SEMI-BLIND CHANNEL ESTIMATION AT THE RECEIVER FOR STEERED-STS TRANSMIT ANTENNA ARCHITECTURE IN CDMA2000

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Abstract

We address the problem of pilot and data channels mismatch in a steered space time spreading (SSTS) system by means of advanced signal processing at the mobile receiver. The main idea is to apply the SSTS antenna array architecture and transmission algorithm and involve both the pilot and traffic signals in channel estimation at the receiver. Two semi-blind algorithms are compared with the conventional pilot-based receiver in a cdma2000-1x RC3 environment by means of simulations with variable speed and spatial angular spread, taking into account the main imperfections such as estimation error for the correlation coefficient at the reverse link, calibration error, different pilot powers for diversity antennas, etc. It is demonstrated that the semi-blind algorithms outperform the conventional receiver in the low- and medium-speed area.

1. INTRODUCTION

In wireless communications, both widely and closely spaced antenna arrays offer some performance benefits in different environments. In the first case, open-loop transmit diversity schemes like the space-time spreading (STS) technique can be applied [1,2]. In the second case, different beamforming techniques can be exploited to utilize the available channel information. An antenna array architecture has been proposed and investigated in [3], which combines the advantages of both these antenna array structures. This antenna architecture employs a combination of transmit diversity and beamforming. It can be realised by spacing two groups of antennas far apart. Antennas within each group are spaced closely together.

For two antennas in a group, the potential gain over the conventional transmit diversity system is 3 dB in flat fading and with a narrow spatial angle spread. Generally, the aperture gain of an adaptive beamforming system is a function of the coherence of the closely spaced transmit antennas. Particular applications may suffer from additional imperfections. For example, in the cdma2000 system, no auxiliary pilot signals are transmitted from and steered by the beamforming antennas [3]. The common pilot signal is not steered and steering the data in the presence of large angle spread can cause a mismatch between the pilot and the traffic. Thus, the beamforming gain depends on the coherence of the training data broadcast by the pilot sequence and the steered traffic channel.

In this paper we address the problem of pilot and data mismatch by means of advanced signal processing at the mobile receiver. That includes an iterative semi-blind channel estimation algorithm and its modification based on parametric temporal modelling of the non-stationary propagation channel.

The problem formulation is given in Section 2. The semi-blind receiver algorithms are presented in Section 3. The simulation assumptions including the propagation model used and imperfections considered are summarised in Section 4. The simulation results are shown in Section 5. Section 6 concludes the paper.

2. PROBLEM FORMULATION

The signals $u_i(t)$ transmitted from the first and second groups of antennas and the received statistics $x_j(t)$ can be expressed as follows [3]:

$$u_i(t) = \nu_i \frac{\sqrt{P}}{2}[s_{\mathrm{e}}(t)w_1(t) - s^*_{\mathrm{o}}(t)w_2(t)],\ i = 1, 2, \quad (1)$$

$$u_i(t) = \nu_i \frac{\sqrt{P}}{2}[s^*_{\mathrm{e}}(t)w_2(t) + s_{\mathrm{o}}(t)w_1(t)],\ i = 3, 4, \quad (2)$$

$$x_1(t) = \sum_{i=1}^{2} \nu_i g_i(t) s_{\mathrm{e}}(t) + \sum_{i=3}^{4} \nu_i g_i(t) s_{\mathrm{o}}(t) + e_1(t), \quad (3)$$

$$x_2(t) = -\sum_{i=1}^{2} \nu_i g_i(t) s^*_{\mathrm{o}}(t) + \sum_{i=3}^{4} \nu_i g_i(t) s^*_{\mathrm{e}}(t) + e_2(t), \quad (4)$$

where $s_{\mathrm{o}}(t)$ and $s_{\mathrm{e}}(t)$ are the odd and even transmitted symbols, $w_1(t) = [w(t)\, w(t)]$, $w_2(t) = [w(t)\ -w(t)]$ are the modified Walsh functions, $w(t)$ is the original spreading code, ν_i are the beam steering weights, $g_i(t) = \frac{\sqrt{P}}{2}\tilde{g}_i(t)$, $\tilde{g}_i(t)$ are the propagation channels, P is the power evenly distributed among the four antennas and e_j are independent noise sources associated with the demodulating Walsh codes.

