On the Use of Implicit Pilots for Channel Estimation with OFDM Modulations

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Abstract - A coherent detection of OFDM signals (Orthogonal Frequency Division Multiplexing) allows good performances in severely time-dispersive channels. However, coherent receivers require good channel estimates, a difficult task when the channel impulse response is long and/or when space diversity techniques are employed.

Usually, the channel estimates are obtained with the help of pilot symbols multiplexed with data symbols, which reduces the useful bit rate, decreasing the spectral efficiency of the systems. To avoid this, we consider the use of implicit pilots instead of multiplexed pilots. Since the interference levels between data and pilots might be very high, we propose an iterative receiver with joint detection and channel estimation.

Our performance results show that our technique allows the use of low-power pilots, with performances close to the ones with perfect channel estimation, even when space diversity techniques are employed.[1]

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

I. Introduction

OFDM modulations (Orthogonal Frequency Division Multiplexing) [1] are suitable for high rate transmission over severely time-dispersive channels, since they have good performances and allow low-cost, FFT-based (Fast Fourier Transform) implementations. 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].

Both non-coherent and coherent receivers were proposed for OFDM modulations [4]. Non-coherent receivers do not require channel estimation, but their performance is worse than coherent receiver, provided that accurate channel estimates are available at the receiver. Accurate channel estimation is even more critical when space diversity techniques are employed at the receiver, since SNR (Signal-to-Noise Ratio) associated to each diversity branch can be much lower than the SNR for detection purposes.

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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 [5], [6], [7]. Since the channel impulse response is usually very long, the required channel estimation overheads can be high, namely for fast-varying scenarios and/or for bursty transmission. This leads to a reduction on the useful bit rate, decreasing the spectral efficiency of the systems. A promising method for overcoming this problem is to employ implicit pilots, which are added to the data block instead of being multiplexed with it [8]. This means that we can increase significantly the pilots' density, while keeping the system capacity. In fact, we can even 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.

In this paper, we consider the use of implicit pilots in OFDM systems with space diversity. We propose an iterative receiver structure with joint detection and channel estimation. For the first iteration, the channel associated to each diversity branch is estimated by averaging the received signal (data plus pilots) over several blocks; for the remaining iterations, enhanced the 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 with N subcarriers and N_T time-domain blocks, each one corresponding to an "FFT block" (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 transmitted OFDM signal associated to the frame is

$$s^{Tx}(t) = \sum_{m=1}^{N_T} s_m^{Tx}(t - mT_B), \quad (1)$$

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

$$s_m^{Tx}(t) = \sum_{n=-N_G}^{N-1} s_{n,m}^{Tx} h_T(t - nT_S), \quad (2)$$

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

$$S_{k,m}^{Tx} = S_{k,m} + S_{k,m}^P, \quad (3)$$

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

We will consider a receiver with L-order space diversity. The signal at the receiver input associated to the mth block and the lth receive antenna is sampled and the cyclic prefix is removed, leading to the time-domain block $\{y_{n,m}^{Rx(l)}; n = 0, 1, \ldots, N-1\}$. 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 $\{Y_{k,m}^{Rx(l)}; k = 0, 1, \ldots, N-1\}$, where

$$\begin{aligned} Y_{k,m}^{Rx(l)} &= S_{k,m}^{Tx} H_{k,m}^{(l)} + N_{k,m}^{(l)} = \\ &= (S_{k,m} + S_{k,m}^P) H_{k,m}^{(l)} + N_{k,m}^{(l)}, \end{aligned} \quad (4)$$

with $H_{k,m}^{(l)}$ denoting the overall channel frequency response for the kth frequency of the mth time block, at the lth receive antenna, and $N_{k,m}^{(l)}$ denoting the corresponding channel noise.

