2)}. \quad (4.11)$$

$$\sigma_t^2 \cong \frac{1}{5 + 2\sigma_u^{2(2)}} 2\sigma_u^{2(2)} + \sigma_u^{2(2)}. \quad (4.12)$$

If we consider the premise of the cooperative transmission, high SNR compared to the SNR of the direct link, we have  $\sigma_u^{2(2)} \ll 1.5$  for  $L=1$  and consequently (4.11) may be express as  $\sigma_t^2 \cong \frac{5}{3} \sigma_u^{2(2)}$ .

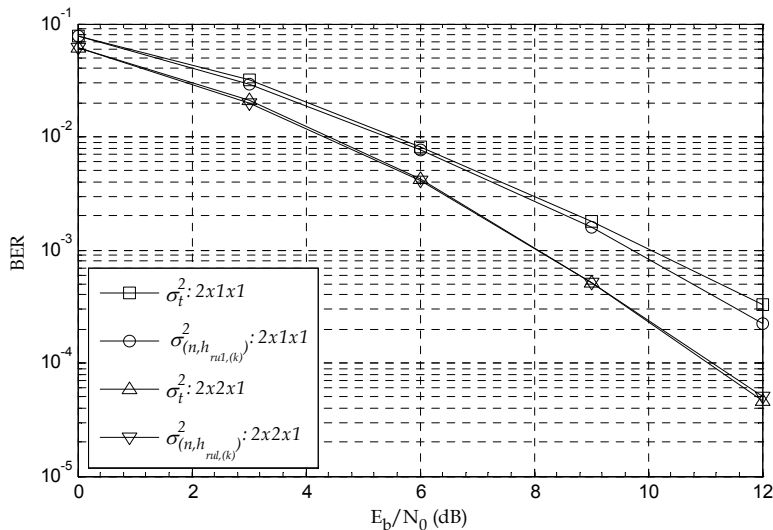


Fig. 10. System performance



To assess the validity of using the averaged noise variance instead of the conditioned one we plot in Fig. 10, the BER versus  $E_b/N_0$  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 5 but similar results were obtained with other models.

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  $\sigma_1^2$  in our schemes.

## 5. Results

### 5.1 Simulation parameters

In order to evaluate the performance of the presented RA schemes we considered a typical scenario, based on LTE specifications (3GPP TS, 2007). In the simulations we used the ITU pedestrian channel model B at speed  $v = 10\text{km/h}$ . 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  $2 \times 1 \times 1$  and  $2 \times 2 \times 1$  scenarios 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 is maintained after transmission. We use the TD-MMSE to estimate all the noise variances as well.

The results are presented in terms of BER and MSE, both as function of  $E_b/N_0$ . The normalized MSE is defined according to (4.9). The MSE performance of the cooperative channel is evaluated by averaging the MSE's of the direct and the relaying channel (Kim et al., 2007). Since the direct channel corresponds to a MISO its MSE is obtained also by averaging the MSE of the B $\rightarrow$ U channels, both normalized. The MSE of the relaying channel corresponds to the MSE of the equivalent channel  $h_{\text{eq}} = \alpha_{(k)} \Gamma_{(k)} h_{\text{rel},(k)}$  which is calculated according to (5.1). Thus the resulting MSE, i.e. the MSE of the cooperative channel, is given by:

$$\text{MSE} = \frac{1}{2} \left( \frac{1}{2} (\text{MSE}_n) + \text{MSE}_{h_{\text{eq}}} \right). \quad (5.1)$$

### 5.2 Performance evaluation

In order to validate the use of the proposed scheme, some channel estimation simulations were performed using the TD-MMSE estimator. Fig. 11 depicts the BER attained with perfect CSI and the TD-MMSE estimator when the RN was employing the proposed pilots. The difference of performance is minimal in most of the cases and in the  $2 \times 1 \times 1$  scheme which is in the worst case this difference is 0.5dB.

Fig. 12 depicts the normalized MSE's performance of the  $2 \times 1 \times 1$  scheme. These results show that the proposed pilot allocation, at the RN, according to Section I.B, 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  $E_b/N_0$  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. According to Fig. 13, this penalty is smaller in the 2x2x1 scheme, since the factor  $\alpha_{(k)}\Gamma_{(k)}$  presents a flatter behavior.

