A Base-Station Antenna Array Receiver for Cellular DS/CDMA with M-ary Orthogonal Modulation

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ABSTRACT

Code division multiple access (CDMA) has been shown to offer higher capacity than the existing FDMA or TDMA systems. In the proposed IS-95 CDMA standard, no pilot is used in the reverse link due to power efficiency considerations. Therefore, an M-ary orthogonal modulation with noncoherent reception is used for this link. Recently, spatial processing with base-station antenna arrays was proposed to reduce co-channel interference and increase system capacity. In this paper, we present a base-station antenna arrays receiver structure for cellular CDMA with M-ary orthogonal modulation and noncoherent RAKE combining in the presence of multipath.

1. INTRODUCTION

In addition to its ability to combat multipath fading and resistance to jamming, Code division multiple access (CDMA) has been shown to offer higher capacity over conventional analog FM systems (AMPS) and the Telecommunication Industry Association (TIA) North American TDMA digital cellular standard IS-54. Capacity estimates show that the CDMA system can provide 20 times more capacity in a given bandwidth than the current AMPS system and 4 times more capacity than TDMA [1, 2]. This increase in system capacity is due to efficient frequency reuse and the exploitation of speech activity.

One approach to further increase the system capacity is the use of spatial processing with base station antenna arrays [3-6]. By using spatial processing at cell sites, we can use optimum directional receive and transmit beams for each user to improve system performance and increase capacity. Such improved spatial processing can be easily incorporated into the proposed CDMA standards IS-95. The increase in system capacity by using antenna arrays in CDMA comes from reducing the amount of co-channel interference from within own cell and neighboring cells. This reduced interference transforms to an increase in capacity. In general, this improvement in the received $\frac{E_b}{N_o}$ can be used to improve other system performance measures such as coverage area and transmitted mobile power.

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In [7] we proposed an antenna array 2D-RAKE receiver that exploits the spatial structure in the received multipath signal in addition to time diversity to provide a more efficient combining of paths. This receiver incorporates a beamstearing processor "front-end" feeding a conventional coherent RAKE combiner, which assumes knowledge of the amplitudes and phases of path gains. This requires the transmission of a pilot signal to obtain good amplitude and phase estimates but, unfortunately, it is a luxury which may not be affordable. Transmitting a pilot in each user's signal, whose power is greater than the data-modulated portion of the signal, reduces efficiency to less than 50%. Instead, either differential phase shift keying (DPSK) which does not require phase coherence or M-ary orthogonal modulation with noncoherent reception should be used. For $H > 8$, where H is the number of orthogonal functions, better noncoherent performance over DPSK can be obtained [8, 9]. Analysis results for CDMA communications systems employing DPSK are reported in[10-13]. The analysis in [13] assumes a base-station antenna array receiver structure. Analysis for DS/CDMA with M-ary orthogonal modulation has appeared in [14-18]

In this paper, we will present an antenna arrays receiver structure for cellular CDMA with M-ary orthogonal modulation with non-coherent RAKE combining. The receiver consists of an antenna array followed by a bank of Hadamard Walsh correlators. The baseband received signal and the post-correlation signal vector are used to estimate the array response vector for each path. The output of the correlators is then fed to an optimum beamformer followed by a noncoherent RAKE combiner. The output of the RAKE combiner is then used to estimate the transmitted data. Assuming that the channel parameters are almost constant over two symbol periods, this estimate is fed back to the array response vector estimation algorithm to determine the winning post-correlation signal vector which corresponds to the actual transmitted Walsh symbol.

This paper is organized as follows. In Section 2, we present the channel and received signal model. In Section 3 we describe the antenna array receiver structure. Numerical and simulation results are presented in Section 4. Finally, Section 5 contains our conclusions and remarks.

