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DESIGN OF EMBEDDED FILTERS FOR INNER-LOOP POWER CONTROL IN WIRELESS CDMA COMMUNICATION SYSTEMS

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ABSTRACT

We study inner-loop power control for mobile wireless communication systems using code division multiple access transmission. We focus on the uplink, *i.e.*, on communication from the mobile- to the base-station, and show how to minimise the variance of the signal-to-interference ratio (SIR) tracking error through incorporation of recursive filters. These filters complement existing power controllers and are designed by using a linear model which takes into account quantisation of the power control signal, dynamics of channel gains, interference from other users, target SIR, and SIR estimation errors. Simulation results indicate that significant performance gains can be obtained, even in situations where the models used for design are only an approximation.

Key Words: Quantised control, power control, code division multiple access, network control, filter design.

I. INTRODUCTION

Power control is a key enabling technology for wireless communications, constituting a versatile means to guarantee quality of service (QoS) to individual users; see, e.g. [1–3] for an introduction to the topic. Power control serves to compensate for variations in the channel gains and trades interference between users of a range of individual mobile stations. Accurate power control becomes especially important in code division multiple access (CDMA) systems, where the same frequency spectrum is shared between users, and multiple-access interference is inherently strong. In the case of cellular systems, power control architectures are typically cascaded two-loop schemes, where an outer loop gives a signal-to-interference ratio (SIR) target signal, which individual mobile stations (MSs) need to track with the help of an inner control loop. Several works, including [4–6], have studied outer loop design from a stochastic control viewpoint. In the present work, we will focus on the inner control loop.

A central issue when designing inner-loop power control systems for CDMA systems is that the bit-rate available for control signaling is very limited to keep the overhead low. Typically, each value is quantised using only one bit; see, e.g., [7]. This limits the achievable control performance significantly. In existing designs, the inner-loop controller simply sends the sign of the difference in estimated and target SIRs, thereby giving an order to increase or decrease the power by one step. A distinct disadvantage with this simple approach is that dynamic effects inherent in the fading, the development of interference, the terminal implementation, and the processing at the base station (BS), *are not exploited in the controller*, a fact that is well known

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to lead to sub-optimal performance. Several approaches have been documented for the design of more advanced inner-loop power control systems. A common underlying theme here has been the adoption of linear models, which are derived around operating points. For example, the articles [8, 9] use robust control techniques, which lead to linear time-invariant controllers followed by a quantiser; [10–13] describe adaptive methods.

The main contribution of the present work is to show how to use filtering to make more efficient use of the signaling resources available for uplink inner-loop power control. In our approach, the control signal is pre-filtered before being quantised at the BS. At the MS, the received signal is then filtered by a second filter, which is the inverse of the first filter. Our designs mitigate the SIR tracking error caused by quantisation effects and take into account dynamics of the channel gain, target SIR, interference from other users and SIR estimation noise. This makes quantisation more efficient whilst, at the same time, ensuring that the control signal is received with minimal distortion. The filters are designed by adapting the networked control systems framework of [14, 15], and seek to optimise *closed loop performance*.

This article uses the wideband code division multiple access (WCDMA) inner-loop power control algorithm as an example. However, the design ideas presented are applicable to CDMA inner-loop power control in general. In fact, a key feature is that the proposal can be used to enhance any existing power control scheme where the output of a linear controller is quantised. This includes current inner-loop uplink WCDMA power control technology and also the approaches presented in [8, 9]. Before proceeding, it is worth emphasizing that, whilst Doppler information can be useful for controller design, the present work is focused on quantization effects. Thus, we have chosen not to include Doppler effects.

The remainder of this paper is organised as follows: In Section II models relevant for controller design are developed. The filter design method is presented in Section III. Simulation studies are included in Section IV. Section V discusses implementation issues. Section VI draws conclusions.

II. CHANNEL FADING, INTERFERENCE AND POWER CONTROL CONSTRAINTS

This section defines models for channel fading and interference as well as inner-loop power control constraints that are relevant for dynamic power control design. More detailed treatments can be found, for example, in [1–3].

