Predictive Downlink Beamforming for Wideband CDMA Over Rayleigh Fading Channels

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Abstract—A new approach to adaptive downlink beamforming to combat fast Rayleigh fading is presented. In this approach, the antennas at the base transceiver station form transmit beam patterns according to the prediction of downlink channels. The channel prediction is a linear prediction based on the autoregressive model, which is downsampled to extend the memory span to a fixed model order. For wideband CDMA downlink, pre-RAKE transmission is employed to achieve the multipath diversity gain. In particular, we combine pseudoinverse DOA beamforming with pre-RAKE transmission to alleviate self interference. The beamforming weights are adjusted within a downlink frame to compensate the predicted fading. We give measures of the prediction and beamforming performance, and evaluate the impact of prediction errors on the downlink. Ray tracing simulations in a 3D urban physical model show that the predictive downlink beamforming outperforms the conventional beamforming over Rayleigh fading channels.

Index Terms—Mobile communication, array signal processing, channel prediction, pre-RAKE.

I. INTRODUCTION

Spatial diversity, obtained through antenna array processing, provides an attractive means to improve the performance of wireless communication systems (see [1], [2] and the references therein). When the base transceiver station (BTS) is equipped with an antenna array, the transmitted signals are properly weighted on each array element in the downlink (base to mobile). The downlink beamforming technique endeavors to enhance the desired signal received at the mobile station (MS) while suppressing the interference. The common interference in a wireless communication cell includes self interference (SI) and multiple-access interference (MAI). In code-division multiple-access (CDMA) systems, the MAI can be effectively reduced by the orthogonal spreading codes [3]. As the MAI is coped with by CDMA despreading, the beamforming effort herein is focused on combating the SI due to multipath fading. The beamforming algorithms that have been developed to date adapt only to slowly varying components in the channel and do not track the Rayleigh fading. However, increased mobility results in fast Rayleigh fading, such that the channel exhibits rapid temporal variation and the received signal may fade within a data frame. This problem is mitigated by the adaptive beamforming approach proposed in this paper, in which the beamforming weights are adjusted according to the predicted channel fading.

Recently, researchers have speculated that the fading process is indeed a deterministic sinusoidal process with time-varying parameters, and can be characterized using a discrete scatter propagation model [4]. A modified root-MUSIC algorithm is used in [5], and an ESPRIT-type algorithm is used in [6] and [7] to estimate the dominant sinusoids that compose the fading channel. However, these subspace-based methods are computationally intensive. The linear predictability of the fading mobile channel is reported in [8] and [9], in which the channel is modeled as a wide-sense stationary autoregressive (AR) process. With an antenna array at the BTS, the vector channel modeling and prediction that use AR models are introduced in [10]. In this paper, AR modeling followed by linear extrapolation is applied to predict the evolution of the vector channel. Once the AR coefficients are obtained, they are applied to the past and present uplink channel vectors to predict the future downlink in time-division duplex (TDD) systems, or applied to the feedback of the past and present downlink channel vectors in frequency-division duplex (FDD) systems. The predicted channel vectors can be exploited to generate downlink beamforming weights. For the purpose of adaptive beamforming, the downsampling of channel vectors to the beamforming rate effectively extends the memory span given fixed model order, and reduces the additional bandwidth for channel feedback in FDD systems.

The channel prediction and beamforming discussed so far are for the flat fading channels. However, wideband systems experience frequency-selective fading, where the multipath delay spread is large with respect to the symbol period. Assuming slow fading conditions, a pre-RAKE transmitter has been proposed for a TDD system in [11]. The pre-RAKE transmitter simplifies the receiver structure while preserving the RAKE performance. In [12], a smart antenna system has combined the advantages of pre-RAKE multipath diversity and transmit antenna spatial diversity. A pre-RAKE transmitter with long range prediction is employed in [13] for systems over rapidly varying multipath fading channels. These pre-RAKE methods utilize the simplified tapped delay line multipath channel model [14], and require the knowledge of the channel state information for every symbol transmitted. However, using the tapped delay line model, the tap number of the channel filter, thus the tap number of the transmit and receive transversal filter, can be formidable. In this paper, a pre-RAKE transmission cooperates with the predictive beamforming for...
wideband signals, and a parametric model is used for the channel. The number of parameters in the pre-transmission combining is moderate. The pre-RAKE beamforming only requires the predicted channel vectors at the beamforming rate, which is much lower than the symbol rate. No channel sample interpolation is needed. This imposes fewer burdens on channel prediction. Furthermore, a pseudoinverse DOA combining is moderate. The pre-RAKE beamforming only requires the predicted channel vectors at the beamforming combining is moderate. The pre-RAKE beamforming only requires the predicted channel vectors at the beamforming rate, which is much lower than the symbol rate. No channel requires the predicted channel vectors at the beamforming rate, which is much lower than the symbol rate. No channel requires the predicted channel vectors at the beamforming rate, which is much lower than the symbol rate. No channel

This paper is organized as follows. In Section II, the space-time data processing of an adaptive antenna system is described through a wideband CDMA uplink. In Section III, the prediction of downlink channels in TDD and FDD systems is presented, along with the discussion on model order selection and prediction performance. In Section IV, the predictive downlink beamforming algorithm for wideband systems is proposed. In Section V, a CDMA system over Rayleigh fading channels is simulated using ray tracing in a 3D urban physical model of downtown Austin, Texas. Numerical results reveal the robustness of the proposed beamforming to rapid channel variations. Finally, conclusions are drawn in Section VI.

