

Enhanced Predictive Up/Down Power Control for CDMA Systems

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Abstract—In this paper we derive an enhanced power control algorithm, fitting into the up/down control scheme, as it is considered in the frequency division duplex (FDD) mode of the current 3GPP standard. Analysis of the classical up/down power control scheme unveils, that with increasing velocities the power control performance degrades, as the fixed step size power control is not able to track the channel fading properly. For the uplink we derive a nonlinear control algorithm generating the up/down power control commands accounting for the future of the channel fading process. Simulations show that this algorithm in combination with perfect future channel state information can partially mitigate the drawbacks of a fixed step-size up/down power control. A prerequisite for predictive power control is the acquisition of the future channel state information. In this paper we deduce a robust and adaptive structure for the prediction of the channel fading process in the context of a power controlled code division multiple access (CDMA) system based on least mean square (LMS) adaptation. Link level simulations show a signal to noise and interference ratio (SINR) gain in terms of the block error rate, enabling a decrease of the target SINR and thus leading to an enhanced spectral efficiency.

I. INTRODUCTION

Effective transmitter power control is essential for high-capacity cellular radio systems, to provide a satisfactory quality of service (QoS) and to cope with the near far problem concerning CDMA systems. The QoS is determined by the SINR at the receiver. Concerning the uplink of the FDD mode of UMTS (Universal Mobile Telecommunications System) [1], the transmit power of the mobiles is controlled by the base station aiming at providing each user with its required SINR. In UMTS, these control actions are implemented by sending commands from the base station to the mobile in order to increase or decrease its transmit power by a fixed amount. We will refer to this kind of power control as *up/down* power control. This power control scheme works linkwise independent as the power control of each link is only based on its own SINR.

Our analysis of the performance of the up/down power control in [2] shows that the power control error variance increases with increasing mobile velocities, as the up/down power control is no longer able to track channel variations due to its fixed stepsize and the limited loop bandwidth of the power control. Both motivate the consideration of future channel state information in the generation of the power control commands. We propose an optimized nonlinear power controller for the uplink taking into account the future channel state information.

A prerequisite for our new prediction based nonlinear controller is the availability of future channel state information. Therefore, we derive a robust and adaptive channel prediction algorithm in the context of a power controlled CDMA system. At this point it has to be noticed, that it will be difficult to

predict the effective channel process - including the transmit power - itself, as the Tx power is controlled, and therefore depends on the prediction result. We derive an algorithm that predicts the physical channel process by removing the effects of alterations in the Tx power. Our algorithm has shown to be robust with regard to non detectable transmit power command (TPC) errors and the Tx power limitation in the mobile. Both of them are not known to the base station.

A further requirement on the channel state prediction is its robustness and adaptivity towards varying channel statistics. Thus we use FIR filters that are adapted by LMS algorithms, leading to robust and adaptive channel state prediction in the context of a power controlled CDMA system.

Link level performance evaluations based on block error rates show a significant SINR gain, that will lead to decreased mean Tx powers and, thus, will enhance the spectral efficiency of the system.

There have already been some contributions on predictive power control [3], [4], [5], [6] all considering only one step future channel prediction. Furthermore, most of them assume the availability of the physical channel fading process - without the Tx power superposed - which is quite problematic, as realistic systems comprise TPC errors and transmit power constraints of the mobile terminals.

The paper is organized as follows. After introducing the system model of the conventional closed loop power control in Section II, the drawbacks of the classical up/down power control algorithm will be pointed out motivating the advantage of prediction based power control. Then a nonlinear power controller based on future channel state information is derived and its performance gain with respect to the classical power control algorithm is shown in Section III. In Section IV channel prediction in the context of a power controlled CDMA system is studied, whereas Section V focuses on the application of the LMS algorithm for the prediction filter adaptation. Section VI shows the performance gain of the new power control algorithm based on link level simulations. Finally, Section VII concludes the paper.

A. Contributions

Our main contributions can be summarized as follows:

- Derivation of an optimized nonlinear controller using future channel states over multiple slots fitting into the 3GPP standard.
- Development of an adaptive prediction algorithm being robust concerning TPC errors and Tx power limitations of the mobile, and thus applicable to real system constraints.
- Performance evaluation of the enhanced algorithm on link level basis in a realistic scenario, enabling estimations of gains in spectral efficiency.

