ENHANCING SYNCHRONIZED CELLULAR FORWARD LINKS BY EXPLOITING SPATIAL-TEMPORAL SIGNATURES USING EXTENDED SUBSPACE APPROACH

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ABSTRACT

Pilot signals are often used to ease mobile channel acquisition and detection. Cochannel signals are usually treated as interference. However, for synchronized cellular CDMA forward links, all cochannel signals from the same base station share a spatial and temporal channel signature. We propose an extended blind subspace approach to exploit this feature to enhance channel estimation. We demonstrate the performance gain by simulating two transmission schemes for MIMO cellular base stations: fixed beam and adaptive beam.

Index Terms— Code division multiaccess, Signal detection, Estimation

1. INTRODUCTION

The forward channel signals from the same base station of cellular networks are usually symbol-synchronized (for single rate networks) or chip-synchronized (for multi-rate networks). To facilitate speedy and accurate channel acquisition, a shared pilot channel is often used. Several blind channel estimation/acquisition methods have been proposed [1, 2, 3, 4, 5]. However, further improvement can be achieved by exploiting the signal and channel structures of a particular situation. For example, in cellular networks, the forward link signals are often synchronized and, to a specific mobile terminal, all cochannel signals share the same temporal and spatial channel impulse response, which we refer to as the channel signature of the mobile. One interesting work of using signal and channel features for blind detection was [6]. In this paper we extend the blind subspace approach to exploit the channel signature sharing feature of synchronized cellular forward links to enhance channel estimation.

For simplicity, we use the single rate CDMA-based cellular base station as an example. Let $\{\theta_i\}$ be the fraction of the total power allocated to channel i and the channel is assigned the Walsh code $w_i(t)$. Further let T_c be the chip duration and T_s the symbol duration. The processing gain is $N = T_s/T_c$. The Walsh code $w_i(t)$ can be written as

$$w_i(t) = \sum_{k=1}^{N} c_i(k)\psi(t - kT_c) \tag{1}$$

where $\psi(t)$ is the pulse shape and $c_i(k)$ are the discrete Walsh sequence. If the discrete information sequence for user i is $d_i(j)$, then the message in waveform is

$$s_i(t) = \sum_j d_i(j) \sum_{k=1}^{N} c_i(k)\psi(t - jT_s - kT_c) \tag{2}$$

where we have not included the long PN code in the model.

The received signal by a mobile user is composed of the signals from the home station as well as those from the neighboring base stations. We denote the former as $s^h(t)$ and latter $s^s(t)$, where $s = 1, 2, \cdots, S$ indicates the set of all interfering neighboring base stations. If all channels involved are flat, the total received signal is

$$r(t) = h_0 s^h(t) + \sum_{s=1}^{S} h_s s^s(t + \tau_s) + n(t) \tag{3}$$

where h_i are the path gains for the flat fading channels.

To establish a discrete model, we assume that the received signal $r(t)$ at the mobile terminal has been demodulated and phase tracked by a phase-locked loop. Although the subspace approach described below could be used to acquire chip timing but we assume that it has been done by some other means to simplify the description.

Passing the demodulated, phase tracked, chip timing acquired received signal through a integrate-and-dump circuitry, we obtain a discrete sequence from the received waveform $r(t)$. However, we do not have the information of the symbol timing and therefore we do not have a decision metric to decide each transmitted symbol yet. For multipath channels, multiple copies of the same transmitted waveform arrive at different chip delays and each of these delays must be estimated.

