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SEMI-BLIND CHANNEL ESTIMATION IN HSDPA SYSTEMS

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ABSTRACT

In this paper we propose a semi-blind channel estimator for High Speed Downlink Packet Access (HSDPA) systems that use multicode transmission to obtain high data rates. In these systems spreading factors are small. Therefore, severe interference caused by multipath propagation may be experienced. The quality of the conventional channel estimates obtained using only pilot signal may not be sufficient. In this paper we propose a semi-blind channel estimator for HSDPA systems. It combines a blind and conventional pilot assisted channel estimate in a novel way. Semi-blind method has larger sample support since the information from both the pilot signal and known spreading codes is used. Consequently, the variance of the estimates is smaller. The performance of the proposed method is compared in simulation to conventional pilot assisted estimator. The proposed method provides significant performance gains over the pilot-based method. Moreover, it allows reducing the pilot power.

1. INTRODUCTION

Demand for higher data rate wireless communication is increasing. 3rd generation systems such as Wideband CDMA (WCDMA) can provide high data rates, see for example 3GPP specification on High Speed Downlink Packet Access (HSDPA) [1]. In this article we propose semi-blind channel estimation method for high rate CDMA systems using multicode transmission. The work extends the work in [5] to HSDPA systems, where in addition to the known codes a pilot signal (spread with different spreading factor) is included to aid the channel estimation. Additionally, the influence of interference due to speech users is taken into account.

In high data rate CDMA systems spreading factors are typically rather small. Hence, the multipath propagation causes severe interference. The accuracy of channel estimation is crucial in such systems. The semi-blind estimation techniques introduced here exploits in a novel way both the pilot signal and knowledge of the HSDPA spreading codes. It is an extension of the blind multicode (MPC) channel estimator [5] to a semi-blind method which can benefit from all known codes regardless of the spreading factor. When the interference is high, the purely blind MPC method is not able to identify the channel. Consequently, it is not able to improve the pilot based estimate. A measure is introduced to evaluate the quality of the blind method. Moreover, a semi-blind combining parameter is defined based on the traces of the post- and pre-despreading covariance matrices. The proposed method yields improved results even with a few HSDPA codes compared to pure pilot-based method especially with small sample support. Therefore it is is an attractive solution to fast fading channels.

The rest of the paper is organized as follows. The HSDPA system model with multicode transmission is described first. In section 3, the blind MPC method is derived. The influence of different spreading factors in the known codes is considered in section 4. Also a novel method to combine the blind estimate with the pilot-based channel estimate is introduced. MMSE equalization is briefly presented in section 5. In Section 6, simulation results using proposed semi-blind channel estimator and a MMSE equalizer in different interference scenarios are presented. The performance gains obtained by the proposed semi-blind method over pilot-only based estimator are significant.

2. SYSTEM MODEL

The system model considered in this paper is based on the 3rd generation wideband CDMA model for high data rates, see [1, 2]. Due to aperiodic (long) codes, each symbol is spread with different code during the observation period. The pth code for nth symbol is denoted by $\mathbf{c}_{np} = [c_{np}(1), \quad \ldots, \quad c_{np}(G)]^T$ where G is the spreading factor. For the frequency selective channel case, we define a code convolution matrix as: $\mathbf{C}_{np} = \text{Toeplitz}\ ([\mathbf{c}_{np}^T, \ \mathbf{0}_{1,L-1}]^T, [c_{np}(1), \quad \mathbf{0}_{1,L-1}])$. This Toeplitz matrix has dimension $G + L - 1 \times L$. Here L is the length of the channel impulse response in chips. Assuming that there are M antennas at the receiver, we can model the code matrix as $\mathcal{C}_{np} = \mathbf{I}_M \otimes \mathbf{C}_{np}$, where $\otimes$ denotes Kronecker product.

The combined multipath multiple receive antenna channel impulse is defined as a $LM \times 1$ vector $\mathbf{h}_n = [\mathbf{h}_{n1}^T \quad \ldots \quad \mathbf{h}_{nM}^T]^T$, where the complex channel coefficients at the mth receive antenna are $\mathbf{h}_{nm}^T = [h_{nm}(1) \quad \ldots \quad h_{nm}(L)]^T$. We are assuming that the channel is constant during the observation period. Therefore the symbol index n may be dropped.