II. CONVENTIONAL CLOSED LOOP POWER CONTROL AND ITS LIMITATIONS

At first we analyze the behavior of the conventional closed loop power control (CLPC) as it is shown in Fig. 1. We consider a CDMA uplink with N users sharing the same physical channel. Using a flat fading model and considering the observation interval to be infinite, the received signal is given by

$$r(t) = \sum_{n=1}^N \sum_{j=-\infty}^{\infty} \sqrt{p_n(t)} b_{n,j} c_{n,j} s(t - jT_c - \tau_n) e^{j\phi_n} + n(t). \quad (1)$$

In this equation

- T_c is the chip duration
- $c_{n,j} \in \{-1, +1\}$ with equal probability is the value of the j th chip of the n th user
- $b_{n,j} \in \{-1, +1\}$ with equal probability represents the value of the bit containing the j th chip $c_{n,j}$. It takes on the same value for M_n successive chips where M_n is the spreading factor of user n .
- $s(t)$ is the chip waveform, which is assumed to be equal for all users, with power $\sigma_s^2 = 1$.
- ϕ_n is the phase offset and τ_n the propagation delay of the signal of user n .
- $p_n(t)$ is the received power of user n which is equal to $a_n(t) \cdot x_n(t)$, if $a_n(t)$ is the link gain, resulting from Rayleigh fading, and $x_n(t)$ the mobile transmit power.
- $n(t)$ is a zero mean complex valued Gaussian noise process with the variance σ_n^2 .

Making the same assumptions as in [7], of uniformly distributed $\tau_k \in [0, T_c]$ and $\phi_n = 0$ ¹, this model yields the following slotwise equation for the SINR of user n , see [7],

$$\gamma_n(k) = \frac{p_n(k)M_n}{\sum_{m \neq n}^N p_m(k) + \sigma_n^2/2} = \frac{a_n(k)x_n(k)M_n}{\sum_{m \neq n}^N a_m(k)x_m(k) + \sigma_n^2/2}. \quad (2)$$

Note, the mobile transmit power is constant over one slot. Due to the circumstance that the power is updated in fixed stepsizes in the logarithmic domain, it is appropriate to apply a system model also in this domain. Otherwise in the linear domain the power update step would depend on the actual transmit power, and thus, the model would become even more complex. Transforming equation (2) into the logarithmic domain results in

$$\Gamma_n(k) = A_n(k) + X_n(k) - I_n(k) \quad (3)$$

where the capital letters denote values in dB. $I_n(k)$ is the interference and noise term for user n , which is equal to

$$I_n(k) = 10 \log_{10} \left(\sum_{m \neq n}^N a_m(k)x_m(k) + \sigma_n^2/2 \right) - 10 \log_{10}(M_n). \quad (4)$$

Fig. 1 shows the conventional power control loop of a single user (for simplification the index n is omitted). At the base station the estimated received SINR $\hat{\Gamma}_n(k)$ is compared to the target SINR Γ_n^{tar} for each user. If $\hat{\Gamma}_n > \Gamma_n^{tar}$, the base station will command the mobile to reduce its power by d dB; if instead $\hat{\Gamma}_n < \Gamma_n^{tar}$ a command to increase the transmit power by d dB will be transmitted to the mobile. The target SINR Γ_n^{tar} is adapted very slowly by an outer loop to guarantee a certain QoS requirement.

¹in [7] it has been shown that this assumption has only little impact on the analysis

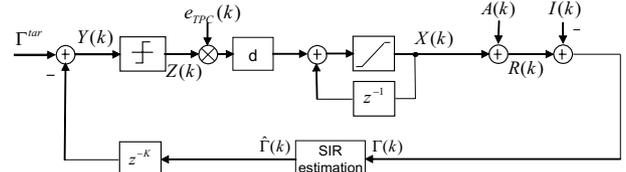


Fig. 1. Conventional closed loop power control

The power control commands transmitted via the downlink to the mobile terminal are corrupted by transmit power control command (TPC) errors $e_{TPC} \in \{-1, 1\}$, so that the mobile station might update its Tx power differently than decided by the base station. In addition, due to the limited transmit power, the mobile can not always react as it is decided by the base station. E.g. in the case the mobile transmits already with its maximal transmit power, a further power up command will not increase the transmit power.

In the next power control group (PCG) the mobile transmits with the adjusted transmit power X_n on the uplink. As we are using a logarithmical model, the received SINR is the result of adding the channel power process A_n to the Tx power and subtracting the interference noise power I_n . The process I_n is the interference noise power process seen by user n , depending on the received signal powers of the other users, the spreading gain of user n and the additive white Gaussian noise (AWGN) as given by equation (4).

From here on we assume a feedback delay of $K = 1$ PCG. Larger feedback delays can be compensated by the Smith predictor [8], that can be used in combination with our approach. It is important to distinguish between the aim of the Smith predictor and the aim of the prediction presented here. The Smith predictor can be used to combat an increased control error variance resulting from 2 slots delay or more. However, it can not speed up the reaction to external disturbances. Our approach is aiming to consider the future behavior of the channel fading process, which is external in this context.