2. THE PROPOSED SUBSPACE APPROACH

When multiple antennas are available at the transmitter site, we can either use fixed weights to form fixed radiation patterns, as shown in Fig. 1, or adjustable weights to form adaptive beams according to user requests, as shown in Fig 2. The former is simple to implement and the burden on feedback channel is low: limited to a few bits to choose the serving

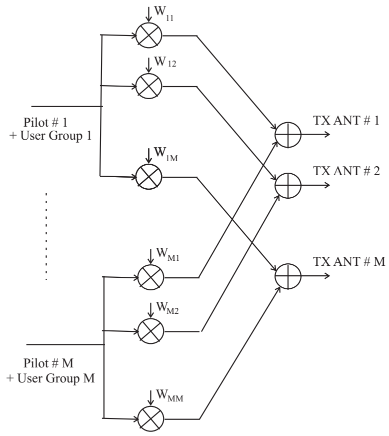

Fig. 1. Scheme 1: Fixed Beams with M Orthogonal Pilots

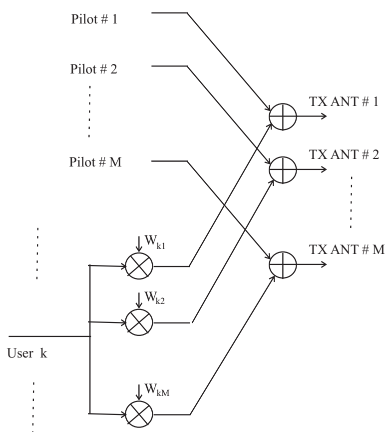

Fig. 2. Scheme 2: Adaptive Beam Transmission with M Pilots (Each for one Antenna)

beams. The latter not only complicates the transmission but also requires heavy feedback: $M-1$ channel vectors must be sent through the feedback channel.

Estimating channel using subspace approach does not require the symbol timing. Even the chip timing can also be estimated if necessary. We refer interested readers to [7] for further details. Our focus is on utilizing the channel sharing feature between the desired signal and the intracell interferences for the synchronized forward links to enhance the performance of channel estimation.

2.1. Subspace Approach for Fixed Beam Scheme

We assume that each pilot controls a fixed subset of the channelization codes and allocates one from the subset whenever a user requests. The initial handshaking process is always initiated from the user side. The mobile user's decision is based on the measured strength of the pilots. The information symbols are spread with the channelization code agreed upon and then modulated by the pilot code before transmission.

For the forward link of synchronized CDMA networks, all user signals from the same base station share the same channel, that is, they experience the same channel impulse response in both time and space dimensions. Therefore, the intra-cell interferers all carry the information of the same channel structure as the desired signal does.

The pilot *m* and all user channels served by the pilot can be written in the following chip discrete vector form for symbol *i*,

$$\mathbf{s}^m(i) = \sum_{k=0}^{K_m} \theta_k^m b_k(i) \mathbf{c}_k^m \tag{4}$$

where $\mathbf{c}_k^m$ is the kth traffic channel served by pilot *m*, $b_k(i)$ is the ith bit of the kth user. When no one uses a channel, the power allocation factor θ_k^m is zero.

Let the channel be $\mathbf{h}_k^m = [h_0, h_1, \cdots, h_{N-1}]$, where h_i is the path gain at delay i, and the channel code for user k $\mathbf{c}_k^m = [c_k^m(0), c_k^m(1), \cdots, c_k^m(N-1)]$. Define the following matrices,

$$\hat{U}_k^m = \begin{pmatrix} c_k^m(N-1) & c_k^m(N-1) & \cdots & c_k^m(0) \\ 0 & c_k^m(N-1) & \cdots & c_k^m(1) \\ 0 & 0 & \cdots & c_k^m(2) \\ & & \cdots & \\ 0 & 0 & \cdots & c_k^m(N-1) \end{pmatrix} \tag{5}$$

$$\check{U}_k^m = \begin{pmatrix} 0 & 0 & \cdots & 0 & 0 \\ c_k^m(1) & 0 & \cdots & 0 & 0 \\ c_k^m(2) & c_k^m(1) & \cdots & 0 & 0 \\ & & \cdots & & \\ c_k^m(N-2) & c_k^m(N-3) & \cdots & c_k^m(1) & 0 \end{pmatrix} \tag{6}$$

