# Influence of Channel Estimation on the Performance of a coherent DS-CDMA System

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*Abstract*— **In this paper the uplink of a coherent direct sequence code division multiple access (DS-CDMA) system will be investigated. The channel estimation is carried out via a control channel which is transmitted simultaneously to the data channel. This concept is very closely related to the CODIT[1]-CDMA concept. In the channel estimation an averaging process is involved, which for long averaging periods suppresses the noise influence efficiently. On the other hand, long averaging intervals lead to systematic errors in the estimated channel coefficients due to the time-variant behavior of the channel. Taking into account these effects, a closed-form analysis of the data channel symbol error rate is presented. This analysis allows to predict the system's performance and helps to optimize the parameter settings for the channel estimation.**

*Keywords*— **DS-CDMA, Channel Estimation, Coherent Detection**

## I. Introduction

DIRECT sequence code division multiple access has become one of the favorite candidates for future mobile radio communication systems. Both coherent and noncoherent modulation techniques have been adopted in several system proposals. A coherent system requires a channel estimation which is able to follow the time-variant mobile radio channel. Usually test sequences are used for the purpose of channel estimation which, on the other hand, increase the necessary $\bar{E}_b/N_0$ for a desired system performance. From the viewpoint of system optimization as little energy as possible should be used for the channel estimation.

In section 2 the uplink of a coherent DS-CDMA system model is introduced. Section 3 presents the detection of its control channel and the channel estimation procedure. The channel estimation is characterized with respect to deterministic and stochastic effects. In section 4 the detection of the data channel is described and the effects of imperfect channel estimation are analyzed. A closed form expression for the symbol error rate in the data channel is derived. Section 5 illustrates the method by an example and compares it to simulation results. The optimization of the system design is treated in section 6. Furthermore a comparison to a noncoherent CDMA-concept is included. The conclusions are drawn in section 7.

## II. The uplink

As mentioned, the system under investigation is very similar to the CODIT-system which is described in detail in [1] and [2]. In this section we will summarize its concept for the uplink and introduce our nomenclature. Every mobile station transmits a data- and a control signal to the base station. The control signal is used to adapt the configuration of the data channel to a service or demand of the radio network. The corresponding binary control- and information data signals with individual rates are denoted by $d_C(i)$ and $d_D(j) \in \{-1, +1\}$, respectively. They are coded separately into the symbol sequences $b_C(\xi)$ and $b_D(\eta)$ with symbol rates $1/T_{hC}$ and $1/T_{hD}$. In addition, the control channel symbols are encoded differentially

$$b_C^{\Delta}(\xi) \;=\; b_C^{\Delta}(\xi-1) \cdot b_C(\xi) \tag{1}$$

which allows a differential coherent demodulation at the receiver without knowledge of the channel coefficients. For the sake of simplicity we will describe the system by a chip tap representation in the complex baseband. Both channels operate with the same chip rate $1/T_c$. The spreading factors $S_{P,C} = T_{hC}/T_c$ and $S_{P,D} = T_{hD}/T_c$ depend on the given symbol rates in the control- and the data channel. The corresponding pseudo noise (PN) spreading sequences $PN_C(k)$ and $PN_D(k) \in \{\frac{1}{\sqrt{2}}(1+j), \frac{1}{\sqrt{2}}(1-j), \frac{1}{\sqrt{2}}(-1+j), \frac{1}{\sqrt{2}}(-1-j)\}$ are independent and user specific. Both PN sequences can be partitioned into consecutive sections $\boldsymbol{h}_C(\xi),\ \xi = 0, 1, \dots$ and $\boldsymbol{h}_D(\eta),\ \eta = 0, 1, \dots$ according to the length of the associated symbol periods as illustrated in **Fig. 1**. In vector representation the transmit signal of the control channel is then given by

$$PN_C(k) \quad PN_C(k+1)$$

$$PN_C: \;\dots\; \underbrace{\square\square\square\square\square\square\square}_{\boldsymbol{h}_C(\xi)}\underbrace{\square\square\square\square\square\square\square}_{\boldsymbol{h}_C(\xi+1)} \;\dots$$

$$PN_D: \;\dots\; \underbrace{\square\square\square\square\square}_{\boldsymbol{h}_D(\eta)}\underbrace{\square\square\square\square\square}_{\boldsymbol{h}_D(\eta+1)} \;\dots$$

