INVERSE CLOSED-LOOP POWER CONTROL FOR CDMA WIRELESS SYSTEMS

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ABSTRACT

In this work, two algorithms for closed loop power control in CDMA wireless systems are proposed. These algorithms aim to minimize the power control error by predicting the inverse of the channel fading. For this purpose, a channel fading prediction scheme is also proposed.

1. INTRODUCTION

In CDMA systems, all the mobile stations (MS) use the same bandwidth for transmission and thus inter-user interference occurs. The signal received by the base station (BS) from a near MS dominates that received from a far MS. This phenomenon is known to as the *near-far effect.*

The objective of power control is to control the transmission power of the mobile units so that the capacity of the overall system is maximized. Power control reduces inter-user interference by overcoming the near-far effect, which results in capacity increase of the overall CDMA system. It also minimizes the power consumption of the mobile units. Instead of using a fixed maximum power by the mobile station, it will now use an adaptive transmission power based on the power control requirements.

Figure 1 shows a block diagram of the conventional CLPC. The transmission power $P_t(t)$ used by the MS is attenuated by the channel fading $\phi(t)$. At the BS, the received power $P_r(n)$ is measured and compared to a desired fixed power level P_d. The error $e_a(n)$ is given by

$$e_a(n) = P_d - P_r(n). \tag{1}$$

Equivalently, we can write

$$e_a(n) = P_d - \phi(n)P_t(n-1). \tag{2}$$

The power error $e_a(n)$ is quantized using a one-bit quantizer to produce the power command bit $b(n)$ scaled by half the step-size of the quantizer Δ, i.e.,

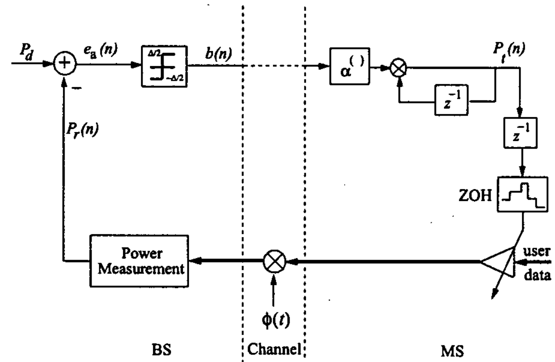

Fig. 1. *Conventional closed loop power control.*

$$b(n) = \frac{\Delta}{2}\text{sign}[e_a(n)]. \tag{3}$$

This PCB is transmitted to the MS. The mobile station then increments or decrements its transmission power by a fixed amount (in dB). This process is expressed as

$$P_t(n) = \alpha^{b(n)}P_t(n-1) \tag{4}$$

where α is a constant (usually $1 < \alpha < 3$).

The tracking ability of CLPC is usually slow compared to the variations in the channel fading. This is especially evident at high speed mobile units. Many variations to CLPC have been proposed in the literature to overcome this problem. Most of these variations rely on adapting the step size of the quantizer so as to cope with the fast variations in the channel fading [1, 2]. In this work, we develop two new techniques for CLPC, in which the BS directly predicts the channel fading and then feeds it (in coded form) to the MS. New prediction and coding schemes are developed for this purpose.

2. OVERSAMPLED CHANNEL PREDICTION

In the next section, we propose two methods for CLPC that require a prediction for the channel power fading $\phi(n)$. In this section, we propose a method for predicting $\phi(n)$. The method is

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based on oversampling the received power variations at the BS. Then, an NLMS-based adaptive predictor is used to estimate the channel fading one-step ahead. For this purpose, we assume that the BS knows the transmission power $P_t(n)$ of the MS at each time instant, which is usually the case in CLPC.

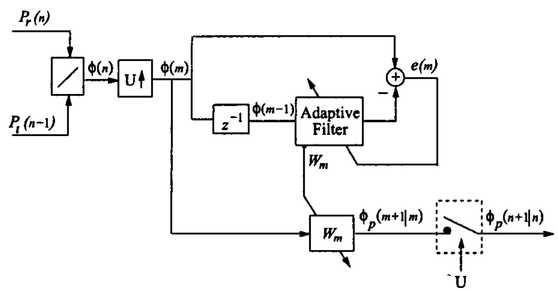

Fig. 2. *Power fading prediction.*

Figure 2 shows the structure of the proposed prediction method. The measured received power $P_r(n)$ is divided by $P_t(n-1)$ to get the power attenuation $\phi(n)$, i.e.,

$$\phi(n) = \frac{P_r(n)}{P_t(n-1)}. \tag{5}$$

The signal $\phi(n)$ is then up-sampled by a factor of U resulting in $\phi(m)$, where m refers to the oversampling index. This can be achieved by increasing the sampling rate of the received power and by assuming that the transmission power is constant between two consecutive samples of $P_t(n)$.

