Joint Detection and Channel Estimation for OFDM Signals with Implicit Pilots

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Abstract - Accurate channel estimation is mandatory for the performance of OFDM modulations (Orthogonal Frequency Division Multiplexing). For this purpose, pilot symbols and/or training sequences are usually multiplexed with data symbols. To avoid the spectral degradation associated to multiplexed pilots, the use of implicit pilots (i.e., pilots superimposed to data) was recently proposed. However, the interference levels between data and pilots might be very high, leading to performance degradation.

In this paper we consider OFDM schemes where the channel estimation is based on implicit pilots. To overcome the difficulties inherent to the interference levels between pilots and data, we propose an iterative receiver with joint detection and channel estimation.

Our performance results show that we can achieve performances close to the ones with perfect channel estimation, even when low-power pilots or sort frames are employed[1].

Index Terms: OFDM, channel estimation, implicit pilots, iterative receivers.

I. Introduction

It is widely recognized that OFDM modulations (Orthogonal Frequency Division Multiplexing) [1] are suitable for broadband wireless systems. For this reason they were selected for digital broadcast systems and wireless networks [2]; they are also being considered for UTRA long term evolution [3].

Since coherent detection is usually employed, we need an accurate channel estimation at the receiver. Typically, the channel estimates are obtained with the help of training symbols that are multiplexed with the data symbols, either in the time domain or in the frequency domain [4], [5], [6].

Typically, the channel impulse response can be very long. Therefore, the required channel estimation overheads can be high, especially in fast-varying scenarios and/or for bursty transmission, leading to a decrease on the system capacity. Since the efficient use of the available bandwidths is of crucial importance for any communication system, it would be desirable to reduce the overheads for channel estimation purposes. A promising method for overcoming this problem is to use implicit pilots, which are added to the data block instead of being multiplexed with it [7]. This means that we can increase significantly the pilots' density, while keeping the system capacity. In fact, we can have a pilot for each data symbol.

However, the interference levels between the data symbols and pilots might be high. This means that the channel estimates are corrupted by the data signal, leading to irreducible noise floors (i.e., the channel estimates can not be improved beyond a given level, even without channel noise). Moreover, there is also interference on the data symbols due to the pilots, leading to performance degradation. Naturally, we need to increase the average power of the transmitted signal, although this is also valid for multiplexed pilots.

In this paper, we consider the use of implicit pilots in OFDM systems. We propose an iterative receiver structure with joint detection and channel estimation. For the first iteration, the channel is estimated by averaging the received signal (data plus pilots) over several blocks; for the remaining iterations, enhanced channel estimates are obtained by considering the data symbols as extra pilots. For the estimation and detection phases of each iteration we remove the undesirable signal (pilots or data) using the most updated version of it.

This paper is organized as follows. The system considered in this paper is introduced in sec. II, while sec. III describes the proposed receiver structure. A set of performance results is presented in sec. IV and sec. V is concerned with the conclusions of this paper.

II. System Description

A. Transmitted Signals

In this paper we consider a frame structure with N_T time-domain blocks, each one corresponding to an "FFT block", and N subcarriers. It is assumed that the channel is almost invariant within the frame. We have a regular grid of pilots, with pilot separation ΔN_T in the time domain and ΔN_F in the frequency domain. The number of pilots per frame is

$$N_P^{Frame} = \frac{N}{\Delta N_F} \cdot \frac{N_T}{\Delta N_T}. \quad (1)$$

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The transmitted OFDM signal associated to the frame is

