JOINT DETECTION AND CHANNEL ESTIMATION FOR MIMO SYSTEMS WITH SC-FDE MODULATIONS

João Carlos Silva[1], Rui Dinis[2], Nuno Souto[1]
[1] ISCTE / Instituto de Telecomunicações, Lisboa, Portugal,
joao.carlos.silva@lx.it.pt, nuno.souto@lx.it.pt
[2] ISR/IST, Lisboa, Portugal, rdinis@ist.utl.pt

ABSTRACT
SC modulation (Single-Carrier) with FDE (Frequency-Domain Equalization) allows excellent performance in severely time-dispersive channels, provided that accurate channel estimates are available at the receiver. For this purpose, pilot symbols and/or training sequences are usually multiplexed with data symbols, which lead to spectral degradation. As an alternative, we can use implicit pilots (i.e., pilots superimposed to data).
In this paper we consider MIMO SC-FDE systems where the channel estimation is based on either explicit or implicit pilots, for comparison purposes. An iterative receiver with joint equalization, turbo decoding and channel estimation was employed for optimum results, and to reduce the high interference levels between data and pilots (for the implicit pilots). The main differences between the different schemes are discussed and the performance results show that the use of the proposed techniques for channel estimation yield excellent results.

KEY WORDS
MIMO, Channel estimation, FDE, single-carrier modulations.

1. Introduction

Due to the lower envelope fluctuations of the transmitted signals (and, implicitly a lower PMEPR (Peak-to-Mean Envelope Power Ratio)), SC-FDE schemes are especially interesting for the uplink transmission (i.e., the transmission from the mobile terminal to the base station) [1], [2].
A promising IFDE (Iterative FDE) technique for SC-FDE, denoted IB-DFE (Iterative Block Decision Feedback Equalizer), was proposed in [3]. This technique was later extended to diversity scenarios [4]. These IFDE receivers can be regarded as iterative DFE receivers with the feed-forward and the feedback operations implemented in the frequency domain. Since the feedback loop takes into account not just the hard decisions for each block but also the overall block reliability, error propagation is reduced. Consequently, IFDE techniques offer much better performance than non-iterative methods [3], [4]. Within these IFDE receivers the equalization and channel decoding procedures are performed separately (i.e., the feedback loop uses the equalizer outputs instead of the channel decoder outputs). However, it is known that higher performance gains can be achieved if these procedures are performed jointly. This can be done by employing turbo equalization schemes, where the equalization and decoding procedures are repeated in an iterative way [5]. Although initially proposed for time-domain receivers, turbo equalizers also allow frequency-domain implementations [6].
In order for the above schemes to operate correctly, good channel estimates are required at the receiver. Typically, these channel estimates are obtained with the help of pilot/ training symbols that are multiplexed with the data symbols, either in the time domain or in the frequency domain [7]. Overhead due to training symbols for channel estimation can be high, leading to decrease of system capacity, especially in fast-varying scenarios and/or high MIMO orders. A promising technique to overcome this problem is to use implicit training or implicit pilots, also called superimposed pilots, where the training block is added to the data block instead of being multiplexed with it [8]-[11]. This means that we can increase significantly the density of pilots (to the maximum extent of one pilot per data symbol), with zero pilot overhead, although with an increase in power. [12] provides a general framework for several approaches to low or zero pilot overhead in the context of OFDM. In one approach, periodic pilot sequences are added to data symbols in the time domain for single carrier systems [8], [11], [13], or in the frequency domain for OFDM systems [9]. The power level of the added pilots is chosen to minimize error rate degradation due to channel estimation errors and to loss of data power. The interference to pilots (and therefore to channel estimates) from data can be mitigated by time-averaging over many pilot sequence repetitions [8], [11]. Once channel estimates are obtained in this way, pilots are subtracted from the received signal prior to equalization and data detection. Improved channel estimation and data detection performance can be obtained with iterative joint maximum likelihood or quasi-maximum likelihood data detection and channel estimation procedures [11], [13].
In this paper, we consider the use of both multiplexed and implicit pilots. For the implicit case, non-data-dependent pilots were used; i.e. using the first approach mentioned above. We propose iterative receiver structures with joint channel estimation and detection. Unlike the iterative schemes of [11] and [13], our schemes do not employ the relatively complex Viterbi algorithm to jointly estimate channel and data – however, they incorporate iterative frequency domain equalization (either IB-DFE or turbo

