LINEAR PRECODING IN W-CDMA SYSTEMS BASED ON S-CPICH CHANNEL ESTIMATION

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ABSTRACT

In this article we analyze the impact of channel estimation by means of the *secondary common pilot channel* (S-CPICH) on linear precoding in W-CDMA systems. The pilot sequences on the S-CPICH are transmitted over a fixed grid of beams, causing a systematic channel estimation error at the receiver. Consequently, the adapted rake receiver coefficients will be corrupted due to the use of the S-CPICH, which will lead to an incoherent superposition of the signal components when conventional adaptive space time precoders are employed. Assuming partial *channel state information* (CSI) at the transmitter we show how the effect of S-CPICH based channel estimation can be reduced or even removed, resulting in an enhanced system performance.

1. INTRODUCTION

Mobile communication systems face the distortion of the transmitted signal by the channel, interference and noise. The conventional approach to deal with these imperfections is receive signal processing, where an adaptive filter at the receiver equalizes the signal and reduces the influences of interference and noise. The aim for low power consumption and thus low numerical complexity at the *user equipment* (UE) and the use of multi sensor antenna arrays lead to linear precoding, where a space time transmit filter adapts to the combination of channel and a given, but known receive filter of reduced complexity, which is independent of the transmit filter. Note, that unlike in joint optimization, where transmit and receive filter are designed in one common procedure, the filter design of the linear precoders directly depends upon the applied receive processing.

Assuming full *channel state information* (CSI) at the transmitter [1], [2] and [3] proposed linear precoding schemes following the matched filter, the zero forcing and the MMSE principle, respectively. Moreover [4] showed, how linear pre-equalization can be applied to a system with only partial CSI at the transmitter, when the UE employs a *maximum ratio combining* (MRC) rake receiver. These approaches are still based on the full and perfect CSI at the receiver, an assumption, that is reasonable as long as the channel estimation is performed on a *primary common pilot channel* (P-CPICH). This P-CPICH allows the UE to estimate the complex path coefficients and thus to correctly adapt its rake coefficients.

An obstacle, that seemed to stall the proposed precoding solutions is the use of S-CPICH channel estimation. The use of this S-CPICH causes a systematic estimation error at the UE, which is not regarded by the conventional linear precoding schemes. The use of conventional precoding solutions will thus result in an incoherent superposition of the receiver outputs and therefore they are not applicable, when S-CPICH channel estimation is used. The following paragraphs will provide an adapted approach, that considering the mentioned adulteration restores the applicability of linear precoding schemes.

For clear understanding and a proper problem statement we first aim for an explicit model of the W-CDMA system in Section 2 before we turn to the derviation of the new linear precoding scheme in Section 3. Finally, the evaluations in Section 4 show the regained applicability and the superiority of our proposal with respect to the state of the art solution.

2. SYSTEM MODEL

For the remaining part of this article, let us examine the downlink of a K user system, with N_a transmit antennas and spatio temporal transmit filters of order L, described by the coefficient vectors $\boldsymbol{p}_{k,l} \in \mathbb{C}^{N_\mathrm{a}}$. The K channels

$$\boldsymbol{h}_k[n] = \sum_{q=0}^{Q} \boldsymbol{h}_{k,q}\delta[n - \nu_{k,q}] \tag{1}$$

are described by their $Q + 1$ components $\boldsymbol{h}_{k,q}$ that arrive with the temporal delays $\nu_{k,q}$. As section 4 confirms, it is sufficient to consider the dominant eigencomponent $\boldsymbol{a}_{k,q} \in \mathbb{C}^{N_\mathrm{a}}$ of $\mathrm{E}\left[\boldsymbol{h}_{k,q}\boldsymbol{h}_{k,q}^{\mathrm{H}}\right]$ only. Thus, each of these channel vectors can be written as

$$\boldsymbol{h}_{k,q} = \rho_{k,q}\boldsymbol{a}_{k,q} \quad \in \mathbb{C}^{N_\mathrm{a}} \tag{2}$$

with the complex scalar path attenuation $\rho_{k,q}$. The UEs all have one receive antenna and an $F + 1$ finger MRC rake receiver.