The signal estimates can be found using the decision statistic and the overall channel estimates $\hat{h}_1 = \sum_{i=1}^{2} \nu_i g_i$ and $\hat{h}_2 = \sum_{i=3}^{4} \nu_i g_i$:

$$\hat{s_{\mathrm{e}}}(t) = f\{\hat{h}_1(t)^* x_1(t) + \hat{h}_2(t) x_2^*(t)\}, \tag{5}$$

$$\hat{s_{\mathrm{o}}}(t) = f\{\hat{h}_2(t)^* x_1(t) - \hat{h}_1(t) x_2^*(t)\}, \tag{6}$$

where $f\{\cdot\}$ is a decision function for the given constellation.

In the particular case of coherent wavefronts at the closely spaced antennas, e.g. $g_2 = e^{-j\Theta} g_1$ and $g_4 = e^{-j\Theta} g_3$, and the ideal beam steering weights $\nu_1 = \nu_3 = 1$ and $\nu_2 = \nu_4 = e^{j\Theta}$, the overall channels for the diversity antennas are: $h_1 = \sum_{i=1}^{2} \nu_i g_i = 2g_1$ and $h_2 = \sum_{i=3}^{4} \nu_i g_i = 2g_3$. Only in this particular case the overall channels are directly related to the diversity antenna channels, supported by the pilot signals. If these assumptions are violated because of any reason, a mismatch between the data and pilot channels appears. Then, direct use of the pilot-based estimates of the diversity channels instead of the overall channel estimates $\hat{h}_1 = \hat{g}_1$ and $\hat{h}_2 = \hat{g}_3$ in (5), (6) causes a performance degradation.

A robust solution for the beamforming weights is discussed in [3], addressing the mismatch problem at the transmitter. (Having characterized the level of mismatch using a reverse-link-based estimate of the correlation coefficient between two beamforming antennas, transmit weights are chosen to minimize the receiver BER.) Our goal in this paper is to address the mismatch problem at the receiver by means of using the pilot and data signals to estimate the overall channels in (3), (4) and apply them at the receiver (5), (6).

3. RECEIVER ALGORITHMS

Semi-blind iterative channel estimation is well known in both coded- and uncoded-data cases [5,6 and others]. It is based on a block of data received in quasi-stationary conditions, e.g. in the situation where the propagation channel does not change significantly over the whole block. Some of the signal features, which belong to the whole block of data, e.g. finite alphabet or constant modulus properties, are normally used to formulate a semi-blind criterion. The pilot or training symbols are employed as a part of a semi-blind criterion and/or for initialization [7 and others].

Generally, the efficiency of semi-blind techniques depends on the size of the block of data. In high speed applications, where the channel remains practically stationary only over one or a few symbols, semi-blind techniques are normally not effective. On the contrary, low speed applications are suitable for these techniques.

Taking into account that a wide range of speeds and convolutionally or turbo encoded data occur in a cdma2000 system [4], our target is the low- to medium-speed range and we consider semi-blind processing techniques for the coded signals. In Section 3.1 we apply a general Iterative Channel Estimation (ICE) algorithm to the SSTS system. In section 3.2 we propose a modified ICE algorithm with parametric channel modelling (ICEM) in the time domain.

First of all, let us formalize the SSTS receiver operation. We assume that all N_f symbols in each frame are divided into K blocks of $L = N_f/K$ symbols and the diversity channel $\mathbf{g} = [g_1, g_3]$ is estimated over each block using the conventional pilot-based procedure.

For convenience we denote the SSTS receiver operation as:

$$\hat{\mathbf{b}} = \mathrm{SSTS}\{x(n), \hat{\mathbf{h}}(n), n = 1 \dots N_f\}, \tag{7}$$

where $\hat{\mathbf{b}}$ is the vector of uncoded information bits, $x(n)$ is the received signal formed from the decision statistic (3), (4) and $\hat{\mathbf{h}}(n)$ is an estimate of the overall channel. Additionally to equations (5), (6), the SSTS operation includes the standard demodulation-deinterleaving-decoding (DMIC) procedure $\hat{\mathbf{b}} = \mathrm{DMIC}\{[\hat{s}_1(1) \dots \hat{s}_1(N_f)]\}$. Then, the conventional receiver can be presented as:

$$\hat{\mathbf{b}}_{\mathrm{SSTS}} = \mathrm{SSTS}\{x(n), \hat{\mathbf{g}}(n), n = 1 \dots N_f\}, \tag{8}$$

where $\hat{\mathbf{g}}(n) = [\hat{g}_1, \hat{g}_3]$ is the pilot-based estimate for the 1st and 3rd channels.