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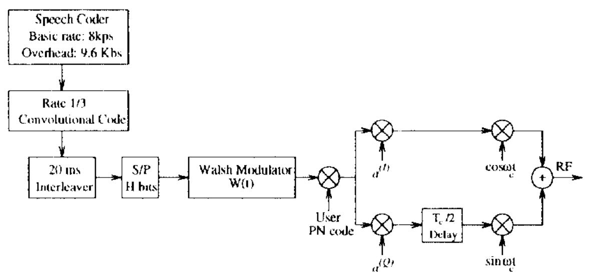

Figure 1: Mobile Transmitter Block Diagram

2. SIGNAL AND CHANNEL MODEL

The mobile transmitter block diagram is shown in Figure 1. The binary data at the output of the interleaver are grouped into groups of $J = \log_2 H$ bits and each group is mapped into one of H orthogonal binary sequences $W(t)$, known as Hadamard-Walsh functions. The resulting signal is then spread using the user's long PN code $c_i(t)$. The signal is further multiplied in both I and Q channels by the short PN codes $a^{(I)}(t)$ and $a^{(Q)}(t)$ respectively. The PN modulated Q channel signal is delayed by half a chip period $T_c/2$. The two spread signals are up-converted to radio frequency for transmission. Note that before up converting to RF, the signal in the I and Q channels is an offset QPSK (OQPSK) signal.

Let N be the number of users in the system. We assume average power control, *i.e.* we assume that each user adjusts its transmitted power such that its received average power at the output of the incoherent RAKE combiner at the base-station is kept constant. Then we can write the signal transmitted by the i-th mobile as

$$s_i(t) = \psi_i \sqrt{P_i} \left(W^{(h)}(t) c_i(t) a^{(I)}(t) \cos(\omega_c t) + W^{(h)}(t - T_o) \right.$$
$$\left. c_i(t - T_o) a^{(Q)}(t - T_o) \sin(\omega_c t) \right) \quad 0 \le t \le T_w \qquad (1)$$

where P_i is the transmitted power per symbol, T_w is the symbol period, ω_c is the carrier angular frequency, T_o is the time offset between the I and Q channels ($T_c/2$), and finally ψ_i is Bernoulli random variable that models the voice activity of the user (we assume that a user will be on with probability ν and will be off with probability $1-\nu$). $W^{(h)}(t)$ is the hth orthogonal Walsh function, $h = 1, \cdots, H$. For simplicity of notation, we shall denote the product of the user PN code and the I or Q channel PN code as

$$a_i^{(I)}(t) = c_i(t) a^{(I)}(t) \quad \text{and} \quad a_i^{(Q)}(t) = c_i(t) a^{(Q)}(t)$$

Note that we shall assume that each of the PN codes is a random sequences, that is we assume that

$$a_i^{(I)}(t) = \sum_{k=-\infty}^{\infty} a_{i,k}^{(I)} p(t - kT_c) \qquad (2)$$

$$a_i^{(Q)}(t) = \sum_{k=-\infty}^{\infty} a_{i,k}^{(Q)} p(t - kT_c) \qquad (3)$$

where $a_{i,k}^{(I)}$ and $a_{i,k}^{(Q)}$ are assumed to be i.i.d. random variables taking values ± 1 with equal probability, and $p(t)$ is the chip pulse shape, which can be any time-limited waveform. Here we assume that $p(t)$ is rectangular although our results can be easily extended for any time-limited waveform.

The complex low pass equivalent representation of the scalar multipath fading channel as seen by the ith user can be written as [8]

$$h_i(t) = \rho_i(t) \sum_{l=1}^{\mathcal{L}_i} \alpha_{l,i}(t) \delta(t - \tau_{l,i}) e^{-j\phi_{l,i}(t)} \qquad (4)$$