B. Channel Estimation

Let us first assume that $S_{k,m} = 0$, 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:

$$\tilde{H}_{k,m}^{(l)} = \frac{Y_{k,m}^{Rx(l)}}{S_{k,m}^P} = H_{k,m}^{(l)} + \frac{N_{k,m}^{(l)}}{S_{k,m}^P} = H_{k,m}^{(l)} + \epsilon_{k,m}^{H(l)}. \quad (5)$$

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

$$E[|\epsilon_{k,m}^{H(l)}|^2 | S_{k,m}] = E[|N_{k,m}^{(l)}|^2] E\left[\frac{1}{|S_{k,m}^P|^2}\right] = \frac{E[|N_{k,m}^{(l)}|^2]}{E[|S_{k,m}^P|^2]}, \quad (6)$$

(it is assumed that $|S_{k,m}^P|$).

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 $\{\tilde{H}_{k,m}^{(l)}; k = 0, 1, \ldots, N-1\}$ = DFT $\{\tilde{h}_{n,m}^{(l)} = \hat{h}_{n,m}^{(l)} w_n; n = 0, 1, \ldots, N-1\}$, where $w_n = 1$ for $0 \leq n \leq N_G - 1$ and 0 otherwise and $\{\hat{h}_{n,m}^{(l)}; n = 0, 1, \ldots, N-1\}$ = IDFT $\{\hat{H}_{k,m}^{(l)}; k = 0, 1, \ldots, N-1\}$. 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., $S_{k,m} \neq 0$ for the pilots. In the following we will assume that

$$E[|S_{k,m}|^2] = 2\sigma_D^2 \quad (7)$$

and, for the frequencies that have pilots,

$$E[|S_{k,m}^P|^2] = 2\sigma_P^2 \quad (8)$$

Clearly, we will have interference between data symbols and pilots. This leads to performance degradation for two reasons:

(1) The data symbols produce interference on pilots, which might lead to inaccurate channel estimates. This effect is negligible if $\sigma_D^2 << \sigma_P^2$.

(2) The pilots produce interference on data symbols, which might lead to performance degradation (even if the channel estimation was perfect). This effect is negligible if $\sigma_D^2 >> \sigma_P^2$.

We can have accurate channel estimates with low-pilots (i.e., with $\sigma_P^2 << \sigma_D^2$), provided that we average the pilots over a large number of blocks. 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. Once we have an accurate channel estimate, we can remove the pilots from the received signal and detect the data symbols.

If the channel impulse response (and the cyclic prefix of each FFT block) has $N_G = NT_G/T$ 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 Naturally, the number of pilots per frame is

$$N_P^{Frame} = \frac{N}{\Delta N_F} \cdot \frac{N_T}{\Delta N_T}. \quad (9)$$

This means that we have a pilot multiplicity or 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

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

where

$$\sigma_N^2 = \frac{1}{2} E[|N_{k,m}|^2] \tag{12}$$

and the SNR associated to data symbols is $SNR_{data} = \sigma_D^2/\sigma_N^2$. For moderate and high SNR values,

$$SNR_{est} \approx N_R \frac{\sigma_P^2}{\sigma_D^2}. \tag{13}$$

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. Decision-Directed Channel Estimation

In this section we present an iterative receiver with decision-directed channel estimation for OFDM schemes 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 estimation/detection procedure is as follows:

(1) We first obtain the L channel frequency response estimates

$$\tilde{H}_k^{(l,1)} = \frac{1}{N_T} \sum_{m=1}^{N_T} \frac{Y_{k,m}^{Rx(l)}}{S_{k,m}^P},\ l = 1, 2, \ldots, L \tag{14}$$

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

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

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

$$\tilde{S}_{k,m}^{(1)} = \frac{\sum_{l=1}^L Y_{k,m}^{(l,1)} \hat{H}_k^{(l,1)*}}{\sum_{l=1}^L |\hat{H}_k^{(l,1)}|^2}, \tag{15}$$

are generated.

