Joint Common-Dedicated Pilots Based Estimation of Time-Varying Channels for W-CDMA Receivers

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Abstract—We address the problem of joint common-dedicated pilots downlink user dedicated channel estimation for W-CDMA receivers, in particular in the presence of dedicated channel transmit beamforming. The optimal dedicated channel estimate is derived via a three steps procedure, consisting of building brute FIR dedicated and common channel estimates on a per-slot basis exploiting the dedicated and common pilots respectively, building a refined unbiased minimum mean square error dedicated channel estimate by optimally combining the previous dedicated and common channels estimates, further refining the obtained dedicated channel estimate by Wiener filtering across slots. Adaptive filtering implementation is addressed for the implementation of the optimal Wiener filtering. The proposed channel estimation technique is suited for structured multipath channels where the FIR channel estimates resulting from the previous three steps procedure are optimally approximated by a multipath model in every slot.

I. INTRODUCTION

We address the problem of downlink user dedicated channel estimation for W-CDMA receivers in the presence of dedicated channel transmit beamforming. The UMTS standard [9] provides in the downlink a user dedicated physical channel (DPCH) consisting of a dedicated physical control channel (DPCCH) time multiplexed with the dedicated physical data channel (DPDCH), consisting of dedicated pilot and data symbols respectively, and a common pilot channel (CPICH) that continuously transmits pilot symbols. Both DPCCH and CPICH pilots can be used for channel estimation purposes. Mostly channel estimation techniques for W-CDMA assume a discrete multipath wide-sense stationary uncorrelated-scattering (WSS-US) sparse channel model where each multipath component is characterized by a time-varying complex coefficient and a delay [6]. Sparse structured channel estimation techniques have been proposed for W-CDMA receivers generally based either on the dedicated pilots (see e.g. [1], [2] and references therein) or on the common pilots (see e.g. [10]). Both approaches are inherently sub-optimal as they do not exploit the whole a priori information available at the receiver. On the one hand the channel estimation accuracy when relying only on the DPCCH is limited by the reduced number of dedicated pilots per slot. Moreover the lack of

known dedicated pilots during the DPDCH period prevents from directly tracking the channel variations in the presence of fast fading. To overcome this problem several authors suggested the use of interpolation [5], [7], or Wiener filtering, e.g. [1]–[3], between consecutive channel estimates. On the other hand the classical channel estimation approach based only the CPICH common pilots can be easily implemented in order to track fast fading variations, although it neglects the a-priori information available from the DPCCH and the common structure shared by the the common and dedicated channels. Only a few authors [4], [5], [8] have proposed solutions for path-wise dedicated channel estimation aiming at exploiting both dedicated and common pilots, assuming a perfect a-priori knowledge of the paths' delays and identical dedicated and common channels though. Moreover, those methods cannot support channel estimation in the presence of dedicated transmit beamforming. The UMTS standard already envisages the use of transmit beamforming for the dedicated channel at the base station to improve the system capacity, but not for the CPICH which is normally broadcasted to all intra-cell (or intra-sector) users. Thus, as seen from the mobile terminal receiver, the channel associated with the DPCH is in general different from the one associated with the CPICH. A channel estimation technique relying only on the dedicated pilots would work also in that case, although with all the previously described limitations. The problem of building a dedicated channel estimate by jointly exploiting the common and dedicated pilots would generally require the knowledge of the transmit beamforming parameters (i.e. the beamforming weight vector, antenna array responses corresponding to the excited angles, and their related statistics). Even in the absence of transmit beamforming the offset between the transmit powers assigned to the DPCCH and CPICH is an unknown parameter which needs to be estimated for the mobile terminal receiver to jointly exploit the common and dedicated pilot chips information. In this paper we solve the problem of time-varying dedicated channel estimation by optimally exploiting all known sources of information, i.e. by relying on both common and dedicated pilots, assuming the paths' delays as well as the beamforming parameters unknown.

II. CHANNEL AND SIGNAL MODEL

The channel is assumed to follow the WSS-US [6]. The time-varying channel impulse response in the presence of