$$s^{Tx}(t) = \sum_{l=1}^{N_T} s_l^{Tx}(t - lT_B), \quad (2)$$

with T_B denoting the duration of each block. The lth block has the form

$$s_l^{Tx}(t) = \sum_{n=-N_G}^{N-1} s_{n,l}^{Tx} h_T(t - nT_S), \quad (3)$$

with T_S denoting the symbol duration, N_G denoting the number of samples at the cyclic prefix and $h_T(t)$ is the adopted pulse shaping filter. As usual, $\{s_{n,l}^{Tx}; n = 0, 1, \ldots, N-1\}$ = IDFT $\{S_{k,l}^{Tx}; k = 0, 1, \ldots, N-1\}$, where $S_{k,l}^{Tx}$ is the symbol transmitted at the kth subcarrier, and $s_{-n,l^{Tx}} = s_{N-n,l}^{Tx}$ (i.e., the first N_G are the cyclic extension of $\{s_{n,l}^{Tx}; n = 0, 1, \ldots, N-1\}$). Clearly, $T_S = T/N$ and $T_B = T(N + N_G)/N$. The frequency-domain symbols to be transmitted are given by

$$S_{k,l}^{Tx} = S_{k,l} + S_k^P, \quad (4)$$

where $S_{k,l}$ is the data symbol transmitted by the kth subcarrier of the lth block, selected from a given constellation under an appropriate mapping rule, and $\{S_{k,l}^P; k = 0, 1, \ldots, N-1\}$ the block of implicit pilots.

The signal at the receiver input is sampled and the cyclic prefix is removed, leading to the time-domain block $\{y_{n,l}^{Rx}; n = 0, 1, \ldots, N-1\}$. If the cyclic prefix is longer than the overall channel impulse response then the corresponding frequency-domain block, obtained after an appropriate size-N DFT operation, is $\{Y_{k,l}^{Rx}; k = 0, 1, \ldots, N-1\}$, where

$$Y_{k,l}^{Rx} = S_{k,l}^{Tx} H_{k,l} + N_{k,l} = (S_{k,l} + S_{k,l}^P) H_{k,l} + N_{k,l}, \quad (5)$$

with $H_{k,l}$ denoting the overall channel frequency response for the kth frequency of the lth time block and $N_{k,l}$ denoting the corresponding channel noise.

B. Channel Estimation

Let us first assume that $S_{k,l} = 0$, i.e., there is no data overlapping the training block, as in conventional schemes. In that case, we could estimate the channel frequency response as follows:

$$\tilde{H}_{k,l} = \frac{Y_{k,l}^{Rx}}{S_{k,l}^P} = H_{k,l} + \frac{N_{k,l}}{S_{k,l}^P} = H_{k,l} + \epsilon_{k,l}^H. \quad (6)$$

The channel estimation error $\epsilon_{k,l}^H$ is Gaussian-distributed, with zero-mean and

$$E[|\epsilon_{k,l}^H|^2 | S_{k,l}] = E[|N_{k,l}|^2] E\left[\frac{1}{|S_{k,l}^P|^2}\right] = \frac{E[|N_{k,l}|^2]}{E[|S_{k,l}^P|^2]}, \quad (7)$$

since for $|S_{k,l}^P|$ is constant $E[1/|S_{k,l}^P|^2] = 1/E[|S_{k,l}^P|^2]$.

Since the channel impulse response is shorter than the cyclic prefix, which is just a fraction of the block duration, we could employ training blocks that are shorter than the standard data blocks. Alternatively, we could use the enhanced channel estimates $\{\tilde{H}_{k,l}; k = 0, 1, \ldots, N-1\}$ = DFT $\{\tilde{h}_{n,l} = \hat{h}_{n,l} w_n; n = 0, 1, \ldots, N-1\}$, where $w_n = 1$ if the nth time-domain sample is inside the cyclic prefix and 0 otherwise and $\{\hat{h}_{n,l}; n = 0, 1, \ldots, N-1\}$ = IDFT $\{\hat{H}_{k,l}; k = 0, 1, \ldots, N-1\}$. In this case, the SNR at the channel estimates is improved by a factor N/N_G.

