Chip Tap Model of a Mobile Radio Link for CDMA-Applications

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Abstract

In this paper an efficient model for a mobile radio link is presented and analized. Some basic phenomena of a multipath propagation situation are exploited sucessfully, while using a statistical based channel description. The resulting model includes correlations of echo pathes due to filtering with transmit and receive filters.

Because of the high system bandwidth of a CDMA system with respect to low Doppler frequencies, it is possible to introduce some simplifications to the channel model. A whole transmission line with transmitter, channel model and coherent or noncoherent receiver structure has been simulated and the bit error behavior has been compared to some theoretical results.

The chip tap model of a mobile radio link approves as powerfull tool for simulation purposes.

1 Introduction

A mobile data transmission line with respect to direct sequence CDMA application shall be analyzed. The background for our investigations is the future standard UMTS (Universal Mobile Telecommunications Services) for mobile speech, video and data transmission, which will be set up in the turn of the century.
Due to the spreading mechanism of a DS-CDMA system, the necessary system clock rate is usually very high compared with the data rate. The Qualcomm CDMA System e.g. uses a chip frequency of $f_c = 1.2288\,MHz$, while the data rate is $9.6\,kbit/s$. The resulting high bandwidth makes necessary an adequate high sampling rate. The computational effort can be limited, when no oversampling is used. In the following a chip tap model of the time variant radio channel will be developed.

2 Channel Modeling

To describe the time varying channel we use a statistical approach. The assumption of a GWSSUS process (gaussian wide sense stationary uncorrelated scattering) for the channel impulse response $c_{RL}(\tau, t)$ is widely accepted. The stationarity supposition is seen to be valid, when only a local receive area is taken into consideration. The uncorrelated scattering assumption can be understood as a consequence of the actual high carrier frequencies. Two pathes which arrive at the receive antenna with a relative delay of a few centimeters will cause complete uncertainty of their mutual phase relations, so that they can be treated as uncorrelated signals. it can be shown, that this assumptions lead to the following relation [Pro89].

$$E\{\, c_{RL}(\tau, t)^* \, c_{RL}(\tau', t') \,\} \;=\; \phi_c(\tau, \Delta t)\, \delta(\tau - \tau') \tag{1}$$

The fourier transformation of $\phi_c(\tau, \Delta t)$ leads to the scattering function $P_s(\tau, f_d)$. Due to equation 1 the scattering function represents a complete channel description.

$$P_s(\tau, f_d) \;=\; \int_{-\infty}^{\infty} \phi_c(\tau, \Delta t)\, e^{-j\,2\pi\, f_d\, \Delta t} d\Delta t \tag{2}$$

It indicates the dispersion of the channel energy in the direction of time (delay spread) and frequency (Doppler spread) and can be interpreted directly as two dimensional probability density function [Hoe92]. It can be separated in two product terms, while assuming independent processes for the Doppler shift and the path delays. We obtain the delay power spectrum and the Doppler power spectrum:

$$P_\tau(\tau) \;=\; \int_{-\infty}^{\infty} P_s(\tau, f_d)\, df_d \tag{3}$$

$$P_f(f_d) \;=\; \int_{-\infty}^{\infty} P_s(\tau, f_d)\, d\tau \tag{4}$$

The working group COST 207 has defined some profiles for typical channels. The power delay spectrum should be modeled as exponential decay. The appropriate probability density function is of the type:

$$p(\tau) = \begin{cases} \frac{\beta}{a}\, e^{-\beta\tau} & : \quad 0 \le \tau \le \tau_m \\ 0 & : \quad \text{otherwise} \end{cases} \tag{5}$$

β denotes the decay constant, τ_m is the maximum delay time and the value a is given by

$$a \;=\; 1 - e^{-\beta\tau} \quad . \tag{6}$$

May be that more than one of such clusters have to be considered. The BAD URBAN channel e.g. contains two echo clusters as depicted in figure 1.