3.1. ICE for space-time-convolutional coded signals

Using the introduced notation, the J-iteration ICE algorithm for the SSTS system can be presented as follows:

$$\hat{\mathbf{b}}^0_{\mathrm{ICE}} = \hat{\mathbf{b}}_{\mathrm{SSTS}}, \tag{9}$$

$$\left[\hat{s}^j_1(1) \dots \hat{s}^j_1(L) \dots \hat{s}^j_K(1) \dots \hat{s}^j_K(L)\right] = \mathrm{CIM}\left\{\hat{\mathbf{b}}^{j-1}_{\mathrm{ICE}}\right\}, \tag{10}$$

$$\hat{\mathbf{C}}^j_k = \left[\begin{bmatrix} \hat{s}^j_k(1) & -\hat{s}^{j^*}_k(2) \\ \hat{s}^j_k(2) & \hat{s}^{j^*}_k(1) \end{bmatrix} \dots \begin{bmatrix} \hat{s}^j_k(L-1) & -\hat{s}^{j^*}_k(L) \\ \hat{s}^j_k(L) & \hat{s}^{j^*}_k(L-1) \end{bmatrix}\right], \tag{11}$$

$$\mathbf{x}_k = \left[x(1 + (k-1)L) \dots x(kL)\right], \tag{12}$$

$$\hat{\mathbf{h}}^j_k = \mathbf{x}_k \hat{\mathbf{C}}^{j^*}_k (\hat{\mathbf{C}}^j_k \hat{\mathbf{C}}^{j^*}_k)^{-1},\ k = 1 \dots K, \tag{13}$$

$$\hat{\mathbf{b}}^j_{\mathrm{ICE}} = \mathrm{SSTS}\{x(n), \hat{\mathbf{h}}^j(n),\ n = 1 \dots N_f\}, \tag{14}$$

where $\hat{\mathbf{h}}^j(n) = \hat{\mathbf{h}}^j_k$, $k = 1 \dots K$, $n = 1 + (k-1)L \dots kL$, $j = 1 \dots J$ is the iteration number, $\mathrm{CIM}\{\cdot\}$ is the standard coding-interleaving-modulation operation [4], $\hat{s}^j_k(l)$ is the recovered l-th symbol at the k-th averaging interval of $L = N_f/K$ symbols, $\hat{\mathbf{C}}^j_k$ is the $(2 \times L)$-matrix of the recovered space-time coded signal at the k-th averaging interval.

3.2. ICEM for space-time coded signals

As we mentioned above, the size of the averaging block L has to be selected according to the quasi-stationary condition for the overall channel. Generally, stochastic or deterministic modelling of the non-stationary channel may allow increases in the averaging interval compared to the conventional ICE algorithm, leading to performance improvement. Taking into account that complicated receiver algorithms are not likely to be implemented at mobile terminals, in this paper we apply simple parametric polynomial channel modelling in the time domain to modify the ICE estimator for space-time coded signals.

In the general case, defining the temporal parametric channel model as $\mathbf{h}(\mathbf{F}, n)$, where $\mathbf{F}$ is the vector of parameters, ICEM for space-time coded signals can be described as follows:

$$\hat{\mathbf{b}}^0_{\text{ICEM}} = \hat{\mathbf{b}}_{\text{SSTS}}, \tag{15}$$

$$\left[\hat{s}^j_1(1)\ldots\hat{s}^j_1(L)\ldots\hat{s}^j_K(1)\ldots\hat{s}^j_K(L)\right] = \text{CIM}\left\{\hat{\mathbf{b}}^{j-1}_{\text{ICEM}}\right\}, \tag{16}$$