The performance of conventional CLPC is limited for at least two reasons. In case of fast and deep fading of the wireless channel, it cannot track the channel due to the limited and fixed step size. According to [1] the step size d can be switched at a low rate between $\{1, 2, 3\}$ dB. In this work we will assume d being equal to 1 dB. Fig. 2 shows the resulting second order statistics of the control error process Y_n and the Tx power X_n in case of different velocities. Obviously in case of medium to high velocities the standard deviation of the control error increases.

Furthermore, the CLPC creates a noisy response known as *granular noise* when the fading is smooth or minimal, thus the power control error standard deviation does not become zero even in case of low velocities.

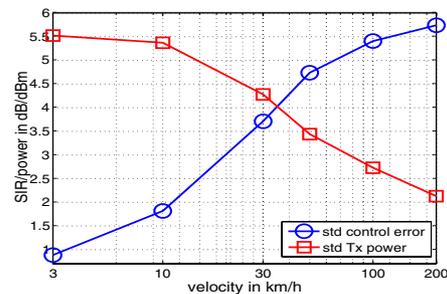


Fig. 2. Standard deviation of control error and Tx power vs. velocity for conventional CLPC, 1-path Rayleigh fading, no TPC errors

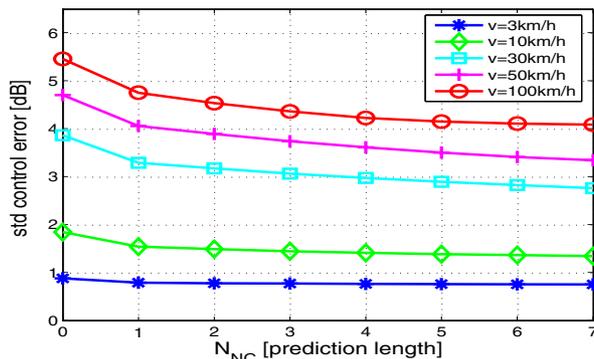


Fig. 3. Performance of the new prediction based power controller in case of perfect future channel knowledge for various velocities ($N_{NC} = 0$ corresponds to conventional CLPC)

III. ENHANCED NONLINEAR PREDICTION BASED POWER CONTROLLER

The characteristics of the conventional CLPC shown in Section II give the motivation to look for enhanced prediction based power control algorithms, still fitting into the 3GPP standard. Obviously, it would be possible to decrease the power control error in case of deep fades, if the power controller would start increasing the Tx power already in advance of a deep fade. The general idea is to decrease the control error variance by generating a TPC sequence accounting for the future channel fading states. A reasonable optimization criterion is to choose the transmit power commands minimizing the sum over the squared control errors over a predefined time horizon N_{NC} , i.e. to minimize the following expression

$$S = \sum_{l=1}^{N_{NC}} Y^2(k+l) = \sum_{l=1}^{N_{NC}} [Y(k) - (A(k+l) - A(k)) - (X(k+l) - X(k))]^2 \quad (5)$$

where the transmit power $X(k)$ can only change stepwise following the constraint

$$X(k+1) = X(k) + d \cdot Z(k) = X(k) \pm d. \quad (6)$$

In (5) the control error for each future time instant $k+l$ with $l \in \{1, \dots, N_{NC}\}$ is calculated by adding the change of the channel power A and the variation in transmit power X with respect to the current time instant. Here we make the assumption that the interference noise power is constant. This would hold in case of perfect power control. In case of a realistic power control the interference will be time varying. Nevertheless, our assumption is reasonable, as this variation can be assumed as relatively small due to despreading. The Tx power $X(k)$ can be chosen with respect to equation (6), so that considering a prediction horizon of N_{NC} time steps, the controller has to compare the metric S of $2^{N_{NC}}$ sequences \mathbf{Z} . Then the controller will choose the sequence $\mathbf{Z}_{opt} = \arg \min_{\mathbf{Z}} S(\mathbf{Z})$. The first entry $Z(k) = \mathbf{Z}_{opt}(1)$ is now transmitted to the mobile terminal. This procedure will be executed at each power control slot.

Fig. 3 shows the standard deviation of the control error for different velocities over the prediction horizon, based on the availability of perfect future channel knowledge. The case $N_{NC} = 0$ corresponds to conventional CLPC. Obviously the standard deviation of the control error can be significantly decreased by using prediction based power control. The gain rises for increasing velocities. For very low velocities no significant gain can be obtained as in this case conventional power control can track the channel variations relatively good.