2.1. Channel Estimation via the CPICH

As channel estimation on the dedicated channels only provides a low number of low power training symbols, CPICHs have been introduced. The key problem now is the effect of the used CPICH on the rake receiver coefficients. Initially, the P-CPICH allows the UE to estimate the full vector channel $\boldsymbol{h}_{k,q} \in \mathbb{C}^{N_\mathrm{a}}$, as every antenna element radiates its own pilot sequence. From (2) the UE can extract the complex path attenuation $\rho_{k,q}$ and adapt its rake coefficients in an MRC sense as $v_{k,q} = \rho_{k,q}^*$.

As the transmit power necessary for P-CPICHs increases with N_a, S-CPICHs have been introduced for $N_\mathrm{a} > 2$ to make use of the available antenna gain. Within, the different S-CPICH sequences are transmitted via N_b fixed beams $\boldsymbol{b}^{(1)}, \ldots, \boldsymbol{b}^{(N_\mathrm{b})}$. The UE therefore has to pick one of these pilots to estimate the scalar channel substitute:

$$h_k^{\mathrm{subs}}[n] = \sum_{q=0}^{Q} \rho_{k,q} \underbrace{\boldsymbol{b}_k^{\mathrm{T}} \boldsymbol{a}_{k,q}}_{\alpha_{k,q}} \delta[n - \nu_{k,q}], \tag{3}$$

where $\boldsymbol{b}_k \in \{\boldsymbol{b}^{(1)}, \ldots, \boldsymbol{b}^{(N_\mathrm{b})}\}$ is the S-CPICH beam used for channel estimation. Only given a scalar channel substitute instead of the full vector channel, the UE is not able to extract $\rho_{k,q}$ from the product of $\rho_{k,q}\boldsymbol{b}_k^{\mathrm{T}}\boldsymbol{a}_{k,q}$ and thus adjusts its rake coefficients to

$$v_{k,f} = \boldsymbol{b}_k^{\mathrm{H}} \boldsymbol{a}_{k,f}^* \rho_{k,f}^* = \alpha_{k,f}^* \rho_{k,f}^*. \tag{4}$$

Note that $\alpha_{k,q} \in \mathbb{C}$ in general is a complex parameter. Thus the corruption of the rake coefficients will cause an incoherent superposition of the different rake finger signals if the precoding vectors $\boldsymbol{p}_{k,l} \in \mathbb{C}^{N_\mathrm{a}}$ do not fulfill:

$$\arg\left\{\sum_{l=0}^{L} \boldsymbol{p}_{k,l}^{\mathrm{T}} \boldsymbol{a}_{k,q}\right\} = \arg\left\{\boldsymbol{b}_k^{\mathrm{T}} \boldsymbol{a}_{k,q}\right\}. \tag{5}$$

The trivial solution of (5) is $\boldsymbol{p}_{k,l} = d_{k,l}\boldsymbol{b}_k$, $\forall l$ with the arbitrary real scalars $d_{k,l}$. This transmission via the S-CPICH

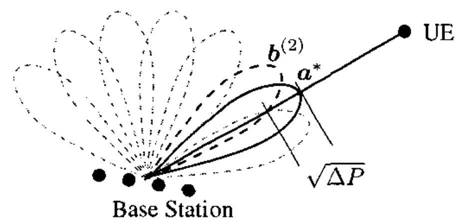

Fig. 1. Optimum beamforming solution (solid) versus use of the S-CPICH beam (dashed)

beam used for channel estimation indeed will ensure a coherent superposition at the rake output. However, it leads to a severe power loss, as only a part of the antenna gain is exploited. Figure 1 schematicly depicts a single user example in a flat fading environment with $N_\mathrm{b} = 7$ S-CPICH beams visualizing the power loss ΔP of S-CPICH beamforming $\boldsymbol{p} = \boldsymbol{b}^{(2)}$ versus the respective optimum transmit matched filter solution $\boldsymbol{p} = \boldsymbol{a}^*$.