equalization) within the iterative channel estimation and detection/decoding framework; an introductory work has been done by the authors in [16] for the implicit pilot case. Like [8], and unlike [11], the channel is estimated before the first iteration by averaging the received signal (data plus training) over several blocks. For the remaining iterations, enhanced channel estimates are obtained by considering the data symbols as an "extended" training. For the estimation and detection phases of each iteration we remove the undesirable signal (training or data) using the most updated version of it. Another problem associated with implicit training is that, by adding training to data signals the envelope fluctuations of the transmitted signals are increased. This is especially important when a low-PMEPR, SC-based transmission is intended. We examine the effect of the added training signals, at various relative powers, on the PMEPR.

This paper is organized as follows. The system considered in this paper is introduced in sec. II and sec. III describes the proposed channel estimation procedure. A set of performance results and the conclusions are presented in sec. IV.

2. System Description

We consider SC-FDE modulation. The l^{th} transmitted block has the form

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

with T_S denoting the symbol duration, N_G denoting the number of samples at the cyclic prefix, $h_T(t)$ representing the adopted pulse shaping filter, and $s_{n,l}^{ntx}$ denotes the length-*N* data block to be transmitted from the *ntx* transmit antenna. After passing the signal to the frequency domain, the implicit pilots can be added. It is better to add them in the frequency domain, since most of the processing is done there.

The transmitted sequences are thus given by

$$X_{k,l}^{ntx} = S_{k,l}^{ntx} + S_{k,l}^{pilot} \quad (2)$$

where, $S_{k,l}^{ntx}$ is the data symbol transmitted by the k^{th} subcarrier (out of a total of *N*) of the l^{th} FFT block and $S_{k,l}^{pilot}$ is the corresponding implicit pilot. Assuming only one user, the data bits are passed through a turbo coder, after which they are submitted to rate matching (taking into account the use of FFTs for faster processing, the antenna multiplexer and block partitioning). All of the antennas will transmit a part of the message (if multiple users were to be employed, we could assign an antenna per user). The data bits are partitioned into blocks and the cyclic prefix is added to each block, so that the total size is a power of 2, for efficient use of the FFT. The considered frame structure for the SC-FDE system with *N* carriers is the same as in [16].

The transmission of pilot symbols superimposed on data will clearly result in interference between them. To reduce the mutual interference and achieve reliable channel estimation and data detection we propose a receiver capable of jointly performing these tasks through iterative processing. The structure of the proposed iterative receiver is described in [16]; the signal, which is considered to be sampled and with the cyclic prefix removed, is converted to the frequency domain after an appropriate size-*N* FFT operation. If the cyclic prefix is longer than the overall channel impulse response, the *nrx* receive antenna is given as:

$$R_{k,l,nrx} = \sum_{ntx=1}^{Ntx}\left(\left(S_{k,l,ntx} + S_{k,l,ntx}^{Pilot}\right)H_{k,l,ntx,nrx} + N_{k,l,nrx}\right) \quad (3)$$

with $H_{k,l,ntx,nrx}$ denoting the overall channel frequency response for the k^{th} frequency of the l^{th} time block between the *ntx* transmit and *nrx* receive antenna, and $N_{k,l,nrx}$ denoting the corresponding channel noise. Before entering the equalization block, the pilot symbols are removed from the sequence resulting