The received signal is sampled at chip rate and stacked to a vector of length $M(G + L - 1)$. The received signal due to P multicode signals transmitted from one base station may now be written as:

$$\mathbf{y}(n) = \sum_{p=1}^{P} \rho_p \, \mathcal{C}_{np} \mathbf{h} \, s_p(n) + \text{ISI} + \mathbf{v}(n), \qquad (1)$$

where ρ_p^2 is the transmitted power for the pth code and $\mathbf{v}(n)$ is noise term. The transmitted symbols, $s_p(n)$, are assumed to be independent and identically distributed, such that $E\{s_p(n)s_k(a)^*\} = \delta(p-k)\delta(n-a)$, where $\delta(n)$ is the Dirac delta function. The inter-symbol interference (ISI) term is defined similarly as in [5]. When different spreading factors are used, the smallest G is used as a common spreading factor. The symbols spread with longer spreading codes can be considered as N consecutive symbols with the common spreading factor. For example, one pilot symbol corresponds to 16 HSDPA symbols, since spreading factor for pilot signal is $G_c = 256$ and for HSDPA symbols $G = 16$. Codes are scaled such that $|\mathbf{c}_p^H \mathbf{c}_p| = 1$. Additionally due to orthogonality of the codes $\mathbf{c}_p^H \mathbf{c}_r = 0$, if $r \neq p$. This holds regardless of the spreading factors if the shorter code p is repeated G_r/G_p times, i.e. $[\mathbf{c}_p;\ \mathbf{c}_p;\ \ldots]^H \mathbf{c}_r = 0$

3. BLIND CHANNEL ESTIMATION

In this section we briefly present the blind MPC method and define a quality measure of the blind estimate with different number of known codes and unknown interfering users.

The despread received symbol can be expressed as follows:

$$\hat{\mathbf{x}}_p(n) = \mathcal{C}_{np}^H \mathbf{y}(n) = \mathcal{C}_{np}^H \mathcal{C}_{np} \mathbf{h}\ \rho_p s_p(n) + \mathcal{C}_{np}^H \tilde{\mathbf{v}}(n), \qquad (2)$$

where $\tilde{\mathbf{v}}(n) = \mathbf{v}(n) + \mathrm{ISI} + \sum_{r \neq p}^{P} \mathcal{C}_{n,r} \mathbf{h}\ \rho_k s_r(n)$. A blind channel estimate, $\hat{\mathbf{h}}_b$, up to complex phase ambiguity can be obtained as the principal eigenvector of a difference matrix, see [4, 5]:

$$\hat{\mathbf{R}}_\triangle = \hat{\mathbf{R}}_X - \beta \hat{\mathbf{R}}_Y \approx \gamma \mathbf{h}\ \mathbf{h}^H, \qquad (3)$$

where $\hat{\mathbf{R}}_X$ and $\hat{\mathbf{R}}_Y$ are the signal covariance matrix estimates before and after despreading, β is a scaling factor and γ denotes a measure of the quality of the blind estimate. The covariance matrix before despreading can be estimated over G chips, by $\hat{\mathbf{R}}_Y = \frac{1}{G} \sum_{n=1}^{G} \mathbf{y}(n)\mathbf{y}^H(n)$. The post-despreading covariance matrix is estimated as $\hat{\mathbf{R}}_{x_p} = \frac{1}{N} \sum_{n=1}^{N} \hat{\mathbf{x}}_\mathbf{p}(n) \hat{\mathbf{x}}_\mathbf{p}^H(n)$. This estimate is averaged over N symbols. Assuming that all P transmitted codes have equal spreading factor G and they are all known, we can estimate $\hat{\mathbf{R}}_X$ as a sum over all codes p, i.e. $\sum_p \hat{\mathbf{R}}_{x_p}$. Further, if the signal powers are equal, the covariance matrix is $\hat{\mathbf{R}}_X \approx P\ \hat{\mathbf{R}}_{x_p}$. In this case, the scaling is simply $\beta = P$.

To evaluate γ we write the difference matrix $\hat{\mathbf{R}}_\triangle$ in more general case with interference present due to unknown codes. Let us assume that the received signal is a composition of P known codes with spreading factor G_p and K unknown codes with spreading factor G_k and one control signal with parameters ρ_c^2 and G_c. For simplicity, we assume that the P known codes have equal powers ρ_p^2 and the K unknown codes have equal powers ρ_k^2. The extension to more complicated system structures is straight forward.