transmit beamforming through a Q antennas array, and a single antenna mobile receiver, can be represented according to a spatio-temporal multipath-cluster model (see e.g. [11] and references therein). Multipaths belonging to the same cluster p are assumed to have approximately the same delay τ_p, the same nominal propagation angle θ_p, and the same nominal angular spread σ_{ϕ_p}. Each multipath component is further characterized by a time-varying complex channel coefficient $\xi_{j_p}(t)$, modeling the scattering near the mobile terminal and by an angle deviation $\phi_{j_p}(t)$ from the nominal angle θ_p. Let $\boldsymbol{a}(\theta)$ and M_p denote the antenna array response vector in the direction θ and the number of excited directions around θ_p respectively. Then an *effective* array response vector can be defined for each cluster p as $\boldsymbol{a}_p(t) = \sum_{j_p=1}^{M_p} \boldsymbol{a}(\theta_p + \phi_{j_p}(t))\xi_{j_p}(t)$. Given the transmit beamforming vector $\boldsymbol{w}$ and defining $c_{d,p}(t) = \boldsymbol{w}^H \boldsymbol{a}_p(t)$ the dedicated channel impulse response can be written as follows

$$h_d(t,\tau) = \sum_{p=0}^{P-1} c_{d,p}(t)\psi(\tau - \tau_p) \qquad (1)$$

where P denotes the number of significant path clusters, or, more simply, significant *paths*, $\psi(\tau)$ represents the pulse-shape filter and $(\cdot)^H$ denotes Hermitian transpose. In the above channel model we implicitly distinguished the *fast varying* parameters, $\xi_{j_p}(t)$, $\phi_{j_p}(t)$, and $c_{d,p}(t)$, from the *slow varying* parameters θ_p, σ_{ϕ_p}, and τ_p which are considered approximately constant over the observation time-window.

In the absence of transmit beamforming, i.e. when the signal is transmitted through an omni-directional antenna, the previous channel model (1) reduces to

$$h_c(t,\tau) = \sum_{p=0}^{P-1} c_{c,p}(t)\psi(\tau - \tau_p) \qquad (2)$$

where $c_{c,p}(t)$ and $c_{d,p}(t)$ are generally correlated. The impulse responses h_c and h_d expressed by (2) and (1) are referred to as the common and dedicated channel respectively.

The receiver is assumed to sample M times per chip the low-pass filtered received signal. Stacking the M samples per chip period in vectors, the discrete-time representation of both common and dedicated channel at chip rate takes the form $\boldsymbol{h}_l = [h_{1,l} \dots h_{M,l}]^T$, which represents the vector of the samples of the overall channel, including pulse shape, propagation channel and receive filters, where the superscript $(\cdot)^T$ denotes transpose. Assuming the overall channel to have a delay spread of N chip periods the dedicated and common channel impulse responses take the form $\boldsymbol{h}(n) = \boldsymbol{\Psi}\boldsymbol{c}(n)$ where $\boldsymbol{h} = [\boldsymbol{h}_1^T \dots \boldsymbol{h}_N^T]^T \in \mathcal{C}^{MN \times 1}$, $\boldsymbol{c}(n) = [c_1(n) \dots c_P(n)]^T \in \mathcal{C}^{P \times 1}$ are the complex path amplitudes and the temporal index n relates to the time instant at which the time-varying channel is observed. The assumption of fixed delays τ_p's over the observation window, yields to a constant pulse-shape convolution matrix $\boldsymbol{\Psi} \in \mathcal{R}^{MN \times P}$ given by

$$\boldsymbol{\Psi} = \boldsymbol{\Psi}(\tau_1, \cdots, \tau_P) = [\boldsymbol{\psi}(\tau_1) \dots \boldsymbol{\psi}(\tau_P)]$$

where $\boldsymbol{\psi}(\tau_p)$ represents the sampled version of the pulse shape filter impulse response delayed by τ_p. The evolution of the

complex path amplitudes variations is modeled as an autoregressive (AR) process of order sufficiently high to characterize the Doppler spectrum. Matching only the channel bandwidth with the Doppler spread leads to a first-order AR(1) model of the form $\boldsymbol{c}(n) = \rho\,\boldsymbol{c}(n-1) + \sqrt{1-\rho^2}\Delta\boldsymbol{c}(n) = \frac{\sqrt{1-\rho^2}}{1-\rho q^{-1}}\Delta\boldsymbol{c}(n)$ so that, $\boldsymbol{\Psi}$ being constant, we obtain

$$\begin{aligned} \boldsymbol{h}(n) &= \rho\boldsymbol{h}(n-1) + \sqrt{1-\rho^2}\Delta\boldsymbol{h}(n) \\ &= \frac{\sqrt{1-\rho^2}}{1-\rho q^{-1}}\Delta\boldsymbol{h}(n) \end{aligned} \qquad (3)$$