$$\left(Y_{k,l,nrx}\right)^{(q)} = R_{k,l,nrx} - \sum_{ntx=1}^{Ntx}\left(S_{k,l,ntx}^{Pilot}\left(\hat{H}_{k,l,ntx,nrx}\right)^{(q)}\right) \quad (4)$$

where $\left(\hat{H}_{k,l,ntx,nrx}\right)^{(q)}$ are the channel frequency response estimates and q is the current iteration. The equalized samples are then simply computed as

$$\left(\hat{S}_{k,l,ntx}\right)^{(q)} = \frac{\left(\hat{H}_{k,l,ntx,nrx}\right)^{(q)^*}\left(Y_{k,l,ntx}\right)^{(q)}}{\left|\hat{H}_{k,l,ntx,nrx}{}^{(q)}\right|^2} \quad (5)$$

Where

$$\left(Y_{k,l,ntx}\right)^{(q)} = \sum_{nrx}\left(\left(Y_{k,l,nrx}\right)^{(q)} - \sum_{ntx1\neq ntx}\left(S_{k,l,ntx1}^{Pilot}\left(\hat{H}_{k,l,ntx1,nrx}\right)^{(q)}\right)\right) \quad (6)$$

The sequences of the equalized samples are then passed through the IFFT, block grouping, demodulated and passed through the channel decoder. Each channel decoder has two outputs. One is the estimated information sequence and the other is the sequence of log-likelihood ratio (LLR) estimates of the code symbols. These LLRs are passed through the Decision Device which outputs either soft-decision or hard decision estimates of the code symbols. These estimates enter the Transmitted Signal Rebuilder which performs the same operations of the transmitter (coding, modulation). The reconstructed symbol sequence can then be used for improving the channel estimates, as will be explained next, for the subsequent iteration

3. Channel Estimation using Pilots

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, the channel frequency response is:

$$\hat{H}_{k,l} = \frac{Y_{k,l}}{S_{k,l}^{TS}} = H_{k,l} + \frac{N_{k,l}}{S_{k,l}^{TS}} = H_{k,l} + \epsilon_{k,l}^{H} \quad (7)$$

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

$$E\left[\left|\epsilon_{k,l}^{H}\right|^2 S_{k,l}\right] = E\left[\left|N_{k,l}\right|^2\right] E\left[\frac{1}{\left|S_{k,l}^{TS}\right|^2}\right] \quad (8)$$

Since the power assigned to the training block is proportional to $E\left[\frac{1}{\left|S_{k,l}^{TS}\right|^2}\right]$ and $E\left[\left|S_{k,l}^{TS}\right|^2\right]$, the training blocks should be constant and equal to $\left|S_{k,l}^{TS}\right|^2 = 2\sigma_T^2$ for all *k*. On the other hand, if we want to minimize the envelope fluctuations of the transmitted signal $\left|s_{n,l}^{TS}\right|$ should also be constant. This can be achieved by employing Chu sequences, which have both $\left|s_{n,l}^{TS}\right|$ and $\left|S_{k,l}^{TS}\right|$ constant [17].

3.1 Estimation Algorithm with Implicit Pilots

To obtain the frequency channel response estimates the receiver applies the following steps in each iteration: Data symbols estimates are removed from the pilots. The resulting sequence becomes

$$\left(\tilde{R}_{k,l,nrx}\right)^{(q)} = R_{k,l,nrx} - \sum_{ntx=1}^{Ntx}\left(\left(\hat{S}_{k,l,ntx}\right)^{(q-1)}\left(\widehat{H}_{k,l,ntx,nrx}\right)^{(q-1)}\right) \quad , \qquad (9)$$

where $\left(\hat{S}_{k,l,ntx}\right)^{(q-1)}$ and $\left(\widehat{H}_{k,l,ntx,nrx}\right)^{(q-1)}$ are the data and channel response estimates of the previous iteration. This step can only be applied after the first iteration. In the first iteration we set $\left(\tilde{R}_{k,l,nrx}\right)^{(1)} = R_{k,l,nrx}$