Let us consider now the use of implicit pilots, i.e., $S_{k,l} \neq 0$ for the subcarriers with pilots. In the following we will assume that

$$E[|S_{k,l}|^2] = 2\sigma_D^2 \quad (8)$$

and, for the frequencies that have pilots,

$$E[|S_{k,l}^P|^2] = 2\sigma_P^2 \quad (9)$$

Clearly, we will have interference between data symbols and pilots. On the one hand, the data symbols produce interference on pilots, which might lead to inaccurate channel estimates; therefore, we should have $\sigma_D^2 << \sigma_P^2$. On the other hand, the pilots produce interference on data symbols, which might lead to performance degradation (even if the channel estimation was perfect), requiring $\sigma_D^2 >> \sigma_P^2$.

To overcome these problems, we can employ pilots with relatively low power (i.e., $\sigma_P^2 << \sigma_D^2$) and average the pilots over a large number of blocks so as to obtain accurate channel estimates. This is very effective since the data symbols have usually zero mean and different data blocks are uncorrelated. Naturally, there are limitations on the length of this averaging window, since the channel should be constant within it (not to mention the associated delays). Once we have an accurate channel estimate, we can detect the data symbols, eventually removing first the signal associated to the pilots.

Let us assume a frame structure with N_T time-domain blocks, each with N subcarriers. If the cyclic prefix of each FFT block has N_G samples we will need N_G equally spaced frequency-domain pilots for the channel estimation. For pilot spacings in time and frequency ΔN_T and ΔN_F, respectively, the total number of pilots in the frame is given by (1). This means that we have a pilot redundancy of

$$N_R = \frac{N_P^{Frame}}{N_G} = \frac{N}{N_G \Delta N_F} \cdot \frac{N_T}{\Delta N_T}. \quad (10)$$

Therefore, the SNR associated to the channel estimation procedure is

$$SNR_{est} = \frac{N_R \sigma_P^2}{\sigma_N^2 + \sigma_D^2} = N_R \frac{\sigma_P^2}{\sigma_D^2} SNR_{data} \frac{1}{1 + SNR_{data}}, \quad (11)$$

where

$$\sigma_N^2 = \frac{1}{2} E[|N_{k,l}|^2] \quad (12)$$

and the SNR associated to data symbols is given by

$$SNR_{data} = \frac{\sigma_D^2}{\sigma_N^2}. \quad (13)$$

For moderate and high SNR values,

$$SNR_{est} \approx N_R \frac{\sigma_P^2}{\sigma_D^2}. \quad (14)$$

To avoid significant performance degradation due to channel estimation errors, SNR_{est} should be high. This could be achieved with $\sigma_P^2 << \sigma_D^2$, provided that $N_R >> 1$.

III. Joint Detection and Channel Estimation

In this section we present a receiver with joint detection and channel estimation for OFDM with implicit pilots. The receiver structure is depicted in fig. 1. Without loss of generality it is assumed that there is a pilot for each subcarrier of each block of the frame, i.e., $\Delta N_F = \Delta N_T = 1$, leading to $N_P^{Frame} = NN_T$ and a pilot multiplicity or redundancy of $N_R = N_P^{Frame}/N_G = NN_T/N_G$.

The principles behind this receiver are the following:

(1) We first obtain the channel frequency response estimate

$$\tilde{H}_k^{(1)} = \frac{1}{N_T}\sum_{l=1}^{N_T}\frac{Y_{k,l}^{Rx}}{S_k^P}, \tag{15}$$

where $\{Y_{k,l}^{Rx}; k = 0,1,\dots,N-1\}$ denotes the lth received frequency-domain block ($l = 1,2,\dots,N_T$).

(2) This channel estimate is enhanced by ensuring that the corresponding impulse response has duration N_G, i.e., we use the channel estimation $\{\hat{H}_k^{(1)}; k = 0,1,\dots,N-1\}$ = DFT $\{\hat{h}_n^{(1)} = \tilde{h}_n^{(1)} w_n; k = 0,1,\dots,N-1\}$, where $\{\tilde{h}_n^{(1)}; k = 0,1,\dots,N-1\}$ = IDFT $\{\tilde{H}_k^{(1)}; k = 0,1,\dots,N-1\}$.