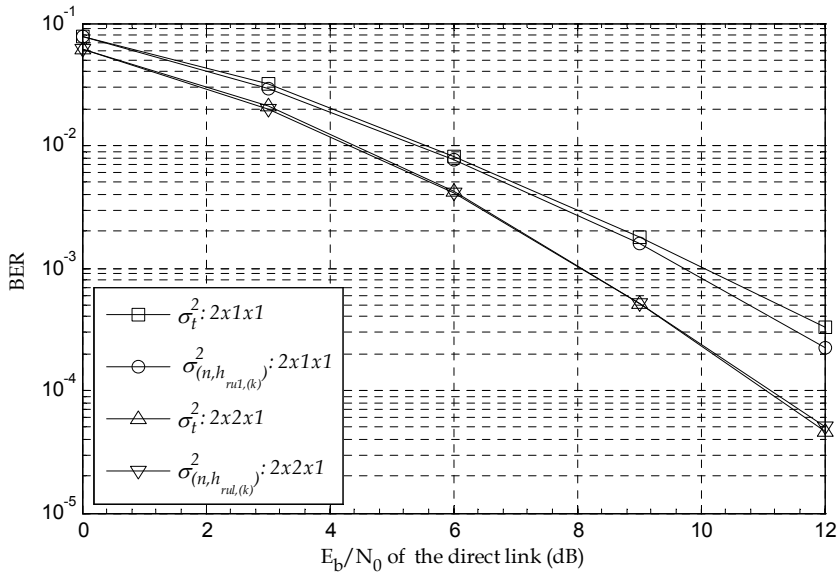


Fig. 11. System performance: RA 2x1x1 and RA 2x2x1 schemes

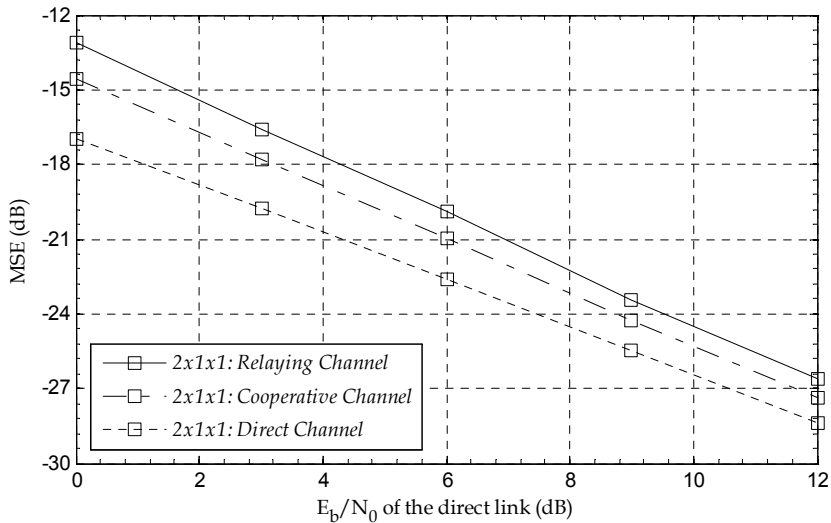


Fig. 12. Channel estimation MSE performance: RA 2x1x1 scheme

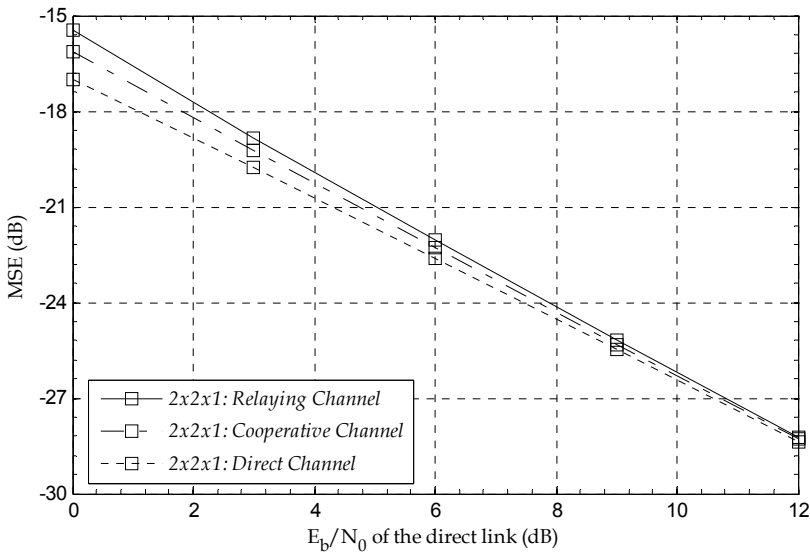


Fig. 13. Channel estimation MSE performance: RA  $2 \times 2 \times 1$  scheme

## 6. Conclusion

In this chapter we considered two problems of channel estimation in a scenario where spatial diversity provided by SFBC is complemented with the use of a half-duplex relay node employing the EF protocol. The channel estimation scheme was based on the TD-MMSE which led to a significant complexity reduction when compared to its frequency domain counterpart. We proposed a scheme where the estimates of the  $B \rightarrow R$  link are inserted in the pilot positions in the  $R \rightarrow U$  transmission. For the estimation of the equivalent channel, *i.e.*  $B \rightarrow R \rightarrow U$ , at the destination 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 equivalent channel. In the RA  $2 \times 1 \times 1$  scheme is shown that in the asymptotic case of high SNR, and equal noise statistics at the relay and destination the penalty in the estimation equivalent channel is 2.2dB relatively to the case of a direct link using the same pilot density. This difference in performance is smaller in the RA  $2 \times 2 \times 1$  scheme since the equivalent channel presents a flatter behaviour. 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.

## 7. Acknowledgments

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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