where q^{-1} denotes the delay operator such that $q^{-1}y(n) = y(n-1)$ and ρ represents the AR process forgetting factor. Since the Doppler spread is the same the model (3) applies to both $\boldsymbol{h}_d(n)$ and $\boldsymbol{h}_c(n)$. The variance of k-th component $h_{c,k}(n)$ of $\boldsymbol{h}_c(n)$ is $\sigma^2_{h_{c,k}} = \sigma^2_{\Delta h_{c,k}} = \boldsymbol{\psi}_k \boldsymbol{D}_c \boldsymbol{\psi}_k^H$ where $\boldsymbol{\psi}_k$ denotes the k-th line of $\boldsymbol{\Psi}$ and $\boldsymbol{D}_c = \text{diag}(\sigma^2_{\Delta c_1}, \dots, \sigma^2_{\Delta c_{c,P}})$. Notice that $\sigma^2_{c_{c,p}} = \sigma^2_{\Delta c_{c,p}}$. Similarly the variance of k-th component $h_{d,k}(n)$ of $\boldsymbol{h}_d(n)$, is $\sigma^2_{h_{d,k}} = \sigma^2_{\Delta h_{d,k}} = \boldsymbol{\psi}_k \boldsymbol{D}_d \boldsymbol{\psi}_k^H$ where $\boldsymbol{D}_d = \text{diag}(\sigma^2_{\Delta c_{d,1}}, \dots, \sigma^2_{\Delta c_{d,P}})$.

III. JOINT COMMON-DEDICATED PILOTS BASED DEDICATED CHANNEL ESTIMATION

In the sequel we shall derive a joint dedicated-common pilots based dedicated channel estimation, suited to operate in the presence of dedicated channel transmit beamforming, assuming that neither the path delays τ_p's, nor the beamforming parameters are known. The proposed approach consists of a three steps procedure where firstly slot-wise LS FIR dedicated and common channel estimates $\hat{\boldsymbol{h}}_c(n)$ and $\hat{\boldsymbol{h}}_d(n)$ are computed based on the a priori knowledge of the common and dedicated pilot chips. Then for each k-th element of $\hat{\boldsymbol{h}}_c(n)$ and $\hat{\boldsymbol{h}}_d(n)$ with $k = 0, ..., MN - 1$ a refined estimate $\hat{\hat{h}}_{d,k}(n)$ of $h_{d,k}(n)$ is built by optimally combining the corresponding LS estimates $\hat{h}_{c,k}(n)$ and $\hat{h}_{d,k}(n)$. Then successive estimates $\hat{\hat{h}}_{d,k}(n)$ of $h_{d,k}(n)$ are temporally filtered in order to exploit the temporal correlation due to the finite Doppler spread generating an improved estimate $\tilde{h}_{d,k}(n)$. As a fourth step the refined FIR estimate $\tilde{\boldsymbol{h}}_d(n) = [\tilde{h}_{d,0}(n) \dots \tilde{h}_{d,MN-1}(n)]^T$ can be used to build a path-wise channel estimate by finding the delays $\boldsymbol{\tau} = [\tau_1 \dots \tau_P]^T$ and paths coefficients $\boldsymbol{c}_d = [c_{d,1} \dots c_{d,P}]^T$ solving the fitting problem

$$\min_{\boldsymbol{\tau}, \boldsymbol{c}_d} \|\tilde{\boldsymbol{h}}(n) - \boldsymbol{\Psi}\boldsymbol{c}_d(n)\|^2 \qquad (4)$$

e.g. via the recursive early-late algorithm [2]. Since in this case the delay estimation problem is considered as an instantaneous channel estimation problem, the delay estimation has to come as the last step of the whole channel estimation process.

Alternatively a more *classical* strategy can be pursued, where one would first *sparsify* the channel impulse response under the assumption of resolvable discrete-multipath component, by estimating the paths' delays τ_p's according to a long term averaged power delay profile. Once the delays are known, namely the matrix $\boldsymbol{\Psi}$ is known, the same previous three steps procedure can be applied in order to estimate $\boldsymbol{c}_d(n)$. Due to

lack of space we shall not detail this second approach in this paper.

A. Least Square Dedicated and Common Channel Estimation

We assume dedicated pilot chips are sent in every user slot during transmission. Let us define $S_d(n) = \mathsf{S}_d(n) \otimes \mathbf{I}_M$, where $\otimes$ denotes the Kronecker product, as the block Hankel matrix comprising the dedicated pilot chip sequence intended for the user of interest in slot n. Similarly we refer to $S_c(n) = \mathsf{S}_c(n) \otimes \mathbf{I}_M$ as the block Hankel matrix containing the common pilot chip sequence in slot n. Let $Y(n)$ be the received signal samples vector corresponding to slot n. The LS unstructured FIR common and dedicated channel estimates FIR are given by