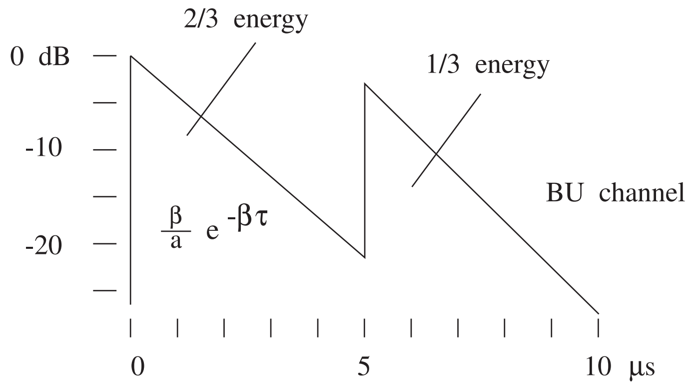

Figure 1: Delay power profile of the BAD URBAN channel

The spread of energy in the frequency domain is caused by the movement of the vehicle, or to be more precise by the involved Doppler effect (see [Kam92]). The maximum Doppler frequency is determined by the relation of vehicle speed and carrier frequency.

$$f_{d\,max} = v\,\frac{f_0}{c_0} \quad , \tag{7}$$

c_0 denotes the speed of light.
The Qualcomm CDMA system uses carrier frequencies in the range of $900\,MHz$, so that a vehicle speed of around $50\,km/h$ results in a Doppler frequency of $f_d = 41\,Hz$. The appropriate power spectrum has the characteristic form of a "bath-tub" with the maximum Doppler frequency as the limit frequency. It is called Jakes spectrum. For simulation purposes it can be modeled in some different ways.

a) Filtering of a white gaussian process with filter $H(j\omega)$

b) Superposition of many complex exponential oscillations

   (i) Amplitude distribution $\sim$ Doppler spectrum (Jakes, [Jak74])

   (ii) Frequency distribution $\sim$ Doppler spectrum (Schulze, [Sch89])

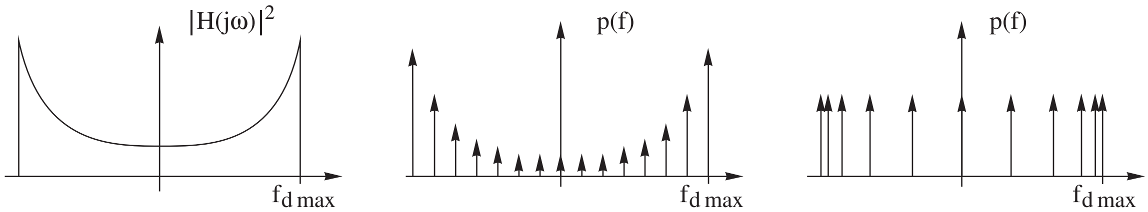

Figure 2: Different ways to model the Jakes spektrum

Figure 2 illustrates these possibilities. Due to the high chip rate the numerical effort for the filter approach becomes enormous. It can be kept to a realistic order when a multirate implementation is taken into consideration.

The second and third proposal start at a complete different point of view. By accumulation of an adequate number of identical distributed processes a gaussian process is approximated. The third approach follows this philosophy most consequent. Furthermore ist can be interpreted graphically. Every signal path from each direction is built from an individual exponential oscillation with fixed magnitude in this model. This fact is related directly to the physical phenomena of the receive situation. Figure 3 shows this interdependence. To construct an actual realization of the statistical process, the