(3) The pilots are removed from the received frequency-domain blocks, leading to the blocks $\{Y_{k,l}^{(1)} = Y_{k,l}^{Rx} - \hat{H}_k^{(1)} S_k^P; k = 0,1,\dots,N-1\}$ and the N_T blocks of equalized samples (one for each block of the frame),

$$\tilde{S}_{k,l}^{(1)} = \frac{Y_{k,l}^{(1)}\hat{H}_k^{(1)*}}{|\hat{H}_k^{(1)}|^2}, \tag{16}$$

are generated.

(4) The equalized blocks are submitted to a decision device so as to obtain the average values of the transmitted symbols $\{\overline{S}_{k,l}^{(2)}; k = 0,1,\dots,N-1\}$ that will be used in the next iteration.

(5) For the second iteration, the pilots are removed from the received blocks and the average values of the data symbols will be used as training symbols for obtaining the channel frequency response estimate

$$\tilde{H}_k^{(2)} = \frac{\sum_{l=1}^{N_T} Y_{k,l}^{(1)}\overline{S}_k^{(1)*}}{\sum_{l=1}^{N_T}|\overline{S}_k^{(1)}|^2}. \tag{17}$$

(6) As in (2), an enhanced channel estimate $\{\hat{H}_k^{(2)}; k = 0,1,\dots,N-1\}$ = DFT $\{\hat{h}_n^{(2)} = \tilde{h}_n^{(1)} w_n; k = 0,1,\dots,N-1\}$, where $\{\tilde{h}_n^{(2)}; k = 0,1,\dots,N-1\}$ = IDFT $\{\tilde{H}_k^{(2)}; k = 0,1,\dots,N-1\}$, is computed.

(7) Repeat steps (3) to (6), for each iteration of the receiver.

The average values associated to the data symbols are given by

$$\overline{S}_{k,l}^{(i)} = \tanh\left(\frac{L_{k,l}^{I(i)}}{2}\right) + j\tanh\left(\frac{L_{k,l}^{Q(i)}}{2}\right), \tag{18}$$

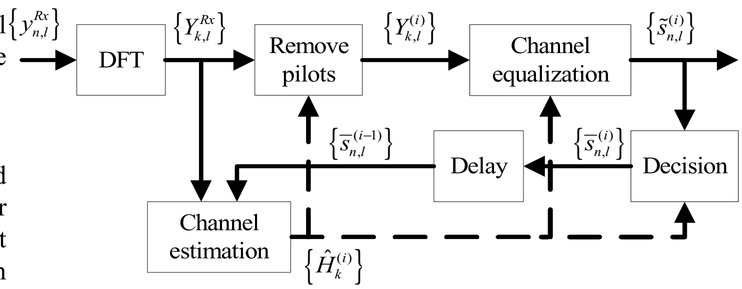

Fig. 1. Receiver structure.

where the LLRs (LogLikelihood Ratios) of the "in-phase bit" and the "quadrature bit", associated to $S_{n,l}^{I(i)}$ and $S_{n,l}^{Q(i)}$, respectively, are given by

$$L_{k,l}^{I(i)} = \frac{2|\hat{H}_k^{(i)}|^2}{\sigma_N^2}\tilde{S}_{k,l}^{I(i)} \tag{19}$$

and

$$L_{k,l}^{Q(i)} = \frac{2|\hat{H}_k^{(i)}|^2}{\sigma_N^2}\tilde{S}_{k,l}^{Q(i)}, \tag{20}$$

respectively.

The log-likelihood values can computed on a symbol-by-symbol basis (i.e., we do not need to perform the channel decoding in the feedback loop). As an alternative, we can employ the channel decoder outputs instead of the uncoded "soft decisions" in the feedback loop. In this case, a SISO channel decoder (Soft-In, Soft-Out) is employed in the feedback loop. The SISO block, that can be implemented as defined in [9], provides that the LLRs of both the "information bits" and the "coded bits". The input of the SISO block are LLRs of the "coded bits" at the equalizer output, given by (19) and (20).