$$\hat{h}_d(n) = \arg\min_{h_d} \|Y(n) - S_d(n)h_d(n)\|^2$$
$$\hat{h}_c(n) = \arg\min_{h_c} \|Y(n) - S_c(n)h_c(n)\|^2 \qquad (5)$$

that, if the pilot chips can be models as i.i.d. random variables, yield to

$$\hat{h}_d(n) \approx \beta_d^{-1} S_d^H(n)Y(n)$$
$$\hat{h}_c(n) \approx \beta_c^{-1} S_c^H(n)Y(n) \qquad (6)$$

where β_d and β_c represent the dedicated and common pilot chip sequences total energies respectively. Let us model the received signal as $Y(n) = S_d(n)h_d(n) + S_c(n)h_c(n) + V(n)$, where $V(n) \sim \mathcal{N}_C(\mathbf{0}, \sigma_v^2 \mathbf{I})$ denotes interference-plus-noise. Then referring to $\hat{e}_{c,k}(n)$ and $\hat{e}_{d,k}(n)$ as the common and dedicated channel LS estimation errors, by assuming the dedicated and common pilot chips to be uncorrelated and invoking the whiteness of the pilot chip sequences, the errors covariance matrices can be approximated as follows

$$C_{\hat{e}_d \hat{e}_d} \approx \beta_d^{-1} \sigma_v^2 \mathbf{I}_{MN} \quad \text{and} \quad C_{\hat{e}_c \hat{e}_c} \approx \beta_c^{-1} \sigma_v^2 \mathbf{I}_{MN} \qquad (7)$$

while for the errors mutual covariance matrix we have $C_{\hat{e}_c \hat{e}_d} \approx \mathbf{0}$.

B. Optimal Combining of LS Channel Estimates

Let $\hat{h}_k(n) = [\hat{h}_{d,k}(n) \ \hat{h}_{c,k}(n)]^T$ denotes the vector of the LS estimates of the kth elements of the dedicated and common pilot channel FIR responses at slot n, i.e.

$$\hat{h}_k(n) = \begin{bmatrix} \hat{h}_{d,k}(n) \\ \hat{h}_{c,k}(n) \end{bmatrix} = \begin{bmatrix} h_{d,k}(n) \\ h_{c,k}(n) \end{bmatrix} + \begin{bmatrix} \hat{e}_{d,k}(n) \\ \hat{e}_{c,k}(n) \end{bmatrix} \qquad (8)$$

In the light of the models introduced in section II a general channel model in the presence of dedicated transmit beamforming leads to

$$h_{c,k}(n) = \alpha_k h_{d,k}(n) + x_{c,k}(n)$$

where $\alpha_k h_{d,k}(n)$, represents the short-term linear minimum mean square error (MMSE) estimate of $h_{c,k}(n)$ on the basis of $h_{d,k}(n)$ and $x_{c,k}(n)$ represents the associated estimation error. Then a refined estimate can be obtained as

$$\hat{\hat{h}}_{d,k}(n) = f_k \hat{h}_k(n) \qquad (9)$$

by optimal combining of common and dedicated LS channel estimates. In order not to introduce bias for the processing

in the next steps we shall determine f as the linear unbiased MMSE (UMMSE) filter, i.e. by solving for all k's the optimization problem

$$\min_{f_k} \mathrm{E}|h_{d,k}(n) - f_k \hat{h}_k(n)|^2 \quad \text{s.t.} \quad f_k[1 \ \alpha_k]^T = 1 \qquad (10)$$

To determine the filter f_k we derive the analytical expression of the covariance matrix $R_{\hat{h}_k \hat{h}_k} = \mathrm{E}\hat{h}_k(n)\hat{h}_k^H(n)$

$$R_{\hat{h}_k \hat{h}_k} = \begin{bmatrix} r_{11} & r_{12} \\ r_{21} & r_{22} \end{bmatrix} =$$
$$\sigma_{h_{d,k}}^2 \begin{bmatrix} 1 \\ \alpha_k \end{bmatrix} \begin{bmatrix} 1 \\ \alpha_k \end{bmatrix}^H + \begin{bmatrix} \sigma_{\hat{e}_{d,k}}^2 & 0 \\ 0 & \sigma_{\hat{e}_{c,k}}^2 + \sigma_{x_{c,k}}^2 \end{bmatrix} \qquad (11)$$