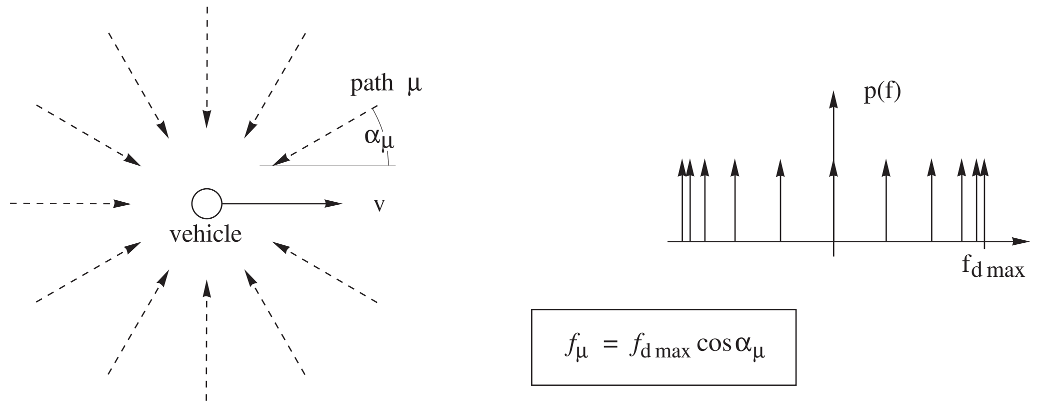

Figure 3: Physical receive situation and probability density function

related pathes are set up by a Monte Carlo experiment during the initialization phase of the simulation. Both, Doppler frequencies $f_{d,\nu}$ and path delays τ_ν are concerned. An efficient realization of this parameter generation is obtained by nonlinear transformations of a uniformly distributed process $x \in [0,1]$ to the desired density functions.
The principle is easy to motivate. Assumed is a receive situation where all angles of incidence are of equal probability. The appropriate Doppler frequencies should be calculated from the relation

$$f_\nu \; = \; f_{dmax} \, \cos(2\pi \, x_\nu) \quad , \tag{8}$$

where x_ν indicates an actual value of the uniformly distributed process x. The echo delays can be calculated with a similar expression, based on new values of the process x.

$$\tau_\nu = -\frac{1}{\beta} \ln(1 - a \, x_{\nu+N_{ges}}) \tag{9}$$

Furthermore some initial phases should be set up randomly by

$$\psi_\nu = x_{\nu+2\,N_{ges}} \, 2\pi \quad , \tag{10}$$

again using new values of the process x. The baseband impulse response of the channel model results in

$$c_{RL}(\tau, t) \; = \; \frac{1}{\sqrt{N_{ges}}} \sum_{\nu=1}^{N_{ges}} e^{\omega_\nu t + \psi_\nu} \, \delta(\tau - \tau_\nu) \tag{11}$$

The factor $\frac{1}{\sqrt{N_{ges}}}$ guarantees an average gain of one for the channel energy. If more than one cluster of echo delay times has to be considered, the fraction of channel enrgy of each cluster determines the belonging fraction of pathes with respect to the total number of pathes, which have to be related to the respective cluster. For the BAD URBAN channel 2/3 of the pathes are set up in the first cluster and 1/3 are set up in the second one.

As we pointed out, the chip frequency is reasonable high with respect to the data rate. For simulation purposes the sampling rate has to be chosen even higher to fulfil the Nyquist theorem. On the other hand a properly designed data transmission system should have a symbol period which is much shorter than the Doppler period, so that the time variance becomes weak. In this situation a chip tap model contains a sufficient statistic of the transmitted data signal [For72].
Therefore we want to develope a complete chip tap model of the whole transmission line, which is depicted in figure 4.

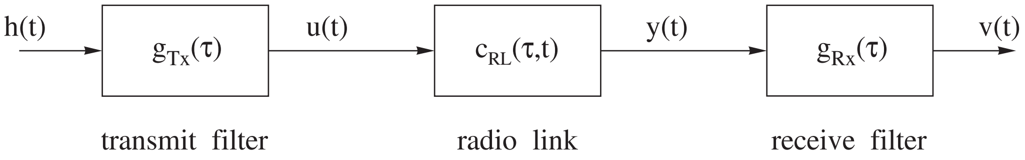

Figure 4: Transmission line with transmit and receive filters

The over all channel impulse response $c(\tau, t)$ is given by

$$c(\tau, t) \;=\; [g_{Tx}(\tau) * c_{RL}(\tau, t)] \;*\; g_{Rx}(\tau) \tag{12}$$