II. SPACE-TIME DATA MODEL

Consider K active MSs in a wireless communication cell. These K MSs communicate simultaneously with a BTS using a common uplink carrier frequency and a common downlink carrier frequency. The BTS has an M-element antenna array and each MS has one antenna, hence a vector channel between the BTS and an individual MS [10]. During uplink, the baseband signal received at the BTS can be expressed as

\[ x(t) = \sum_{k=1}^{K} \sum_{i=1}^{L_k} a_{k,i}^{(u)}(t) \sqrt{p_k^{(w)}(t)} s_k(t) + n(t) \]  

where superscript \((u)\) denotes uplink, and

\( L_k \) total number of resolvable paths from MS \( k \);

\( a_{k,i}(t) \) time-varying channel vector of the \( i^{th} \) path from MS \( k \);

\( \tau_{k,i} \) time delay of the \( i^{th} \) path from MS \( k \);

\( p_k(t) \) transmit power of MS \( k \);

\( s_k(t) \) transmitted signal of MS \( k \);

\( n(t) \) receiver background noise vector, each element of which is an independent AWGN with one-sided spectrum density \( N_0 \).

Assume that the BTS employs a RAKE receiver with its fingers synchronized to each resolvable path. A resolvable path consists of a cluster of temporally unresolved path components, which may have different directions of arrival (DOA). Suppose there are \( L_k \) unresolved subpaths which contribute significantly to the \( i^{th} \) path, the channel vector \( a_{k,i}(t) \) is given by

\[ a_{k,i}(t) = \sum_{j=1}^{L_k} \alpha_{k,i,j}(t) v(\theta_{k,i,j}) \]  

where \( v(\theta_{k,i,j}) \) is the array response associated with DOA \( \theta_{k,i,j} \), and \( \alpha_{k,i,j}(t) \) is the complex amplitude as (omitting \( k \)

\[ \alpha_{j}(t) = \rho_j e^{j2\pi(f_j t - (f_c + f_j) \tau_j)} \]  

where \( f_c \) is the carrier frequency, \( \rho_j \) and \( \tau_j \) are the complex gain and the path delay of the \( j^{th} \) subpath, respectively. The Doppler frequency \( f_j \) is caused by the motion of the mobile or the scatterers. It is given by \( f_j = f_c v cos(\psi_j) \), where \( v \) is the relative speed of the mobile, \( C \) is the speed of electromagnetic propagation, and \( \psi_j \) is the angle between the direction of departure of the \( j^{th} \) subpath at the mobile and the mobile moving direction. The channel parameters \( \theta_{k,i,j} \), \( \rho_j \) and \( \tau_j \) are assumed to be fixed over a short period of time containing several hundreds of symbols, during which the Rayleigh fading contributes dominantly to channel variation compared with path loss or shadowing effects. The Rayleigh fading is a result of destructive combining of the multipath components due to the non-zero Doppler frequencies in \( \{ \alpha_{j}(t) \} \). All Doppler frequencies \( \{ f_j \} \) are assumed to remain constant over the time of interest. The transmitted signal \( s_k(t) \) depends on the information-bearing bit stream \( b_k(n) \) and the modulation waveform \( g_k(t) \) as

\[ s_k(t) = \sum_{n} b_k(n) g_k(t - nT_s) \]  

where \( T_s \) is the symbol period. The waveform \( g_k(t) \) is the convolution of the pulse-shaping filter and the CDMA orthogonal spreading code, which can be scrambled by a pseudo random long code [3]. The variation of the channel vector \( a_{k,i}(t) \) over a symbol period \( T_s \) is assumed to be negligible.

Suppose a matched filter with perfect synchronization is present at each finger of the RAKE receiver, then the sampled output of the \( i^{th} \) finger is

\[ y_{k,i}(n) = \int_{(n-1)T_s + \tau_{k,i}}^{nT_s + \tau_{k,i}} g_k(t - nT_s - \tau_{k,i}) x(t) dt \]

\[ = \sqrt{G p_k a_{k,i}^H(n)} b_k(n) + \tilde{n}_i(n) \]  

where \( G \) is the processing gain, \( \tilde{n}_i(n) \) is the undesired component consisting of SI, MAI and Gaussian noise. Given that the channel vectors are known to the receiver, the matched filter outputs can be coherently summed up using maximal ratio combining (MRC) as

\[ r_k(n) = \sum_{i=1}^{L_k} a_{k,i}^H(n) y_{k,i}(n) \]

\[ = \sqrt{G p_k b_k(n)} \sum_{i=1}^{L_k} ||a_{k,i}(n)||^2 + \eta_k(n) \]  

where \( H \) denotes conjugate transpose, and \( \eta_k(n) = \sum_{i=1}^{L_k} a_{k,i}^H(n) \tilde{n}_i(n) \). We assume that the CDMA modulation waveforms of the interfering users or with different delays appear as mutually uncorrelated noise, and the interference is uncorrelated between the antennas. With such assumptions that \( \tilde{n}_i(n) \) is temporally and spatially white, MRC maximizes the signal-to-noise-ratio (SNR) at the receiver [16].
III. DOWNLINK CHANNEL PREDICTION

Let us consider the asymmetry between the uplink channel and the downlink channel. The path delay $\tau_{k,i}$ of the downlink channel does not differ from that of the uplink channel. However, in TDD systems, the downlink channel vector $a_{k,i}^{(d)}(t)$ can vary drastically from the uplink channel vector $a_{k,i}^{(u)}(t)$ over a period of duplex time under fast Rayleigh fading, and the signal may exhibit deep fades within a frame. In FDD systems, not only does the channel vector vary rapidly, but also the uplink and downlink channels have independent fading. Therefore, the prediction of future downlink channel is necessary at the BTS for adaptive transmission over fast Rayleigh fading channels.