Having an estimate of the matrix $R_{\hat{h}_k \hat{h}_k}$, e.g. by temporal averaging, we can apply the *covariance matching* criterion so that $\sigma_{h_{d,k}}^2 = r_{11} - \sigma_{\hat{e}_{d,k}}^2$, $\alpha_k = r_{21}/(r_{11} - \sigma_{\hat{e}_{d,k}}^2)$, (i.e. α_k has the same phase as r_{21}), where the following bound $|\alpha_k| \leq \sigma_{h_{c,k}}/\sigma_{h_{d,k}} = \sqrt{(r_{22} - \sigma_{\hat{e}_{c,k}}^2)/(r_{11} - \sigma_{\hat{e}_{d,k}}^2)}$ can be used in actual estimation. Furthermore it results $\sigma_{x_{c,k}}^2 = r_{22} - \sigma_{\hat{e}_{c,k}}^2 - |r_{21}|^2/(r_{11} - \sigma_{\hat{e}_{d,k}}^2)$, so that the optimal UMMSE filter is given by

$$f_{k,\mathrm{UMMSE}} = ([1 \ \alpha_k^*]R_{\hat{h}_k \hat{h}_k}^{-1}[1 \ \alpha_k]^T)^{-1}[1 \ \alpha_k^*]R_{\hat{h}_k \hat{h}_k}^{-1}$$
$$= ([1 \ \alpha_k^*]R^{-1}[1 \ \alpha_k]^T)^{-1}[1 \ \alpha_k^*]R^{-1} \qquad (12)$$

where

$$R = \begin{bmatrix} \sigma_{\hat{e}_{d,k}}^2 & 0 \\ 0 & \sigma_{\hat{e}_{c,k}}^2 + \sigma_{x_{c,k}}^2 \end{bmatrix}$$

By denoting $\hat{\hat{e}}_{d,k}$ the estimation error after UMMSE combining its variance is readily given by

$$\sigma_{\hat{\hat{e}}_{d,k}}^2 = ([1 \ \alpha_k^*]R^{-1}[1 \ \alpha_k]^T)^{-1}$$
$$= ([1 \ \alpha_k^*]R_{\hat{h}_k \hat{h}_k}^{-1}[1 \ \alpha_k]^T)^{-1}[1 \ \alpha_k^*] - (r_{11} - \sigma_{\hat{e}_{d,k}}^2)$$
$$= \frac{\sigma_{\hat{e}_{d,k}}^2(\sigma_{\hat{e}_{c,k}}^2 + \sigma_{x_{c,k}}^2)}{\sigma_{\hat{e}_{d,k}}^2 |\alpha_k|^2 + \sigma_{\hat{e}_{c,k}}^2 + \sigma_{x_{c,k}}^2} \qquad (13)$$

In order to estimate the LS channel estimation error variances we observe that $\sigma_{\hat{e}_{d,k}}^2 = \sigma_{\hat{h}_{d,k}}^2$ and $\sigma_{\hat{e}_{c,k}}^2 = \sigma_{\hat{h}_{c,k}}^2$ for impulse response samples k at which $h_{d,k}(n) = h_{c,k}(n) \equiv 0$. Hence both $\sigma_{\hat{e}_{d,k}}^2$ and $\sigma_{\hat{e}_{c,k}}^2$ can be estimated from $\hat{h}_{d,k}$ and $\hat{h}_{c,k}$ respectively at delays k larger than the channel delay spread. To this end one may overestimate the delay spread, and exploit the tail of the channel estimate to obtain an unbiased estimate of $\sigma_{\hat{e}_{d,k}}^2$ and $\sigma_{\hat{e}_{c,k}}^2$ by long term temporal averaging for delays $k > MN$ at which $h_{d,k} = h_{c,k} \equiv 0$ can be assumed. Alternatively, in order not to increase the length of the channel impulse response to be estimated, we can estimate $\sigma_{\hat{e}_{d,k}}^2$ and $\sigma_{\hat{e}_{c,k}}^2$ from the $\hat{h}_{d,k}$ and $\hat{h}_{c,k}$ with smallest variance as detailed in [2].

1) Optimal Combining of LS Channel Estimates in the Absence of Beamforming: In the absence of dedicated transmit beamforming it results $\alpha_k = \alpha \in \mathcal{R}$, $\alpha \geq 0$, independent of k and $x_{c,k}(n) = 0 \ \forall k$. Then the optimal UMMSE filter f_k can

still be derived by accounting for the different structure of the matrix $\boldsymbol{R}_{\hat{h}_k\hat{h}_k}$, i.e.

$$\boldsymbol{R}_{\hat{h}_k\hat{h}_k} = \begin{bmatrix} 1 \\ \alpha \end{bmatrix} \begin{bmatrix} 1 \\ \alpha \end{bmatrix}^H + \begin{bmatrix} \sigma_{\hat{e}_{d,k}}^2 & 0 \\ 0 & \sigma_{\hat{e}_{c,k}}^2 \end{bmatrix} \quad (14)$$

from which α and $\sigma_{\hat{h}_{d,k}}^2$ can be estimated by approximated non-orthogonal joint-diagonalization, more specifically by joint LDU factorization in which the parameter α is in common, along the lines of [12]. The details are left out here for lack of space.