IV. CHANNEL PREDICTION

As it has been shown in the previous section, future channel state information can significantly improve the power control performance. In this section we discuss the prediction of the channel fading process in the context of a power controlled system. Obviously, the channel fading power $A(k)$ is not directly accessible as the received signal power $R(k)$ is a superposition of the channel fading power and the transmit power $X(k)$. Furthermore, the base station does not exactly know the transmit power $X(k)$ of the mobile terminal, as the TPCs transmitted by the base station are corrupted by errors and the transmit power limitation of the mobile can hinder the mobile to increase its Tx power as commanded by the base station. Thus, we have to derive a channel fading state prediction accounting for these characteristics.

The first idea is to implement an algorithm for direct TPC error detection. However, analysis shows that the detection of TPC errors at a symbol SINR of 0dB for a power update stepsize $d = 1\text{dB}$ - reflecting a typical point of operation - is not feasible.

Lacking any possibility of detecting TPC errors, the most reasonable estimate of the Tx power at the base station $\hat{X}(k)$ is to assume that the mobile always reacts as commanded in the base station. Following this assumption we arrive at the power control system structure shown in Fig. 4. There the conventional power controller has been replaced by the new nonlinear prediction based power controller introduced in Section III. This controller requires the current control error $Y(k)$ and estimates of the current and the predicted channel powers $\hat{A}(k) \dots \hat{A}(k+N_{NC})$ - all in the logarithmical domain - as input signals.

The block *Mobile Sim* estimates the Tx power of the mobile terminal $\hat{X}(k)$, assuming that the mobile always reacts as commanded by the base station.

In Fig. 4 the channel is presented in the linear domain, thus the transmit power $X(k)$ is transferred to a linear amplitude $x(k)$. The channel fading process $a(k)$ is multiplied by $x(k)$ and then additive white Gaussian noise $n(k)$ and the interfering signals $i(k)$ are added, yielding the received signal $r(k)$. All signals are shown at slot rate, i.e. spreading, scrambling, modulation and also the averaging over the symbols of one power control slot is not presented here. Here we assume for the calculation of $r(k)$, that all symbols of previous slots are detected correctly and thus averaging is performed over all 10 symbols of the Dedicated Physical Control Channel (DPCCH) [1], whereas in the current power control slot only 4 pilot symbols are available for averaging.

The block *1st stage channel estimator* (Fig. 5) calculates a first estimate of the channel process, named $\check{\mathbf{U}}(k) = [\check{U}_0(k), \check{U}_1(k), \dots, \check{U}_{C-1}(k)]$ with

$$\check{U}_i(k) = \check{R}(k-i) - [\hat{X}(k-i) - \bar{X}(k)], \quad i = 0 \dots C-1, \quad (7)$$

where C is the length of the prediction filters and

$$\bar{X}(k) = \frac{1}{C} \sum_{l=0}^{C-1} \hat{X}(k-l) \quad (8)$$

is the mean of the Tx power over C slots. Due to averaging over 4 symbols in the current slot in contrast to 10 symbols in the preceding slots, the first entry of $\check{\mathbf{U}}(k)$ has a lower SINR than the other entries. The vector $\check{\mathbf{U}}(k)$ can

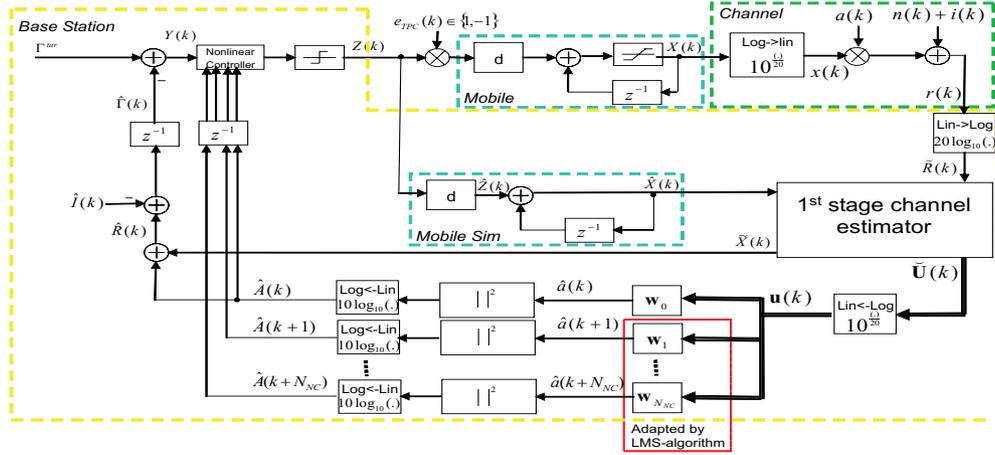


Fig. 4. Prediction based closed loop power control

be understood as a biased estimate of the complex channel process $a(k) \dots a(k-C+1)$ transferred into the logarithmical domain. The normalization of the Tx power with $\tilde{X}(k)$ is necessary in order to limit the dynamic range of the process $\tilde{\mathbf{U}}(k)$, and hence avoid fixed point implementation problems. An accumulating offset between $\hat{X}(k)$ and $X(k)$ due to TPC errors and the effect of the Tx power limiter might lead to an increasing dynamic range otherwise.