where $\alpha_{l,i}$, $\tau_{l,i}$, and $\phi_{l,i}$ are the amplitude, time delay, and phase shift of the l-th multipath component. $\mathcal{L}_i$ is the number of paths received from the ith user. The variable ρ_i models the effects of path loss and log-normal shadowing, *i.e.* $10 \log_{10} \rho_i^2 \sim \mathcal{N}(-e \cdot 10 \log r_i, \sigma_s^2)$, where e is the pass loss exponent, r_i is the distance between the ith user and the cell site, and σ_s varies between 6-12 dB depending on the degree of shadowing. The phase term $\phi_{l,i}$ is assumed to be uniformly distributed over $[0, 2\pi]$, the path amplitude $\alpha_{l,i}$ is assumed to follow a filtered Rayleigh distribution, and the delay $\tau_{l,i}$ is assumed to be uniformly distributed over $[0, \tau_{\max}]$, where $\tau_{\max}$ is the maximum delay spread of the channel. We assume that each distinguishable multipath components (*i.e.* those separated by more than T_c from one another) will actually be themselves a linear combination of several indistinguishable paths of varying amplitudes which arrive at the base-station within $\pm\Delta_{l,i}/2$ at angle $\varphi_{l,i}$. This will cause the fading signal in some path to be not perfectly correlated from one antenna to another. The fading correlation depends on the spacing between antenna elements, the angle $\varphi_{l,i}$, and the scattering angle $\Delta_{l,i}$. We assume that the multipath channel parameters vary slowly as compared to the symbol duration, so that they are nearly constant over several symbol periods. In the current IS-95 standard, the spread signal bandwidth is very small compared to the carrier frequency and, therefore, the narrowband assumption for antenna arrays is valid. This enables modeling time delays due to propagation across the array as phase shifts. Then the complex lowpass equivalent of the vector multipath channel as seen by the ith user with respect to the cell site antenna array can be written as [19]

$$\mathbf{h}_i(t) = \rho_i(t) \sum_{l=1}^{\mathcal{L}_i} \delta(t - \tau_{l,i}) \mathbf{a}_{l,i} \qquad (5)$$

where $\mathbf{a}_{l,i}$ is the $M \times 1$ response vector of the cell site antenna array to signals in the l-th path from the ith user and M is the number of antennas. The elements of $\mathbf{a}_{l,i}$ are correlated complex zero mean Gaussian random variables. The correlation between the elements of $\mathbf{a}_{l,i}$ is described by the model in [20].

At the base station, the received signals from all users are down converted to baseband. After the down-converter, we can write the complex baseband received signal vector for the i-th user as

$$\mathbf{x}_i = \rho_i\sqrt{P_i}\psi_i \sum_{l=1}^{\mathcal{L}_i} \Big(W^{(h)}(t-\tau_{l,i})a_i^{(I)}(t-\tau_{l,i})+ jW^{(h)}(t-T_o-\tau_{l,i})a_i^{(Q)}(t-T_o-\tau_{l,i})\Big) \cdot(\cos\theta_{l,i}+j\sin\theta_{l,i})\,\mathbf{a}_{l,i} \quad (6)$$

where $\theta_{l,i}=\omega_c\tau_{l,i}$. The total received signal at the cell site is the sum of all users' signals plus noise and is given by

$$\mathbf{x}(t)=\sum_{i=1}^{N}\sum_{l=1}^{\mathcal{L}_i}\mathbf{x}_{l,i}(t)+\mathbf{n}(t) \quad (7)$$

where $\mathbf{n}(t)=\mathbf{n}_c(t)+j\mathbf{n}_s(t)$ is the additive Gaussian noise vector with zero mean and covariance

$$E\{\mathbf{n}(t_1)\mathbf{n}^*(t_2)\}=\sigma^2\mathbf{I}\cdot\delta(t_1-t_2) \quad (8)$$

where σ^2 is the noise power per antenna. Equation 8 implies that the noise is both ***temporally*** and ***spatially*** white. $\mathbf{n}_c(t)$ and $\mathbf{n}_s(t)$ are low pass Gaussian random processes.