C. Optimal Wiener Filtering of Channel Estimates

Once the optimal combining of LS common and dedicated channel estimates has been performed via UMMSE filtering (12), optimal temporal causal Wiener filtering can be applied over successive estimates $\hat{\tilde{h}}_{d,k}(n)$ (9) in order to further refine the dedicated channel estimate. The dedicated channel estimate after UMMSE combining, $\hat{\tilde{h}}_{d,k}(n) = h_{d,k}(n) + \hat{e}_{d,k}(n)$, is such that the post-combining estimation error $\hat{e}_{d,k}(n)$ is mutually uncorrelated with $h_{d,k}(n)$, $\hat{e}_{d,k}(n)$ and $\hat{e}_{d,j}(n)$ are mutually uncorrelated for any $k \neq j$, and the variance of $\hat{e}_{d,k}(n)$ is independent of k while it depends on the mobile velocity (i.e. on the Doppler spread), on the channel power, and on the signal-to-interference-plus noise ratio (SINR). Then given $\hat{\tilde{h}}_d(n) = [\hat{\tilde{h}}_{d,0}(n) \ldots \hat{\tilde{h}}_{d,MN-1}(n)]^T$ the refined estimate $\tilde{h}(n)$ is of the form $\tilde{h}_d(n) = H(q)\hat{\tilde{h}}_d(n)$ where $H(q)$ represents the optimal Wiener filter (of unlimited order). For every k-th component of the channel estimate we can write $\tilde{h}_{d,k}(n) = H_k(q)\hat{\tilde{h}}_{d,k}(n)$ where

$$H_k(q) = \frac{1}{S_{\hat{\tilde{h}}_{d,k}\hat{\tilde{h}}_{d,k}}^+(q)} \left\{ \frac{S_{h_{d,k}h_{d,k}}(q)}{S_{\hat{\tilde{h}}_{d,k}\hat{\tilde{h}}_{d,k}}^-(q)} \right\}^+ \quad (15)$$

where $S_{xx}(q)$ denotes the power spectral density (PSD) of x, $\{\cdot\}^+$ and $\{\cdot\}^-$ denote the causal and anti-causal parts respectively, and $S_{xx}(q) = S_{xx}^+(q)S_{xx}^-(q)$ is the spectral factorization of $S_{xx}(q)$ in its causal minimum-phase factor and in its anti-causal maximum-phase counterpart. Assuming an AR(1) model for the channel coefficients as in (3), the PSD of $\hat{\tilde{h}}_{d,k}(n)$ is given by

$$\begin{aligned} S_{\hat{\tilde{h}}_{d,k}\hat{\tilde{h}}_{d,k}}(q) &= S_{h_{d,k}h_{d,k}}(q) + \sigma_{\hat{e}_{d,k}}^2 \\ &= \sigma_{\hat{e}_{d,k}}^2 \frac{a_k(1 - b_k q^{-1})(1 - b_k q)}{(1 - \rho q^{-1})(1 - \rho q)} \end{aligned}$$

where, by defining the channel estimation signal-to-noise ratio (SNR) $J_k = \sigma_{h_{d,k}}^2/\sigma_{\hat{e}_{d,k}}^2$ and $\bar{J}_k = 1 + \rho^2 + (1 - \rho^2)J_k$,

$$b_k = \frac{1}{2\rho}\left(\bar{J}_k - \sqrt{\bar{J}_k^2 - 4\rho^2}\right) \text{ and } a_k = \rho/b_k \quad (16)$$

Thus

$$S_{\hat{\tilde{h}}_{d,k}\hat{\tilde{h}}_{d,k}}^+(q) = \sigma_{\hat{e}_{d,k}} \sqrt{a_k} \frac{1 - b_k q^{-1}}{1 - \rho q^{-1}}$$

$$S_{\hat{\tilde{h}}_{d,k}\hat{\tilde{h}}_{d,k}}^-(q) = \sigma_{\hat{e}_{d,k}} \sqrt{a_k} \frac{1 - b_k q}{1 - \rho q}$$

so that

$$\left\{ \frac{S_{h_{d,k}h_{d,k}}(q)}{S_{\hat{\tilde{h}}_{d,k}\hat{\tilde{h}}_{d,k}}^-(q)} \right\}^+ = \left\{ \frac{\gamma_k q}{1 - b_k q} + \frac{\beta_k}{1 - \rho q^{-1}} \right\}^+ \quad (17)$$

$$= \frac{\beta_k}{1 - \rho q^{-1}}$$

where

$$\beta_k = \frac{(1 - \rho^2)\sigma_{h_{d,k}}^2}{\sigma_{\hat{e}_{d,k}} \sqrt{a_k}(1 - b_k \rho)} \quad \text{and} \quad \gamma_k = \beta_k b_k$$