Fig. 1. Illustration of vectors $\boldsymbol{h}_C(\xi)$ and $\boldsymbol{h}_D(\eta)$

$$\boldsymbol{z}_C(\xi) \;=\; \sqrt{\rho_C}\, b_C^{\Delta}(\xi)\, \boldsymbol{h}_C(\xi) \quad . \tag{2}$$

For the data channel the corresponding relationship is

$$\boldsymbol{z}_D(\eta) \;=\; \sqrt{\rho_D}\, b_D(\eta)\, \boldsymbol{h}_D(\eta) \quad . \tag{3}$$

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[1]CODIT: **C**ode **Di**vision **T**estbed, research project within the RACE program (Research and technology development in Advanced Communications technologies in Europe)

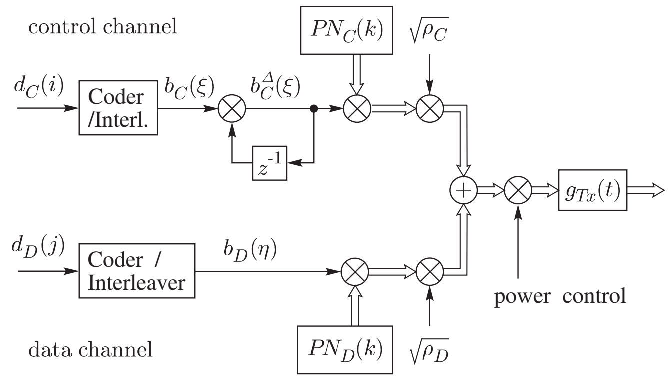


Fig. 2. Uplink transmitter

The factors $\sqrt{\rho_C}$ and $\sqrt{\rho_D}$ split up the transmit power to the data and the control channel with

$$\rho_C \;+\; \rho_D \;=\; 1 \quad . \tag{4}$$

The corresponding transmitter is depicted in **Fig. 2**.

## III. Control channel detection and channel estimation

The control channel and the data channel are detected in separate RAKE[2] receivers. First the DBPSK modulated control channel is demodulated differentially coherently in each RAKE path and a channel estimation is carried out. The stored received signal is used later (delay $\chi_2$) for a coherent detection of the data channel symbols in the second RAKE receiver (see **Fig. 3**).

We will start with the analysis of the control channel detection with emphasis on the channel estimation. The received signal in the $\mu$-th RAKE path of the control channel RAKE receiver can be written as

$$\boldsymbol{s}_{C,\mu}(\xi) \;=\; \boldsymbol{z}_C(\xi) \;+\; \boldsymbol{r}_{C,\mu}(\xi) \quad , \tag{5}$$

where $\boldsymbol{r}_{C,\mu}(\xi)$ consists of all signal components, which interfere with the control channel signal in the $\mu$-th path. The correlation with $\frac{1}{\sqrt{\rho_C}E_{hC}}\boldsymbol{h}_C(\xi)$, where $E_{hC}$ denotes the energy of the control channel symbols $\boldsymbol{h}_C(\xi)$, yields the echo path correlations

$$u_{C,\mu}(\xi) \;=\; b_C^{\Delta}(\xi)\, c_\mu(\xi) \;+\; \frac{1}{\sqrt{\rho_C}\,E_{hC}}\, \boldsymbol{h}_C^*(\xi)\,\boldsymbol{r}_{C,\mu}(\xi) \tag{6}$$

where $c_\mu(\xi)$ denotes the channel coefficient in the $\mu$-th RAKE path at time $\xi$. After the differentially coherent detection of these echo correlations and decoding, we assume that due to the powerful code used for the control channel, the control data $\hat{d}_C(i)$ is decided almost error free.

From a reencoding of the decided data $\hat{d}_C(i)$ we obtain code symbols $\hat{b}_C^{\Delta}(\xi)$ which should be equal to the transmitted code symbols $b_C^{\Delta}(\xi)$. As the detection of the control channel is not the key issue of this paper and to simplify the presentation we will assume from now on that the control channel data is decided correctly by the receiver or known by other means (test data), i.e. $\hat{b}_C^{\Delta}(\xi) = b_C^{\Delta}(\xi)$. The channel estimation is then performed by the following averaging procedure over $M_c$ control channel symbols

$$\hat{c}_\mu(\xi) \;=\; \frac{1}{M_c} \sum_{\ell=-\frac{M_c-1}{2}}^{\frac{M_c-1}{2}} \hat{b}_C^{\Delta}(\xi+\ell)\, u_{C,\mu}(\xi+\ell) \tag{7}$$

$$= \underbrace{\frac{1}{M_c} \sum_{\ell=-\frac{M_c-1}{2}}^{\frac{M_c-1}{2}} c_\mu(\xi+\ell)}_{=:\,\tilde{c}_\mu(\xi)} + \underbrace{\frac{1}{M_c\sqrt{\rho_C}\,E_{hC}} \sum_{l=\xi-\frac{M_c-1}{2}}^{\xi+\frac{M_c-1}{2}} \hat{b}_C^{\Delta}(l)\, \boldsymbol{h}_C^*(l)\, \boldsymbol{r}_{C,\mu}(l)}_{=:\,n_{C_1,\mu}(\xi)} \quad . \tag{8}$$