3. RECEIVER MODEL

The block diagram of the antenna array receiver is shown in Figure 2. It has a 2D-RAKE structure where several multipath components are tracked in both time and space. After down-converting to baseband, the outputs of the LPF are fed into a bank of H Hadamard-Walsh correlators shown in Figure 3. The pre-correlation and post-correlation signal vectors $\mathbf{x}(t)$ and $\mathbf{z}_{l,i}^{(n)}$ are used to estimate the array response vector for the l multipath component of the ith user assuming that the nth Walsh symbol was transmitted, where $n=1,\cdots,H$. In [7] we showed that the array response vector of the lth multipath component of the ith user (and hence the optimum beamforming weights) can be estimated from the covariances of $\mathbf{x}(t)$ and $\mathbf{z}_{l,i}^{(n)}$. The post-correlation vectors $\mathbf{z}_{l,i}^{(n)}$ are then fed to an optimum beamformer followed by an incoherent RAKE combiner. The beamformer and the incoherent RAKE combiner are shown in Figures 3 and 4. Note that, at this stage we have no knowledge which vector $\mathbf{z}_{l,i}^{(n)}$ corresponds to the true transmitted Walsh symbol $W^{(h)}(t)$. Here we rely on the assumption that the vector multipath channel seen by the ith user remains constant over several symbol periods. Hence the previous estimate of the array response vector is used in the beamforming.

The decision variables $z_i^{(1)},\cdots,z_i^{(H)}$ at the output of the incoherent RAKE are then fed to M-ary decoder, deinterleaver, and Viterbi convolutional decoder. Also, to estimate the array response vector that will be used in the beamformer for the next symbol, the index $\hat{h}$ of the $\max(z_i^{(h)})$ is then used to select the post-correlation signal vector $z_i^{(\hat{h})}$.

Without loss of generality, let us assume that the 1st user is the desired user and let $\tau_{k,1}$ be the time delay of the kth tracked multipath which is assumed to be estimated perfectly and $k=1,\cdots,\mathcal{L}_1$. Define

$$\tilde{c}_{l,i}^{(n)}(t-\tau_{l,i})=\Big(W^{(n)}(t-\tau_{l,i})a_i^{(I)}(t-\tau_{l,i})+ jW^{(n)}(t-T_o-\tau_{l,i})a_i^{(Q)}(t-T_o-\tau_{l,i})\Big) \quad (9)$$

Then, we can write the nth correlator array output for the kth tracked multipath component of the first user as

$$\mathbf{z}_{k,1}^{(n)} = \frac{1}{\sqrt{T_w}}\int_{\tau_{k,1}}^{\tau_{k,1}+T_w}\mathbf{x}(t)\tilde{c}_{k,1}^{(n)}(t-\tau_{k,1})dt \quad (10)$$

$$= \mathbf{d}_{k,1}^{(n)}+\mathbf{y}_{k,1}^{(n)} \quad ;\text{ if } \quad n=h \quad (11)$$

$$= \mathbf{y}_{k,1}^{(n)} \quad ;\text{ if } \quad n\neq h \quad (12)$$

where

$$\mathbf{d}_{k,1}^{(n)} = \rho_1\sqrt{T_wP_1}(\cos\theta_{k,1}+j\sin\theta_{k,1})\mathbf{a}_{k,1} \quad (13)$$

$$\mathbf{y}_{k,1}^{(n)} = \mathbf{y}_{k,1}^{I,(n)}+j\mathbf{y}_{k,1}^{Q,(n)} \quad (14)$$

and

$$\mathbf{y}_{k,1}^{I,(n)} = \mathbf{I}_{k,1}^1+\mathbf{I}_{k,1}^2+\mathbf{I}_{k,1}^3+\mathbf{N}_{k,1}^I, \quad (15)$$

$$\mathbf{y}_{k,1}^{Q,(n)} = \mathbf{Q}_{k,1}^1+\mathbf{Q}_{k,1}^2+\mathbf{Q}_{k,1}^3+\mathbf{N}_{k,1}^Q, \quad (16)$$