2.1 Channel fading

We will first concentrate on a single MS connected to a single BS. The channel gain is, in general, time varying and can be modeled (in discrete-time ℓ) via

$$\bar{g}(\ell) = \bar{g}^{\mathrm{ls}}(\ell)\,\bar{g}^{\mathrm{ss}}(\ell), \quad \ell \in \mathbb{N}, \tag{1}$$

where $\bar{g}^{\mathrm{ls}}(\ell)$ refers to *large-scale fading*, whilst $\bar{g}^{\mathrm{ss}}(\ell)$ denotes *small-scale fading*. It is convenient to rewrite (1) in dB as

$$g(\ell) = g^{\mathrm{ls}}(\ell) + g^{\mathrm{ss}}(\ell), \quad \ell \in \mathbb{N}. \tag{2}$$

Large-scale fading represents the average signal power attenuation. It is caused by distance-dependent propagation loss and by electromagnetic shadowing by large objects. It can be modeled via

$$\bar{g}^{\mathrm{ls}}(\ell) = D(\bar{r}(\ell))^{-\alpha}\,\bar{S}(\ell),$$

where D is a constant, $\bar{r}(\ell)$ is the distance between the MS and the BS, α is the path loss exponent and $\bar{S}(\ell)$ encompasses shadow fading effects. Shadow fading effects are spatially correlated and can be modelled as a stochastic process; see, e.g., [16]. Small-scale fading is caused by small changes in MS position and MS surroundings. Here, we adopt the stochastic modeling approach and focus on the power channel gain at slot level [This means that effects of fading at chip level need to be aggregated to produce slot level power channel gains.] A convenient way to model this is by means of autoregressive moving average (ARMA) models; see, e.g., [17–19].

2.2 Interference

We will next consider a single BS which provides service to K MSs and denote the channel gain at time $\ell \in \mathbb{N}$ between the MS k and the BS via $\bar{g}_k(\ell)$. Thus, if each MS k transmits with power $\bar{p}_k(\ell)$, then, at time ℓ, the BS will receive a signal of power

$$\sum_{k=1}^{K} \bar{g}_k(\ell)\,\bar{p}_k(\ell) + \bar{n}(\ell),$$

where $\bar{n}(\ell)$ denotes thermal noise.

In a CDMA system, all mobile stations share the same frequency band. Thus, multiple-access interference will produce occasional transmission errors and affect QoS to users. To examine the uplink communication link for MS k, we define the received power of user k via

$$\bar{H}_k(\ell) \triangleq \bar{g}_k(\ell)\,\bar{p}_k(\ell)$$

and the interference from other users and noise via

$$\bar{I}_k(\ell) \triangleq \sum_{j \neq k,\, 1 \leq j \leq K} \theta_j \bar{g}_j(\ell) \bar{p}_j(\ell) + \bar{n}(\ell), \tag{3}$$

where θ_j depends, *inter alia*, on the cross-correlation between the codes used by MSs j and k.

A useful measure of the perceived signal quality for user k is the SIR, defined (in dB) by:

$$\begin{aligned}\gamma_k(\ell) &\triangleq \varphi_k + H_k(\ell) - I_k(\ell)\\ &= \varphi_k + g_k(\ell) + p_k(\ell) - I_k(\ell),\end{aligned} \tag{4}$$

where φ_k is the spreading factor. The latter is a known constant and, thus, does not influence the control system design. Without loss of generality, in the sequel, we set $\varphi_k = 0$.

2.3 Uplink power control in WCDMA

To control MS power, the BS sends power control command signals, say $\{w_k\}$. Upon receipt, the MSs update their power levels, say $\{p_k\}$, accordingly. Uplink power control in WCDMA is carried out via a cascade architecture, where a slow outer loop provides target SIR trajectories, say $\{\gamma_k^\star\}$, to fast inner loops. The inner loops have a sampling frequency of 1.5 kHz, which is sufficient for small scale fading at moderate speeds. In the inner power control loops, the interference from other users and noise, see (3), are regarded as exogenous signals. The situation is shown in Fig. 1. To take account of possible errors in the power control signal transmission, in this figure we have also included a down-link channel noise process, η_k.

To limit the bit-rate overhead, control commands are constrained to a small number of bits. The associated quantization effects affect the achievable tracking performance and form the main motivation for the present work. In existing designs, see [7], the kth inner-loop controller sends power increments

$$w_k(\ell) = Q_k(\gamma_k^\star(\ell) - \hat{\gamma}_k(\ell)), \quad \ell \in \mathbb{N}, \tag{5}$$

where $\hat{\gamma}_k(\ell)$ is an SIR estimate, and $Q_k(\cdot)$ is a scalar quantiser. Typically, each value $w_k(\ell)$ needs to be expressed via 1 bit, in which case the range of $Q_k(\cdot)$ has only two elements [7].