The channel frequency response estimates is computed using a moving average with size W ,whilst at the same time removing the pilots, as follows (data is considered of zero mean):

$$\left(\widehat{H}_{k,l,ntx,nrx}\right)^{(q)} = \frac{1}{W}\sum_{l'=l-\lfloor W/2\rfloor}^{l+\lfloor W/2\rfloor}\frac{\left(\tilde{R}_{k,l',nrx}\right)^{(q-1)}}{S_{k,l',ntx}^{Pilot}} \qquad (10)$$

After the first iteration the data estimates can also be used as pilots for channel estimation refinement. This is especially useful if the spacing of pilot symbols in the time domain is $\Delta N_T > 1$. The respective channel estimates are computed as

$$\left(\widetilde{H}_{k,l,ntx,nrx}\right)^{(q)} = \frac{\left(Y_{k,l,nrx}\right)^{(q-1)}\left(\hat{S}_{k,l,ntx}\right)^{(q-1)^*}}{\left|\left(\hat{S}_{k,l,ntx}\right)^{(q-1)}\right|^2} \qquad (11)$$

These channel estimates are enhanced by ensuring that the corresponding impulse response has a duration N_G. This is accomplished by computing the time domain impulse response of (10) and (11) through $\{ \left(\tilde{h}_{i,l}\right)^{(q)}$; i = 0, 1, …, N-1} = IDFT{ $\left(\tilde{H}_{k,l}\right)^{(q)}$; k = 0, 1, …, N-1} (zeros can be used for the missing carriers if $\Delta N_F > 1$, in order to perform a "FFT-interpolation"), followed by the truncation of this sequence according to $\{ \left(\hat{h}_{i,l}\right)^{(q)} = w_i\left(\tilde{h}_{i,l}\right)^{(q)}$; i = 0, 1, …, N-1} with $w_i = 1$ if the i^{th} time domain sample is inside the cyclic prefix duration and $w_i = 0$ otherwise. The final frequency response estimates are then simply computed using { $\left(\hat{H}_{k,l}\right)^{(q)}$; k = 0, 1, …, N-1}= DFT{ $\left(\hat{h}_{i,l}\right)^{(q)}$; i = 0, 1, …, N-1}* ΔN_F .

3.2 Data-aided Estimation Algorithm with Multiplexed Pilots

Data-aided estimation was also employed for algorithms with multiplexed pilots, in order to be able to use a lesser amount of pilots and promote bandwidth efficiency. To obtain the frequency channel response estimates, the receiver applies the following steps:

In the first iteration, the channel estimates are simply computed from the pilots. Since only the first and last blocks were used for multiplexed pilots, the remaining block's estimates may be found via a linear interpolation.

In the second and posterior iterations, the process of getting the new channel estimates is simple, admitting now that the estimated data bits are our new pilots. Of course that this procedure will not yield good results if the estimates are wrong. Before doing this however, it is necessary to remove the transmit signal interference for each receive antenna, in order to obtain a channel estimate for a pair of transmit-receive antennas. This is done assuming the channel and data estimates of the previous iteration.

$$\ddot{\mathrm{H}}_{\mathrm{ntx,nrx}} = \frac{\left(R_{k,l,nrx} - \sum_{n=1,n\neq \mathrm{ntx}}^{Ntx}\left(\left(\hat{S}_{k,l,n}\right)^{(q-1)}\left(\widehat{H}_{k,l,n,nrx}\right)^{(q-1)}\right)\right)\cdot\left(\hat{S}_{k,l,ntx}\right)^{(q-1)^*}}{\left|\left(\hat{S}_{k,l,ntx}\right)^{(q-1)}\right|^2} \ (12).$$

The channel estimates are now processed through the DFT and IDFT, in order to guarantee that the corresponding impulse response has a duration N_G.