A. Channel Prediction in TDD Systems

In TDD systems with the same carrier frequency, radio channels for uplink and downlink are reciprocal. The uplink channel vector $a_{k,i}^{(u)}(t)$ is extracted from the received signal at the BTS and can be used to predict the future downlink channel. Hereafter, the subscripts $k$ and $i$ are omitted with the understanding that the $i$th path to MS $k$ is under consideration. Assuming perfect detection of the uplink symbol bit $b(n)$, the BTS can extract the uplink channel vector at symbol rate from the matched filter output $y(n)$ by applying

$$b^*(n)y(n) = \sqrt{G_{bk}}a(n) + u(n)$$

(7)

where $u(n) = b^*(n)a(n)$ is the interference-plus-noise component. In this paper, the $K$ MSs are assumed to transmit with equal and constant power during uplink, therefore the estimate $\hat{a}(n)$ of $a(n)$ can be derived from (7) by normalizing the gain. If the MSs transmit at different power levels to compensate the near-far effect in CDMA systems, the estimate can be normalized to $\sqrt{p_b}a(n)$, which will not affect the beamforming result. According to (2) and (3), the channel vector $a(n)$ can be written as

$$a(n) = \sum_{j=1}^{L_s} v(\theta_j) e^{-j2\pi(f_j T_s + f_l)T_s} e^{j2\pi f_j n T_s}$$

(8)

Clearly, the estimate of $a(n)$ is of harmonic form that consists of a superposition of complex exponentials in noise. As a result, sharp peaks are the predominant feature of the power spectrum of the channel estimates. This justifies an all-pole model for the channel. Furthermore, as revealed in (8), the signals received at each of the $M$ BTS antennas, though combined with different phases, experience the same Doppler frequency shifts. Therefore, the coefficients of the all-pole models are the same for channel estimates at each antenna. Assuming wide-sense stationary, we model the channel vector using a $p^{th}$ order Gaussian autoregressive (AR($p$)) process as

$$\hat{a}(n) = - \sum_{l=1}^{p} q_l^* \hat{a}(n-l) + e(n)$$

(9)

where $\{q_l \mid l = 1, \ldots, p \}$ are the AR coefficients, and $e(n)$ is a white Gaussian sequence. $e(n)$ is also assumed spatially white. The locations of the sharp peaks in spectrum correspond to the poles near the unit circle as $z_j' = e^{j2\pi f_j T_s}$, which are determined by the Doppler frequencies of the $p$ dominant subpaths. We use $q^*$ and $z'$ to distinguish them from the AR coefficients and channel poles described in the following.

For the purpose of downlink beamforming, it only needs to predict the downlink channel vectors at the time instants when the beamforming weights are adjusted. As the rate of adjusting beamforming weights is much lower than the symbol rate, the uplink channel samples can be decimated to form the observation interval for downlink prediction. Consequently, the memory span for prediction can be much longer given a fixed model order. The decimator consists of a downsampler preceded by a low-pass filter, which functions as an anti-aliasing filter. We bandlimit the uplink channel samples to a frequency band which includes the expected Doppler frequencies and downsample the resulting data to the beamforming rate. The beamforming rate is usually higher than twice the maximum Doppler frequency. The downsampled version of the Gaussian AR($p$) process can be written as

$$\hat{a}(n) = - \sum_{l=1}^{p} q_l \hat{a}(n - lL_b) + e(n)$$

(10)

where $L_b$ is the beamforming symbol span, within which the data will be transmitted using fixed weights, hence the beamforming rate $f_B = \frac{1}{T_{DB}}$. The AR coefficients $q_p = [1, q_1, \ldots, q_p]^T$ can be determined as the least square solution to the linear equations [17]

$$A q_p = 0$$

(11)

where matrix $A$ is constructed as

$$\begin{bmatrix}
\hat{a}(N) & \hat{a}(N-L_b) & \cdots & \hat{a}(N-pL_b) \\
\hat{a}(N-1) & \hat{a}(N-1-L_b) & \cdots & \hat{a}(N-1-pL_b) \\
\vdots & \vdots & \ddots & \vdots \\
\hat{a}(1+pL_b) & \hat{a}(1+(p-1)L_b) & \cdots & \hat{a}(1) \\
\hat{a}^*(1) & \hat{a}^*(1+L_b) & \cdots & \hat{a}^*(1+pL_b) \\
\vdots & \vdots & \ddots & \vdots \\
\hat{a}^*(N-pL_b) & \hat{a}^*(N-(p-1)L_b) & \cdots & \hat{a}^*(N)
\end{bmatrix}$$

with $^*$ denoting complex conjugate. \{$\hat{a}(1), \hat{a}(2), \ldots, \hat{a}(N)$\} are the estimated uplink channel vectors after low-pass filtering, where $N$ is restricted by the total number of symbols within one uplink frame. As shown in the structure of $A$, $p+1$ channel vectors with a gap of $L_b$ symbol periods between any adjacent two are selected to form $M$ linear equations. There are $N-pL_b$ groups of such $p+1$ channel vectors as the first vector of each group can be chosen from $\hat{a}(1)$ to $\hat{a}(N-pL_b)$. The forward and backward linear prediction is applied to ensure the Toeplitz structure of the normal equations along with the guaranteed stability of the AR model [17]. Given the AR coefficients, the downlink channel vectors can be derived from the current uplink using a tapped-delay-line configuration, which makes one-step linear prediction via an FIR filter as

$$\hat{a}(n) = - \sum_{l=1}^{p} q_l \hat{a}(n - lL_b)$$

(12)
Suppose that \( \hat{a}(N) \) is the last estimated channel vector of the uplink frame, the downlink channel vectors \( \{a(n)\} \) at \( n = N + L_b, N + 2L_b, \ldots \) are to be predicted for beamforming. The one-step predictor gives \( \hat{a}(N + L_b) \) from the observed uplink data. However, for \( \{\hat{a}(n)\} \) at \( n = N + 2L_b, N + 3L_b, \ldots \), the previously predicted values are needed to feed the FIR filter. This causes error propagation in the prediction of future downlink channel; this effect is included in the simulation.

### B. Channel Prediction in FDD Systems

In order to proceed with the linear prediction described above, both the AR coefficients and the previous and current channel vectors are required by the prediction filter. In FDD systems, the instantaneous channels of uplink and downlink are uncorrelated when the carrier frequency separation is larger than the channel coherence bandwidth [14]. Therefore, the uplink channel vectors \( \hat{a}^{(u)}(n) \) can not be extrapolated to approximate the future downlink channel vectors \( a^{(d)} \). Nevertheless, the prediction coefficients of the downlink channel model can be obtained by modifying the AR coefficients of the uplink channel model.