At the kth MS, the power signal p_k is computed by passing the received signal, *i.e.*, $w_k + \eta_k$, through an integrator with a time delay of $d_1 \geq 1$ WCDMA slots (each 2/3 ms long) and through a limiter. At the BS, SIR estimates are obtained with a delay of $d_2 \geq 1$ slots. We denote the associated estimation noises via $\{\varepsilon_k\}$. The resulting control loop is depicted in Fig. 2, where β is the integrator gain and where the saturation block at the mobile station models limitations on available MS power. To encompass control algorithms presented in [8, 9], in Fig. 2 we have also included an additional linear-time invariant controller, namely $C_k(z)$. Current technology, see (5), amounts to setting $C_k(z) = 1$.

In the following section, we will show that the power control architecture in Fig. 2 can be enhanced through the use of filters which are embedded at the base-station and at the mobile-station. These filters will be designed to mitigate quantisation effects on tracking performance.

III. EMBEDDED FILTERING FOR INNER-LOOP POWER CONTROL

Sending power increments (as in Fig. 2) reduces bit-rates, if signals are approximately constant. Nevertheless, alternative strategies should be examined. The key point we will explore is that prior knowledge of dynamic effects associated with $\{\gamma_k^\star\}$, $\{g_k\}$, $\{I_k\}$, $\{\varepsilon_k\}$, and $\{\eta_k\}$ can be used to minimise the performance degradation which results from the bit-rate limitation imposed on the control commands. For that purpose, we propose to embellish the configuration in Fig. 2 by incorporating pre- and post-filters $F_k(z)$ and $F_k^{-1}(z)$, as shown in Fig. 3.

We will next show how the pre- and post-filters can be designed to minimise the variance of the tracking error caused by quantisation. We will study a situation where the controllers $C_k(z)$ have already been designed (e.g., by using the linear design techniques described in [8, 9]). In order not to alter this nominal control design, the filters are chosen to be perfect-reconstruction pairs, *i.e.*, we set

$$F_k^{-1}(z) F_k(z) = 1. \tag{6}$$

In what follows, we consider the control loop for user k and will therefore drop the subscript k in all signals. A key ingredient of this work is to neglect saturation effects and to base the controller design on a linear approximation of the quantiser as follows:

Assumption 1. The quantiser output is related to its input v via

$$Q(v) = v + q,$$

where q is arbitrarily distributed zero-mean white noise with variance Φ_q. This variance is related to that of v via:

$$\Phi_q = \left(\frac{1}{\lambda}\right) \Phi_v, \tag{7}$$

where λ is the *signal-to-noise ratio of the quantiser*.

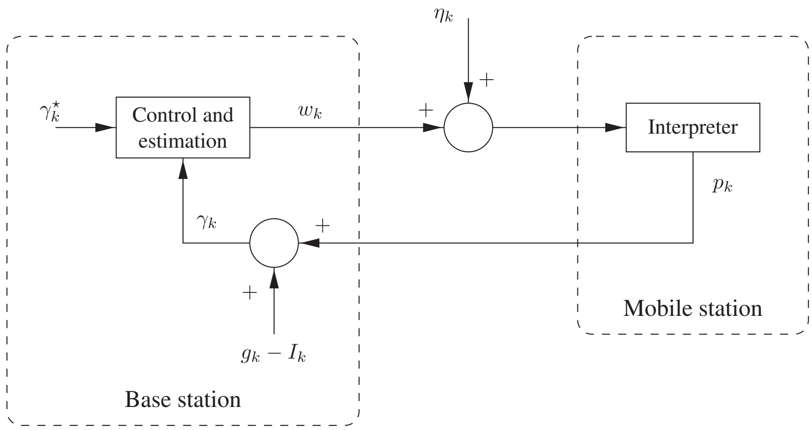

Fig. 1. Inner-loop uplink power control system.