2.2. Instantaneous S-CPICH System Model

Using the above result we can analyze the decision variable at the k^{th} UEs. Assuming a rake receiver with the coefficients $v_{k,f} = \alpha_{k,f}^* \rho_{k,f}^*$, (cf. Eqn. 4) and a spreading sequence $c_k[0], \ldots, c_k[\chi - 1]$ of length χ, the component of the decision signal, that propagated over the q^{th} channel path and the f^{th} rake finger reads as

$$\begin{aligned} \hat{s}_{k,q,f}[n] = & \sum_{j=0}^{\chi-1} c_k^*[j] \alpha_{k,f}^* \rho_{k,f}^* y_{k,q}[n + f + j] \\ & + \sum_{j=0}^{\chi-1} c_k^*[j] \alpha_{k,f}^* \rho_{k,f}^* \eta_{k,f}[n + f + j]. \end{aligned} \tag{6}$$

Within, $\eta_{k,f}[n]$ denotes the respective noise portion and $y_{k,q}[n]$ is the corresponding noiseless receive signal component after pre-equalization and transmission over $\rho_{k,q}\boldsymbol{a}_{k,q}$:

$$y_{k,q}[n] = \rho_{k,q} \boldsymbol{a}_{k,q}^{\mathrm{T}} \sum_{i=1}^{K} \sum_{l=0}^{L} \boldsymbol{p}_{i,l} s_i[n - l - q]. \tag{7}$$

2.3. Average S-CPICH System Model

The complex path coefficients $\rho_{k,q}$ are not reciprocal in the up- and downlink of W-CDMA and therefore can not be estimated from the uplink at the base station. In order to still apply linear pre-equalization to W-CDMA systems, we follow [4] and aim for an approximation of $\hat{s}_{k,q,f}[n]$ by a longterm equivalent $\hat{u}_{k,q,f}[n]$, that only depends upon longterm parameters like the path power $\sigma_{k,q}^2 = \mathrm{E}\left[|\rho_{k,q}|^2\right]$ and $\boldsymbol{a}_{k,q}$. With such an average system model, we can derive linear precoding solutions, that are computable for the base station even in W-CDMA scenarios, as they do not require the knowledge of the fading coefficients $\rho_{k,q}$.

In detail, if we call for a power preserving approximation of Eqns. (6) and (7), according to

$$\mathrm{E}[|\hat{u}_{k,q,f}[n]|^2] = \mathrm{E}[|\hat{s}_{k,q,f}[n]|^2], \tag{8}$$

we ensure, that the components of the approximated decision signal $\hat{u}_{k,q,f}[n]$ have the same power as their exact instantaneous equivalents $\hat{s}_{k,q,f}[n]$. Substituting equations (6) and (7) into (8) and solving for $\hat{u}_{k,q,f}[n]$ yields:

$$\begin{aligned} \hat{u}_{k,q,f}[n] = & \kappa_{q,f} \sigma_{k,f} \alpha_{k,q} \sum_{j=0}^{\chi-1} c_k^*[j] \tilde{y}_{k,q}[n + f + j] \\ & + \alpha_{k,f} \sigma_{k,f} \sum_{j=0}^{\chi-1} c_k^*[j] \eta[n + f + j]. \end{aligned} \tag{9}$$

with the longterm equivalent of the component $y_{k,q}$:

$$\tilde{y}_{k,q}[n] = \sigma_{k,q} \boldsymbol{a}_{k,q}^{\mathrm{T}} \sum_{i=1}^{K} \sum_{l=0}^{L} \boldsymbol{p}_{i,l} s_i[n-l-q]. \qquad (10)$$