Due to the time variant behaviour of $c_{RL}(\tau, t)$ the setting of the parenthesis becomes obligatory. But it can be shown that for $f_c \gg \frac{1}{T_d}$ – or equivalent for short transmitter filters compared to the Doppler period – the 'Doppler modulation' of the transmit filter can be neglected. With this simplification which is usually valid for CDMA applications, we have

$$c(\tau, t) \;=\; c_{RL}(\tau, t) * [g_{Tx}(\tau) * g_{Rx}(\tau)] \tag{13}$$

The resulting filter after convolution of transmit and receive filter will be denoted as $g_{TR}(t)$. This yields the following band limited channel impulse response $c(\tau, t)$.

$$c(\tau, t) \;=\; \frac{1}{\sqrt{N_{ges}}} \sum_{\nu=1}^{N_{ges}} e^{\omega_\nu t + \psi_\nu}\, g_{TR}(\tau - \tau_\nu) \tag{14}$$

Sampling of this impulse response with $\tau = \kappa\, T_c$ and $t = k\, T_c$ results in

$$c(\kappa, k) \;=\; \frac{1}{\sqrt{N_{ges}}} \sum_{\nu=1}^{N_{ges}} e^{\Omega_\nu k + \psi_\nu}\; g_{TR}(\kappa\, T_c - \tau_\nu) \quad . \tag{15}$$

In this representation any path delays are allowed. In particular they needn't to be equispaced in the time domain.
Although some approximations have been introduced, the essential demands are satisfied successfully with this model. So correlations of the echo pathes due to the filtering are

taken in account (the uncorrelated scattering assumption is only valid for the not band limited channel model). Furthermore a change of the channel – as necessary for frequency hopping – can be done easily without any simulation delay and without any reasonable extra computational effort.