With these definitions and neglecting the time varying terms [1] and noise, we can write the following approximations for the first elements of the covariance matrices :

$$\hat{\mathbf{R}}_Y(1,1) \approx B(|h_1|^2 + |h_2|^2 + |h_3|^2 + \ldots) = B|\mathbf{h}|^2$$
$$\hat{\mathbf{R}}_X(1,1) \approx C|h_1|^2 + A(|h_2|^2 + |h_3|^2 + \ldots),$$

where $h_1 = \mathbf{h}(1)$ is the first channel tap etc. and

$$B = \rho_c^2 \tfrac{1}{G_c} + \rho_p^2 \tfrac{P}{G_p} + \rho_k^2 \tfrac{K}{G_k}$$
$$C = P\rho_p^2$$
$$A = PB.$$

[1] That is temporal variation due to aperiodic codes.

Similar derivation may be done for all the elements of the covariance matrices. From equation (3), we get $\gamma = C - A$, since when the known codes have equal spreading factors clearly $\beta = P$. In the next section we'll extend the MPC method to unequal spreading factors. Note that B is the power of the transmitted chip sequence.

In Figure 1 we have plotted γ, which gives a measure for the quality of the blind MPC method. We assume knowledge of P codes with $G_p = 16$, and vary the number of interfering codes and pilot power. For example, if we know 8 HSDPA codes and have $K = 40$ interfering speech users, γ for the HSDPA codes is 1.16 with 5% pilot power and 0.78 with 10% pilot power. The values for $P = 8$ are marked in the Figure with '+'. A large value of γ indicates robustness against the time variations caused by aperiodic spreading codes. When γ is small (positive or negative) the blind MPC is not able to identify the channel, since the diagonal elements of the difference matrix given in equation (3) can also have negative values. In this case the negative values have similar magnitudes as the positive values and are due to time varying the terms.

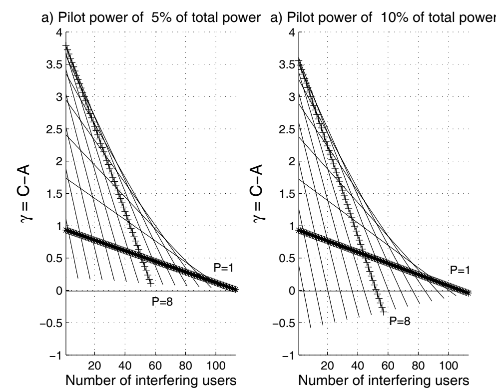

Fig. 1. *The quality measure, $\gamma = C - A$ of the blind MPC method against number of interference users with P known codes ($G_p = 16$). One control signal with $G_c = 256$ and different number of speech users, $G_k = 128$ are considered as interference. The control signal power was set to 5% in figure a) and 10% in figure b) for each scenario and $\rho_p^2 = \rho_k^2 = 1$ for all the other signals.*

The quality γ depends on the signal powers, system load and number of known and unknown codes. The mobile terminal does not necessary have knowledge of all these transmit parameters needed to evaluate γ. An simple estimate, which is proportional to γ and channel energy is obtained by

$$\tau = \mathrm{trace}(\hat{\mathbf{R}}_\triangle) \approx \gamma |\mathbf{h}|^2. \qquad (4)$$

Even in a case when we don't know $|\mathbf{h}|^2$ we can use τ to find the cases when $\gamma < 0$. Additionally, this measure can be used when multiple codes with different spreading factors are combined to form a blind MPC estimate. This extension will be considered in the next section.

4. SEMI-BLIND CHANNEL ESTIMATION

In WCDMA downlink systems a strong pilot signal is transmitted to enable channel estimation. Additionally, if the value of γ for the other known codes is large enough, we can improve the pilot-based channel estimate with the blind MPC method. In this section we will propose three different semi-blind methods to combine the known HSDPA codes with the pilot signal to provide more accurate channel estimation. First the blind MPC method is extended to benefit from codes with different spreading factors. With this method no extra knowledge of system parameters is needed. Next, we combine the pilot-based channel estimate with the blind MPC. The combination is done in a similar fashion as in the first method. However, some estimate of the pilot signal power is needed for proper scaling of the combination. The third method is conventional combining of two estimates. Now the scaling parameters can be determined similarly as in the first two cases, but additionally some threshold parameter is needed to limit the influence of the blind estimate in cases when γ is small.