IV. Performance Results

In this section we present a set of performance results concerning the OFDM schemes with implicit pilots and the proposed receivers with joint detection and channel estimation. The frame has N_T FFT-blocks, each with $N = 512$ data symbols selected from a QPSK constellation under a Gray mapping rule (similar results were observed for other values of N, provided that $N >> 1$). There is an implicit pilot for each symbol of each FFT-block, i.e., $\Delta N_T = \Delta N_F = 1$. The channel impulse response is characterized by the PDP (Power Delay Profile) of fig. 2, with uncorrelated Rayleigh fading on the different paths (similar results were observed for other severely time-dispersive channels). The channel is assumed to be invariant within the frame duration. The duration of the useful part of the data blocks (N symbols) is 1μs and the cyclic prefix has duration 0.125μs. Linear power amplification is considered at the transmitter and perfect synchronization is assumed at the receiver. The channel encoder is a rate-1/2 turbo code [8] based on two identical recursive convolutional codes characterized by $G(D) = [1 \quad (1+D^2)/(1+D+D^2)]$. A random interleaver is employed within the turbo encoder and the coded bits are also interleaved before being mapped into a QPSK constellation and distributed by the symbols of the frame.

Unless otherwise stated, our performance results are expressed as a function of E_b/N_0, where N_0 is the one-sided power spectral density of the noise and E_b is the energy of the transmitted bits (i.e., the degradation due to the useless power spent on the cyclic prefix (about 0.5dB, in our case) is not included). Since we are considering a rate-1/2 channel encoder, the energy of the corresponding information bits is 3dB higher.

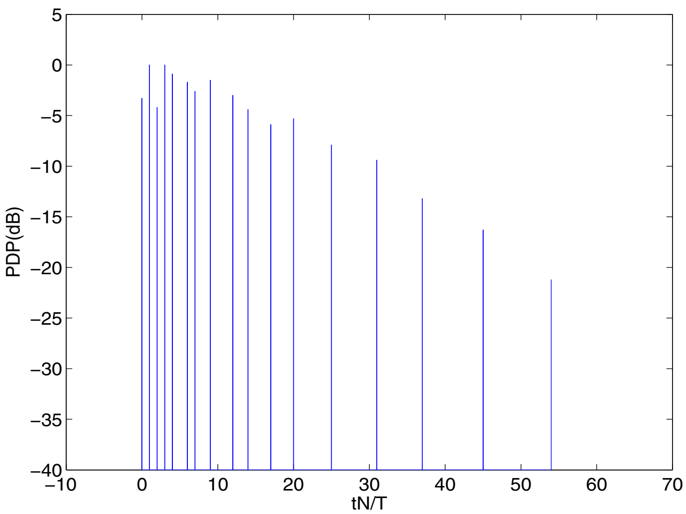

Fig. 2. Adopted PDP.

Let us consider first a conventional OFDM receiver where the channel estimation is made from the implicit pilots (i.e., just the first iteration). The turbo decoder has 12 iterations. Fig. 3 shows the BER performance for $N_T = 12$ and different values of

$$\beta_P = \sigma_P^2/\sigma_D^2. \tag{21}$$

We also included the performance with perfect channel estimation (and $\sigma_P^2 = 0$). Clearly, channel estimates based only on low-power pilots can be very poor, leading to significant performance degradation (the performance degradation is already high for $\beta_P = 1/4$). From this figure it might seem that we should spent significant power on the pilots. However, if we express the performance as a function of the total power (i.e., including the power spent on the pilots) instead of the power associated to data symbols (this corresponds to an additional degradation of

$$10\log_{10}((\sigma_P^2 + \sigma_S^2)/\sigma_S^2) = 10\log_{10}(1+\beta_P), \tag{22}$$

it is clear that the best performance is achieved for relatively moderate pilot powers, as depicted in fig. 4.