The contribution from the kth user to the received discrete signal is

$$\mathbf{r}_k^m(i) = \hat{U}_k^m \mathbf{h}_k^m \theta_k^m b_k^m(i-1) + \check{U}_k^m \mathbf{h}_k^m \theta_k^m b_k^m(i) \tag{7}$$

from two adjacent symbol bits, $b_k^m(i-1)$ and $b_k^m(i)$. The received signal at symbol interval i is the sum of contributions from all users served by all pilots from a single base station.

$$\mathbf{r}(i) = \sum_{m=1}^{M} \sum_{k=0}^{K_m} \mathbf{r}_k^m(i) + \mathbf{n}(i) \tag{8}$$

By choosing an arbitrary sampling time, we obtain a sequence of received signals from the output of the integrate-and-dump circuitry. The received signal contribution can be rewritten from Eqn. 7 as

$$\mathbf{r}_k^m(i) = [\hat{U}_k^m \mathbf{h}^m \; \check{U}_k^m \mathbf{h}^m] \begin{pmatrix} \theta_k^m b_k^m(i-1) \\ \theta_k^m b_k^m(i) \end{pmatrix} \tag{9}$$

and all users from the same pilot beam would be

$$\mathbf{r}^m(i) = [U_1^m \mathbf{h}^m \ \cdots \ U_{Km}^m \mathbf{h}^m] \begin{pmatrix} \mathbf{b}_1^m(i) \\ \vdots \\ \mathbf{b}_{Km}^m(i) \end{pmatrix} = U^m \mathbf{b}^m(i), \tag{10}$$

$U_k^m = [\hat{U}_k^m \mathbf{h}^m \ \check{U}_k^m \mathbf{h}^m]$, $\mathbf{b}_k^m(i) = [\theta_k^m b_k^m(i-1) \ \theta_k^m b_k^m(i)]^T$ and K_m is the total number of users under the pilot beam m.

The final received signal is the sum of contributions from all users of every pilot beams plus the channel noise,

$$\begin{aligned} \mathbf{r}_k(i) &= \sum_{m=1}^{M} \mathbf{r}^m(i) + \mathbf{n}(i) & (11) \\ &= [U^1 \ U^2 \ \cdots \ U^M] \begin{pmatrix} \mathbf{b}^1 \\ \mathbf{b}^2 \\ \vdots \\ \mathbf{b}^M \end{pmatrix} + \mathbf{n}(i) & (12) \\ &= U\mathbf{b}(i) + \mathbf{n}(i). & (13) \end{aligned}$$

The correlation R of the received sequence is

$U \mathrm{diag}[(\theta_{k=1}^{m=1})^2, (\theta_{k=1}^{m=1})^2, \cdots, (\theta_{k=K_m}^{m=M})^2, (\theta_{k=K_m}^{m=M})^2] U^\dagger + \sigma_n^2 \mathbf{I}$

By SVD, $R = V\Lambda V^\dagger$ and $V = [V_s \ V_n]$, V_s and V_n form the bases of the signal space and noise space respectively. The column's space of U is orthogonal to the noise space and therefore, let the projection operator be P_n,

$$P_n U = V_n V_n^\dagger U = [\mathbf{0} \ \mathbf{0} \ \cdots \ \mathbf{0}]^T \tag{14}$$

The columns of U relate to the channel impulse responses by linear transformations, $\mathbf{u}_i = U_k^m \mathbf{h}^m$. For a specific pilot beam m, its channel impulse response $\mathbf{h}^m$ relates to the noise subspace V_n by

$$V_n V_n^\dagger U_k^m \mathbf{h}^m = \mathbf{0} \tag{15}$$

for $k = 0, \ 1, \ 2, \ \cdots, \ K_m$, where $k = 0$ denotes the pilot of the beam. When the noise is not present, any single one of the above equations can uniquely determine the beam channel response as long as $V_n V_n^\dagger U_k^m$ has rank-deficiency one. However, two main factors make it not the case in reality: the noise effect and the numerical error from the SVD computation. Both of the two terms can be considered as zero mean Gaussian variables. Therefore, the more equations we have, the better the accuracy the estimated channel impulse response.