The Gaussian AR(p) process in (10) can be expressed as

\[
\begin{bmatrix}
\hat{a}^{(u)}(n) \\
\hat{a}^{(u)}(n - L_b) \\
\vdots \\
\hat{a}^{(u)}(n - pL_b)
\end{bmatrix} q^{(u)}_p = e(n)
\] (13)

Therefore, the estimates of channel vectors \( \hat{a}^{(u)}(n) \) can be modeled as the output of an all-pole linear filter driven by white noise \( e(n) \). The transfer function of this linear filter is given by

\[
H(z) = \frac{1}{A(z)} = \frac{1}{1 + \sum_{i=1}^{p} q^{(u)}_i z^{-i}}
\] (14)

The AR coefficients \( \{q^{(u)}_i, i = 1, \ldots, p\} \) and the channel poles \( \{z^{(u)}_i, i = 1, \ldots, p\} \) are directly related, since \( \{q^{(u)}_i\} \) and \( \{z^{(u)}_i\} \) are the polynomial parameters and the roots of \( A(z) \), respectively. The channel poles are given by

\[
z^{(u)}_i = e^{j2\pi f^{(u)}_i L_b T_s} = e^{j2\pi f^{(u)}_i / f_B}
\] (15)

Recall that the Doppler frequency \( f^{(u)}_i = f^{(d)}_i v / c \cos(\psi_i) \). The relative speed \( v \) and \( \psi_i \) at the MS do not change from uplink to downlink, whereas the carrier frequencies \( f^{(u)}_i \) and \( f^{(d)}_i \) are different in FDD. So the Doppler frequency is different in uplink and downlink, hence the difference in uplink and downlink channel poles. Fast fading channels encountered in practice exhibit Doppler spreads on the order of \( 100 - 200 \) Hz [18]. Usually the beamforming rate \( f_B \gg f_i \). For example, adjusting beamforming weights 8 times evenly within a 2.5 ms transmission frame results in a beamforming rate of 3.2 KHz. Therefore, with moderate changes in Doppler frequencies, there is no \( 2\pi \)-phase ambiguity of the channel poles on the complex plane. Let \( \Phi(\phi) \) denote the phase of a complex value, such that \( \Phi(z_i) = 2\pi f^{(u)}_i / f_i \). The uplink channel poles can be transformed to the downlink channel poles by modifying the phase components as

\[
\Phi(z^{(d)}_i) = \frac{f^{(d)}_i}{f^{(u)}_i} \Phi(z^{(u)}_i)
\] (16)

Finally, the AR coefficients of the downlink channel \( q^{(d)}_p \) are derived from the downlink channel poles \( \{z^{(d)}_i, i = 1, \ldots, p\} \).

In addition, in a practical personal communication system with FDD implementation, the carrier frequencies are high compared with the carrier frequency separation. For carrier frequencies up to several gigahertz and the carrier separation tens or hundreds of megahertz, \( f^{(d)}_i / f^{(u)}_i \approx 1 \). Therefore, with the assumption that each element of the BTS array has identical antenna patterns at uplink and downlink frequencies, the AR coefficients of the uplink channel model can be directly applied to downlink channel prediction without any modification.

As mentioned, one issue that complicates channel prediction in FDD is the lack of direct measurement of current downlink channel at the BTS. MS probing-feedback can be implemented to provide current downlink channel vectors [19]. The feedback of the downlink channel has a time delay for current BTS transmission, which justifies the necessity of channel prediction. As the pre-RAKE transmission is applied along with fading compensation on each branch (in the following section), the MS can only extract a single fade-compensated channel which is the composite of several individually weighted paths. Therefore, a pilot signal with omnidirectional beamforming weights needs to be inserted in the downlink, assuming that the MSs can synchronize to the periodic arrivals of the pilot. The pilot insertion rate and the probing feedback rate can be as low as the beamforming rate, because only the channel vectors at these positions are required by the linear predictor. Many researches have been conducted on the pilot-assisted channel estimation [20] and modulation [21], and on the effects of quantized or imperfect feedback [22], [23]. The discussion on downlink pilot and feedback channel is beyond the scope of this paper. In the simulation, we assume perfect channel estimation at the MSs and error-free feedback.

### C. Model Order Selection

If \( \{\hat{a}(1), \hat{a}(2), \ldots, \hat{a}(N)\} \) are estimated to construct matrix \( A \) of the linear equations (11), the model order \( p \) is restricted to \( p \leq (N - 1)/L_b \). Suppose that, in (13), \( e(n) \) is drawn i.i.d. by the complex multivariate Gaussian distribution of zero mean and covariance matrix \( C_e = \sigma_e^2 I \). The model order \( p \) is selected, therefore in the range \( 0 < p \leq (N - 1)/L_b \), to minimize the total squared prediction error as

\[
p^* = \arg\min_{0 < p \leq (N - 1)/L_b} \sum_{n=pL_b+1}^{N} ||e(n)||^2
\] (17)