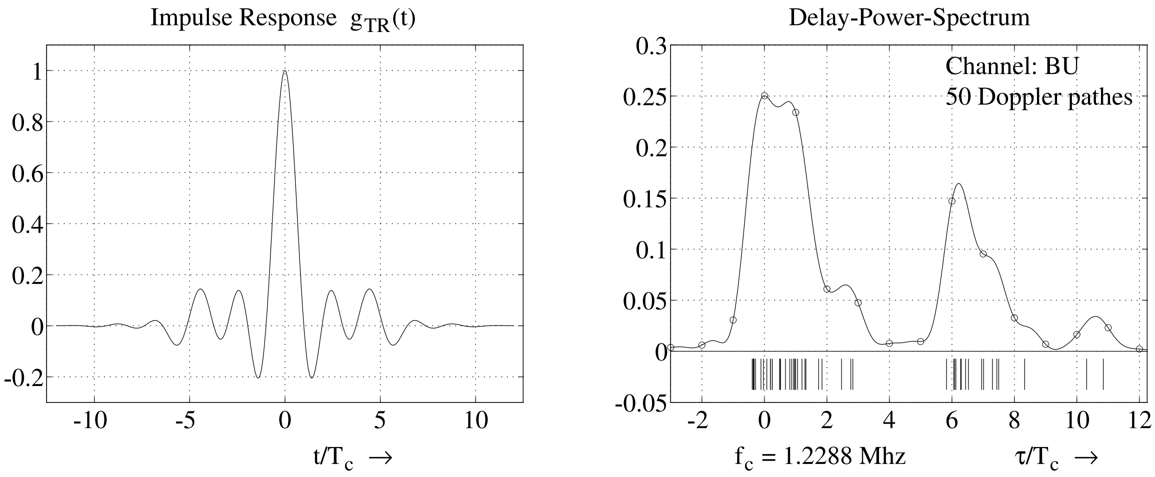

Figure 5: Filter response and delay power spectrum

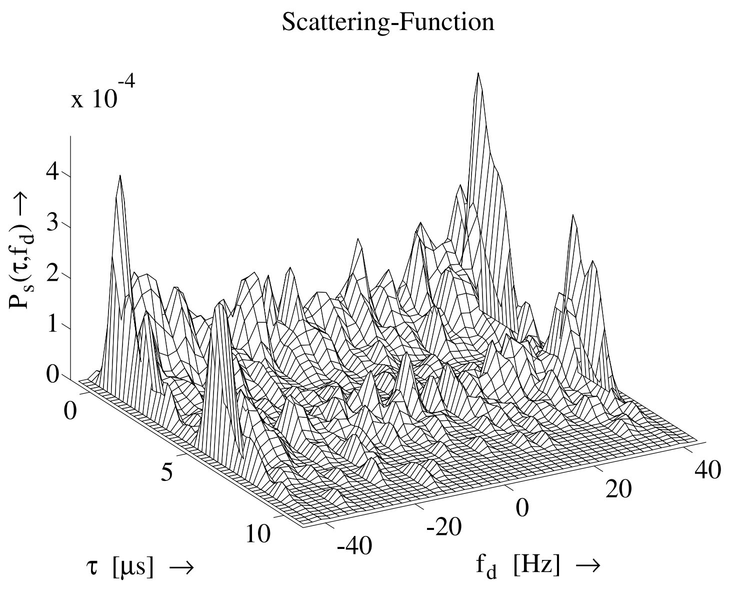

Figure 6: Scattering function

Figure 5 shows on the left hand side an arbitrary linear phase, equal ripple, filter $g_{TR}(t)$ which has been designed with steep flanks in the transition area by a Remez approach. On the right side of the figure the delay power spectrum of an actual realization of a BAD URBAN channel with respect to COST 207 is presented. $N_{ges} = 50$ echo pathes have

been used. The vertical lines below the delay power spectrum indicates in a symbolic manner the echo times, on which the echo delays τ_ν have been set to. The circles mark some values on the chip tap grid. The time axis has been shifted arbitrarily, so that its orgin is aligned to the maximum value of the delay power spectrum.
Figure 6 illustrates the scattering function of the channel model, which has been obtained by averaging 20 Monte Carlo experiments, each with $N_{ges} = 50$ echo pathes.

The presented channel model has been implemented as C program in the environment of data stream driven simulator (COSSAP). For each echo time the appropriate filter coefficients, which talk over to the determined chip tap grid, are calculated off line. During the simulation only this precalculated coefficients are used. By means of this the channel impulse response is generated directly at the chip frequency as illustrated in figure 7. The

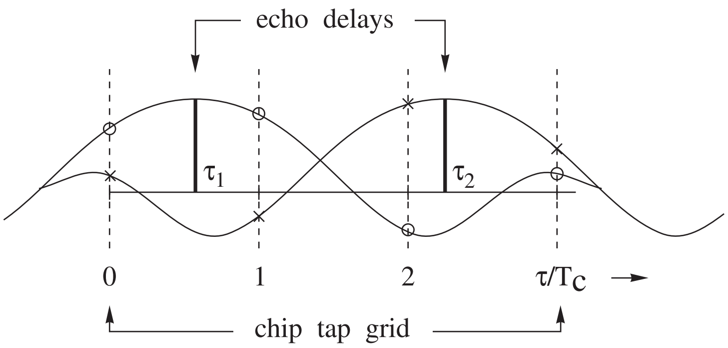

Figure 7: Generation of the chip tap baseband impulse response

Doppler period is very long compared to the chip period, or equivalent the channel is slow time variant. Because of this it is not necessary to construct the channel impulse response completely at every time k. An interpolation in direction of the time k as depicted in figure 8, reduces the numerical effort significantly.
The following example is based on a chip frequency of $f_c = 1.2288\,MHz$ and a maximum Doppler shift of $41\,Hz$. We assume 96 supporting values within one Doppler period, at which the channel impulse response has to be evaluated completely. The channel coefficients between these correct ones are obtained from a quadratic interpolation. By this the frequency of complete channel evaluations can be reduced to the discrete times $t = \Delta k\, T_c$, where Δk may be given by:

$$\Delta k \;=\; \frac{\text{Doppler period}}{96\,\text{supporting values}\cdot T_c} \;=\; \frac{1.2288\,MHz}{96\cdot 41\,Hz} \;=\; 312 \qquad (16)$$

It was found, that the difference of the output signals of a non interpolating channel module compared to the interpolating one remains in the region of 10^{-4}, where of course the number of supporting values (here 96) can be used to fix the trade off between accurancy and computational effort.

3 Simulation of a CDMA transmission line with coherent Detection

The combined structure of transmitter and channel, including the receive filter, is shown in figure 9 as chip tap model. The filter length of L_e taps results from the maximum echo delay T_m of the mobile channel and the impulse duration $T_{g_{TR}}$ of the filter $g_{TR}(t)$.

$$L_e = \left\lfloor \frac{T_m + T_{g_{TR}}}{T_c} \right\rfloor \tag{17}$$

With respect to the computational effort a short impulse response $g_{TR}(t)$ should be aspired.

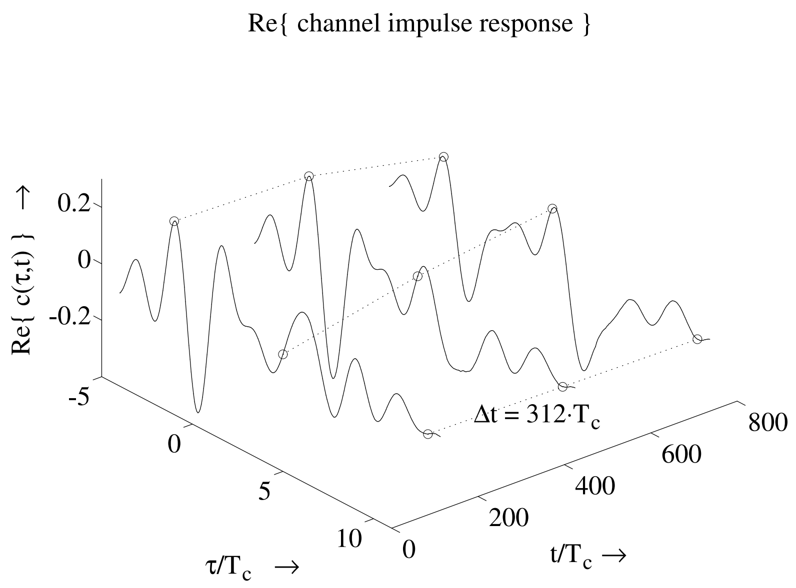

Figure 8: Interpolation of channel impulse responses

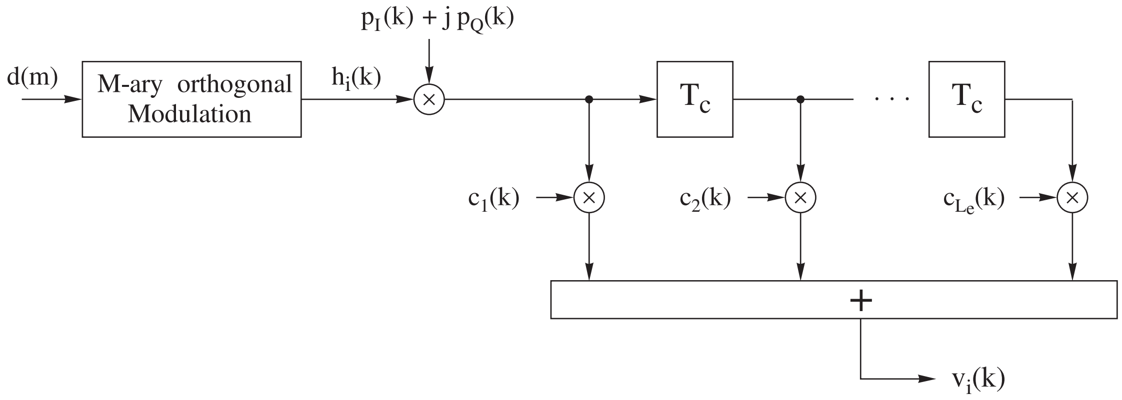

Figure 9: Baseband model of transmitter and channel

With respect to the Qualcomm CDMA proposal we also use a 64 ary orthogonal Modulation with Walsh functions. A transmitted symbol is chosen by $K = 6$ bit. The pseudo noise sequences $p_I(k),\ p_Q(k) \in \{-1, 1\}$ are approximately orthogonal and specific to the users. We combine them to one complex PN sequence $p_{IQ}(k) = p_I(k) + j\, p_Q(k)$. In every symbol interval another section of the sequence is valid. The 'quadrature spreading' of the transmitted signal with the complex PN sequence produces 256 Chips per Walsh symbol. The processing gain equals therefore $G_p = 256/6 \approx 43$.