In the second and posterior iterations, the resulting channel estimate is blended with the estimate from the previous iteration, with a specific weight for the iteration at hand (weights of 10%, 20% and 30% were used).

$$\widetilde{H} = \mathrm{IW}^{(q)}\ddot{\mathrm{H}} + \left(1 - \mathrm{IW}^{(q)}\right)\widehat{\mathrm{H}}^{(q-1)} \qquad (13)$$

The resulting estimates are passed through a moving average filter, in order to cancel out some of the noise effect and to provide continuity.

$$\left(\widehat{H}_{k,l,ntx,nrx}\right)^{(q)} = \frac{1}{W}\sum_{l'=l-\lfloor W/2\rfloor}^{l+\lfloor W/2\rfloor}\left(\widetilde{H}_{k,l',ntx,nrx}\right)^{(q)} \qquad (14)$$

For all interpolations, the channel estimates are enhanced by ensuring that the corresponding impulse response has a duration N_G.

4. Numerical Results & Conclusion

The number of carriers employed was N=256, each carrying a QPSK data symbol. Each information stream was encoded with a variable block size per antenna, yielding a deterministic number of 256-bit blocks after the FFT conversion, as depicted in Table 1. For the implicit case, the overall block size was of 2880 bits, whereas for the multiplexed pilots case, it was 720 bits per antenna – this way we had a fixed number of blocks per antenna for the multiplexed pilots case, and the same overall amount of bits for the implicit case, in order to avoid coding gains. The multiplexed pilots case used an extra block dedicated for channel estimation. The channel impulse response employed is characterized an exponential PDP (Power Delay Profile) with 32 symbol-spaced taps and normalized delay spread 8 $\sum_{t_S=0}^{31} 10\log_{10}\left(\frac{e^{(t-t_S)}}{8}\right)$. A symbol duration of T_s=260ns was used. The channel encoders were rate-1/2 turbo codes based on two identical recursive convolutional codes with two constituent codes characterized by G(D) = [1 $(1+D^2+D^3)/(1+D+D^3)$]. A random interleaver was used within the turbo encoders. At the receiver 9 turbo decoding iterations were employed for the conventional receiver (i.e. one receiver iteration) while 3 receiver iterations each with 3 turbo decoding iterations were applied in the iterative scheme.

	Siso	Mimo 2x2	Mimo 4x4
Implicit	23	12	6
Multiplexed	6		

Table 1. Number of 256 bit blocks.

Most of the BER (Bit Error Rate) results presented next will be shown as a function of E_b/N_0, where E_b is the average data bit energy and N_0 is the single sided noise power spectral density. For channel estimation purposes, the moving average window size used was W=9, considering different values of power ratio β_P. The figures combine the use of perfect channel estimation, with estimation using multiplexed pilots (using $\Delta N_F = ntx$ and $\Delta N_T = N_T$) and estimation using implicit pilots (using $\Delta N_F = ntx$ and $\Delta N_T = 1$). Unless otherwise stated, the power of the pilots is taken to be the same of the mean symbols' power throughout the block, else it will be given by the ratio between pilots' powers and data symbols' powers, as

$$\beta_P = E\left[\left|S_{k,l}^{Pilot}\right|^2\right] \Big/ E\left[\left|S_{k,l}\right|^2\right]. \quad (15)$$

The channel block is considered constant as long as

$$f_d \times T_{chip} < 5\% \quad (16)$$

where f_d is the Doppler frequency and T_{chip} is the chip duration. Figure 1 portrays perfect-channel results for the conditions of the implicit setting of Table 1, and will be used for comparison purposes. Note that the high MIMO orders yield excellent results due to high diversity and good equalization receiver.

Figure 2 portrays results for the use of plain-multiplexed pilot estimation – notice that there is a larger difference between low and high speeds, easily explained by the extrapolation error caused between pilot-exclusive blocks. In Figure 3, the difference of using data-aided estimation is verified. The fact of using the data for refining the channel estimation after the first iteration proves crucial for obtaining close-to-perfect results. Figure 4 exemplifies that the data-aided estimation is only of value for speeds above a certain order, though. Lower speeds may actually be prejudiced by poor data estimation compared to good initial channel estimates.