Note, that $\rho^*_{k,f}$ in the first line of (6) and $\rho^*_{k,q}$ in (7) are identical for $f = q$. As furthermore $\mathrm{E}[|\rho_{k,q}|^4] = 2\sigma^4_{k,q} \neq \sigma^4_{k,q}$ for complex Gaussian random variables $\rho_{k,q}$, we introduced the factor $\kappa_{q,f} = \sqrt{2}$ for $f = q$ and $\kappa_{q,f} = 1$ else. Using the commutativity of the scalar product in Eqn. (10), the average model of Eqns. (9) and (10) can be written in vector matrix form as:

$$\hat{u}_{k,q,f}[n] = \sum_{i=1}^{K} \boldsymbol{p}_i^{\mathrm{T}} \boldsymbol{X}_{k,q,f}(\alpha_{k,f}) \boldsymbol{s}_i[n] + \tilde{\eta}_{k,f}[n+f].$$

The vector $\boldsymbol{p}_i$ is obtained by stacking all $L+1$ weighting vectors for user i according to

$$\boldsymbol{p}_i = [\boldsymbol{p}_{i,0}^{\mathrm{T}}, \ldots, \boldsymbol{p}_{i,L}^{\mathrm{T}}]^{\mathrm{T}},$$

and the vector $\boldsymbol{s}_i[n] \in \mathbb{C}^M$ contains all $M = L + Q + F + \chi$ samples of s_i that influence $\hat{u}_{k,q,f}[n]$ through the chain of transmit filter, channel, rake receiver, and decorrelator. Moreover the matrix $\boldsymbol{X}_{k,q,f}(\alpha_{k,f}) \in \mathbb{C}^{N_a(L+1)\times M}$ is defined as:

$$\boldsymbol{X}_{k,q,f}(\alpha_{k,f}) = \kappa_{q,f} \sigma_{k,f} \sigma_{k,q} \alpha_{k,f} \boldsymbol{A}_{k,q,f} \boldsymbol{C}_k.$$

Within, the matrix $\boldsymbol{A}_{k,q,f} \in \mathbb{C}^{N_a(L+1)\times L+Q+F+1}$ describes the spatio temporal structure of the channel as

$$\boldsymbol{A}_{k,q,f} = \begin{bmatrix} \boldsymbol{0}_{F+q-f\times N_a(L+1)} \\ \boldsymbol{1}_{L+1} \otimes \boldsymbol{a}_{k,q}^{\mathrm{T}} \\ \boldsymbol{0}_{Q+f-q\times N_a(L+1)} \end{bmatrix}^{\mathrm{T}},$$

while the Toeplitz matrix $\boldsymbol{C}_k \in \mathbb{C}^{L+Q+F+1\times L+Q+F+\chi}$ contains the vector $[c_k^*[\chi-1], \ldots, c_k^*[0]]$ in its rows and denotes the deconvolution with the spreading code.

3. LINEAR PRECODING SOLUTIONS

The key idea of linear precoding in S-CPICH environments is the estimation of the rake corrupting factors

$$\alpha_{k,q} = \boldsymbol{b}_k^{\mathrm{T}} \boldsymbol{a}_{k,q} \qquad (11)$$

at the base station. Employing the reciprocity of the W-CDMA channel with respect to its longterm properties, the base station always can estimate the channel eigenvectors $\boldsymbol{a}_{k,q}$ from uplink measurements and use it to reconstruct the UEs choice of $\boldsymbol{b}_k$. Using Eqn. (11) the transmitter has full knowledge about the corruption of the rake coefficients and can include it in the optimization of its precoding vectors and finally avoid an incoherent summation of the rake output.