The second stage of channel estimation and prediction is performed by the following vector multiplication

$$\hat{a}(k+l) = \mathbf{w}_l \cdot \mathbf{u}(k), \quad l = 0 \dots N_{NC}, \quad (9)$$

where $\mathbf{w}_0 \dots \mathbf{w}_{N_{NC}}$ are the appropriate weighting vectors and $\mathbf{u}(k) = 10^{\frac{\tilde{\mathbf{U}}(k)}{20}}$. The coefficients $\mathbf{w}_0 \dots \mathbf{w}_{N_{NC}}$ can be precalculated by the Wiener-Hopf equation, if the channel statistics are available or can be adapted, like proposed in Section V. Then the absolute squares of these biased estimates and predictions of the complex channel weight process $\hat{a}(k+l)$ are used to calculate a biased estimate of the current and predictions of the future channel powers in the log-domain $\hat{A}(k) \dots \hat{A}(k+N_{NC})$, required by the nonlinear controller.

The current received signal power $\hat{R}(k)$ is calculated by adding the biased channel power estimate $\hat{A}(k)$ and the biased Tx power estimate $\tilde{X}(k) = \hat{X}(k) - \bar{X}(k)$. The estimate $\hat{R}(k)$ is not biased, as the bias in $\hat{A}(k)$ and $\tilde{X}(k)$ cancel out each other. Therefore $\hat{R}(k)$ is used to calculate the control error $Y(k)$. The bias in $\hat{A}(k) \dots \hat{A}(k+N_{NC})$ is irrelevant to the nonlinear controller, as in equation (5) only differences $\hat{A}(k+l) - \hat{A}(k)$, with $l = 1 \dots N_{NC}$, are evaluated.

In case there are TPC errors or the Tx power limitation takes effect, the estimate of the Tx power $\hat{X}(k)$ will differ from the actual transmit power $X(k)$. As it is not possible to estimate this error, its effect on the system will be evaluated. First we want to show that a difference $E = \hat{X}(l) - X(l)$ that is constant for $l = (k-C+1) \dots k$, with C as the length of the

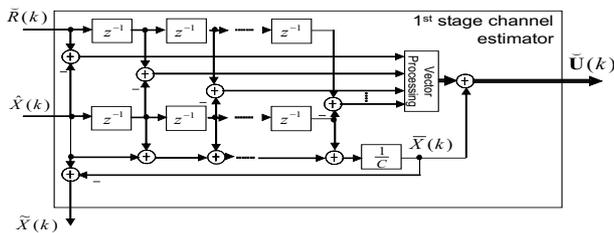


Fig. 5. 1st stage channel estimator

estimation and prediction filters, will not have any effect on the performance of the predictive power control. Therefore, with equation (5) we get

$$\begin{aligned} S &= \sum_{l=1}^{N_{NC}} \left[\Gamma^{\text{tar}} - (\tilde{X}(k) + \hat{A}(k) - \hat{I}(k)) - (\hat{A}(k+l) - \hat{A}(k)) \right. \\ &\quad \left. - (\hat{X}(k+l) - \hat{X}(k)) \right]^2 \\ &= \sum_{l=1}^{N_{NC}} \left[\Gamma^{\text{tar}} - (X(k) + \log_{10} |\mathbf{w}_0 10^{\tilde{\mathbf{U}}}| - \hat{I}(k)) - \right. \\ &\quad \left. (\log_{10} |\mathbf{w}_l 10^{\tilde{\mathbf{U}}'}| - \log_{10} |\mathbf{w}_0 10^{\tilde{\mathbf{U}}'}|) - (X(k+l) - X(k)) \right]^2 \quad (10) \end{aligned}$$

where $\tilde{\mathbf{U}}'(k) = [\tilde{U}'_0(k), \tilde{U}'_1(k), \dots, \tilde{U}'_{C-1}(k)]$ with

$$\tilde{U}'_i(k) = \check{R}(k-i) - X(k-i), \quad i = 0 \dots (C-1), \quad (11)$$

being the filter input vector in case of no TPC errors, no Tx power limitation and no offset $\tilde{X}(k)$. Obviously the factor E and also the normalization $\tilde{X}(k)$ nullify and thus constant differences between the actual Tx power X and its estimate \hat{X} do not change the system behavior. This is an important characteristic, as E can become relatively large due to accumulating TPC errors and the effect of the Tx power limiter.