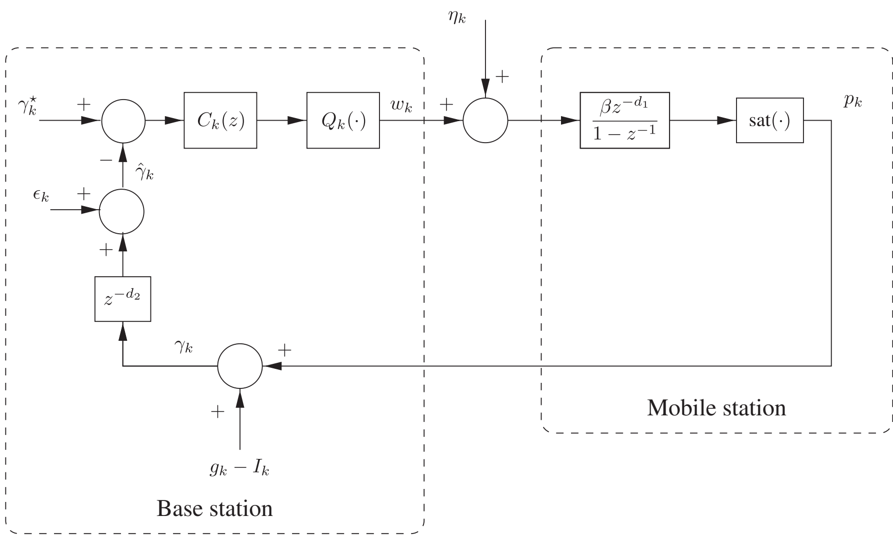

Fig. 2. WCDMA inner-loop uplink power control system with controller $C_k(z)$.

In the above model (see also [13, 20–22]) λ depends upon the distribution of v and on the output set of $Q(\cdot)$ and should not be mistaken for the SIR in (4).

Given the quantisation model adopted, and since pre- and post-filters are taken as the inverse of each other (see (6)), direct calculations yield that, in the absence of saturation effects, the SIR γ in the loop depicted in Fig. 3 is characterised via:

$$\gamma = \phi(z)((1-z^{-1})(g-I)+\beta z^{-d_1}F^{-1}(z)(\eta+q)) \\ +\phi(z)\beta z^{-d_1}C(z)(\gamma^\star-\varepsilon),$$

where

$$\phi(z) \triangleq (1-z^{-1}+\beta z^{-d}C(z))^{-1},$$

$$d \triangleq d_1+d_2.$$

Thus, the tracking error defined via

$$y \triangleq \gamma^\star - \gamma$$

satisfies:

$$y = \beta\phi(z)z^{-d_1}C(z)\,\varepsilon - \phi(z)(1-z^{-1})(g-I) \\ -\phi(z)\beta z^{-d_1}F^{-1}(z)\,(\eta+q) \\ +\phi(z)\big(1-z^{-1}+\beta C(z)(z^{-d}-z^{-d_1})\big)\gamma^\star. \quad (8)$$

We note that the filters introduced do not alter the nominal design relationships between $\gamma^\star$, g, I, and ε and the tracking error y. On the other hand, filtering does

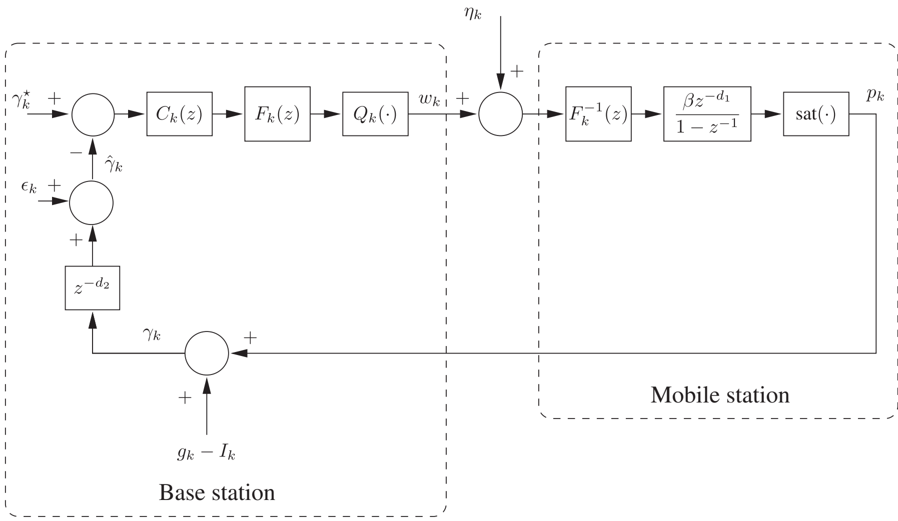

Fig. 3. Proposed WCDMA inner-loop uplink power control system with embedded filters $F_k(z)$ and $F_k^{-1}(z)$.

influence the impact that the quantiser and η have on tracking performance.

To design the filters $F(z)$ and $F^{-1}(z)$, we will proceed as is common for power control design (see, e.g. [8, 11, 12]) and adopt a linear model for the power control loop. Linear models describe *incremental variables* and are derived at an operating point. The actual bias of the signals becomes irrelevant. Furthermore, we will neglect errors in the power control signal transmission. Accordingly, we introduce the following assumption:

Assumption 2. The (incremental) signals $\gamma^\star$, g, I and ε are zero mean and mutually independent. In addition, $\eta = 0$.