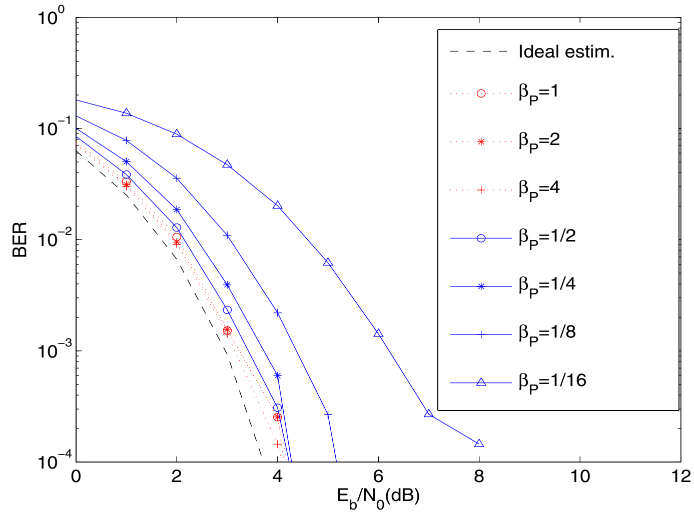

Fig. 3. BER performance for conventional OFDM receivers (one iteration).

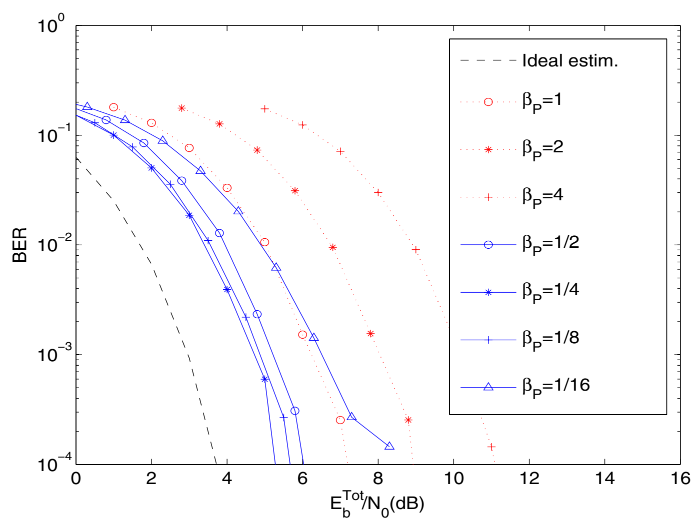

Fig. 4. As fig. 3, but with the BER a a function of the total E_b/N_0 (i.e., including the power spent on the pilots).

Let us consider now the proposed iterative receivers with joint detection and channel estimation. The receiver has 4 iterations with detection and channel estimation procedures. For each detection/estimation iteration we perform 3 iterations of the turbo decoder. To speed up the decoding procedure, the extrinsic values of the decoding procedure of the previous detection/estimation iteration are stored and used as *à priori* information for the next decoding procedure. The results with perfect channel estimation are obtained with 12 iterations of the turbo decoder.

Figs. 5 to 7 concern different values of N_T, as well as different values of σ_D^2/σ_P^2.[2] From these figures we can conclude that it is possible to have excellent performances, close to the ones with perfect channel estimation, even for pilots with relatively low power. As expected, the performances are much better for larger frames (higher values of N_T), allowing $\beta_P = 1/16$ for $N_T = 12$. For smaller frames β_P, should be slightly higher (about 1/8 for $N_T = 4$ and 1/4 for $N_T = 2$).