It remains to examine the effect if E is not constant but changes over the estimation/prediction filter length C . Fig. 6 displays the standard deviation of the power control error for different velocities over the prediction range N_{NC} . Obviously, the performance gain due to prediction nearly rests the same with TPC errors, showing that our approach is quite robust. Naturally, the power control error always increases due to TPC errors, as a Tx power change into the wrong direction can not be compensated by an enhanced power control.

V. ESTIMATION AND PREDICTION FILTER ADAPTATION

Now, we want to discuss the calculation of the estimation and prediction filter coefficient vectors $\mathbf{w}_0 \dots \mathbf{w}_{N_{NC}}$. In case the channel statistics are known, the optimal choice, assuming linear filtering, is to apply Wiener filters. In this case the filter coefficients have to be precalculated according to

$$\mathbf{w}_L = \mathbf{R}_{\mathbf{u}'}^{-1} \mathbf{r}_{\mathbf{u}' a_L} \quad (12)$$

where $\mathbf{R}_{\mathbf{u}'}$ is the autocorrelation matrix of the filter input process $\mathbf{u}'(k)$ under the assumption that the mobile always reacts as commanded by the base station, meaning that there

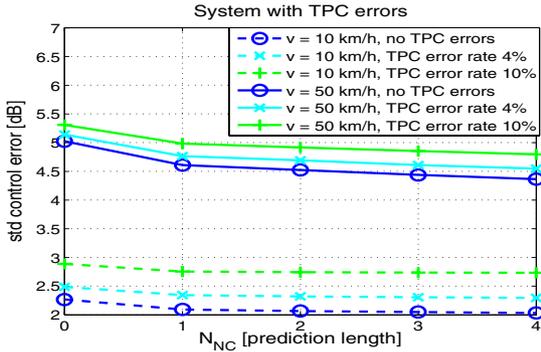


Fig. 6. Effect of undetected TPC errors on the power control error (realistic channel prediction based on Wiener filters, 1-path Rayleigh fading, interference noise power assumed to be known, symbol SINR 0dB, filterlength 20 slots)

is no difference between $X(k)$ and $\hat{X}(k)$, and that there is no offset ($\tilde{X}(k) = 0$), i.e

$$u'(k) = a(k) + \frac{n(k)}{x(k)}. \quad (13)$$

The correlation $\mathbf{r}_{\mathbf{u}'a_L}$ is the correlation between $u'(k)$ and the desired response $a(k+L)$ with L as index for the prediction range in power control slots ($L = 0$ for the estimation filter). The problem concerning the use of Wiener filters is, that usually the channel statistics are unknown and change over time. Thus we examine the application of adaptive filters.

We consider the use of the LMS algorithm for the adaptation of the prediction filter [9]. The LMS algorithm requires an error signal for iteratively adapting the filter coefficients according to

$$\mathbf{w}_L(k+L+1) = \mathbf{w}_L(k) + \mu \mathbf{u}(k) \cdot e_{LMS}^*(k) \quad (14)$$

where $*$ denotes the complex conjugate and

$$e_{LMS}(k) = u_1(k+L+1) \cdot 10^{\frac{\tilde{X}(k) - \tilde{X}(k+L+1)}{20}} - \hat{a}(k+L), \quad (15)$$

with $u_1(k+L+1)$ being the second entry of

$$\mathbf{u}(k+L+1) = [u_0(k+L+1), u_1(k+L+1), \dots, u_{C-1}(k+L+1)].$$

It would also be possible to use the first entry of $\mathbf{u}(k+L)$ instead, but as the first entry of the vector \mathbf{u} has always a lower SINR than the following entries (only 4 pilot symbols in the current slot are used for averaging, whereas all symbols of one slot are used for averaging in the preceding slots), it is advantageous to wait one further slot with the filter adaptation.

The correction term $10^{\frac{\tilde{X}(k) - \tilde{X}(k+L+1)}{20}}$ is required as the adaptation has to be based on the same offset $\tilde{X}(k)$, that has been used for the calculation of the filter input vector $\mathbf{u}(k)$, leading to the filter output $\hat{a}(k+L)$. The filter is adapted with a delay of $L+1$ power control slots, where L is the prediction length of the filter. Fig. 7 shows the configuration of the prediction filter adaptation. It remains to determine the length C of the prediction filters and the adaptation constant μ . Simulations

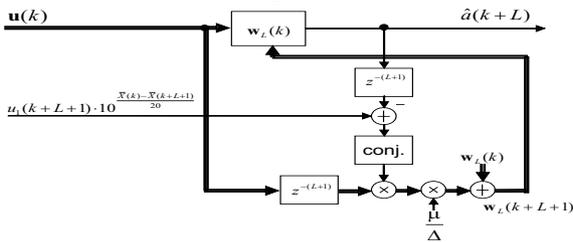


Fig. 7. LMS prediction filter adaptation

show that it is reasonable to choose a filter length of $C = 20$ and an adaptation constant of $\mu = 0.001$, ensuring the stability of the LMS adaptation and leading to adequate prediction error variances and adaptation times.