Several simulations were run for different power levels of implicit pilots. For a certain E_s/N_0, there is an optimum value is a direct relationship between pilot power and diversity order. With a density of 100% (SISO), the optimum pilot power is -6dB, whereas for lower densities, higher powers are needed. Notice also from Figure 8 that the window size plays an important role as well. In this case, the worst scenario of v=200km/h was considered, and thus small window sizes were preferred. In figure 5, simulations were run with the optimum results from Table 2 for the *v=200km/h* and with higher window sizes for the *v=100km/h* case (value of 45 being the highest). It can be seen from the results that they are within 2dBs of the optimum case, with the natural advantage over the multiplexed-pilots estimation case of not using pre-allocated slots for channel estimation – the implicit pilot estimation case allows for comparable results using the full bandwidth for data transmission – and a small amount of power for implicit pilot estimation.

Setting	Window	Rel.P.	Density
1x1	13	-6	100%
2x2	13	-3	50%
4x4	45	0	25%

Table 2. Best settings for implicit pilot estimation

Acknowledgements

This paper was supported in part by FCT (pluriannual funding and project U-BOAT PTDC/EEA-TEL/67066/2006).

References

[1] A. Gusmão, R. Dinis, J. Conceição, and N. Esteves, "Comparison of Two Modulation Choices for Broadband Wireless Communications", Proc. IEEE VTC Spring, pp. 1300–1305, May 2000.

[2] D.Falconer, S.Ariyavisitakul, A.Benyamin-Seeyar and B.Eidson, "Frequency Domain Equalization for Single-Carrier Broadband Wireless Systems", IEEE Comm. Mag., Vol. 4, No. 4, pp. 58–66, April 2002.

[3] N.Benvenuto and S.Tomasin, "Block Iterative DFE for Single Carrier Modulation", IEE Electronic Letters, Vol. 39, No.19, September 2002.

[4] R. Dinis, A. Gusmao, and N. Esteves, "On Broadband Block Transmission over Strongly Frequency-Selective Fading Channels", Wireless 2003, Calgary, Canada, July 2003.

[5] M. Tuchler, R. Koetter and A. Singer, "Turbo Equalization: Principles and New Results", IEEE Trans. on Comm., Vol.50, May 2002.

[6] M. T¨uchler, J. Hagenauer, "Turbo Equalization Using Frequency Domain Equalizers,", Allerton Conf., Oct. 2000.

[7] P. Hoher, S. Kaiser, and P. Robertson, "Pilot-Symbol-Aided Channel Estimation in Time and Frequency", IEEE Communication Theory Mini-Conference (CTMC), IEEE GLOBECOM97, pp. 90-96, 1997.

[8] A. Orozco-Lugo, M. Lara and D. McLernon, "Channel Estimation Using Implicit Training", IEEE Trans. on Sig. Proc., Vol. 52, No. 1, Jan. 2004.

[9] C. Ho, B. Farhang-Boroujeny and F. Chin, "Added Pilot Semi-Blind Channel Estimation Scheme for OFDM in Fading Channels", IEEE GLOBECOM'01, Nov. 2001.

[10] C. Lam, D. Falconer, F. Danilo-Lemoine and R. Dinis, "Channel Estimation for SC-FDE Systems Using Frequency Domain Multiplexed Pilots", IEEE VTC'06(Fall) , Sep. 2006.

[11] X.Meng, J. Tugnait and S. He, "Iterative Joint Channel Estimation and Data Detection Using Superimposed Training: Algorithms and Performance Analysis", IEEE Transactions on Vehicular Technology, Vol. 56, No. 4, pp. 1873–1880, July 2007.

[12] S. Ohno and G. Giannakis, "Optimal Training and Redundant Precoding for Block Transmissions with Application to Wireless OFDM", IEEE Trans. on Communications, Vol. 50, No. 12, pp. 2113–2123, Dec., 2002.