2.2. Subspace Approach for Adaptive Beam

For adaptive beam transmission, M pilots are transmitted independently from the M transmit antennas as shown in Fig. 2. The contribution from the M pilots would be

$$\begin{aligned} \mathbf{r}^P(i) &= \sum_{p=1}^{M} \left[\hat{U}_p^P \mathbf{h}^p b_p^P(i-1) + \check{U}_p^P \mathbf{h}^p b_p^P(i) \right] & (16) \\ &= U^P \mathbf{b}^P(i). & (17) \end{aligned}$$

The subspace approach to estimate the individual channels can also be applied for the pilot component of the signal. However, it is not directly applicable to the users' signals.

For user k, assuming that θ_k is the power allocation factor, the contribution of user k to the received signal is

$$\begin{aligned} \mathbf{r}_k(i) &= \theta_k \sum_{p=1}^{M} w_k^p (\hat{U}_k \mathbf{h}_p b_k(i-1) + \check{U}_k \mathbf{h}_p b_k(i)) & (18) \\ &= \sum_{p=1}^{M} [w_k^p \hat{U}_k \mathbf{h}_p \ \ w_k^p \check{U}_k \mathbf{h}_p] \begin{pmatrix} \theta_k b_k(i-1) \\ \theta_k b_k(i) \end{pmatrix} & (19) \\ &= \sum_{p=1}^{M} U_k^p \mathbf{b}_k = \left(\sum_{p=1}^{M} U_k^p \right) \mathbf{b}_k & (20) \end{aligned}$$

where $U_k^p = [w_k^p \hat{U}_k \mathbf{h}_p \ \ w_k^p \check{U}_k \mathbf{h}_p]$ and $\mathbf{b}_k = \begin{pmatrix} \theta_k b_k(i-1) \\ \theta_k b_k(i) \end{pmatrix}$.

The total received signal includes contributions from the pilots, the K users' signals and the channel noise,

$$\begin{aligned} \mathbf{r}(i) &= \mathbf{r}^P(i) + \sum_{k=1}^{K} \mathbf{r}_k(i) + \mathbf{n}(i) \\ &= [U^P \ \sum_{p=1}^{M} U_1^p \ \cdots \ \sum_{p=1}^{M} U_K^p] \begin{pmatrix} \mathbf{b}^P(i) \\ \mathbf{b}_1(i) \\ \vdots \\ \mathbf{b}_K(i) \end{pmatrix} + \mathbf{n}(i) \\ &= U\mathbf{b}(i) + \mathbf{n}(i) & (21) \end{aligned}$$

The projection of these columns into the noise space would be the origin of the noise space.

$$V_s V_s^\dagger \hat{U}_P^p \mathbf{h}^p = 0, \tag{22}$$

$$V_s V_s^\dagger \check{U}_P^p \mathbf{h}^p = 0, \tag{23}$$

$$V_s V_s^\dagger \left(\sum_{p=1}^{M} w_k^p \hat{U}_k \mathbf{h}^p \right) = 0, \tag{24}$$

$$V_s V_s^\dagger \left(\sum_{p=1}^{M} w_k^p \check{U}_k \mathbf{h}^p \right) = 0, \tag{25}$$

for $p = 1, \ 2, \ \cdots, \ M$ for the first two equations and $k = 1, \ 2, \ \cdots, \ K$ for the second two equations.

The first two equations can be solved for the channel impulse response $\mathbf{h}^p$ following the usual subspace approach. However, the last two involve linear combinations of all the channel responses, which makes it impossible to estimate the channel by subspace approach directly.