Figure 10 demonstrates the system performance with coherent detection without channel coding. The solid line represents the theoretical bit error rate of an AWGN channel. The abscissa designates the white noise power, which corresponds to a number of interfering users. The circles point out some simulation results when using the time variant channel BAD URBAN with respect to COST 207. The energy of 11 significant chip tap echos has been corrected, so that their sum equals the value one during the simulation. The correct complete channel energy has been recognized to deviate only slightly from this value for a properly designed filter $g_{TR}(t)$. So this method describe ideal power control. The crosses illustrates the simulation results for a time invariant 8 tap channel. Each of the pathes carries 1/8 of the complete channel energy. In all cases the receiver has been supplied with the correct channel coefficients (ideal channel estimation). The performance of the transmission system degrades only little due to the multipath propagation.

An M-ary orthogonal modulation enables a superior performance with respect to the necessary E_b/N_0 in the case of an AWGN channel. In case of multipath propagation the orthogonality of time shifted versions of the transmitted symbols generally will fail. The scrambling of the transmitted signals with the PN sequences can maintain this orthogo-

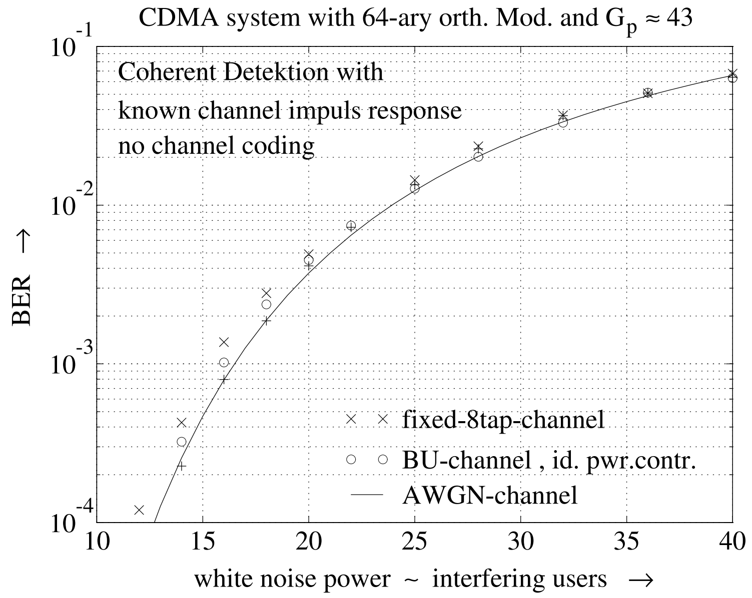

Figure 10: BER performance with coherent detection

nality demand approximately, as can be seen from the simulation results. By means of this the advantage of an M-ary orthogonal modulation can be exploited successfully.
It is interesting to recognize the better performance of the BU channel compared to the fixed 8 tap channel, although the BU channel is represented by 11 taps. The 8 tap channel carries the same energy level on each tap. Because of the better BER performance, we must conclude that the energy of the BU channel is concentrated to less than 8 taps in the average, so that the degradation due to the multipath propagation becomes smaller.