[13] K. Josiam and D. Rajan, “Bandwidth Efficient Channel Estimation Using Super-Imposed Pilots in OFDM Systems”, IEEE Trans. Wireless Comm., Vol. 6, No. 6, pp. 2234–2245, June, 2007.
[14] M. Ghogho, D. McLernon, E. A-Hernandez and A. Swami, “Channel Estimation and Symbol Detection for Block Transmission Using Data-Dependent Superimposed Training”, IEEE Signal Processing Letters, Vol. 12, No. 3, pp. 226–229, March 2005.
[15] J. Tugnait and S. He, “Doubly-Selective Channel Estimation Using Data-Dependent Superimposed Training and Exponential Bases Models”, Proc. 40th Annual Conf. on Information Sciences and Systems, 2006.
[16] J. Silva, N. Souto, R. Dinis, “Efficient Channel Estimation for Iterative MIMO SC-FDE Systems”, IEEE VTC 2008 Fall, Calgary, Canada, 21-24 September 2008
[17] D. Chu, “Polyphase Codes with Good Periodic Correlation Properties”, IEEE Trans. Inform. Theory, Vol. 18, No. 4, pp.531-532, July 1972.

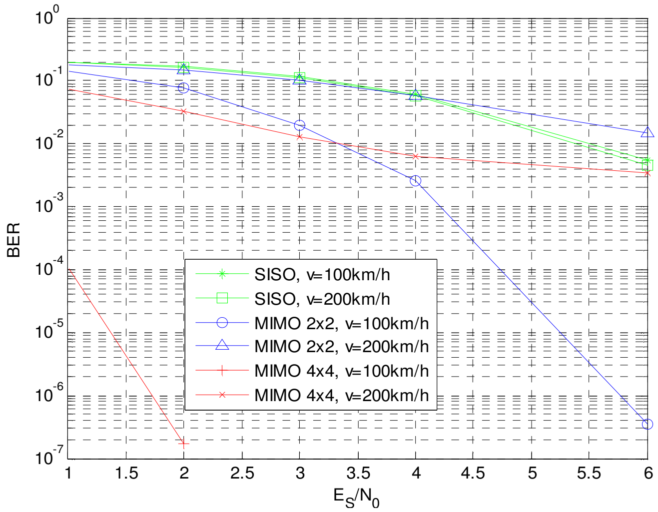

Figure 1. Perfect channel Estimation.

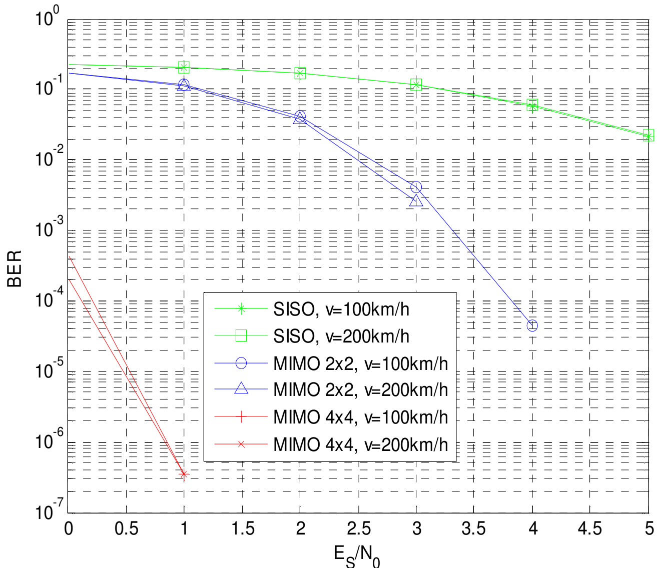

Figure 2. Channel estimation with multiplexed pilots.