The signal $\phi(m)$ is then passed through a delay as shown in Figure 2. The delayed samples $\phi(m-1)$ are fed into an adaptive filter of order M. The output of the adaptive filter is compared to $\phi(m)$. The comparison error is fed back to the adaptive filter for online training. The taps of the adaptive filter, W_m, extract the correlation between the fading samples. The tap values are carried out online and used to adapt the taps of an FIR filter as shown in the figure. The input to this FIR filter is $\phi(m)$ and its output is the prediction of $\phi(m+1)$ denoted by $\phi_p(m+1|m)$. This signal is then down-sampled by the factor U to produce the required prediction value

$$\phi_p(n+1|n) \approx \phi(n+1). \tag{6}$$

The normalized-LMS algorithm is used here, where the taps W_m are updated according to the rule

$$W_{m+1} = W_m + \frac{\mu}{\delta + ||\mathbf{u}_m||^2}\mathbf{u}_m^*(\phi(m) - \mathbf{u}_m W_m) \quad (M \times 1) \tag{7}$$

where the regression vector $\mathbf{u}_m$ contains the latest M samples of $\phi(m-1)$ and the notation $||.||^2$ denotes the Euclidean norm. The constant μ is the step-size of the adaptive filter and δ is an arbitrary small positive number.

The performance of this predictor is dependent on many factors such as the filter type, order, and step-size. Furthermore, the oversampling factor U plays a significant role in the performance of the predictor since it increases the correlation between the samples of $\phi(m)$. It should be noticed here that increasing U will introduce noise in the measured $P_r(n)$ resulting in degradation in performance. This usually sets an upper limit for choosing U. We found through simulations that $U \leq 5$ is an acceptable choice.

Figure 3 shows the prediction mean square error (MSE) versus the step-size μ. The MSE can be further reduced by increasing the oversampling factor U. In Figure 4, the MSE is shown as a function of U for different f_D's and for $\mu = 1.2$.

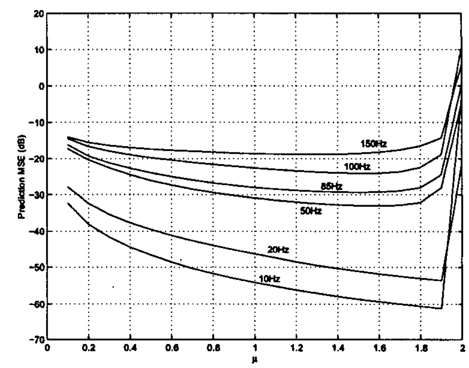

Fig. 3. *Effect of the step-size μ on prediction MSE for different Doppler frequencies.*

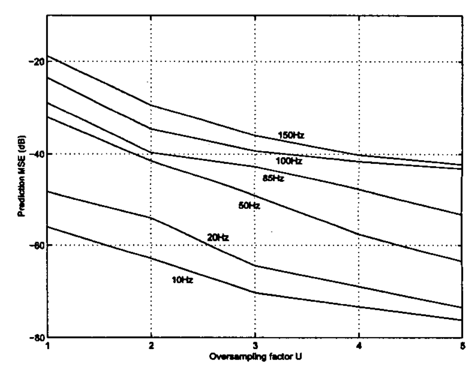

Fig. 4. *The prediction MSE as a function of the oversampling factor U for different Doppler frequencies.*

3. INVERSE POWER CONTROL

We now describe two algorithms for power control that rely on inverse control ideas. The basic principle here is not to provide the MS with commands to increase or decrease its power according to how far its transmitted power is from the reference level point. Instead, the idea is for the BS to estimate what the transmitted power should be for the next period of time and to provide this value directly (in coded form) to the MS.