$$\hat{\mathbf{F}}^j_k = \arg\min_{\mathbf{F}} \sum_{l=1}^{L} |x(l+(k-1)L) - \mathbf{h}(\mathbf{F}, l)\hat{\mathbf{c}}^j_k(l)|^2, \tag{17}$$

$$\hat{\mathbf{b}}^j_{\text{ICEM}} = \text{SSTS}\{x(n), \hat{\mathbf{h}}^j(n),\, n = 1\ldots N_f\}, \tag{18}$$

where $\hat{\mathbf{h}}^j(n) = \mathbf{h}(\hat{\mathbf{F}}^j_k, l)$, $l = 1\ldots L$, $k = 1\ldots K$, $n = 1+(k-1)L\ldots kL$, $\mathbf{x}_k$ is defined as in (12) and $\hat{\mathbf{c}}^j_k(l)$ is the l-th column on matrix $\hat{\mathbf{C}}^j_k$ defined in (11).

In the polynomial model case, the general model $\mathbf{h}(\mathbf{F}, n)$ can be specified as follows:

$$\mathbf{h}(\mathbf{F}, n) = \sum_{m=0}^{M} \mathbf{f}_m n^m, \tag{19}$$

where $\mathbf{f}_m = [f_{1m}\, f_{2m}]$, $\mathbf{F} = [\mathbf{f}_0 \ldots \mathbf{f}_M]$ is the $(2 \times M)$-vector of parameters and M is the model order.

Then, equation (17) can be replaced with the following ones:

$$\hat{\mathbf{F}}^j_k = \mathbf{x}_k \tilde{\mathbf{C}}^{j^*}_k (\tilde{\mathbf{C}}^j_k \tilde{\mathbf{C}}^{j^*}_k)^{-1}, \tag{20}$$

$$\tilde{\mathbf{C}}^j_k = \begin{bmatrix} \hat{\mathbf{C}}^j_k \mathbf{T}_0 \\ \vdots \\ \hat{\mathbf{C}}^j_k \mathbf{T}_M \end{bmatrix}, \tag{21}$$

$$\mathbf{T}_m = \text{diag}([1^m\, 2^m \ldots L^m],\, m = 0\ldots M, \tag{22}$$

$$\hat{\mathbf{h}}^j(n) = \sum_{m=0}^{M} \hat{f}^j_{km} l^m, \tag{23}$$

where $l = 1\ldots L$, $k = 1\ldots K$, $n = 1+(k-1)L\ldots kL$.

One can see that the ICE algorithm (9)-(14) is a particular case of the zero-order ($M = 0$) polynomial ICEM algorithm (15)-(23).

4. SIMULATION ASSUMPTIONS

Physical configuration: The four-element antenna configuration from [3] was used at the base station, i.e. 10λ distance between two pairs of beamforming antennas and $\lambda/2$ distance between the two antennas in any pair. The mobile was randomly placed such that its azimuthal angle with respect to the antenna array perpendicular was uniformly distributed in the range $\pm 60°$. This is consistent with a tri-sectored cell.

Channel Model: Reusing the model from [3], the desired spatial correlation between the channels corresponding to the four antennas is achieved by distributing a single set of scatterers in a 2-D Gaussian distribution about the mobile. More detail is given in [3].

Transmissions: Spreading Rate 1 (1.2288 Mchip/s) and RC3 mode (as described in [4]) were used with a data rate of 9600 kb/s and 20 ms frames. The carrier frequency was from the PCS band (1900 Mhz) with a separation of 80 MHz between the forward and reverse links.

Uplink correlation coefficient estimation: The average forward-link correlation coefficient ρ between the two antenna elements in any beam-forming pair was estimated by calculating the equivalent correlation coefficient in the reverse link taking into account (a) the difference in carrier frequencies between the reverse- and forward-link, (b) the noise and interference in the reverse link, (c) calibration errors in the reverse-link RF chains, and (d) reverse-link estimation errors arising from the random instantiations of the spatial modelling deviating from the long-term average. The received pilot SINR levels in the reverse link were calculated from the target SINR specified in [8].