Let \( q_1, \ldots, q_p, \sigma_e^2 \) comprise the AR(p) parameter vector \( \Theta(p) \). Regarding the prediction errors as spatially white, independent vectors with zero mean, the joint probability density is given by

\[
f(e(pL_b+1), \ldots, e(N); \Theta(p)) = \prod_{i=pL_b+1}^{N} \frac{1}{(\pi\sigma_e^2)^{M}} e^{-\frac{1}{\sigma_e^2} e^{H}(i)e(i)}
\] (18)

where \( M \), the dimension of the vector, is the number of antennas at the BTS. The optimum model order \( p^* \) can be found
by selecting the model which minimizes the standard Akaike information criterion (AIC) [24], or the minimal description length criterion (MDL) [25]. After some manipulation, these criteria are found as

\[
\text{AIC}(p) = 2M(N - pL_b)(\ln(\pi r_{00} - r_0^H R^{-1} r_0)) + 1 + 2(2p + 1)
\]

\[
\text{MDL}(p) = M(N - pL_b)(\ln(\pi r_{00} - r_0^H R^{-1} r_0)) + 1 + \frac{2p + 1}{2} \ln(N - pL_b)
\]

where

\[
r_0 = [r_{10}, \ldots, r_{p0}]^T, \quad R = \begin{bmatrix} r_{11} & r_{12} & \cdots & r_{1p} \\
r_{21} & r_{22} & \cdots & r_{2p} \\
\vdots & \vdots & \ddots & \vdots \\
r_{p1} & r_{p2} & \cdots & r_{pp} \end{bmatrix},
\]

and

\[
r_{kl} = \frac{1}{M(N - pL_b)} \sum_{i = pL_b + 1}^{N} \hat{a}^H(i - kL_b)\hat{a}(i - lL_b).
\]

D. Prediction Performance Metrics

Assume that each MS transmits at unit power level, that is, \(p_k = 1\) for \(k = 1, 2, \ldots, K\). Let \(\mathbf{a}(d)\) and \(\mathbf{a}^{(d)}\) denote the true and the predicted downlink channel vectors, respectively. The normalized root mean square error (RMSE) of channel prediction is defined as

\[
\text{RMSE}(n) = \sqrt{\frac{\text{E}||\mathbf{a}(d)(n) - \mathbf{a}^{(d)}(n)||^2}{\text{E}||\mathbf{a}(d)(n)||^2}}
\]

(21)

We use this metric to measure the quality of channel prediction.

The time-variation of the vector channel can be measured by the relative amplitude change and the relative angle change as defined in [10]. The relative amplitude change of the channel vectors is usually small, because it is quite unlikely that all vector components vanish simultaneously. However, the relative angle change of the channel vectors can range from 0 to 1 within a transmission frame over fast Rayleigh fading, with “0” indicating two vectors parallel and “1” indicating two vectors orthogonal [10]. When beamforming is the application of vector channel prediction, we incorporate the relative amplitude with the relative angle between the true and the predicted channel vectors, and define the prediction error ratio \(\gamma(n)\) as

\[
\gamma(n) = \frac{||P_{\mathbf{a}(d)(n)}\mathbf{a}(d)(n)||}{||\mathbf{a}(d)(n)||} = \frac{||\mathbf{a}(d)^H (n)\mathbf{a}(d)(n)||}{||\mathbf{a}(d)(n)||^2}
\]

(22)

where \(P_{\mathbf{a}(d)(n)}\) denotes the orthogonal projection onto vector \(\mathbf{a}(d)(n)\). As shown later, in the single-path case with fading compensation, the received signal power at the MS is proportional to \(\gamma^2(n)\). We use this metric to evaluate the impact of prediction errors on downlink beamforming.

IV. PREDICTIVE DOWNLINK BEAMFORMING

The predicted downlink channel vectors \(\{\hat{a}_{k,i}(d)(n), n = N + L_b, N + 2L_b, \ldots\}\) can be used to generate BTS antenna weights that enhance the desired signal at each MS. The transmit beam pattern changes at time instants \(n = N + L_b, N + 2L_b, \ldots\) within one downlink frame, such that the MS reception is robust to channel variation under fast Rayleigh fading. In wideband systems, the predicted channel vectors \(\{\hat{a}_{k,i}(d)(n)\}\) of all paths are incorporated in beamforming to achieve MRC at the receiver.

A. Pre-RAKE Transmission with Beamforming

RAKE receivers have been used in wideband systems to achieve multipath diversity gain, and conventionally they are implemented at both the BTS and the MS. In order to preserve the simplicity of MS receiving, the BTS employs the pre-RAKE transmission. The RAKE combining takes effect in the BTS transmission, and each MS requires only a single receiver finger. With the transmit weight vector \(w_{k,i}\) being assigned to the \(i^{th}\) path reaching MS \(k\), the BTS transmits a signal that consists of copies of user data with the time leads corresponding to the path delays as

\[
s^{(d)}(t) = \sum_{k=1}^{K} \sum_{i=1}^{L_k} w_{k,i}^H a^{(d)}(t + \tau_{k,i})e^{j2\pi f_c \tau_{k,i}}
\]

(23)

where \(s^{(d)}(t)\) is the desired signal of MS \(k\), and \(\tau_{k,i}\) is the delay of the \(i^{th}\) path reaching MS \(k\). The cophasing term \(e^{j2\pi f_c \tau_{k,i}}\) in (23) is negligible, because \(e^{j2\pi f_c \tau_{k,i}}\approx 1\) for practical Doppler frequencies and multipath delays. Another term, \(e^{j2\pi f_c \tau_{k,i}}\), in radio transmission with carrier frequency \(f_c\), is included in the predicted channel vector \(\hat{a}_{k,i}(d)\). As \(\{w_{k,i}\}\) are generated according to \(\{\hat{a}_{k,i}(d)\}, \{w_{k,i}^H\}\) fulfill cophasing. Therefore, the received signal at MS \(k\) is given by

\[
x_k(t) = \sum_{i=1}^{L_k} s^{(d)}(t - \tau_{k,i})a^{(d)}_{k,i}(t) + n_k(t)
\]

(24)

where \(D_k\), \(S_k\), and \(A_k\) are the desired signal part, the SI, and the MAI, respectively, and \(n_k(t)\) is the zero-mean AWGN with one-sided spectrum density \(N_0\).