3.1. Algorithm 1: Direct Inverse CLPC (DI-CLPC)

A block diagram of the proposed scheme is shown in Figure

5. The power control process works as follows. The BS measures the received power $P_r(n)$ from the bit stream arriving at its end. Then, the MS transmission power $P_t(n-1)$ and $P_r(n)$ are fed to the prediction block, which produces $\phi_p(n+1|n)$. The BS estimates the transmission power that should be used by the MS as

$$\widehat{P}_t(n) = \frac{P_d}{\phi_p(n+1|n)}. \quad (8)$$

It can be easily shown from (2) that this value "minimizes" the power error $e_a(n)$. This information is to be transmitted to the MS. Since we are limited by the power bit rate, $\widehat{P}_t(n)$ should be coded to meet this rate.

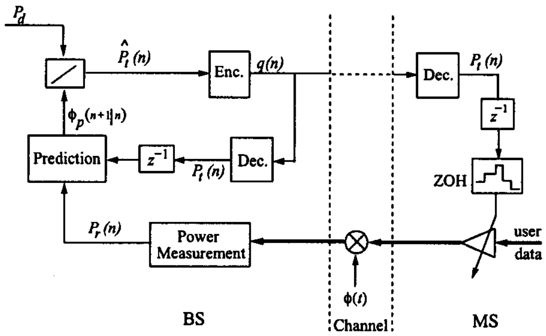

Fig. 5. *Block diagram of the Direct Inverse CLPC algorithm. The prediction scheme of Figure 2 can be used here.*

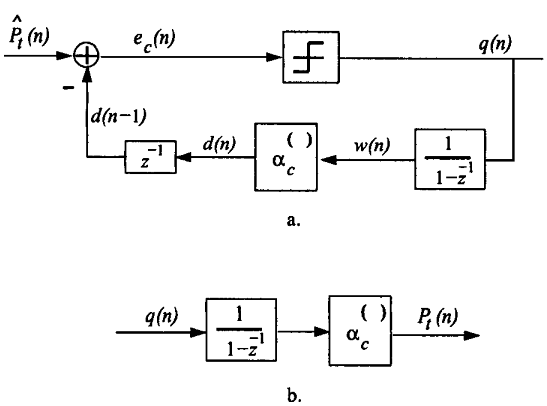

Fig. 6. *Coding scheme used in Direct Inverse CLPC algorithm. a. Encoder b. Decoder.*

The coding scheme used to transmit $\widehat{P}_t(n)$ could be the one developed in [3]. This coder exhibits strong tracking, good stability, and high dynamic range. Figure 6 shows a block diagram of this coding scheme, the encoder and decoder are shown in parts a and b respectively. The equations describing the dynamics of the coder are:

$$\begin{aligned} e_c(n) &= \widehat{P}_t(n) - d(n-1), \qquad d(0) = d_0 \\ q(n) &= \text{sign}[e_c(n)] \end{aligned}$$

Initialization:
Choose the desired received power P_d.
Choose the coding parameters α_c and $d(0)$.
Choose the prediction parameters:
Filter order, μ, and U.

For every CLPC time sample $n > 0$ do:

BS.

1. Measure $P_r(n)$ from the received sequence.
2. Knowing $P_t(n-1)$, estimate $\phi(n)$.
3. Evaluate $\phi_p(n+1|n)$.
4. Code the power data $\widehat{P}_t(n) = \frac{P_d}{\phi_p(n+1|n)}$.
5. Send the coded data $q(n)$ to the MS.

MS

6. Extract $q(n)$ from the received data.
7. Use $q(n)$ to decode the power data $d(n)$.
8. Set $P_t(n) = d(n)$.

Table 1. *Summary of the Direct Inverse CLPC (DI-CLPC) algorithm.*

$$\begin{aligned} w(n) &= w(n-1) + q(n), \quad w(0) = 0 \\ d(n) &= \alpha_c^{w(n)}. \end{aligned} \quad (9)$$

In this algorithm, the term α_c denotes the coding exponent. The DI-CLPC algorithm is summarized in Table 1.