The variance of the noise term $n_{C_1,\mu}(\xi)$ is proportional to $1/M_c$. For a time-invariant channel $\tilde{c}_\mu(\xi)$ equals $c_\mu(\xi)$ and a large value of $M_c$ would suppress the noise efficiently. However, a typical mobile radio channel is time-variant and a large averaging period (large $M_c$) would lead to a tracking problem with respect to the instantaneous value of $c_\mu(\xi)$. So we must find a trade-off between noise suppression and tracking of the channel coefficient.

The following analysis of the value $\tilde{c}_\mu(\xi)$ is described for an $L$-path Rayleigh channel. Although this kind of channel model simplifies the scenarios of real world radio transmission to some extent, it covers the most essential properties of mobile radio channels, like time-variance and frequency-selectivity. Furthermore, it allows easy analytical analysis which becomes much more difficult for more complex model definitions like the well-known GSM[3] channel models (see [4]) or even more general ones, like the e.g. Nagakami-m distribution based channel models [3]. Due to this, the $L$-path Rayleigh channel has become a familar reference to evaluate performance aspects of mobile radio transmission systems (see [2], [10]). It should be noted that the following analysis of the channel estimation is not restricted to the "pure" $L$-path Rayleigh model in principle. Especially non-equal average path strengths and other Doppler probability densities can be taken into account easily.

For the pure $L$-path Rayleigh channel, the $L$ diversity paths (echo components) have equal average strengths $E\{|c_\mu(\xi)|^2\} = 1/L$. The channel coefficient of the $\mu$-th echo path $c_\mu(\xi+\ell)$, where $\xi$ is a temporarily fixed time reference and $\ell$ is a time offset which varies during the averaging process, can be modelled as

$$c_\mu(\xi+\ell) \;=\; \sqrt{\frac{1}{L}}\, \frac{1}{\sqrt{N_\mu}} \sum_{\nu=0}^{N_\mu-1} e^{j\theta_{\mu,\nu}(\xi)}\, e^{j\Omega_{\mu,\nu}\ell} \quad . \tag{9}$$

[2]RAKE receivers extract the signal components of several echos separately, like the fingers of a garden rake [8].

[3]GSM: Global System for Mobile Communication

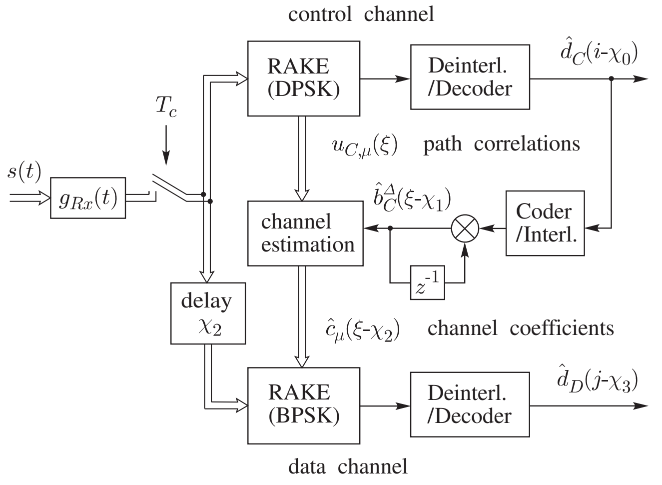


Fig. 3. Base station receiver

This realization is a straight forward adaption of the general stochastic model for a mobile radio channel by H. Schulze ([9],[5]) to the $L$-path Rayleigh case. $N_\mu$ denotes a "large" number of micro echo paths (e.g. $N_\mu = 50$). $\theta_{\mu,\nu}(\xi) \in [0, 2\pi]$ represent the phases at time $\xi$ which construct the "current" channel coefficient $c_\mu(\xi)$. The values $\Omega_{\mu,\nu} = 2\pi f_{d,\nu} T_{hC}$, where $T_{hC}$ denotes the control channel symbol duration, contain the Doppler frequencies $f_{d,\nu} \in [-f_{d_{\max}}, +f_{d_{\max}}]$ and therefore the influence of the time variance on the channel coefficient. The parameters $\theta_{\mu,\nu}(\xi)$ and $f_{d,\nu}$ are chosen by a Monte Carlo experiment. The maximum Doppler frequency $f_{d_{\max}}$ is determined by