4 Noncoherent Detection

We look at the same transmission structure as in the last section. The noncoherent receiver performs correlations on the significant chip tap echos with all possibly transmitted symbols. These correlation results are square law combined to the detection variables. The coherent structure is known to exploit any signal component, which carries energy, to improve the BER performance in the case of ideal channel estimation. Using noncoherent detection, energy is collected with any path, but on the other hand the combining loss will increase [Pro89].
We have chosen the following strategy to select adequate pathes. The path with maximum path strength is selected first. Any other path is only taken into consideration, if its path strength overcomes a relative threshold value of 20% with respect to the strongest detected path. We didn't recognize a sensitive dependency on the value of the threshold, which makes this philosophy very general.
Alternatively we considered a receiver which uses a fixed number of strongest pathes.

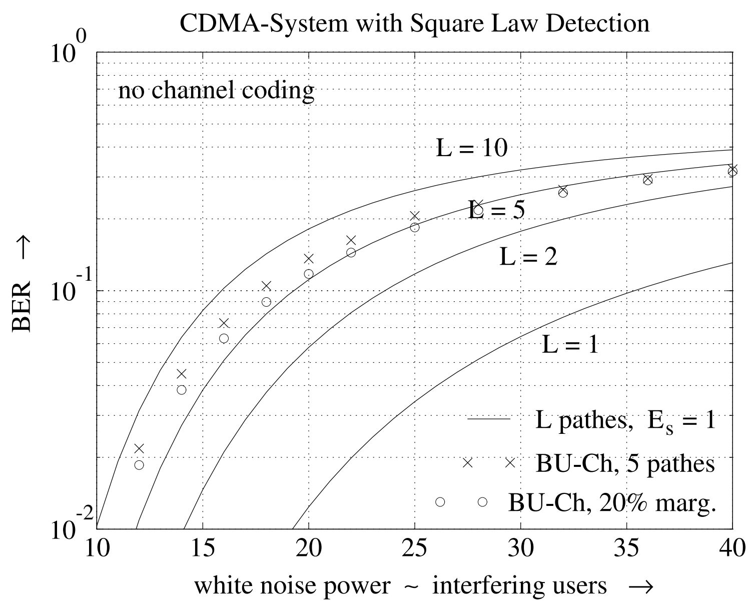

Figure 11: BER performance with noncoherent detection

With this strategy a number of 5 pathes leads to the best bit error rate with the BU channel. We didn't recognize significant changes, when varying the number of pathes. With both strategies only a portion less than one of the channel energy can be exploited for the detection. The simulation results for the last described method are represented by the crosses in the figure 11.

For comparisation we have also inserted some theoretical bit error rates for square law detection. The assumption for this L-path curves is that the considered L pathes carry the complete channel energy. The noise is considered with variance one per interfering user.

The theoretical BER curve for 5 pathes is comparable to the simulated one. The little difference is easy to understand, if we remember the fact that the theoretical analysis exploits the complete channel energy with the 5 pathes, while the simulated receiver can't follow this assumption.
The noncoherent receiver with threshold path selection has a similar performance. We can assume that this result represents a kind of optimal curve for the BU channel, when no channel coding is done. Nonideal power control and ignorance of the channel impulse response will impair the BER performance.

It can be summerized, that a chip tap model of the mobile radio link is an adequate tool for efficient simulations of some essential aspects of a CDMA transmission system.

Acknowledgement
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