3.1. Wiener Filter

This section will use the derived signal model to evolve an S-CPICH solution of the W-CDMA *transmit Wiener filter* (TxWF) as proposed in [4]. Unlike its receive counterpart, the TxWF allows for a real valued gain $\beta \in \mathbb{R}$ at the rake output and thus results from the minimization of the modified *mean squared error* (MSE):

$$\epsilon(\boldsymbol{p}_1, \ldots, \boldsymbol{p}_K, \beta) = \sum_{k=1}^{K} \mathrm{E}\left[\left\|\boldsymbol{u}_k[\chi m] - \beta^{-1}\hat{\boldsymbol{u}}_k[\chi m]\right\|_2^2\right],$$

with $\boldsymbol{u}_k[\chi m]$ and $\hat{\boldsymbol{u}}_k[\chi m]$ being column vectors obtained by stacking $\hat{u}_{k,q,f}[\chi m]$ and the desired values

$$u_{k,q,f}[\chi m] = \begin{cases} \sqrt{2}\sigma^2_{k,q} d_k[\chi m] & f = q, \\ 0 & f \neq q, \end{cases}$$

for $\{q, f\} = \{0,0\}, \ldots, \{Q,0\}, \{0,1\}, \ldots, \{Q,F\}$. With the constraint of limited transmit power, the identifying optimization problem reads:

$$\{\boldsymbol{p}_{\mathrm{WF},1}, \ldots, \boldsymbol{p}_{\mathrm{WF},K}\} = \arg \min_{\boldsymbol{p}_1,\ldots,\boldsymbol{p}_K} \epsilon(\boldsymbol{p}_1, \ldots, \boldsymbol{p}_K, \beta),$$
$$\text{s.t.: } \sum_{k=1}^{K} \sigma_s^2 \|\boldsymbol{p}_k\|_2^2 = E_{\mathrm{tr}}. \qquad (12)$$

Note that $\hat{\boldsymbol{u}}_k[\chi m]$ and thus the MSE depends on the rake corruption $\boldsymbol{\alpha}_k$, which now is considered by the optimization in (12). The solution to the above problem is known from [4] and given as:

$$\boldsymbol{p}_{\mathrm{WF},k} = \beta_{\mathrm{WF}} \left(\boldsymbol{X}^* \boldsymbol{X}^{\mathrm{T}} + \gamma \frac{\sigma_\eta^2}{E_{\mathrm{tr}}} \mathbf{1} \right)^{-1} \boldsymbol{X}^* \boldsymbol{q}_k, \qquad (13)$$

where the real scalar β_{WF} is chosen to fulfill the constraint in Eqn (12). Stacking the matrices $\boldsymbol{X}_{k,q,f}(\alpha_{k,f})$ and the vectors

$$\boldsymbol{q}_{i,q,f} = \begin{cases} \sqrt{2}\sigma_{i,q} & \text{for } i = k \wedge q = f, \\ 0 & \text{else}, \end{cases} \qquad (14)$$

for $\{k,q,f\} = \{1,0,0\}, \ldots, \{1,Q,0\}, \ldots, \{1,Q,F\}, \ldots$ $\ldots, \{K,Q,F\}$ yields $\boldsymbol{q}_k$ and $\boldsymbol{X}$, which contains the influence of the used S-CPICH beam on the TxWF result.

4. EVALUATION

Finally let us evaluate the performance of the TxWF in the downlink of an UMTS W-CDMA system with the following setting: QPSK Modulation, $K = 4$, $\chi = 4$, $N_a = N_b = 8$. The employed channel model has $Q+1 = 4$ block Rayleigh fading channel paths per user with relative mean path powers of 0, −3, −6 and −9 dB, i.e. a 3GPP CASE 3 power delay profile. The data tranmission over the S-CPICH *grid* *of beams* (GoB), i.e. the beam $b^{(i)}$ used for channel estimation, is serving as a reference. The employed weights $b^{(i)}$ are phased array vectors corresponding to the GoB *directions of departure* (DOD) $\pm 52.5°$, $\pm 37.5°$, $\pm 22.5°$, $\pm 7.5°$.