As stated above, the zero mean assumptions are justified by the fact that the integration in the MS makes the controller design rely on the incremental variables. The assumption of mutual independence of the other variables is justified by the fact that they are generated by different physical processes. It is worth noting that the same underlying assumptions are behind most other CDMA power control schemes, although they may not be stated explicitly. In this work a stochastic controller design approach is followed, which is the reason why Assumption 2 becomes visible.

Equation (8) shows that the variance of the component of the tracking error which arises from q is quantified by the cost function J in:

$$J \triangleq \Phi_q \left(\frac{\beta^2}{2\pi} \int_{-\pi}^{\pi} |\phi(e^{j\omega}) F^{-1}(e^{j\omega})|^2 \, \mathrm{d}\omega \right). \tag{9}$$

Theorem 1 characterises optimal filter pairs.

Theorem 1. Suppose that Assumptions 1 and 2 hold. Then the filters which minimise J defined in (9) have frequency responses:

$$|F_k^{\pm 1}(e^{j\omega})| = \left(\frac{1}{B_k \sqrt{|C_k(e^{j\omega})(e^{j\omega}-1)\Delta_k(e^{j\omega})|}} \right)^{\pm 1}, \tag{10}$$

where B_k is a positive real number, and $|\Delta_k(e^{j\omega})|^2$ denotes the power spectral density of

$$\delta_k \triangleq \gamma_k^\star + g_k + I_k + \varepsilon_k. \tag{11}$$

Proof. From Fig. 3 and with $\eta = 0$, it follows that the input to the quantiser is given by

$$\begin{aligned} v &= F(z)C(z)(1-z^{-1})\phi(z)(\gamma^\star - \varepsilon - z^{-d_2}(g-I)) \\ &\quad - z^{-d}\beta\phi(z)C(z)q. \end{aligned}$$

The variance of v satisfies:

$$\begin{aligned} \Phi_v &= \frac{1}{2\pi}\int_{-\pi}^{\pi} |F(e^{j\omega})\psi(e^{j\omega})\Delta(e^{j\omega})|^2 \, \mathrm{d}\omega \\ &\quad + \frac{\beta^2 \Phi_q}{2\pi} \int_{-\pi}^{\pi} |\phi(e^{j\omega})C(e^{j\omega})|^2 \, \mathrm{d}\omega, \end{aligned} \tag{12}$$

where

$$\psi(z) \triangleq C(z)(1-z^{-1})\phi(z).$$

Now, (7) gives that $\Phi_v = \lambda \Phi_q$, so that (12) yields:

$$\Phi_q = \frac{\tilde{\lambda}^{-1}}{2\pi} \int_{-\pi}^{\pi} |F(e^{j\omega})\psi(e^{j\omega})\Delta(e^{j\omega})|^2 \, \mathrm{d}\omega, \tag{13}$$

where

$$\tilde{\lambda} \triangleq \lambda - \frac{\beta^2}{2\pi}\int_{-\pi}^{\pi} |\phi(e^{j\omega})C(e^{j\omega})|^2 \,\mathrm{d}\omega.$$

Substitution of (13) into (9) provides

$$J = \frac{\beta^2}{\tilde{\lambda}}\left(\frac{1}{2\pi}\int_{-\pi}^{\pi} |\mathscr{A}(e^{j\omega})|^2 \,\mathrm{d}\omega\right) \cdot \left(\frac{1}{2\pi}\int_{-\pi}^{\pi} |\mathscr{B}(e^{j\omega})|^2 \,\mathrm{d}\omega\right), \tag{14}$$

where

$$\begin{aligned} \mathscr{A}(e^{j\omega}) &\triangleq |F^{-1}(e^{j\omega})\phi(e^{j\omega})| \\ \mathscr{B}(e^{j\omega}) &\triangleq |F(e^{j\omega})\psi(e^{j\omega})\Delta(e^{j\omega})|. \end{aligned} \tag{15}$$

Equation (14) allows one to quantify the impact of the quantiser on closed-loop power control performance. It depends upon the filter pair used. Performance can be optimised by choosing the filter pairs to minimise J. For that purpose, we recall the Cauchy–Schwarz inequality, see, e.g., [23], which, when applied to (14), gives

$$\begin{aligned} J &\geq \frac{\beta^2}{\tilde{\lambda}}\left|\frac{1}{2\pi}\int_{-\pi}^{\pi} \mathscr{A}(e^{j\omega})\mathscr{B}(e^{j\omega}) \,\mathrm{d}\omega\right|^2 \\ &= \frac{\beta^2}{\tilde{\lambda}}\left(\frac{1}{2\pi}\int_{-\pi}^{\pi} |\psi(e^{j\omega})\phi(e^{j\omega})\Delta(e^{j\omega})| \,\mathrm{d}\omega\right)^2, \end{aligned}$$

where we have used (15). Since the functions $\mathscr{A}(e^{j\omega})$ and $\mathscr{B}(e^{j\omega})$ are nonnegative everywhere, the bound is achieved if and only if $\mathscr{A}(e^{j\omega})/\mathscr{B}(e^{j\omega})$ is constant $\forall\omega \in [-\pi, \pi)$, *i.e.*, if and only if (10) holds. □