$$f_{d_{\max}} = v \, \frac{f_0}{c_0} \quad , \tag{10}$$

where $v$ denotes the vehicle speed, $f_0$ the carrier frequency and $c_0$ the speed of light [8]. The probability density function (pdf) $p_{f_d}(f_d)$ of the Doppler frequencies is here assumed to be given by the classical Jakes spectrum [6].

Using (9) it follows after some easy algebra for the term $\tilde{c}_\mu(\xi)$ which has been introduced in (8):

$$\tilde{c}_\mu(\xi) = \sqrt{\frac{1}{L}} \frac{1}{\sqrt{N_\mu}} \sum_{\nu=0}^{N_\mu - 1} e^{j\theta_{\mu,\nu}(\xi)} \underbrace{\frac{\sin(\pi f_{d_{\mu,\nu}} T_{hC} M_c)}{M_c \sin(\pi f_{d_{\mu,\nu}} T_{hC})}}_{=: \varrho_{si}(f_{d_{\mu,\nu}} M_c)} \quad . \tag{11}$$

For a fixed time $\xi$ and a given $c_\mu(\xi)$ the Doppler frequencies remain as an independent stochastic process which determines the value $\tilde{c}_\mu(\xi)$. We introduce the mean value $\bar{\varrho}_{si}(M_c)$ of the process $\varrho_{si}(M_c)$. With the mentioned Jakes pdf for the Doppler frequencies it results from:

$$\bar{\varrho}_{si}(M_c) := \int_{-\infty}^{\infty} p_{f_d}(f_d) \, \varrho_{si}(f_d, M_c) \, df_d \tag{12}$$

$$= \frac{1}{\pi f_{d_{\max}}} \int_{-f_{d_{\max}}}^{f_{d_{\max}}} \frac{1}{\sqrt{1 - (f_d / f_{d_{\max}})^2}} \frac{\sin(\pi f_d T_{hC} M_c)}{M_c \sin(\pi f_d T_{hC})} \, df_d$$

A closed form solution for this integral does not exist, but it can be evaluated numerically. Using this mean value $\bar{\varrho}_{si}(M_c)$ the term $\tilde{c}_\mu(\xi)$ can be rewritten as

$$\tilde{c}_\mu(\xi) = \bar{\varrho}_{si}(M_c) \underbrace{\sqrt{\frac{1}{L}} \frac{1}{\sqrt{N_\mu}} \sum_{\nu=0}^{N_\mu - 1} e^{j\theta_{\mu,\nu}(\xi)}}_{= c_\mu(\xi)}$$
$$+ \underbrace{\frac{1}{\sqrt{L \, N_\mu}} \sum_{\nu=0}^{N_\mu - 1} e^{j\theta_{\mu,\nu}(\xi)} \left[ \varrho_{si}(f_{d_{\mu,\nu}}, M_c) - \bar{\varrho}_{si}(M_c) \right]}_{=: n_{C_2}(\xi)} \quad . \tag{13}$$

The first summand is given by the true channel coefficient and the deterministic factor $\bar{\varrho}_{si}(M_c)$, which is affected by the time variance of the channel. The noise value $n_{C_2}(\xi)$ is built by the superposition of a large number of terms. With the central limit theorem in mind, it approximates a gaussian process. Due to the introduced correction by the mean value $\bar{\varrho}_{si}(M_c)$, the remaining noise $n_{C_2}(\xi)$ has zero mean and minimum variance which is given by:

$$\sigma^2_{n_{C_2}} = \underbrace{\frac{1}{L} \frac{1}{N_\mu} \sum_{\nu=0}^{N_\mu - 1} E\{ | \, e^{j\theta_{\mu,\nu}(\xi)} \, |^2 \}}_{= 1/L}$$
$$\cdot \frac{1}{\pi f_{d_{\max}}} \int_{-f_{d_{\max}}}^{f_{d_{\max}}} \frac{1}{\sqrt{1 - (\frac{f_d}{f_{d_{\max}}})^2}} \left[ \varrho_{si}(f_d, M_c) - \bar{\varrho}_{si}(M_c) \right]^2 df_d \tag{14}$$