(4) The equalized blocks are submitted to a decision device so as to obtain the average values of the transmitted symbols $\{\overline{S}_{k,m}^{(l,2)}; k = 0, 1, \ldots, N-1\}$ 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

$$\tilde{H}_k^{(l,2)} = \frac{\sum_{m=1}^{N_T} Y_{k,m}^{(l,1)} \overline{S}_k^{(1)*}}{\sum_{m=1}^{N_T} |\overline{S}_k^{(1)}|^2}. \tag{16}$$

(6) As in (2), the L enhanced channel estimates $\{\hat{H}_k^{(l,2)}; k = 0, 1, \ldots, N-1\}$ = DFT $\{\hat{h}_n^{(l,2)} = \tilde{h}_n^{(l,1)} w_n; k = 0, 1, \ldots, N-1\}$, where $\{\tilde{h}_n^{(l,2)}; k = 0, 1, \ldots, N-1\}$ = IDFT $\{\tilde{H}_k^{(l,2)}; k = 0, 1, \ldots, N-1\}$, $l = 1, 2, \ldots, L$, are computed.

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

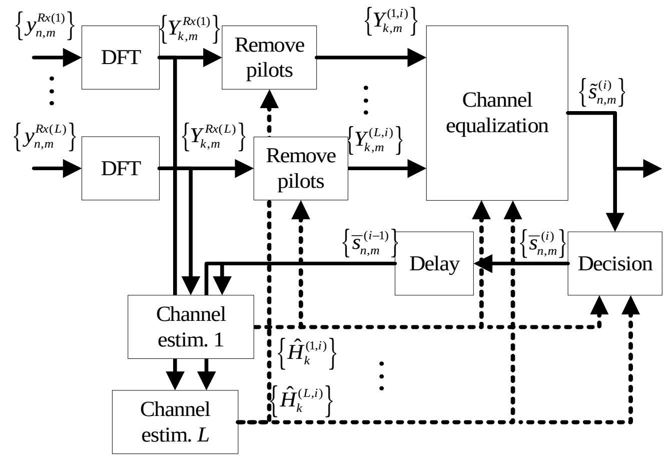

Fig. 1. Receiver structure.

The average values associated to the data symbols are given by

$$\overline{S}_{k,m}^{(i)} = \tanh\left(\frac{L_{k,m}^{I(i)}}{2}\right) + j \tanh\left(\frac{L_{k,m}^{Q(i)}}{2}\right), \tag{17}$$

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

$$L_{k,m}^{I(i)} = \frac{2\tilde{S}_{k,m}^{I(i)}}{\sigma_N^2} \sum_{l=1}^L |\hat{H}_k^{(l,i)}|^2 \tag{18}$$

and

$$L_{k,m}^{Q(i)} = \frac{2\tilde{S}_{k,m}^{Q(i)}}{\sigma_N^2} \sum_{l=1}^L |\hat{H}_k^{(l,i)}|^2, \tag{19}$$

respectively[2].

If we do not perform the channel decoding in the feedback loop, the log-likelihood values can computed on a symbol-by-symbol basis. However, we can improve significantly the performance if the channel decoding procedure is performed before each channel estimation iteration. In this case, a SISO

[2]Since the sums in (18)-(19) cancel the sum in the denominator of (15), they can be removed in both places.

channel decoder (Soft-In, Soft-Out) is employed in the feedback loop. The SISO block, that can be implemented as defined in [11], 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 (18) and (19).

IV. Performance Results

In this section we present a set of performance results concerning channel estimation using implicit pilots for OFDM modulations. The frame has $N_T = 12$ FFT-blocks, each with $N = 512$ subcarriers. The data symbols are selected from a QPSK constellation under a Gray mapping rule (similar results were observed for other values of N, provided that $N >> 1$). $\Delta N_T = \Delta N_F = 1$, i.e., there is an implicit pilot for each symbol of each FFT-block. The receiver has L diversity branches and the channel associated to each diversity antenna is based on the power delay profile type C for HIPERLAN/2 (HIgh PERformance Local Area Network) [12], with uncorrelated Rayleigh fading on the different paths and different diversity branches. The duration of the overall channel impulse response is 12.5% of the duration of the useful part of the block (i.e., $T_G/T = N_G/N = 0.125$). This means that the pilot multiplicity or redundancy is $N_R = 96$. The channel is assumed to be invariant within the frame duration. 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 [10] 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.