Furthermore we define a *spatial scenario* (SPACE) 1, where the users are placed at $\theta_{k,q} = -45°$, $-15°$, $15°$ and $45°$ for $k = 1, \ldots, 4$ respectively. Moreover, each users has one DOD for all its temporal macropaths only, i.e. $\theta_{k,q} =$ const., $\forall q$. A second setting (SPACE 2) where the spatial situation changes with every TTI is defined in [5]. Here the DOD $\theta_{k,0}$ of the first macropaths are realizations of a uniform distribution $\mathcal{U}(-60°, 60°)$ whereas the following ones are distributed normally according to $\mathcal{N}(\theta_{k,0}, 15°)$.

Let us first examine the TxWF in terms of raw *bit error rate* (BER) over the ratio of total transmit power over noise power. Basing on SPACE 1, Fig. 2 compares the TxWF and the S-CPICH GoB for different values of Laplacian *angular spread* (AS). Due to its ability to recover the full antenna

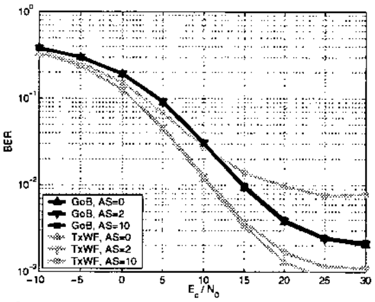

Fig. 2. Mean BER in different SPACE 1 scenarios

gain and its spatial adaptivity the TxWF for AS= $0°$ outperforms the S-CPICH GoB by 3.3 dB and 4.2 dB at 10% and 1% BER respectively. Though designed under the assumption of discrete wave propagation (AS= $0°$), the TxWF is rather robust against increasing spread. A Laplacian AS of $2°$ does not change the performance significantly for the BERs regarded above. Only with the vast spread of $10°$ the TxWF performance for large SNR values falls behind the GoB reference. This robustness can further be improved if more eigencomponents of the channel are included in Eqn. (2).

Note, that the TxWF not only in GoB worst case settings like SPACE 1 yields significant gains. In fact, the TxWF in any scenario outperforms the state of the art solution. Averaging over different spatial settings, Tab. 1 shows the corresponding results for a SPACE 2 scenario and gives mean values for the achieved performance gain.

	AS=0°	AS=2°	AS=10°
BER=10%	0.5 dB	0.5 dB	−0.1 dB
BER= 5%	0.4 dB	0.4 dB	−0.7 dB

Table 1. Mean power advantage of the TxWF in SPACE 2

Extending our evaluations to *coded bit error rate* (CBER), we add a rate $\frac{1}{3}$ turbo code as standardized in 3GPP with an iterative max log decoder to the above system. One code word contains 30 shortterm realizations with 2560 chips each and is perfectly interleaved. Fig. 3 plots the CBER for the 4 UEs separately and shows, that the gains from Fig. 2 result in an effective performance enhancement of more than 3 dB in the given setting. Considering the standardization independent applicability, this is a valuable advantage.

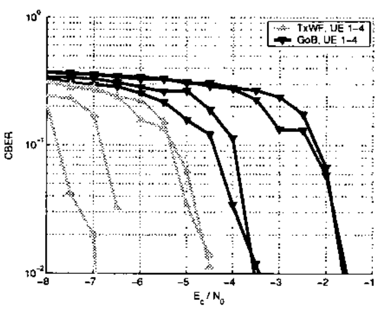

Fig. 3. Results of a coded transmission in SPACE 1, AS=2°

As the TxWF is designed to minimize the sum of MSEs though, it mutually trades performance among the UEs, resulting in different quality of service levels. Two possible solutions to this problem are adaptive modulation schemes, that make use of the resulting performance differences, or the derivation of a *fair* TxWF.

5. CONCLUSION

Linear precoding techniques can be applied to S-CPICH based W-CDMA systems, where they yield performance enhancements of up to 3.5 dB. Moreover, the proposed scheme is standard compliant, making it an attractive solution for UMTS base station manufacturers.

6. REFERENCES

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