In CDMA systems, \(A_k\) can be alleviated by matched filtering at the MS receiver. However, \(S_k\) will produce time-dispersed peaks at the output of the matched filter. In order to
suppress $S_k$, while achieving MRC of the desired signal $D_k$, the downlink beamforming weights $\{w_{k,i}\}$ need to satisfy
\[
\sum_{i=1}^{L_k} w_{k,i}^H(n) a_{k,i}^{(d)}(n) = \sum_{i=1}^{L_k} ||a_{k,i}^{(d)}(n)||^2
\]
\[
w_{k,j}^H(n) a_{k,i}^{(d)}(n) = 0, \quad i \neq j
\]
where $\{a_{k,i}^{(d)}(n)\}$ are the downlink channel vectors at discrete time $n$ occurring at the beamforming rate. In (26), we assume that the number of antennas $M$ is no less than the number of paths $L_k$, and apply the pseudoinverse DOA beamforming [15] to eliminate each component in $S_k$. Based on the predicted downlink channel vectors $\{\tilde{a}_{k,i}(n)\}$, the beamforming weights can be generated as
\[
w_{k,i}(n) = \frac{\beta}{\sum_{i=1}^{L_k} ||\tilde{a}_{k,i}(n)||^2} \frac{||\tilde{a}_{k,i}(n)||^2 (I - P_i) \tilde{a}_{k,i}^{(d)}(n)}{||(I - P_i) \tilde{a}_{k,i}^{(d)}(n)||^2}
\]
where $\beta$ is a constant that takes into account the transmit power constraint. In the following, we assume that $\beta = 1$. The normalization term at time $n$ is the same for all weight vectors assigned to the signal for MS $k$, hence guaranteeing the MRC. We assume that the channel vectors $a_{k,j}$ and $\tilde{a}_{k,j}$ are not close in the vector space if the corresponding paths have different delays. Therefore, the denominator $||(I - P_i) \tilde{a}_{k,i}^{(d)}(n)||^2$ will not be too small that causes processing error.

The dynamic beamforming scheme is as follows. Within a downlink frame, the BTS transmits signal $s^{(d)}(t)$ as in (23) using a set of weight vectors $\{w_{k,i}(n), n = N + L_k, N + 2L_k, \ldots\}$ as in (28) at $t \in \left(\frac{n - \frac{L_k}{2}}{T_s}, (\frac{n + \frac{L_k}{2}}{T_s})\right)$. Note that, at the beginning of the downlink transmission, $t \in (NT_s, (N + \frac{L_k}{2})T_s)$, the weight vectors are generated based on the estimated or fed-back channel vectors $\{a_{k,i}^{(d)}(N)\}$.

B. Performance Analysis

As the SI is suppressed by the proposed beamforming, MS $k$ can synchronize its receiver finger to the desired signal. The output of the matched filter is then given by
\[
y_k(n) = \int_{(n-1)T_s}^{nT_s} g_k(t - nT_s)x_k(t)dt
\]
\[= D_k(n) + S_k(n) + A_k(n) + \eta_k(n)
\]
where $D_k(n)$, $S_k(n)$, $A_k(n)$, and $\eta_k(n)$ are respectively the desired signal, the SI, the MAI, and the noise component of the output of the matched filter. Assuming that the channel vector $a_{k,i}^{(d)}(t)$ is fixed over a symbol period $T_s$, and that the CDMA modulation waveform $g_k(t)$ has negligible overlapping across symbols, we have
\[
D_k(n) = \sqrt{G}b_k(n) \sum_{i=1}^{L_k} w_{k,i}^H(n') a_{k,i}^{(d)}(n)
\]
\[= \sum_{i\neq j}^{L_k} \sum_{i,j}^{L_k} \int_{(n-1)T_s}^{nT_s} g_k(t - nT_s) g_k(t - mT_s + \tau_{k,j} - \tau_{k,i})dt
\]
\[\sum_{i=1}^{K} \sum_{i,j}^{L_k} \int_{(n-1)T_s}^{nT_s} g_k(t - nT_s) g_k(t - mT_s + \tau_{i,j} - \tau_{k,i})dt
\]
\[\sum_{i=1}^{K} \sum_{i,j}^{L_k} \int_{(n-1)T_s}^{nT_s} g_k(t - nT_s) g_k(t - mT_s + \tau_{i,j} - \tau_{k,i})dt
\]
where $n' = N + \left[\frac{n - N + L_k/2}{L_k}\right]L_k$, which takes the values of $N, N + L_k, N + 2L_k, \ldots$. Suppose that the modulation waveforms are the CDMA spreading codes with ±1 chips, therefore $\eta_k(n)$ is a zero-mean Gaussian random variable with variance $T_sN_0$. We assume a low out-of-phase autocorrelation of the modulation waveform, and assume that the modulation waveforms of the interfering users appear as mutually uncorrelated noise. Using the Gaussian approximation, we treat the SI and the MAI as Gaussian random variables, and it is readily shown that they are independent with zero mean [11]. Therefore, only the variances of $S_k$ and $A_k$ are of interest. They are evaluated conditioned on a particular set of $\{a_{k,i}^{(d)}\}$ and the results are averaged over all possible $\{a_{k,i}^{(d)}\}$.