Our performance results are expressed as a function of E_b/N_0 or E_b^{Tot}/N_0, where N_0 is the one-sided power spectral density of the noise, 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) and $E_b^{Tot} = E_b + 10log_{10}((\sigma_P^2 + \sigma_D^2)/\sigma_D^2)$ (dB) is the total bit energy, including the energy spent on the pilots. Since we are considering a rate-1/2 channel encoder, the energy of the corresponding information bits is 3dB higher.

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. 2 shows the BER performance for $L = 1$ (no diversity), 2 and 4 and different values of $\beta_P = \sigma_P^2/\sigma_D^2$. 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 E_b^{Tot}/N_0 (i.e., including the power spent on the pilots) instead of just the power associated to data symbols (this corresponds to an additional degradation of $10\log_{10}((\sigma_P^2 + \sigma_S^2)/\sigma_S^2) = 10\log_{10}(1 + \beta_P))$, it is clear that the power spent on the pilots should not be too high, as depicted in fig. 3. From figs. 2 and 3, it is also clear that the required pilot power is not too different for different values of L.

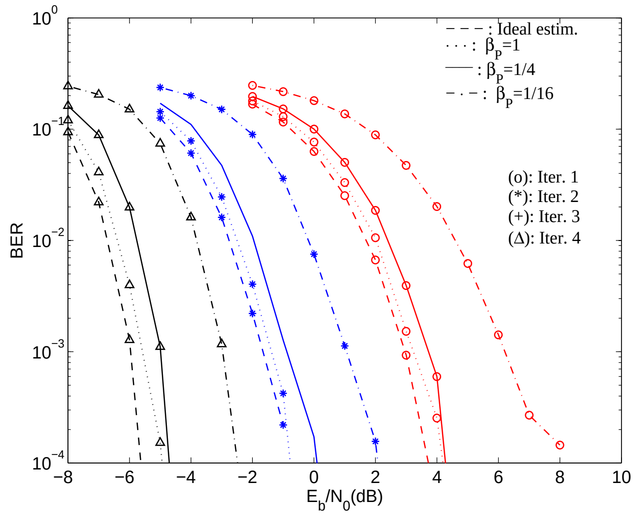

Fig. 2. BER performance for conventional OFDM receivers (one iteration).

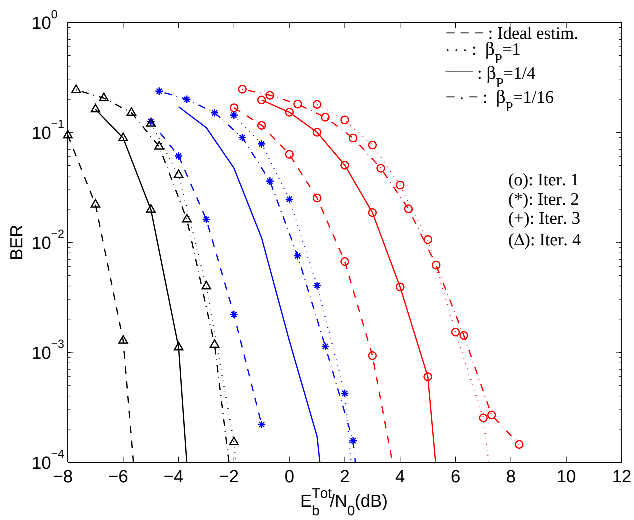

Fig. 3. As fig. 2, but with BER as function of E_b^{Tot}/N_0 (i.e., including the power spent on the pilots).

Let us consider now the proposed iterative receivers with joint detection and channel estimation. The receiver has 4 iterations with channel estimation and detection 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. 4, 5 and 6 concern $L = 1$ (no diversity), 2 and 4, respectively. 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, even for $L = 4$.

It should be pointed out that, although the performance is better for large values of N_T (i.e., larger frames) our technique is still effective for values of N_T as small as 4 or even 2 (naturally, the power spent on the pilots should be higher for smaller values of N_T). This means that we can use implicit pilots for channel estimation even with time-varying channels.