In Section IV we have discussed an offset compensation to limit the dynamic range of the estimated Tx power. We found out, that this offset compensation is not required from the algorithmic point of view, due to the linearity of the processing, see equation (10). Only implementation aspects necessitate the offset compensation. Now, the LMS filter adaptation is no longer a linear processing. Therefore, we are going to examine the effect of a difference $E = \hat{X}(k) - X(k)$ that is constant over the filter length C on the behavior of the LMS filter adaptation. Starting with the filter coefficient update equation

$$\begin{aligned} \mathbf{w}_L(k+L+1) &= \mathbf{w}_L(k) + \mu \cdot \mathbf{u}(k) \cdot e_{LMS}^*(k) \\ &= \mathbf{w}_L(k) + \mu \cdot \mathbf{u}(k) \\ &\quad \cdot \left(u_1(k+L+1) \cdot 10^{\frac{\tilde{X}(k) - \tilde{X}(k+L+1)}{20}} - \mathbf{w}_L(k) \mathbf{u}(k) \right)^* \\ &= \mathbf{w}_L(k) + \mu 10^{\frac{\tilde{X}(k) - E}{10}} \mathbf{u}'(k) (u_1'(k+L+1) - \mathbf{w}_L(k) \mathbf{u}'(k))^* \end{aligned} \quad (16)$$

with $\mathbf{u}'(k) = 10^{\frac{\tilde{X}(k)}{20}} = [u_0'(k), u_1'(k), \dots, u_{C-1}'(k)]$ where the entries are given $u_i'(k) = u_i(k-i)$, see eq. (13). Obviously the error E and the offset $\tilde{X}(k)$ lead to an effective adaptation constant $\mu \cdot 10^{\frac{\tilde{X}(k) - E}{10}}$.

The prediction performance and thus the choice of μ is based on a variance of the filter input process of $\sigma_a^2 = 1$, neglecting the noise component $\frac{n(k)}{x(k)}$, see eq. (13). An alteration of the effective adaptation constant directly influences the adaptation behavior. To avoid this effect, we replace the weight μ by $\mu' = \frac{\mu}{\Delta}$ with

$$\Delta = \frac{1}{\kappa} \sum_{k=1}^{\kappa} [u_1(k)]^2, \quad (17)$$

i.e. an estimate of the second moment of the input process. The averaging length κ is chosen sufficiently large.

Up to now, we have only discussed the adaptive implementation of the prediction filters. Unfortunately it is not possible to use the structure shown in Fig. 7 for the adaptation of the estimation filter \mathbf{w}_0 , due to the fact, that in this case the noise component in $u_1(k+L+1)$ is correlated with the noise component in $\mathbf{u}(k) \cdot z^{-1}$. Thus the filter cannot be appropriately adapted. Therefore we use Wiener filters for the channel estimation, under the design assumption that the channel is characterized by a Jakes spectrum.

Analysis shows that various design mismatches concerning the SINR, the shape of the spectrum and the velocity do not lead to a strong performance degradation, as long as the filter is not designed for a velocity smaller than the actual one. This is reasonable, considering that this filter is working as a simple smoothing filter. Assuming that rough velocity estimates are available, it should be possible to precalculate a set of filters, so that one of those is chosen according to the current velocity of the mobile terminal.

Finally we have described all required components to apply our new prediction based controller. For interference noise power estimation we use an approach running on chip level, as e.g. presented in [10].