The blind MPC method was originally derived assuming all the known codes have the same spreading factor. Yet, for example in WCDMA downlink, the control signals and data signals can have different spreading factors. In this case the post-despreading covariance matrices are calculated with different sample supports. This needs to be taken into account when calculating the difference matrix in equation (3). With equal spreading factors we could simply estimate $\hat{\mathbf{R}}_X = \sum_p \hat{\mathbf{R}}_{x_p}$. However if the spreading factors and signal powers differ, direct combining will not yield very good results in general. Using combining weights for each spreading factor, we can rewrite equation (3) as :

$$\hat{\mathbf{R}}_\Delta = \alpha_h \hat{\mathbf{R}}_{\Delta_h} + \alpha_c \hat{\mathbf{R}}_{\Delta_c} \approx \gamma \mathbf{h}\,\mathbf{h}^H. \tag{5}$$

For simplicity this is written only for two different spreading factors, and index h corresponds to one spreading factor and c to an other. Now β_h and β_c [2] are simply the number of known codes with each spreading factor and $\gamma = \alpha_h\gamma_h + \alpha_c\gamma_c$. One natural way to define the combing weights α_h and α_c is to employ the power of transmitted chips used to estimate the corresponding post-despreading covariance matrix. If these powers are not known, we can estimate their ratio as follows. Notice that $\mathrm{trace}(\hat{\mathbf{R}}_X) \approx C|\mathbf{h}|^2 + (LM-1)A|\mathbf{h}|^2$ and $\mathrm{trace}(\hat{\mathbf{R}}_Y) \approx (LM)B|\mathbf{h}|^2$, which is actually LM times the total received power. With P known codes (with spreading factor G_p) we can write

$$\mu_h \approx \frac{\mathrm{trace}(\hat{\mathbf{R}}_X)}{\mathrm{trace}(\hat{\mathbf{R}}_Y)} \approx \frac{LMBP + P\rho_p^2 - PB}{LMB}, \tag{6}$$

and solve for the transmitted chip power, $P\rho_p^2$,

$$P\rho_p^2 \approx B(\mu_h LM - PLM + P).$$

Similar derivation can be used to calculate the power of the other known codes with different spreading factor. Assuming now for simplicity one control signal, we can estimate the ratio as follows:

$$\alpha = \frac{\alpha_h}{\alpha_c} = \frac{P\rho_p^2}{G_p}\frac{G_c}{\rho_c^2} \approx \frac{ML\mu_h - PML + P}{ML\mu_c - ML + 1}\frac{G_c}{G_p}. \tag{7}$$

The extension to the case with more than two different spreading factors is straight-forward. For example, in a HSDPA system all the control signals could be included to aid the estimation. As seen in Figure 1, it might not be beneficial to use the known codes if γ is small. From equation (4) we estimate $\gamma_h \approx \frac{\tau_h}{|\mathbf{h}|^2}$. If this is a negative number then the corresponding $\hat{\mathbf{R}}_{\Delta_h}$ in equation (5) can not improve the blind estimate. Additionally, if the ratio α is very small then the matrix $\hat{\mathbf{R}}_{\Delta_h}$ does not influence the estimation and can be ignored. Similarly if the ratio α is very large then $\hat{\mathbf{R}}_{\Delta_c}$ does not provide additional gain to the blind estimate.

[2] Consequently $\beta = \alpha_h\beta_h + \alpha_c\beta_c$.

The purely blind MPC method does not take into account the knowledge of the pilot symbols, except in the removal of the phase ambiguity. The second and third semi-blind approaches considered next in this paper combine the blind channel estimate with a pilot symbol based estimate, denoted as $\hat{\mathbf{h}}_p$. In the simulations we have used a conventional pilot-based channel estimate. It can be obtained by: $\hat{\mathbf{h}}_p = \frac{1}{N}\sum_{n=1}^{N}\{\mathcal{C}_{nc}^H \mathbf{y}(n) s_c^*(n)\}/\rho_c^2$, since the transmitted pilot symbols $s_c(n)$ are known. The index c denotes the control channel and N is the number of symbols used in the averaging.