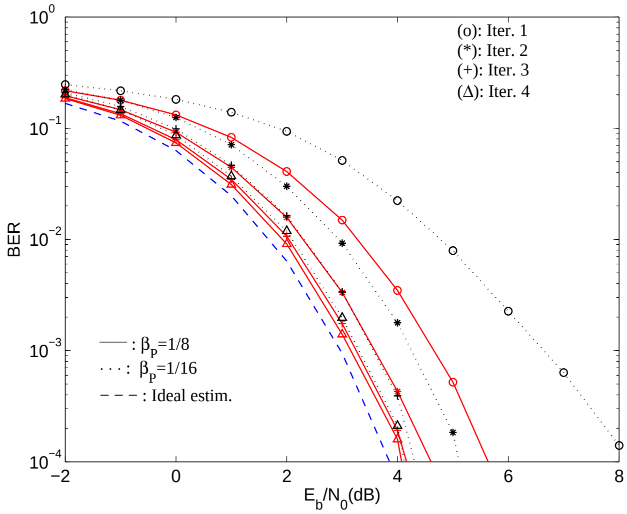

Fig. 4. BER performance for the iterative receiver, when $L = 1$.

V. Conclusions

In this paper we considered channel estimation based on implicit pilots for OFDM schemes with space diversity and coherent receivers. We proposed an iterative receiver with joint detection and channel estimation to cope with the interference levels between data and pilots.

Our performance results showed that our technique allows the use of low-power pilots, with performances close to the ones with perfect channel estimation, even when space diversity techniques are employed.

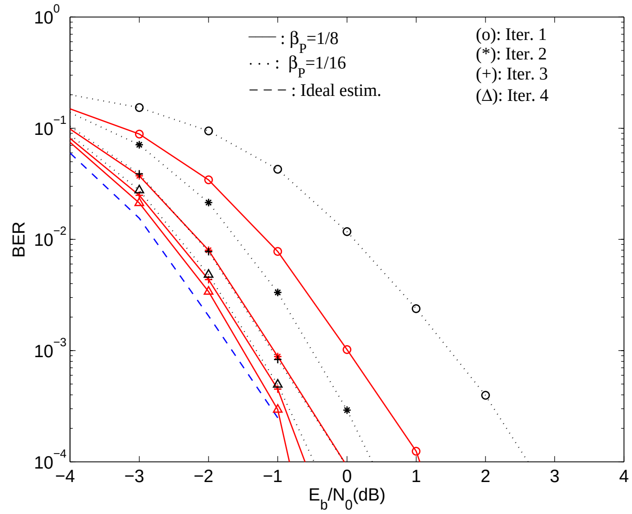

Fig. 5. BER performance for the iterative receiver, when $L = 2$.

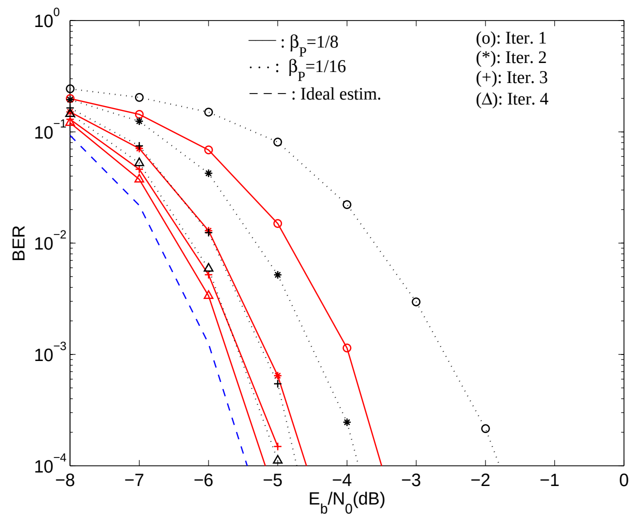

Fig. 6. BER performance for the iterative receiver, when $L = 4$.

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