Substituting (17) in (15), we obtain

$$H_k(q) = \frac{\beta_k}{\sigma_{\tilde{h}_k} \sqrt{a_k}(1 - b_k q^{-1})} \quad (18)$$

leading to

$$\tilde{h}_{d,k}(n) = b_k \tilde{h}_{d,k}(n-1) + \eta_k \hat{\tilde{h}}_{d,k}(n) \quad (19)$$

where

$$\eta_k = \frac{\beta_k}{\sigma_{\hat{e}_{d,k}} \sqrt{a_k}} = \frac{\sigma_{h_{d,k}}^2}{\sigma_{\hat{e}_{d,k}}^2} \frac{(1 - \rho^2)}{\rho(1 - b_k \rho)} b_k \quad (20)$$

When there is no time correlation ($\rho = 0$) over slots, we have $b_k = 0$ and $\eta_k = \frac{\sigma_{h_{d,k}}^2}{\sigma_{h_{d,k}}^2 + \sigma_{\hat{e}_{d,k}}^2}$, so that every channel coefficient is weighted by $\eta_k \leq 1$, weighting between a priori variance information and estimation error. The estimation error variance of $\tilde{h}_{d,k}(n)$ is given by

$$\sigma_{\tilde{e}_{d,k}}^2 = \sigma_{\hat{e}_{d,k}}^2 \left(1 - \frac{\sigma_{\hat{e}_{d,k}}^2}{\sigma_{k,\infty}^2}\right) \quad (21)$$

where $\sigma_{k,\infty}^2 = \sigma_{\hat{e}_{d,k}}^2 a_k$ denotes the infinite order forward prediction error variance. Along the lines of [2] the optimal Wiener filter (19) can be implemented in adaptive fashion by adapting the two coefficients b_k and η_k to match to the Doppler statistics of the channel taps, in order to minimize the mean square error between $\tilde{h}_{d,k}(n)$ and the actual channel $h_{d,k}(n)$. Introducing temporal averaging over slots, with exponential weighting, the RLS adaptation algorithm for the minimization problem stated above can be formulated where the coefficients b_k and η_k are computed by recursive solution

$$\begin{cases} \boldsymbol{R}_n = \lambda \boldsymbol{R}_{n-1} + \text{Re}\left\{ \begin{bmatrix} \tilde{h}_{d,k}(n-1) \\ \hat{\tilde{h}}_{d,k}(n) \end{bmatrix} \begin{bmatrix} \tilde{h}_{d,k}(n-1) \\ \hat{\tilde{h}}_{d,k}(n) \end{bmatrix}^H \right\} \\ \boldsymbol{P}_n = \lambda \boldsymbol{P}_{n-1} - \begin{bmatrix} 0 \\ \sigma_{\hat{e}_{d,k}}^2 \end{bmatrix} \\ \begin{bmatrix} b_k \\ \eta_k - 1 \end{bmatrix} = \boldsymbol{R}_n^{-1} \boldsymbol{P}_n \end{cases}$$

$$(22)$$

where $\text{Re}\{\cdot\}$ denotes the real part, and λ is the forgetting factor. Since there are just two coupled parameters direct 2×2 matrix inversion can be performed instead of using true RLS. The RLS initialization requires only $\boldsymbol{R}_0$ to be different from zero, so we can set it to $\boldsymbol{R}_0 = 10^{-3}\boldsymbol{I}_2$, and $\tilde{h}_{d,k}(0) = \hat{\tilde{h}}_{d,k}(1)$.