This noise is not the effect of interfering user signals, but the result of the time variance and the statistical behavior of the channel. The influence of other users on the channel estimation is given by the noise value $n_{C_1}(\xi)$. We assume that every user is received with the average signal power one, i.e. a power control has eliminated the long term (lognormal) fading. A commonly used simple approximation of the interference process is its characterization by white gaussian noise with variance equal to the power of the interference signal. Hence, the average noise variance of $N_U$ interfering user signals equals $N_U$. We focus on the detection of the $(N_U + 1)$-th user signal which will be denoted as "own signal". The received signal energy is distributed on $L$ paths. The $L - 1$ echo paths of the own control channel signal which interfere with the $\mu$-th control channel path, produce an additional noise variance of approximately $\frac{L-1}{L} \rho_C \, E_{ch}(\xi)$, where $E_{ch}(\xi)$ denotes the current channel energy

$$E_{ch}(\xi) = \sum_{\mu=0}^{L-1} |c_\mu(\xi)|^2 \quad . \tag{15}$$

The own data channel signal results in a noise component of $n_{C_1}(\xi)$ with variance $\rho_D \, E_{ch}(\xi)$. When taking into account the averaging process over $M_c$ control channel symbol periods, we obtain for the variance of $n_{C_1}(\xi)$:

$$\sigma^2_{n_{C_1}} = \frac{1}{M_c \; \rho_C \; E_{hC}} \cdot \Big( N_U + \big[\, \rho_D + \rho_C \, \tfrac{L-1}{L} \,\big] \underbrace{\frac{1}{M_c} \sum_{\ell=-\frac{M_c-1}{2}}^{\frac{M_c-1}{2}} E_{ch}(\xi+\ell)}_{\approx E_{ch}(\xi)} \Big) \quad (16)$$

## IV. Detection of the Data Channel

The received signal in the $\mu$-th path of the data channel RAKE receiver can be written as

$$\boldsymbol{s}_{D,\mu}(\eta) = b_D(\eta)\, c_\mu(\eta) \sqrt{\rho_D}\, \boldsymbol{h}_D(\eta) + \boldsymbol{r}_{D,\mu}(\eta) \quad (17)$$

In every RAKE path $\mu$ the correlations

$$u_{D,\mu}(\eta) := \frac{1}{\sqrt{\rho_D}\, E_{hD}} \boldsymbol{h}^*_D(\eta)\; \boldsymbol{s}_{D,\mu}(\eta) \quad (18)$$

$$= b_D(\eta)\, c_\mu(\eta) + \underbrace{\frac{1}{\sqrt{\rho_D}\, E_{hD}} \boldsymbol{h}^*_D(\eta)\, \boldsymbol{r}_{D,\mu}(\eta)}_{=:\, n_{D,\mu}(\eta)} \quad . \quad (19)$$

are carried out. Analogous to the discussion of the control channel it follows for the variance of the noise term $n_{D,\mu}(\eta)$

$$\sigma^2_{n_D} = \frac{1}{\rho_D\, E_{hD}} \Big( N_U + \big[\rho_C + \rho_D \tfrac{L-1}{L}\big]\; E_{ch}(\eta) \Big) \quad . \quad (20)$$

From the echo path correlations the decision variable $V_D(\eta)$ is constructed by maximum ratio combining:

$$V_D(\eta) = \mathrm{Re}\Big\{ \sum_{\mu=0}^{L-1} \hat{c}^*_\mu(\eta)\; u_{D,\mu}(\eta) \Big\} \quad (21)$$

These decision variables represent corrupted versions of the data channel code symbols $b_D(\eta)$. They are fed to the decoder which produce estimates of the transmitted data bits $d_D(j)$. An analysis including the decoding process is quite complex. So we restrict ourselves to calculating the symbol error rate before decoding. For this purpose we have to decide the soft decision variables $V_D(\eta)$ directly to code symbols (hard decision) and to compare them to the transmitted code symbols. In order to simplify notation we rewrite the decision variable $V_D(\eta)$ like

$$V_D(\eta) = \frac{1}{2} \Big( \sum_{\mu=0}^{L-1} X^*_\mu\, Y_\mu + X_\mu\, Y^*_\mu \Big) \quad (22)$$

where $X_\mu$ represent the channel estimation values $c_\mu$ and $Y_\mu$ represent the data channel correlations. As previously discussed $X_\mu$ and $Y_\mu$ can be interpreted as gaussian random variables with means

$$\bar{X}_\mu = \bar{\varrho}_{si}(M_c)\, c_\mu(\eta) \qquad \bar{Y}_\mu = c_\mu(\eta) \quad (23)$$

and variances

$$\sigma^2_X = \sigma^2_{n_C} = \sigma^2_{n_{C_1}} + \sigma^2_{n_{C_2}} \qquad \sigma^2_Y = \sigma^2_{n_D} \quad (24)$$