3.2. Algorithm 2: Adaptive Direct Inverse CLPC (ADI-CLPC)

In this algorithm, we modify the coding scheme of the DI-CLPC by using an adaptive exponent term α_c, as shown in Figure 7. The purpose of adapting α_c is to cope with large variations in the channel power fading. The adaptation technique used here is

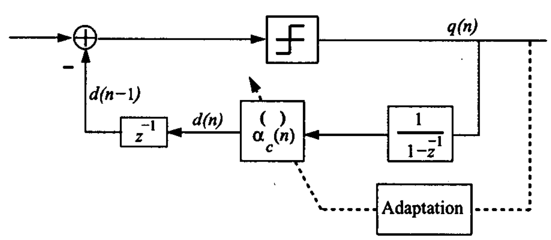

Fig. 7. *Coding scheme used in the ADI-CLPC algorithm.*

$$\alpha_c(n) = \alpha_c(n-1) + \lambda(n)C \quad (10)$$

Initialization:
Choose the desired received power P_d.
Choose the prediction parameters:
Filter order, μ, and U.
Choose the adaptation parameters:
C, α_{max} and α_{min}.

For every CLPC time sample $n > 0$ do:

BS

1. Measure $P_r(n)$ from the received sequence.
2. Knowing $P_t(n-1)$, estimate $\phi(n)$.
3. Evaluate $\phi_p(n+1|n)$
4. Compute $\lambda(n)$ and $\alpha_c(n)$ from (10) and (11).
5. Use $\alpha_c(n)$ to code $\widehat{P}_t(n) = \frac{P_d}{\phi_p(n+1|n)}$
6. Send the coded data $q(n)$ to the MS.

MS

7. Extract $q(n)$ from the received data.
8. Use (10) and (11) to recompute $\alpha_c(n)$.
9. Decode $d(n)$ from $q(n)$ and $\alpha_c(n)$.
10. Set $P_t(n) = d(n)$.

Table 2. *Summary of the Adaptive Direct Inverse CLPC (ADI-CLPC) algorithm.*

where

$$\lambda(n) = \begin{cases} +1 & \text{if } q(n) = q(n-1) \text{ and } q(n-1) = q(n-2) \\ -1 & \text{if } q(n) \neq q(n-1) \\ 0 & \text{otherwise} \end{cases} \quad (11)$$

and

$$\alpha_c(n) = \begin{cases} \alpha_{max} & \text{if } \alpha_c(n) > \alpha_{max} \\ \alpha_{min} & \text{if } \alpha_c(n) < \alpha_{min} \end{cases} \quad (12)$$

with typical values $C = 0.2$, $\alpha_{max} = 2.5$, and $\alpha_{min} = 1.1$. The ADI-CLPC algorithm is summarized in Table 2.

4. SIMULATIONS

The algorithms developed in this study are simulated using Matlab and Simulink.The power bit rate is chosen as 1500Hz. The prediction upsampling factor U, and the desired power P_d are set to 2 and 0dB respectively. The channel is frequency selective Rayleigh fading with two paths and a variable mobile speed. The channel fading data is obtained from Simulink. The standard deviation of the power control error is used as a measure of how well the power control algorithms achieve the desired received power.

Figure 8 shows the PCE performance of the DI-CLPC and ADI-CLPC. The coding parameters $d(0)$ and α_c used in the DI-CLPC algorithm are chosen as 1E-3 and 1.8, respectively. Moreover, the parameters C, α_{min}, α_{max} for the ADI-CLPC algorithm are set to 0.1, 1.1, and 2, respectively. Figure 8 includes also the performance of the conventional CLPC and that of an adaptive CLPC developed in [4], for the sake of comparison. The ADI-CLPC demonstrates the best performance over all other algorithms.

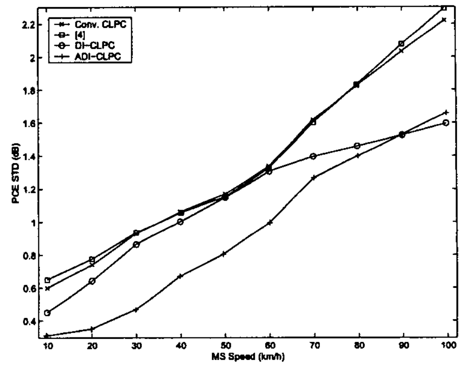

Fig. 8. *Performance of the developed algorithms compared to conventional CLPC and an adaptive CLPC developed in [4].*

5. CONCLUSION

In this work, we proposed two algorithms for closed loop power control. The algorithms are based on two operations, namely, prediction of channel fading and coding of power command information. For this purpose, a new prediction scheme has been proposed and a powerful coder is implemented, resulting in improved power control performance.

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