The filters specified in Theorem 1 minimise the SIR tracking error caused by quantisation (under suitable assumptions). The designs provided take into account dynamics of the channel gain, target SIR, interference from other users, and SIR estimation noise, see (11). These quantities can be estimated, using, for example, techniques akin to those presented in [11, 12, 18].

Interestingly, in special cases, the WCDMA inner-loop power control scheme used in current practice, see Fig. 2, is optimal, according to Theorem 1. This fact is made precise in the following corollary:

Corollary 1. Suppose that Assumptions 1 and 2 are satisfied and that δ_k of (11) fulfills

$$|\Delta_k(e^{j\omega})|^2 = \frac{1}{|C_k(e^{j\omega})|^2|e^{j\omega}-1|^2}. \tag{16}$$

Then the WCDMA inner-loop power control scheme of Fig. 2 minimises J of (9).

Proof. Follows directly from Theorem 1. □

In particular, with current technology, we have $C_k(z) = 1$. This choice is optimal, if δ_k is a random walk process. As will become apparent in Section IV, in situations where (16) does not hold, significant performance gains can be obtained by using a filter pair with a frequency response satisfying (10).

Remark 1. The idea of incorporating pre- and post-filtering around a quantiser is certainly not new; see, e.g., [24, 25] and the references therein. The distinguishing aspect of the situation at hand is that the filters in (10) are designed by respecting the closed-loop nature of the signals. This puts our proposal in the framework of [14, 15] and makes it a novel concept for power control.

Remark 2. The loop in Fig. 3 is stochastic and involves quantization and saturation. It is worth noting that, if $Q_k(\cdot)$ has a bounded range, all the exogenous (incremental) signals have bounded support, and $C(z)$, $F(z)$ and $F^{-1}(z)$ are stable, then the closed loop is practically stable, *i.e.* all signals are bounded.

IV. SIMULATION STUDY

In order to investigate the potential performance gains associated with the proposed filter design method, we will focus on a single control loop and model the (incremental) channel gain sequence g_k, via the ARMA process [The channel model (17) is identified to resemble a typical spectral shape of the channel gains. ARMA models have been used to predict fading phenomena around an operating point, e.g., in [12, 17–19].]

$$g_k(\ell) = \left(\frac{(z+0.95)}{150(z-0.97)(z-0.96)}\right) v_k(\ell), \tag{17}$$

In (17), v_k is zero-mean i.i.d. white Gaussian noise of variance 30; the sampling frequency is chosen as 1.5 kHz. The interference from other users and noise, I_k, is taken as a zero-mean white i.i.d. Gaussian noise process of variance 0.01 and with sampling rate 500 Hz. Delays are set to $d_1 = d_2 = 1$ WCDMA slot. The SIR estimation error, ε_k, is modeled via a zero-mean white i.i.d. Gaussian noise process of variance 0.01 and with sampling rate 1.5 kHz. The SIR reference signal, $\gamma_k^\star$, is modeled as a zero-mean white i.i.d. Gaussian process of variance 0.1 and with sampling rate 50 Hz. I_k and $\gamma_k^\star$ are held constant between sampling instants. Control commands are restricted according to the allowed values $w_k(\ell) \in \{\pm 1\,\mathrm{dB}\}$, $\forall \ell \in \mathbb{N}$.

We will quantify control performance via the empirical error variance over a simulation time of 10 seconds, *i.e.*:

$$V \triangleq \frac{1}{15001} \sum_{\ell=0}^{15000} (\gamma_k^\star(\ell) - \gamma_k(\ell))^2, \tag{18}$$

averaged over 100 Monte Carlo simulations for each loop.

We first examine the current procedure for WCDMA power control as described in Section II with $C_k(z)=1$. The resulting power control performance depends upon the value chosen for β, see Fig. 2. Simulation results for $\beta=0.39$ are included in Fig. 4. As can be seen in Fig. 5, under the simulation conditions described above, choosing $\beta \approx 0.45$ gives the lowest value of V, namely, $V=0.3978$.