3. SIMULATION RESULTS AND CONCLUSIONS

We use the Bartlett estimate of the correlation matrix based on L recent observations,

$$\hat{R} = \frac{1}{L} \sum_{i=1}^{L} r(i) r(i)^\dagger \tag{26}$$

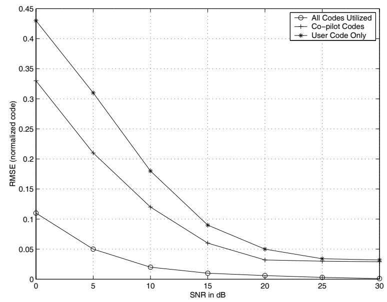

Fig. 3. Root Mean Squared Error of Estimated Normalized Code for User k: Single Cell (No Intercell Interference), 64 Chip Gold Code Set, Transmit Antenna with 4 Elements (4 Pilots), Total Users 16

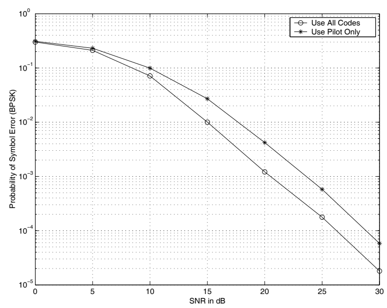

Fig. 4. Comparing Performance of the Two Channel Estimations: Utilizing All Code Sets and the Pilot Code Only. The Scenario is Single Cell (No Intercell Interference), 64 Chip Gold Code Set, Transmit Antenna with 4 Elements (4 Pilots), Total Users 16

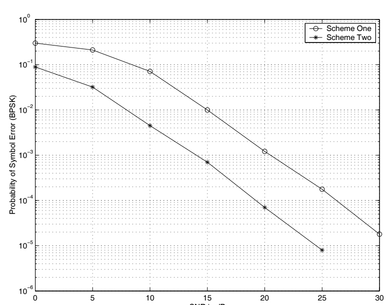

Fig. 5. Comparison of Fixed Beam and Adaptive Beam: Single Cell (No Intercell Interference), 64 Chip Gold Code Set, Transmit Antenna with 4 Elements (4 Pilots), Total Users 16

and decompose it into signal and noise subpace, $\hat{V}_s$, $\hat{V}_n$. Each column of U, for example, $U_k^m h_k^m$, must be orthogonal to the noise subpace V_n, and therefore, the projection should be zero. However, due to the approximation error and noise effect, it is not true in practice for $\hat{V}_n$. Instead, a sensible approach is to minimize the projection in l_2 norm,

$$\hat{h}_k = arg\min_h \left(|\hat{V}_n \hat{V}_n^\dagger \hat{U}_k^m h|^2 + |\hat{V}_n \mathbf{V}_n^\dagger \check{U}_k^m h|^2 \right) \quad (27)$$

$$= arg\min_h h^\dagger M h \quad (28)$$

where $M = [\hat{U}_k^m V_n V_n^\dagger \hat{U}_k^m + \check{U}_k^m V_n V_n^\dagger \check{U}_k^m]$.

The solution of the above optimization problem is well known: h_k is the eigenvector corresponding to the smallest eigenvalue of M.

For a single cell without intercell interference, Fig. 3 shows the improvement in normalized root-mean-squared error of the channel impulse response using from pilot only, all co-pilot users to all co-cell users in the extended subspace channel estimation for fixed beam schems. We have also simulated, in Fig. 4, the probability of bit error of BPSK modulation to compare using all the available code sets with using the pilot only for channel estimation. We observe 2 to 3 dB performance gain over the SNR range above 15 dB.

For the adaptive beam transmission scheme, it is not possible to apply directly the extended subspace approach. However, the performance gain by adaptation is far more than enough to compensate the this loss. Fig. 5 shows 7dB gain for four transmit antennas over the fixed beam scheme. A more thorough investigation of the adaptive beam scheme taking into account of the implementation complexity and robustness to channel estimation error should be further pursued. 0.9

4. REFERENCES

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