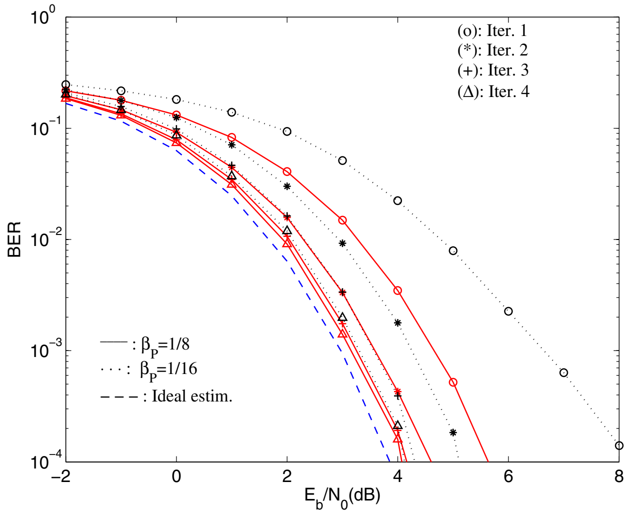

Fig. 5. BER performance for the iterative receiver, when $N_T = 12$.

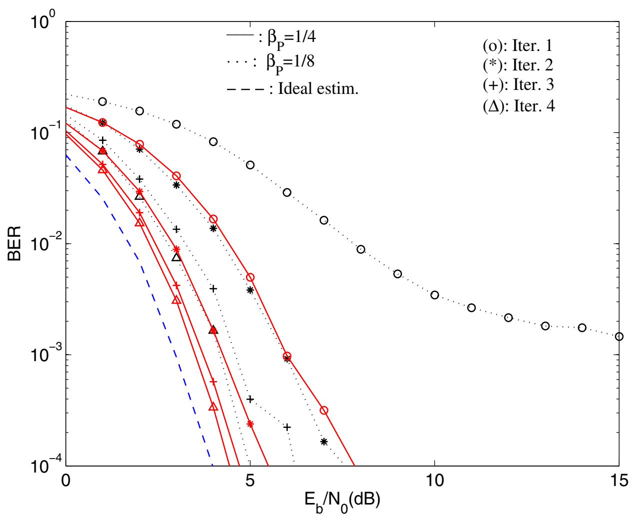

Fig. 6. BER performance for the iterative receiver, when $N_T = 4$.

V. Conclusions

In this paper we considered OFDM systems where the channel estimation is based on implicit pilots. To overcome the difficulties inherent to the interference levels between data and pilots, we proposed an iterative receiver structure with joint detection and channel estimation.

Our performance results show that the use of implicit pilots, combined with the proposed receiver, allows performances close to the ones with perfect channel estimation, even when low-power pilots or sort frames are adopted.

[2] It should be noted that the interleaver depth for the turbo code is $N_T N$. Therefore, the achievable performance (i.e., the performance with perfect channel estimation) is slightly worse for smaller values of N_T.

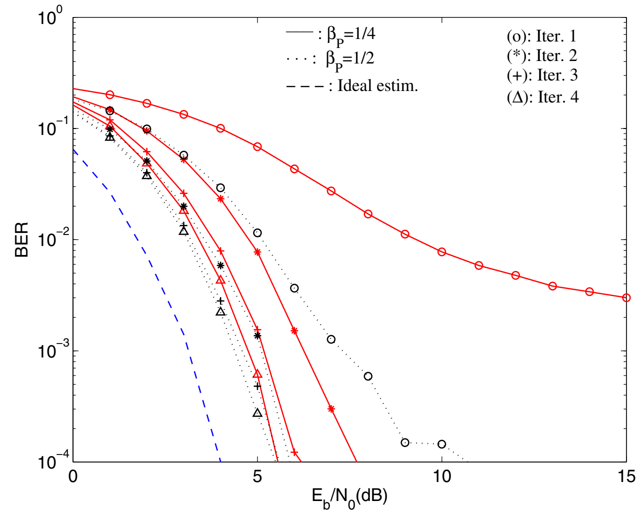

Fig. 7. BER performance for the iterative receiver, when $N_T = 2$.

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