The probability of a symbol error for the quadratic form in (22) can be calculated with the help of appendix 4B.1 of Proakis [8]. The result is

$$P(E_{ch}(\eta), N_U) = Q(a,b) - \frac{1}{2}\, I_0(ab)\, \exp\Big(-\frac{a^2+b^2}{2}\Big) + \frac{\exp\big[-(a^2+b^2)/2\big]}{2^{2L-1}} \cdot \sum_{n=1}^{L-1} I_n(ab) \sum_{\mu=0}^{L-1-n} \binom{2L-1}{\mu} \Big[ \Big(\frac{b}{a}\Big)^n - \Big(\frac{a}{b}\Big)^n \Big] \;,\; L>1 \quad (25)$$

$$P(E_{ch}(\eta), N_U) = Q(a,b) - \frac{1}{2}\, I_0(ab)\, \exp\Big(-\frac{a^2+b^2}{2}\Big) \;,\; L=1 \quad (26)$$

with

$$a^2 = \frac{\big(\, \bar{\varrho}_{si}(M_c)\, \sigma_{n_D} - \sigma_{n_C} \big)^2}{\sigma^2_{n_D}\, \sigma^2_{n_C}} \frac{E_{ch}(\eta)}{2} \quad (27)$$

$$b^2 = \frac{\big(\, \bar{\varrho}_{si}(M_c)\, \sigma_{n_D} + \sigma_{n_C} \big)^2}{\sigma^2_{n_D}\, \sigma^2_{n_C}} \frac{E_{ch}(\eta)}{2} \quad . \quad (28)$$

The $Q$-function is given by

$$Q(a,b) = e^{-\frac{a^2+b^2}{2}} \sum_{n=0}^{\infty} \Big(\frac{a}{b}\Big)^n I_n(ab) \quad , \quad (29)$$

where $I_n(x)$ is the $n$-th-order modified Bessel function of the first kind. The expression $P(E_{ch}(\eta), N_U)$ depends on the instantaneous value of the channel energy $E_{ch}(\eta)$ and the variance $N_U$ of the noise (other users). The average symbol error probability under the influence of the channel statistics of an $L$-path Rayleigh channel results from

$$P_s(\bar{E}_{ch}(\eta), N_U) = \int_0^\infty P(E_{ch}, N_U)\, p(E_{ch})\, dE_{ch} \quad , \quad (30)$$

where the pdf of the channel energy is given by a chi-square distribution with $2L$ degrees of freedom

$$p(E_{ch}) = \frac{1}{(L-1)!\, (\bar{E}_{ch}/L)^L}\; E^{L-1}_{ch}\; e^{-E_{ch}/(\bar{E}_{ch}/L)} \quad . \quad (31)$$

The integral in equation (30) can easily be evaluated numerically.

## V. Illustrations

In this section we will present some results for the data channel symbol error rate of the described CDMA uplink with the parameter settings specified in Table 1.

TABLE I
SETUP OF THE DS-CDMA SYSTEM

| | |
|---|---|
| carrier frequency: | $f_0 = 950\,\text{MHz}$ |
| chip rate: | $f_c = 1.2288\,\text{MHz}$ |
| data channel: | |
| - modulation: | BPSK |
| - detection: | coherent (Maximum Ratio Combining) |
| - spreading factor: | $S_{P,D} = 43$ |
| control channel: | |
| - modulation: | DBPSK |
| - detection: | differentially coherent |
| - spreading factor: | $S_{P,C} = 258$ |

For urban and hilly environments the delay spread of the channel impulse response is typically in the order of $2\ldots5\ \mu s$ [4]. With respect to the chip duration of $0.81\ \mu s$ (1.2288 MHz) we can expect that a number of 4 RAKE paths is a good choice to catch the bulk of the channel energy. Due to this a Rayleigh channel with 4 paths is considered. Its average total channel energy is assumed to be equal to one. Every user signal (data + control channel signal) is assumed to contribute the average noise power one. E.g. for $N_U = 20$ interfering user signals the average SNR is $\bar{E}_s/N_0 = S_{P,D}/N_U = 43/20 = 3.3\,\text{dB}$. It should be pointed out that $\bar{E}_s$ contains both, the energy of a data channel symbol and an associated energy portion of a control channel symbol. First we will compare the closed-form expression for the symbol error rate by means of equation (30) with simulation results. **Fig. 4** shows the data symbol error rate ($SER_D$) with variation of the parameters $\rho_D$ (relative power of the data channel) and $M_c$ (channel estimation over $M_c$ control channel symbols).