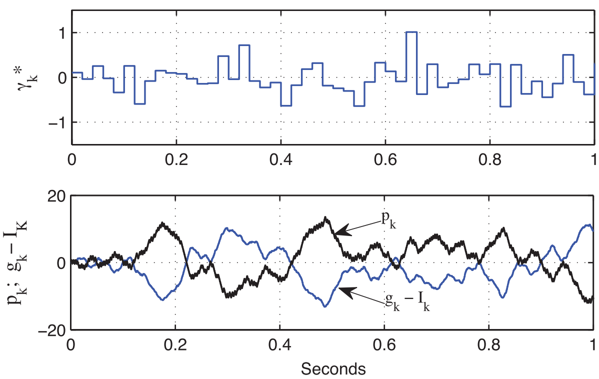

Fig. 4. Current WCDMA power control scheme; see Fig. 2 with $C_k(z)=1$ and $\beta=0.39$.

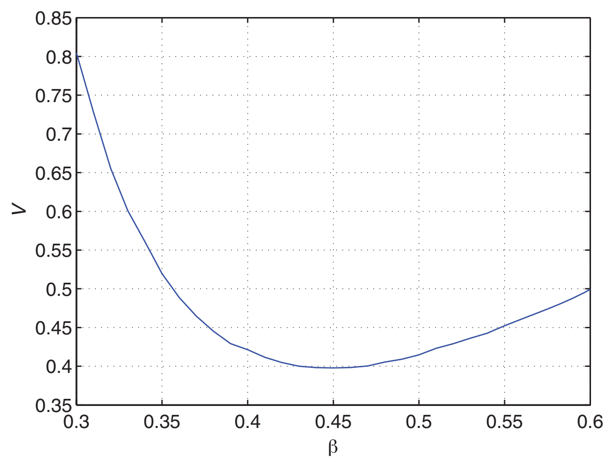

Fig. 5. Error variance V of the current WCDMA power control algorithm as a function of β; see Fig. 2 with $C_k(z)=1$.

We next keep $C_k(z)=1$ and $\beta=0.45$ and synthesise filters following our result in (10). We adopt a simple design model which neglects the effects of ε_k, I_k, and $\gamma_k^\star$ and only uses knowledge of g_k as provided in (17). We furthermore limit filters to be of order 6 to obtain

$$F_k^{-1}(z) = B_k \frac{(z-0.997)(z-0.812)(z-0.363)}{(z+0.838)(z+0.56)(z-0.913)} \cdot \frac{(z+0.914)(z+0.728)(z+0.311)}{(z-0.99)(z-0.634)(z-0.0206)}, \tag{19}$$

which constitutes a rational approximation of (10). Its frequency response is almost identical to that given in (10), and is included in Fig. 6.

The value B_k in (19) is a tuning parameter, which allows the designer to tradeoff *granular* versus *overload*/saturation quantisation noise. To study the effect of B_k on closed-loop performance, we use the simulation model that gave rise to Fig. 4. Results are documented in Fig. 7, which illustrates V as a function of B_k. For comparison purposes, in this figure we have also included, via a dashed line, the value achieved with the current power control strategy (*i.e.*, without filtering). Clearly, filtering, as proposed in the present work, outperforms the current power control strategy if $0.29 \le B_k \le 0.56$. In particular, if $B_k=0.41$, then $V=0.3347$ is obtained. This corresponds to an improvement of about 16% when compared with the existing power control strategy. It is worth noting that the filters in (19) were obtained by using a simplified design model, which differs from the simulation model used.

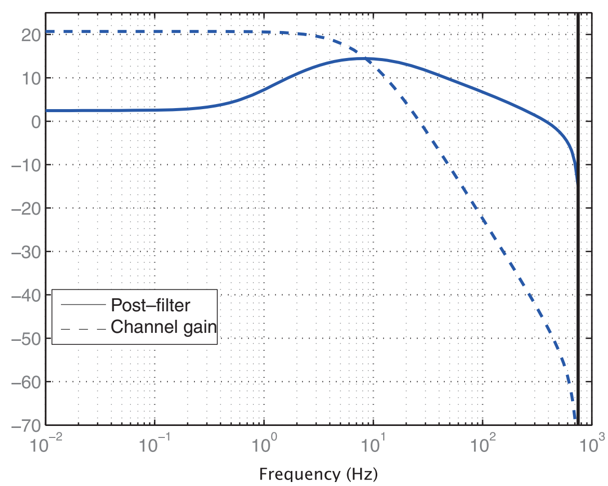

Fig. 6. Frequency response (dB) of the channel gain in (17) (dashed line) and of the post-filter in (19) (solid line).