The received signal-to-interference-pulse-noise-ratio (SINR) of MS $k$ can be written as
\[
\Gamma_k = \frac{|D_k(n)|^2}{\text{var}[S_k(n)] + \text{var}[A_k(n)] + \text{var}[\eta_k(n)]}
\]
\[= \frac{|D_k(n)|^2}{\text{E}[|S_k(n)|^2] + \text{E}[|A_k(n)|^2] + T_sN_0}
\]
\[= \frac{|D_k(n)|^2}{\text{E}[|S_k(n)|^2] + \text{E}[|A_k(n)|^2] + T_sN_0}
\]
\[= \frac{|D_k(n)|^2}{\text{E}[|S_k(n)|^2] + \text{E}[|A_k(n)|^2] + T_sN_0}
\]
where $|D_k(n)|^2 = T_s^2 \sum_{i=1}^{L_k} w_{k,i}^H(n') a_{k,i}^{(d)}(n)^2$. Assume that the symbol bits are uncorrelated between successive bits and between different users. Therefore, given $\{a_{k,i}^{(d)}\}$, the variances can be written as the summation of the second moment of each term as
\[
\text{E} \left[ |S_k(n)|^2 \left| a_{k,i}^{(d)}(n) \right| \right] = \sum_{i\neq j} \sum_{i,j} |w_{k,j}^H(n') a_{k,i}^{(d)}(n)|^2 \chi(\tau_{k,j}, \tau_{k,i})
\]
\[= \sum_{i=1,j \neq k}^{K} \sum_{i,j} |w_{k,j}^H(n') a_{k,i}^{(d)}(n)|^2 \chi(\tau_{i,j}, \tau_{k,i})
\]
where 
\[
\chi(\tau_{l,j}, \tau_{k,i}) = E \left[ \left( \sum_m b_l(m) \int_{-T_s}^{0} g_k(t)g_l(t + n - m)T_s + \tau_{l,j} - \tau_{k,i})dt \right)^2 \right]
\]
\[
= \sum_m E \left[ \left( \int_{-T_s}^{0} g_k(t)g_l(t + n - m)T_s + \tau_{l,j} - \tau_{k,i})dt \right)^2 \right]
\]
(37)

For the MAI \( A_k(n) \) with \( l \neq k \), if \( \text{mod}((\tau_{l,j} - \tau_{k,i}), T_s) = 0 \) and periodic orthogonal codes are used, or if \( \tau_{l,j} - \tau_{k,i} = 0 \) and aperiodic orthogonal codes are used, \( \chi(\tau_{l,j}, \tau_{k,i}) = 0 \). Therefore we have (38), shown at the bottom of the next page. \( \xi \) is an indicator with \( \xi = 0 \) indicating code orthogonality is satisfied, and \( \xi = 1 \) otherwise. For perfect channel prediction and continuously adjusted beamforming, with \( \beta = 1 \) in (28), we have

\[
\Gamma_k|_{\{a_{k,i}^{(d)}(n)\}} = \frac{T_s^2}{\sum_{l=1, l \neq k}^{K} \sum_{i,j} \left| w_{l,j}^{H}(n) a_{k,i}^{(d)}(n) \right|^2 \xi \chi(\tau_{l,j}, \tau_{k,i}) + T_sN_0}
\]
(39)

Assume that \( y_k(n) \) is a Gaussian random variable, it follows that the probability of error of the reception of MS \( k \) conditioned on \( \{a_{k,i}^{(d)}(n)\} \) is given by

\[
Pr\left( e \left| \{a_{k,i}^{(d)}(n)\} \right. \right) = Q(\sqrt{T_k|_{\{a_{k,i}^{(d)}(n)\}}})
\]
(40)

where \( Q(\cdot) \) denotes the standard Q-function. The final probability of error \( Pr(e) \) is evaluated using the Monte Carlo integration over all possible \( \{a_{k,i}^{(d)}(n)\} \).

Recall that the beamforming weights \( \{w_{k,i}(n')\} \) are generated based on the predicted downlink channel vectors, and they are fixed over a time period of \( \{n' - \frac{L}{2T_s}, n' + \frac{L}{2T_s}\} \). The sparsity of weight adjustment and the channel prediction error can cause performance degradation. We consider the effect of prediction errors on downlink beamforming performance in the single-path case. Suppose there is a single resolvable path between the BTS and MS \( k \), the received signal power is \( |D_k(n)|^2 = T_s^2 \gamma^2_k(n) \), and the received SINR conditioned on \( a_{k,i}^{(d)}(n) \) is given by

\[
\Gamma_k|_{a_{k,i}^{(d)}(n)} = \frac{T_s^2 \gamma^2_k(n)}{\sum_{l=1, l \neq k}^{K} \sum_{i,j} \left| w_{l,j}^{H}(n) a_{k,i}^{(d)}(n) \right|^2 \xi \chi(\tau_{l,j}, \tau_{k}) + T_sN_0}
\]
(41)

where \( \gamma_k(n) \) is the prediction error ratio defined in Section III.

V. SIMULATIONS

Simulations are conducted through an electromagnetic solver FASANT [26], [27]. It is a deterministic ray tracing technique based on geometric optics and the uniform theory of diffraction. A computer-aided design (CAD) model of downtown Austin, Texas (Fig. 1) is used as the geometry input of the simulator. It represents a typical urban center with some high-rise (up to about 30 stories) and low-rise buildings. The material properties of the CAD model are: relative permittivity \( \epsilon = 9.0 \), relative permeability \( \mu = 1.0 \), conductivity \( \sigma = 0.1 \) for the building walls, and \( \epsilon = 2.0, \mu = 1.0, \sigma = 0.001 \) for the ground. Fig. 1 shows a vehicular mobile communications system, in which a BTS communicates with four randomly positioned MSs. The BTS is located near the corner of 7th Street and Congress Avenue. It has an 8-element uniform circular antenna array at a height of 20 meter, with a radius of 0.085 meter about half the carrier wavelength. The carrier frequency for uplink is 1.8 GHz. Each BTS array element is a vertically placed omnidirectional dipole antenna. MSs 1 and 2 move along 6th Street, and MSs 3 and 4 move along 8th Street. Each mobile has one vertically placed omnidirectional dipole antenna, which is in non-line-of-sight with the BTS array. The ray tracing outputs of the dominant (in terms of receiving power) paths are shown in Fig. 2 and Fig. 3. The DOA, delay and field strength are indicated for each path viewed at the BTS and at each MS. The path delays are expressed as the excess path lengths in meters. Throughout the simulations, we assume that the strongest resolvable paths of the four MSs are synchronized at the BTS.