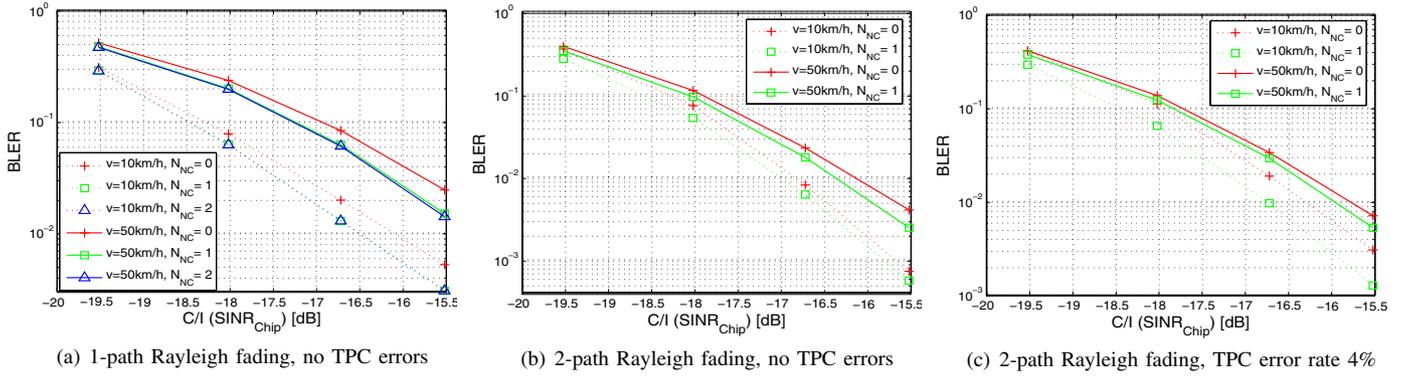


Fig. 8. Performance gain of prediction based power control, $N_{NC} = 0$ corresponds to classical CLPC

VI. PERFORMANCE OF PREDICTION BASED CLPC

In this section we show, how our new approach enhances the ability of the power control to track the time varying channel and hence, the system performance in terms of the block error rate (BLER) is increased. Exemplary we choose an Adaptive Multi Rate (AMR) speech service with the following parameterization

- convolutional code with rate $1/3$, blocklength of 192 bits
- interleaver length 15 power control slots,
- spreading gain $M_d = 64$ for the data channel,
- spreading gain $M_c = 256$ for the control channel,
- amplitude ratio between control and data channel $\beta = \frac{11}{15}$,
- assumption of perfect channel knowledge for decoding,
- complex Gold scrambling sequences, length 38.4Mchips.

Fig. 8(a) shows the BLER on the data channel over the chip to interference and noise ratio (C/I) for conventional power control ($N_{NC} = 0$) in contrast to prediction based power control for a 1-path Rayleigh fading environment. Obviously in case of one step prediction performance gains of up to 0.4dB can be achieved. The additional gain due to multi-step prediction is quite small. It has to be remarked, that this performance gain is achieved without increasing the Tx power in the mean.

In addition Fig. 8(b) depicts the system performance of predictive CLPC in case of a 2-path Rayleigh fading channel, where we assume that both paths fade independently and are combined by means of maximum ratio combining (MRC). The performance gain due to predictive power control is larger for the 1-path Rayleigh fading than for the 2-path Rayleigh fading, as in the latter case less deep fades occur, considering MRC. Obviously our algorithm is adaptive, as the prediction filters adapt to the different channel statistics. Note that the channel statistics after MRC are not longer Gaussian.

Furthermore Fig. 8(c) shows, that this prediction gain can also be achieved in case TPC errors occur on the downlink, showing the robustness of our new algorithm.

The specific gain due to prediction based power control depends strongly on the point of operation of the system. Firstly, the decrease of the control error variance is a function of the velocity, this is a specific characteristic of the power control itself. Secondly, the performance gain on the link level depends also on the point of operation concerning the SINR, the specific code and the interleaver length, determining, how a gain in the control error variance maps into a SINR gain. Depending on the system scenario, e.g. the velocity and the TPC error rate, we have observed performance gains of up to 0.5dB. These gains seem to be quite small, but they are for free, as they are not accompanied by any degradations.

As predictive power control decreases the control error variance, the effective channel comes closer to an AWGN channel, which is optimal concerning the BLER for a given SINR. Thus, every SINR gain achieved by predictive power control enables decreasing the target SINR by the outer loop power control, and thus decreasing the mean Tx power. Hence, these SINR gains result in an increased spectral efficiency.

VII. CONCLUSION

In this paper, we derive a new prediction based power controller using an optimization criterion based on the minimization of the sum over the predicted squared control errors, motivated by deficiencies of conventional closed loop power control. This algorithm shows a significant decrease of the standard deviation of the control error. In a second step we examined the channel state prediction in the context of a power controlled system considering TPC errors and a Tx power limitation in the mobile terminal, arriving at a robust and adaptive structure, using LMS filter adaptation algorithms, and thus requiring no knowledge about the channel statistic. In realistic link level simulations, our new prediction based power control scheme shows improved performance over the conventional scheme in terms of the SINR gain, also in case of realistic assumptions like TPC errors. This SINR gain can be used to decrease the target SINR by the outer loop power control, and hence, the Tx power in the mean, resulting in enhanced spectral efficiency. It is important to note, that this approach is totally compliant with the UMTS standard.

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