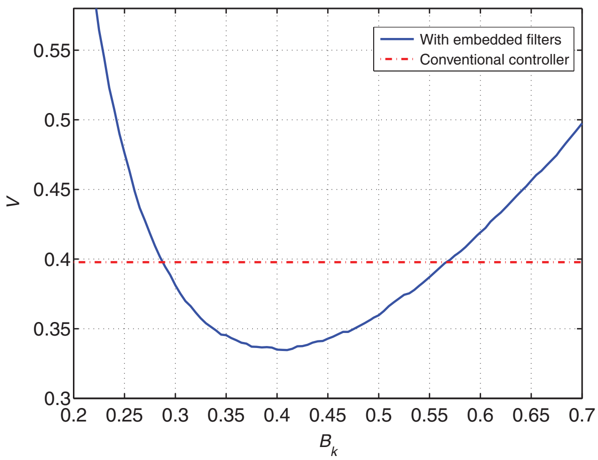

Fig. 7. Error variance V of the current control algorithm (dashed line) and of the proposed method for WCDMA power control (solid line) as a function of B_k; $\beta=0.45$.

To further illustrate the robustness of our design with respect to model uncertainties, we will next change the channel gain in the simulation model from (17) to

$$g_k(\ell)=\left(\frac{(z+0.94)}{150(z-0.98)(z-0.96)}\right)v_k(\ell). \qquad (20)$$

Use of the conventional power control method with $\beta=0.45$ gives $V=0.4718$. On the other hand, the proposed control algorithm with $\beta=0.45$ and with $F_k^{-1}(z)$ as in (19) where $B_k=0.41$ provides $V=0.3621$. Thus, in this case, the performance gained by using the embedded filters is approximately 23%. (Note that the filter in (19) was designed for the channel gain model (17) and not for (20)).

This last result illustrates that the method proposed in the present work gives rise to robust control designs. It is a strong indication that signaling of filter coefficients from the base station to the mobile station as side information may not be required, at least not with a high rate.

V. IMPLEMENTATION ASPECTS

The implementation of the scheme proposed requires knowledge of the models appearing in the design equations of Theorem 1. The nominal controllers $C_k(z)$ are readily determined from knowledge of the loop delays. The spectral densities associated with γ_k, $\gamma_k^\star$, g_k, I_k, and ε_k can all be estimated in the base station. In particular, $\gamma_k^\star$ is easily estimated in the base station, whereas the channel gain g_k is available from baseband channel estimation. It is more difficult to estimate the neighbour cell interference. One approach to solve this problem can be found in [26, 27]. In essence, these algorithms first estimate the sum of neighbour cell interference and the noise power floor, followed by a subtraction of the thermal noise power floor from the sum, thereby obtaining an estimate of the neighbour cell interference. It can be noted that this and related methods are applied in the cellular field nowadays. Finally, the SIR measurement error can also be easily estimated in the base station.

Given the above, all information needed to compute the filters $F_k(z)$ and $F_k^{-1}(z)$ is available in the BS. However, the filter $F_k^{-1}(z)$ is also needed in the MS, see Fig. 3. Two approaches are then possible, of which one was discussed in the simulation study. This first method designs a nominal fixed filter $F_k(z)$ that the MS is using. As illustrated in the simulation study, though simple, this approach can still give significant performance gains. The second method involves sending down-link side information in the form of the filter $F_k(z)$. Here, further work is needed to find the best solution in terms of bandwidth, accuracy and the bearer of the information. It can be noted that, with the development of mobile broadband in WCDMA, opportunities to signal this information exist both in the control and in the user planes. In case of new standards, the information can be built into the power control signaling.

VI. CONCLUSIONS

We have presented a novel concept for fast inner-loop power control for CDMA wireless communication systems. The method amounts to incorporating pre- and post-filters. These filters are designed to mitigate quantisation effects by using a linear quantisation model and taking into account feedback of the quantisation error. Simulation studies indicate that our approach gives rise to robust designs. Performance gains in logarithmic scale are in the order of 10% to 30%, when compared with the current WCDMA control algorithm.

Future work may include studying how much side information on filter coefficients needs to be transmitted to achieve desired performance levels. We foresee that obtaining such a characterisation would require more extensive simulation studies. It would also be of interest to develop power control configurations which take into account both quantization and saturation of the power control signal. For that purpose, quantized model predictive controllers, as presented for example in [28, 29], could be investigated.

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