Finally the delays τ_p and the coefficients $c_{d,p}(n)$ can be estimated from $\tilde{h}_d(n)$ by solving (4).
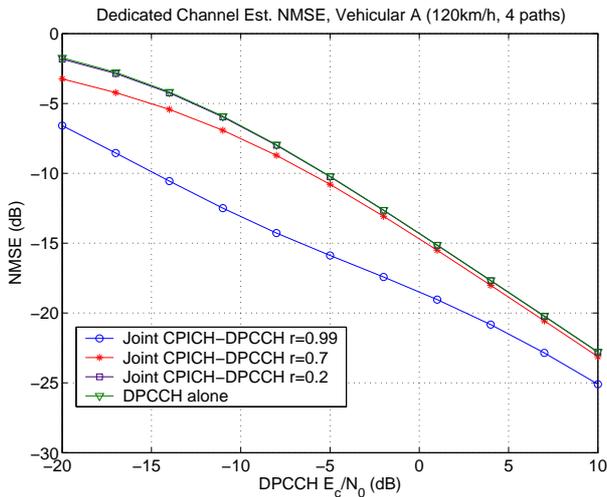
Fig. 1. DPCH channel estimation NMSE vs. DPCCH E_c/N_0, in fast fading channel, for different values of common-dedicated channel correlation (r)
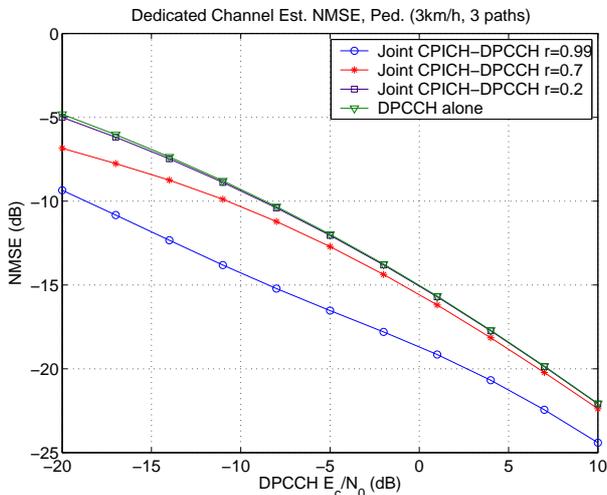

Fig. 2. DPCH channel estimation NMSE vs. DPCCH E_c/N_0, in slow fading channel, for different values of common-dedicated channel correlation (r)

IV. NUMERICAL EXAMPLES

Here we assess the performance of the joint common-dedicated pilots based dedicated channel estimation in the presence of dedicated transmit beamforming in terms of analytical normalized MSE (NMSE) of the overall channel estimate $\tilde{\boldsymbol{h}}_d(n)$ according to (21). The quantities $|\alpha_k|$, $\sigma^2_{h_{d,k}}$, σ^2_v, and $\sigma^2_{h_{c,k}}/\sigma^2_{h_{d,k}}$ are assumed known. We also define the correlation factor $r_k = |\alpha_k|\sigma_{h_{d,k}}/\sigma_{h_{c,k}} \leq 1$. In the light of the previous derivation it is clear that the performances are primarily affected by the correlation between common and beamformed dedicated channel. Therefore for the sake of simplicity in the numerical examples presented here we set $|\alpha_k| = |\alpha_0|$ constant $\forall k$. Without loss of generality we assume the chip energy $E_c = 1$. In addition we set the CPICH and DPCCH pilot chips sequences such that $\beta_c = 2560$ and $\beta_d = 512$ [9], and we simulate a power offset between common and

dedicated channel by setting $\sigma^2_{h_{c,k}}/\sigma^2_{h_{d,k}} = 2$ for all k's so that $r = r_k = |\alpha_0|/\sqrt{2}$. We also denote the DPCCH receive chip SNR $E_c\mathrm{E}|\boldsymbol{h}_d(n)|^2/\sigma^2_v$ as "DPCCH E_c/N_0". Figure 1 shows the dedicated channel NMSE vs. the DPCCH E_c/N_0, for different values of r under the assumption of a four paths fast fading channel (mobile velocity of 120 km/h). Similarly in figure 2 the dedicated channel NMSE vs. the DPCCH E_c/N_0 is shown, for different values of r, under the assumption of a three paths slow fading channel (mobile velocity of 3 km/h). The performances of the dedicated pilots only based channel estimation [2] are shown as well for comparison. As expected, large performance improvements can be achieved by jointly exploiting both dedicated and common pilots rather than only dedicated pilots when the dedicated and the common channels are highly correlated (up to 10 dB in terms of DPCCH E_c/N_0 at NMSE ≈ -10 dB, for $r = 0.99$, in the examples shown here).

V. CONCLUSION

We derived an optimal approach to estimate the user dedicated channel in W-CDMA receivers, jointly exploiting common and dedicated pilots, which is suited to operate in the presence of dedicated transmit beamforming without requiring the prior knowledge of the channel paths delays and/or of beamforming parameters. We showed that significant performance improvements can be achieved by the proposed joint common-dedicated pilots based dedicated channel estimation, with respect to classical dedicated pilots only based channel estimation approaches, particularly when the dedicated and common channels are highly correlated.

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