Now we can write the equation (5), where a blind combination for the difference matrices with two spreading factors was defined, in a semi-blind fashion:

$$\hat{\mathbf{R}}_\Delta = \alpha_h \hat{\mathbf{R}}_{\Delta_h} + \alpha_c \rho_c^2 \hat{\mathbf{h}}_p \hat{\mathbf{h}}_p^H \approx \gamma \mathbf{h}\,\mathbf{h}^H. \tag{8}$$

Note that using the blind MPC in equation (3) with the knowledge of the pilot code only, a performance rather close to the conventional pilot based estimation is achieved.

Simulation in section 6 will show only very minor differences in the BER performance between the methods given in equations (5) and (8). Using (5) makes it easier to benefit from other control signals with spreading factors equal to the pilot signal. On the other hand, using (8) makes it easier to use some improved channel estimate instead of conventional pilot-based method.

A third alternative is to use a conventional combining of a pilot-based and blind channel estimates, $\hat{\mathbf{h}}_p$ and $\hat{\mathbf{h}}_b$, with some weighting

$$\hat{\mathbf{h}}_{sb} = \alpha_p \hat{\mathbf{h}}_p + \alpha_b \hat{\mathbf{h}}_b. \tag{9}$$

Before combining both the estimates are scaled to have unit norm, since by default the blind method gives unit length estimates. It is not straight forward to find the optimal scaling parameters α_p and α_b. The quality of the estimates should be taken into account somehow. One promising solution is to define the ratio between the scaling parameters based on the ratio of the chip powers, similarly to equations (5) and (8). This gives $\frac{\alpha_b}{\alpha_p} = \sqrt{\frac{\alpha}{\rho_c^2}}$, where α is as in equation (7). The BER performance of (**??**) is quite similar to methods in equations (5) and (8) in most cases. Yet, since this scaling does not depend on the system load there is some performance degradation experienced with small values of γ, i.e. when the blind estimate has poor quality. Consequently, some additional threshold value should then be set to define when this conventional semi-blind method is beneficial.

In section 6 we have compared in simulations these three semi-blind methods to conventional pilot-based estimation. The phase ambiguity in the blind estimate is resolved using the phase of the strongest tap of the channel estimate obtained using the pilot only as follows: $\hat{\mathbf{h}}_b = (e^{i(\phi_p - \phi_e)}\hat{\mathbf{h}}_m)$, where ϕ_p is the pilot based phase estimate and ϕ_e the corresponding phase in the principal eigenvector, $\hat{\mathbf{h}}_m$, of difference matrix given in equation (5).

5. MMSE EQUALIZATION

The MMSE equalizer, [3, 6], is defined as follows:

$$\mathbf{f} = (\sigma_d^2 \hat{\mathbf{H}}\hat{\mathbf{H}}^H + \sigma_v^2 \mathbf{I})^{-1}\sigma_d^2 \hat{\mathbf{h}}_D, \tag{10}$$

where $\hat{\mathbf{H}}$ is the channel convolution matrix estimate, size $MF \times (F + L - 1)$, and F is the filter length. $\hat{\mathbf{h}}_D$ is the D'th column of the matrix $\hat{\mathbf{H}}$, D is the delay and σ_v^2 is the noise power. σ_d^2 is the power of the transmitted signal. An alternative derivation of MMSE equalizer using the inverse of signal covariance matrix, $\mathbf{f} = \hat{\mathbf{R}}_{\mathbf{Y}}^{-1}\sigma_d^2\hat{\mathbf{h}}$ requires longer sample support to converge to the optimal solution, see [7]. In general the MMSE equalizer is time varying due to aperiodic spreading codes. But in WCDMA type systems with multiple transmitted codes the time dependency of optimal equalizer is reduced due to code averaging, see [3]. In case all codes are used, with equal spreading factors and equal powers, the time varying terms will vanish.

6. SIMULATIONS

In this section, the performance of the proposed blind and semi-blind channel estimation method (MPC) for SIMO channels is compared to pilot-based method. The used performance measure is the bit error rate (BER) obtained with MMSE equalizer in equation (10).