Pilot transmit power levels: Using a system-level simulator, cdfs were generated for the geometry (SINR) seen at mobiles in an urban macro-cell network. With a path-loss exponent of 35 dB per decade and a log-normal shadow fading standard deviation of 8 dB, the geometry at the 50% point in this cdf is 4 dB—and this is used as a representative value of the SINR. Given that, as assumed here, 15% of total base station power is allocated to the main and diversity pilot channels and that there are 128 chips per pilot symbol, one has a total pilot symbol-level SINR of $\text{SINR}_{\text{pilot}} = 16.8$ dB.

By concentrating on the "typical" SINR, rather than (for example) the worst-case SINR, the results are indicators of the achievable capacity gains.

Antenna calibration: To achieve the desired beam pattern between the two antennas in an antenna pair, there needs to be a precise relationship between the amplitudes and phases of the transmit chain gains. This depends on the accuracy of the built-in calibration mechanisms. These, however, will have a residual error. Here 1° rms phase and 0.25 dB rms amplitude calibration errors were assumed for each antenna's RF path.

Channel estimation: The first order ($M = 1$) model is used for ICEM and one iteration of ICE and ICEM is implemented.

5. SIMULATION RESULTS

Frame Error Rate (FER) performance averaged over 5000 frames is estimated in the typical case ($\text{SINR}_{\text{pilot}} = 16.8$ dB) for variable speed and angular spread. One iteration ($J = 1$) is used for both ICE and ICEM semi-blind techniques. The number of averaging intervals K is selected separately for the conventional and semi-blind receivers to get the best results. The summarized gain results for SSTS over STS and the semi-blind over conventional channel estimation are presented in Figure 1 and 2 respectively for different speed values.

Analysis of the simulation results shows that:

- The semi-blind channel estimation techniques demonstrate up to 1 dB performance improvement over conventional channel estimation at the receiver.
- The highest gain is observed for the low- to medium-speed range. The gain decreases with increased speed, e.g. it falls below 0.5 dB for 30 km/hr.
- First order polynomial ICEM outperforms ICE by 0.1 to 0.3 dB for 2° to 16° of angular spread.

6. CONCLUSIONS

We have addressed the problem of pilot and data channels mismatch in a SSTS system by means of semi-blind signal processing at the receiver. Two semi-blind algorithms have been developed and compared with the conventional pilot-based receiver in a cdma2000 RC3 environment. It has been demonstrated that the semi-blind algorithms outperform the conventional receiver by up to 1 dB in typical cases depending on angular spread and speed. Optimal channel modelling of the non-stationary channel in the cdma2000-1x environment to improve the ICEM performance could be a subject for future study.

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7. REFERENCES

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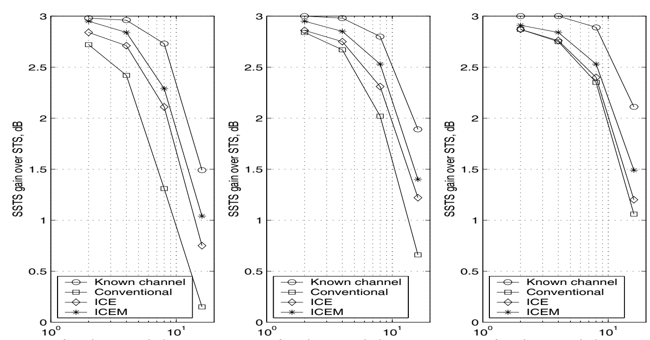

Figure 1: SSTS over STS gain for (3 - 30) km/h

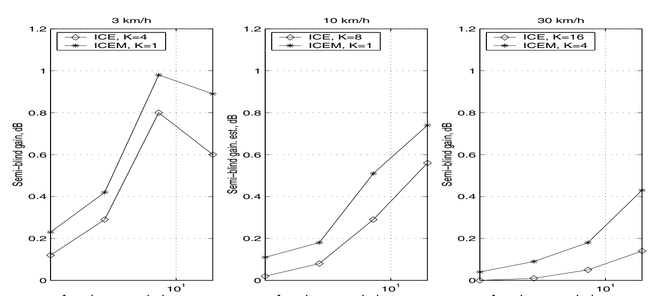

Figure 2: Semi-blind processing gain for (3 - 30) km/h