Two CDMA cases with different data rates are considered. The bit symbols are QPSK modulated, spread with Walsh codes of length 8, and scrambled with a pseudorandom long code. The pulse-shaping filters at the transmitter and the receiver are root raised cosines with roll-off factor 0.35. The system is operated either in TDD mode with uplink and downlink carriers of 1.8 GHz, or in FDD mode with uplink carrier of 1.8 GHz and downlink carrier of 2.0 GHz. In FDD, although the frequency spacing is about 10% of the carrier frequency, we apply the AR coefficients derived from the uplink data directly to downlink channel prediction. Uplink power control is employed to compensate the near-far effect. The receive frame and the transmit frame have equal duration of 2.5 ms.

In Case One, there are 320 consecutive symbols in a data
frame of duration 2.5 ms. With a CDMA spreading gain of 8, the chip rate is 1.024 Mcps, and the propagation length during one chip period is approximately 293 meters. As shown in Fig. 2 and Fig. 3, the delay spreads of the four MSs are mostly within one chip period. Therefore, we use one RAKE finger for BTS receiving, assuming that the finger is synchronized to the strongest resolvable path from each MS.

Assuming perfect symbol detection, we use the detected bits to extract the four uplink channels. In Fig. 4, the MDL criterion is used to determine the orders of the channel models of the strongest resolvable paths. At high channel SNR, the estimated model orders are 1 for MSs 1 and 3, and 2 for MSs 2 and 4. This result agrees with the multipath distribution exhibited in Fig. 3. The model order can be interpreted as the number of dominant paths with different DOAs at the MS, since different DOAs at the MS result in independent Doppler components. In the prediction of downlink channels, we use model order $p = 2$ for each MS.

The data of one uplink frame of 2.5 ms is used to predict the downlink channel over the next 2.5 ms. The antenna beam pattern is adjusted 8 times evenly within a transmission of 2.5 ms, thus the beamforming rate $f_B = 3.2$ KHz. The uplink channel model is downsampled accordingly as in Section III. Fig. 5 shows the normalized RMSEs of the predicted channels at these 8 time instants during future downlink. The RMSE is averaged over the four MSs. It is compared with the RMSE between the current downlink channel and the actual future downlink channel. The reduction in normalized RMSE at both high and low channel SNR shows that the prediction provides sufficient knowledge of future channel variation.

Fig. 6 depicts the system performance in the average bit-error-rate (BER) of MS reception. Each MS moves at a speed of 30 mph. The downlinks are over Rayleigh fading channels with AWGN at the receivers. We compare the predictive beamforming with the conventional beamforming where fixed weights are used over one transmit frame. The fixed weights are generated based on the current uplink channel in TDD systems, or the last downlink channel feedback in FDD systems. As shown in the figure, in both TDD and FDD, the predictive beamforming provides significant performance improvement. At a BER of $10^{-3}$, the predictive beamforming has about 10 dB and 3 dB gain in TDD and FDD, respectively. The performance of the predictive beamforming over fast fading channels approaches the performance of the conventional beamforming over non-fading channels. The larger BER in FDD downlink is due to larger Doppler spread, noting that the downlink carrier is 1.8 GHz in TDD and 2.0 GHz in FDD. The error floor at high channel SNR is due to the fact that, even though the AWGN is small, there is always a certain amount of interference in the simulated multipath channels.

Fig. 7 shows the average BER with MS speed varying from 10 mph to 50 mph. All four MSs move at the same speed and communicate with the BTS in TDD mode. The channel SNR at the BTS equals to 0 dB or 20 dB. As the MS increases its speed, the channel varies more rapidly. When conventional beamforming is used with fixed weights, the system performance degrades. With predictive beamforming, the system performance is robust to MS speed. At high MS speed and channel SNR 20 dB, the BER can be reduced over a magnitude of 10. This is because the rapid channel variation is well predicted and compensated by the timely adjustment of downlink weights.

$$
\Gamma_k = |a^{(d)}_{k,i}(n)|^2 = \frac{T_s^2}{\sum_{i=1}^{L_k} w^H_{k,i}(n') a^{(d)}_{k,i}(n)}^2 \sum_{i \neq j} \left| w^H_{i,j}(n') a^{(d)}_{k,i}(n) \right|^2 \sum_{l=1}^{K} \left| w^H_{l,j}(n') a^{(d)}_{k,i}(n) \right|^2 \xi(\tau_{l,j}, \tau_{k,i}) + T_s N_0
$$
In Case Two, there are 1280 consecutive symbols in a data frame of duration 2.5 ms. The chip rate is 4.096 Mcps, and the propagation length during one chip period is approximately 73 meters. As shown in Fig. 2 and Fig. 3, the delay spreads of MSs 2 and 4 are about 5 chip periods – more than a half symbol period. Therefore, we extract the uplink channel vectors of the two most dominant resolvable paths of each MS, and employ a 2-finger pre-RAKE transmitter for the downlink.

In FDD, downlink pilot and error-free feedback are assumed. We proposed a predictive downlink beamforming approach that effectively combats the fast Rayleigh fading in wireless channels. The channel prediction requires only a moderate increase in complexity, because it is based on the linear prediction using the AR model. The beamforming algorithm is bandwidth-efficient in TDD systems due to its open-loop
Fig. 8. Average BER for QPSK over a Rayleigh fading channel with AWGN. Mobile speed 30 mph. Data rate 4.096 MHz. Pre-RAKE transmission.

Fig. 9. Average BER for QPSK over a Rayleigh fading channel with AWGN. Data rate 4.096 MHz. Pre-RAKE transmission. TDD systems.

nature, and it can be implemented in practical FDD systems with downlink pilots and a feedback channel. For wideband systems, we combined the predictive beamforming with pre-RAKE transmission to enhance the downlink performance, while preserving the simplicity of mobile receiving. Simulation results on a ray-tracing model show performance improvement of the predictive downlink beamforming over conventional beamforming methods for CDMA systems with high mobility.

ACKNOWLEDGMENT
The authors would like to thank Prof. Yih-Fang Huang for constructive discussions, and Editor James K. Cavers and the anonymous reviewers for their insightful comments.

REFERENCES


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