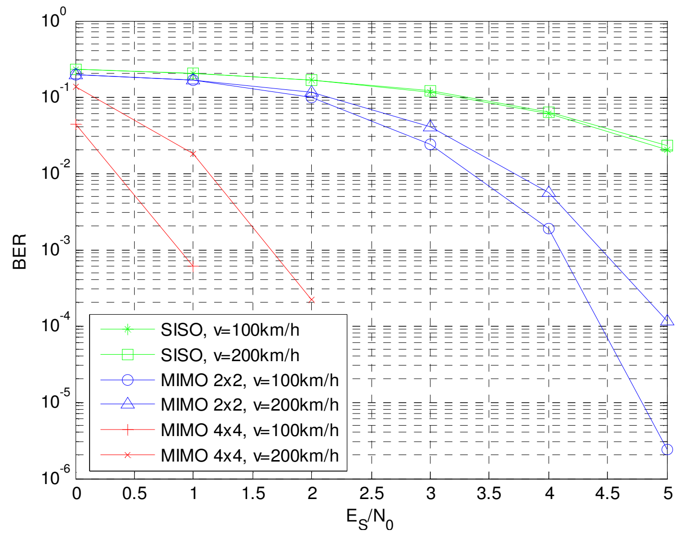

Figure 3. Channel estimation with multiplexed data-aided pilots.

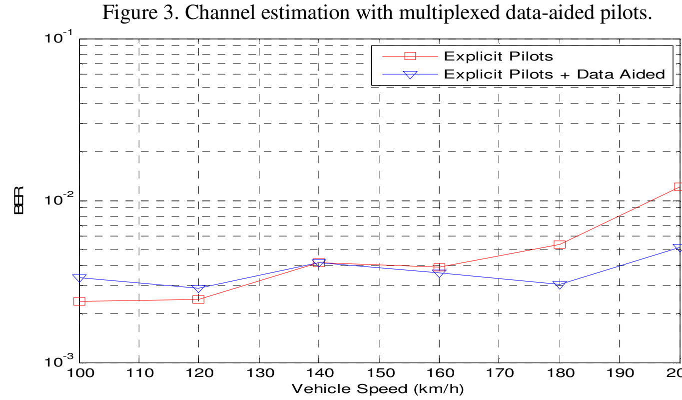

Figure 4. Best speeds for multiplexed-data-aided estimation, SISO E_s/N_0=6dB.

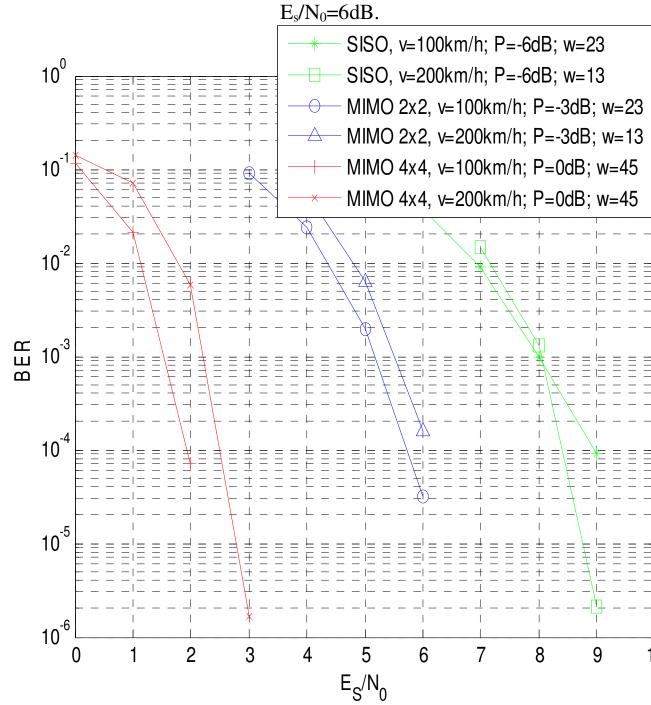

Figure 5. Channel estimation with implicit pilots