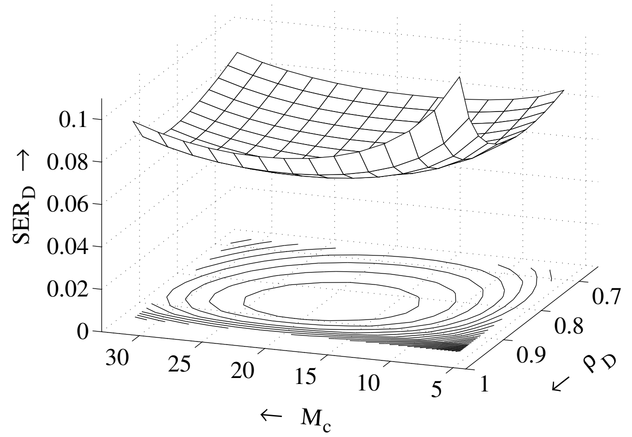


Fig. 4. $SER_D$ closed-form analysis, 4-path Rayleigh fading channel ($N_U = 20$, $v = 120\,\text{km/h}$)

In **Fig. 5** the relative error of the simulated and the calculated data channel symbol error rate is illustrated for isolated points in the $M_c, \rho_D$ plane. The relative error is bounded to 1.5%, which underlines the efficiency of the analytic approach.

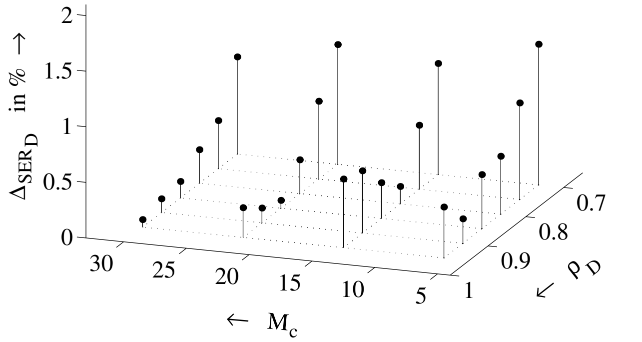


Fig. 5. Relative error for the analytical calculation of $\text{SER}_D$ with respect to simulation results; 4-path Rayleigh fading channel ($N_U = 20$, $v = 120\,\text{km/h}$)

If we again have a look at Fig. 4, we recognize a minimum of the symbol error rate for one pair of $M_c$ and $\rho_D$. The optimum choice for these parameters depends on the $\bar{E}_s/N_0$ and on the vehicle speed as depicted in **Fig. 6**. The quantitative interpretation of this figure might be hard, but a qualitative conclusion can be drawn easily: The SER depends very sensitively on the parameter $M_c$, especially with increasing vehicle speed, whereas the choice of $\rho_D$ is less critical.

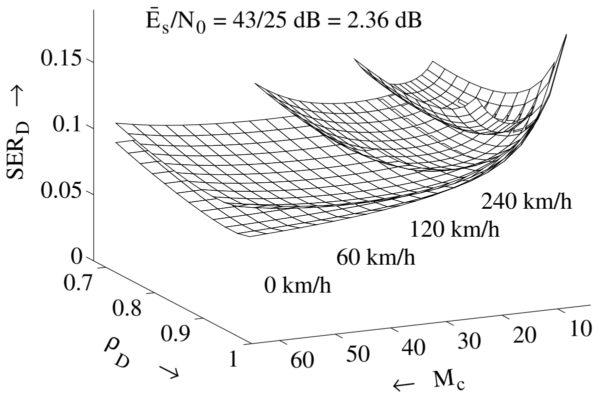


Fig. 6. $SER_D$ for a 4-path Rayleigh channel with different speeds ($N_U = 25$)

## VI. Minimum symbol error rate with channel estimation

We have used a numeric search algorithm to find the optimum combinations $(M_c, \rho_D)$ and the related data channel symbol error rates for the system setup specified in Table 1. In **Fig. 7** the results are depicted for a 4-path Rayleigh channel with $\bar{E}_s/N_0$ and the vehicle speed as parameters. The relative cost of channel estimation can be expressed by the ratio $\rho_C/(\rho_C + \rho_D) = \rho_C = 1 - \rho_D$ which results in approximately $0.5\ldots1.0$ dB (see Fig. 7 for values $\rho_D$) in the presented scenario. The related data channel symbol error rates are shown in **Fig. 8**. It can be found that the SER curves for BPSK with channel estimation degrade by approximately by $1\ldots2\,\text{dB}$ with respect to the ideal BPSK performance (known channel). This excess loss of