The channel coefficients are generated using METRA channel model software based on Vehicular ITUA specifications, see [8, 9]. The average power of the channel taps are defined as $[0\ -1\ -9\ -10\ -15\ -20]$ dB with delay spread of 11 chips. We have combined the channel with transmit and receiver filters, i.e. with two root raised cosine filters with roll off factor 0.22. After filtering the sampling is performed once per chip and 11 chip long channels are used. The equalizer length was $F = 32$. The delay was set to $D = \lfloor (F + L - 1)/2 \rfloor = 21$, see [6]. Two antennas are employed both at receiver assuming spacing of $d = \lambda$. [3] All the shown results are averaged over 200 independent channel realizations.

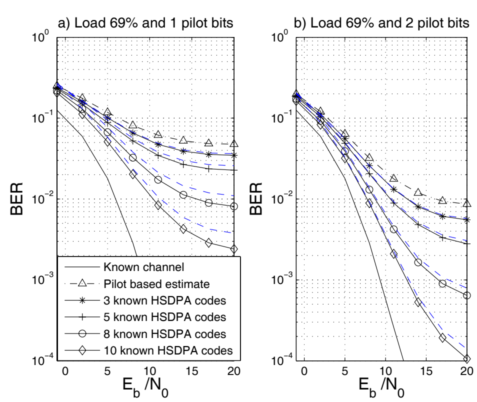

Fig. 2. *Raw BER for $M = 2$, and 69% of the OVSF codes used, e.g. $P = 5$, $K = 48$ and one pilot. The performance of the MMSE equalizer obtained with known channel coefficients, pure pilot based channel estimate and proposed semi-blind method. The pilot power was 10 %, the channel estimates are obtained with one and two pilot bits, see figure a) and b). The dashed lines are estimated signal powers, see equation (7).*

The spreading code is a combination of orthogonal variable spreading factor (OVSF) codes and long Gold codes (complex). Spreading factor of 16 is used for the HSDPA codes. One control signal is simulated with $G_c = 256$ and using spreading code of all ones [2, 1]. Interfering users are considered as speech signals with spreading factor $G_k = 128$. Number of HSDPA signals is denoted with P and number of speech signals as K. All signals are QPSK modulated.

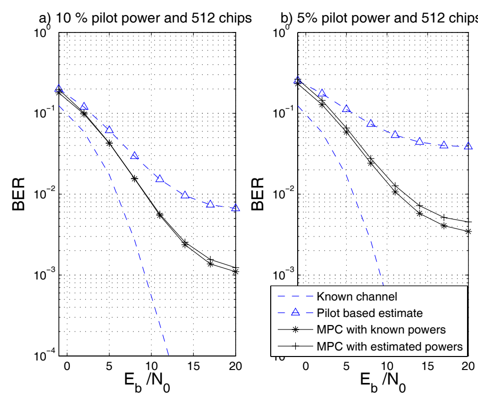

Fig. 3. *Raw BER for $M = 2$, and $P = 5$ and $K = 40$. The performance of the MMSE equalizer obtained with known channel coefficients, pure pilot based channel estimate and proposed semi-blind method. The pilot power was 10% and 5%, see figure a) and b) and the channel estimates are obtained with two pilot bits.*

The influence of number of known codes is shown in Figure 2. We assumed 69% of all the OVSF codes were used, and the number of HSDPA codes and number of interfering codes were varied. The pilot power was set to 10% at each simulation, while the other signals had unit power, i.e. $\rho_p^2 = \rho_k^2 = 1$. The gain obtained with semi-blind estimate compared to pure pilot-based estimate is clear when majority of the codes are know (8 or 10 known codes). With sample support of two pilot bits (512 chips) the gain at BER 10^{-2} is about 3 dB when 3 HSDPA codes are used, while with 8 codes the gain is about 7 dB. The semi-blind estimates are calculated assuming perfect knowledge of the transmitted powers ρ_c^2 and ρ_p^2. The dashed lines close to semi-blind estimates are obtained by estimating α using equation (7). The results are shown only for blind MPC method of equation (5). BER curves with the semi-blind MPC version (8) are similar.

In Figure 3 BER values with two different pilot power specification are shown. The number of known HSDPA codes was $P = 5$, and $K = 40$ speech users were causing the intra-cell interference. All signals except pilot signal have unit power, i.e. $\rho_p^2 = \rho_k^2 = 1$. The performance with the semi-blind channel estimate is reduced only about 2dB at BER 10^{-2} when the pilot signal power is lowered from 10% to 5% of the total transmitted power. Meanwhile the pure pilot-based method has a clear degradation in the performance. The used sample support was 512 chips, i.e. two pilot symbols.

As expected, the power of the known code has a clear impact on the performance of the semi-blind method. In Figure 4 is shown the improvement of increasing the signal power of the HSDPA signal from $\rho_p^2 = 1$ to $\rho_p^2 = 1.5$. The improvement in the BER rate performance is clearly seen. The pilot signal power was in both cases 10% of the total power and the $K = 64$ interfering users had unit power.

Finally, in Figure 5 the different semi-blind methods of equa-

[3] λ is the wavelength.

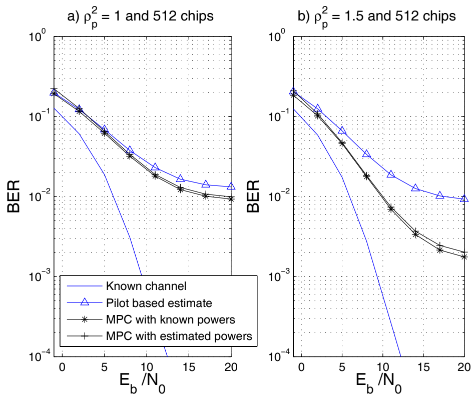

Fig. 4. *Raw BER for $M = 2$, and $P = 5$ and $K = 64$. The performance of the MMSE equalizer obtained with known channel coefficients, pure pilot based channel estimate and proposed semi-blind method. The power of HSDPA signals was $\rho_p^2 = 1$ and $\rho_p^2 = 1.5$, see figure a) and b). The channel estimates are obtained with two pilot bits.*

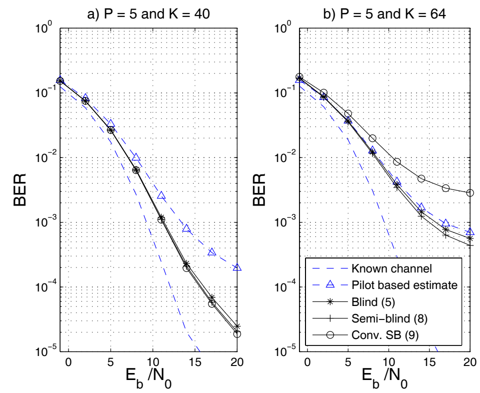

Fig. 5. *Raw BER for $M = 2$, and $P = 5$ and sample support of 5 pilot bits. The performance of the MMSE equalizer obtained with known channel coefficients, pure pilot based channel estimate and proposed semi-blind method. The power of pilot signal was 10% and of all other signals was $\rho_p^2 = 1$. In figure a) $K = 40$ and b) $K = 64$ interfering users were present.*

tions (5), (8), and (9) are compared. The channel estimates are obtained averaging over five pilot symbols. In Figure 5a) all three estimates give equal performance. When the interference is increased, see Figure 5b), the conventional semi-blind combining given in (9) has weaker performance, while the two other methods have similar performance to the pure pilot-based estimator. In these cases the quality measure γ of blind MPC for HSDPA signals are 1.53 and 0.49, respectively.

As a conclusion, it can be said, that the proposed semi-blind methods, see equations (5) and (8) provide clear performance gain over the conventional pilot-based estimation. These methods adjust naturally to the quality of the blind channel estimate, and in the worst case the performance is similar to pure pilot-based estimate. The conventional semi-blind method, given in equation (9), requires a threshold parameter to avoid performance loss when quality measure γ is small. Consequently, either blind MPC of equation (5) or the semi-blind method of equation (8) should be used. Both these methods adjust naturally to γ and provide therefore a simple combining method.

7. CONCLUSIONS

In this paper we introduced a semi-blind channel estimator for HSDPA systems. The proposed multicode principal component method performs better than purely pilot-based estimator, when the pilot signal is combined with other known codes in the estimate. Different spreading factors and transmitted signal powers are taken into account in the semi-blind combining. The combining weights are calculated based on the traces of the post- and pre-despreading covariance matrices. Additionally, an estimate of a quality measure for the blind method is given. Consequently, similar quality channel estimates can be obtained with smaller sample support or reduced pilot power. In this paper we used a block type approach, but alternatively both the pilot channel estimate and covariance matrix estimates could be computed adaptively, e.g with RLS type update. This is left for further study.

Acknowledgements

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