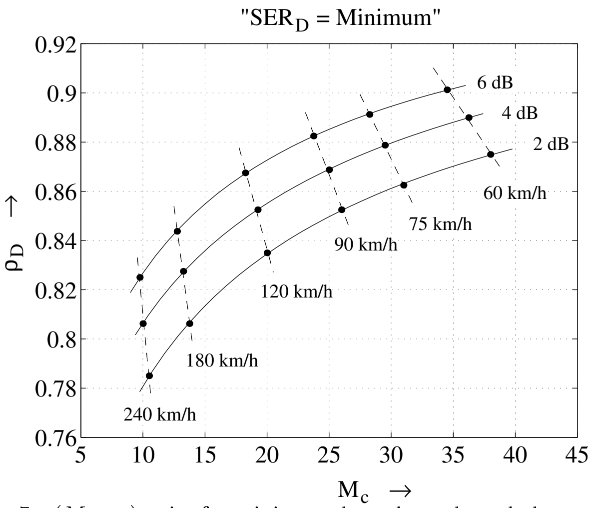


Fig. 7. ($M_c$,$\rho_D$) pairs for minimum data channel symbol error rates presented with $\bar{E}_s/N_0$ or the vehicle speed as parameter for a 4-path Rayleigh channel

1 . . . 1.5 dB compared to the relative cost of channel estimation given by $\rho_C/(\rho_C+\rho_D)$ is due to the fact that the channel estimation is not perfect. In addition, some calculated bit error rates are included for $M$-ary orthogonal modulation with square law detection. This modulation scheme is used in the uplink of the well-known QUALCOMM-CDMA system [7],[11]. It is very interesting to observe that the

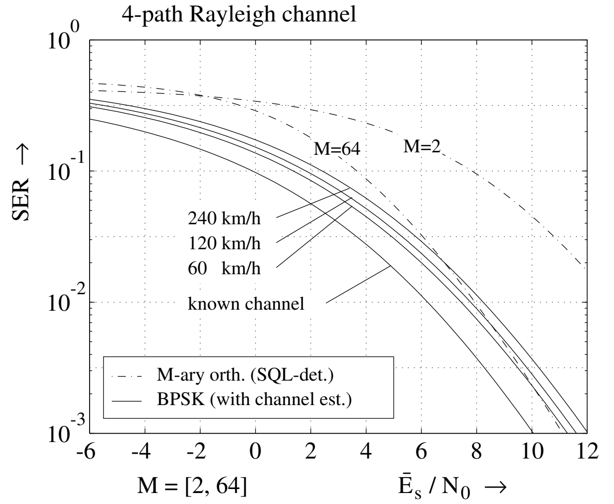


Fig. 8. Data channel symbol error rates for BPSK in correspondence to the optimum choice of parameters $M_c$ and $\rho_D$.
Symbol error rates for 64-ary orthogonal modulation with square law detection. (In the context of $M$-ary modulation ld$M$ data symbols correspond to one modulation symbol. The term symbol error rate is used for the data symbol error rate throughout this paper).

noncoherent detection of the 64-ary orthogonal modulation beats the performance of the coherent system for moderate high values of $\bar{E}_s/N_0$.

## VII. CONCLUSIONS

We have analyzed the uplink of a coherent CDMA-system with separate control- and data channels for each user. The derived closed-form expression for the symbol error rate in the data channel has been verified successfully in comparison to simulation results for an $L$-path Rayleigh channel. The method is open to extensions, e.g. modifications of the channel estimation procedure or more complex channels (non constant delay profiles and other Doppler spectra). The analysis allows to predict the system's performance and to optimize the parameters.

Furthermore it helps to compare different system designs with respect to their performance in a specified environment. We presented a comparison of the discussed coherent CDMA-system to a noncoherent system which is based on $M$-ary orthogonal modulation. We found that the noncoherent system design is able to beat the performance of the coherent system for data channel symbol error rates in the region of $10^{-2}$ to $10^{-3}$. This region becomes interesting for high quality data links with bit error rates in the order of $10^{-6}$ to $10^{-9}$ after decoding. For speech services symbol error rates of $10^{-1}$ are sufficient which qualifies the coherent system for these services.

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