and $\mathbf{I}_{k,1}^1$ is the self-interference due to the I and Q channel spreading codes, $\mathbf{I}_{k,1}^2$ is the self-interference due to other multipath components of the 1st user, $\mathbf{I}_{k,1}^3$ is due to the multiple access interference, and $\mathbf{N}_{k,1}^I$ is due to the AWGN. $\mathbf{Q}_{k,1}^1$, $\mathbf{Q}_{k,1}^2$, $\mathbf{Q}_{k,1}^3$, and $\mathbf{N}_{k,1}^Q$ are defined similarly. Let $E_{w,i}=T_wP_i$ be the ith user's symbol energy. Also, for the kth tracked multipath, let the optimum beamforming weights assuming that the nth Walsh symbol was transmitted be $\mathbf{w}_{k,1}^{(n)}$.For an equal gain combining incoherent RAKE, the nth decision variable of the 1st user, corresponding to the nth Walsh symbol is given by

$$z_1^{(n)}=\sum_{k=1}^{\mathcal{L}_1}|\mathbf{w}_{k,1}^{(n)*}\mathbf{z}_{k,1}^{(n)}|^2 \quad n=1,\cdots,H. \quad (17)$$

Note that to select which post-correlation signal vector should be used in estimating the post-correlation array covariance, a hard decision is made on which Walsh symbol was transmitted while for the data, a symbol-by-symbol M-ary decoder is used [9]. Both approaches yield exactly the same decisions for the M-ary-symbol and both are optimal for an AWGN channel. Since the multiple access interference is not necessarily Gaussian, this decision rule is actually not optimal. However, it is commonly used for simplicity. The primary reason for using the symbol-by-symbol approach for the data is to provide better performance with error-correcting codes by using soft-decoding. Note also that we can not use the output after the convolutional decoder to select the post-correlation signal vector. The reason for this is that we will have to wait for a decision to be made on the current symbol and encode and interleave again. By the time that process is over (which is at least the time of one frame of bits plus decoding time), the channel might have changed and the estimated array response

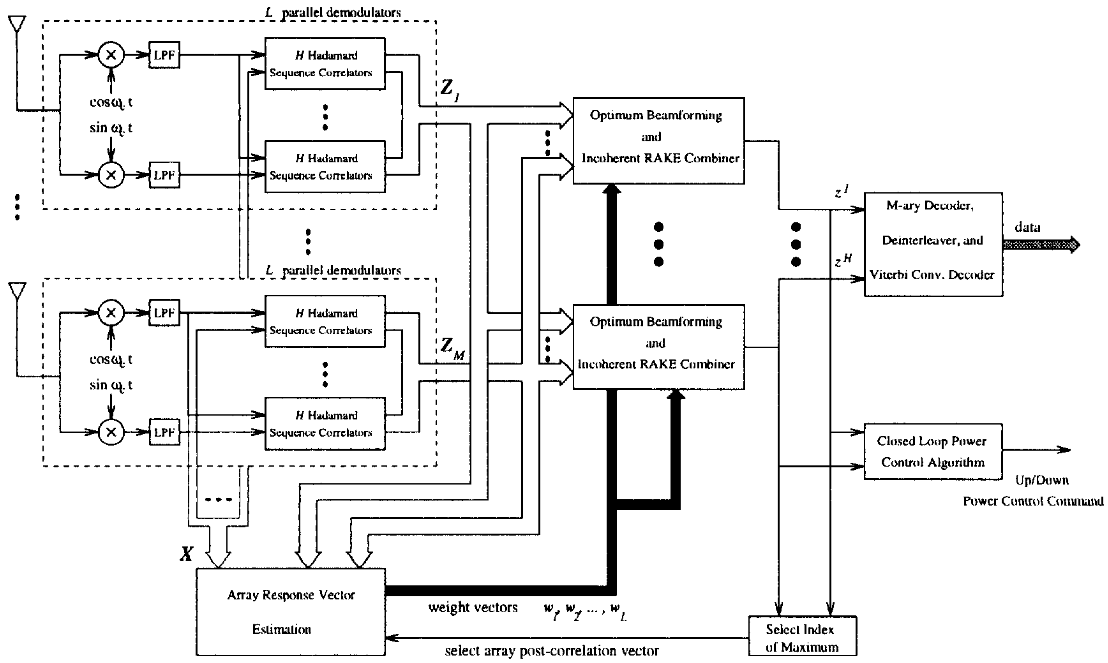

Figure 2: Receiver Block Diagram

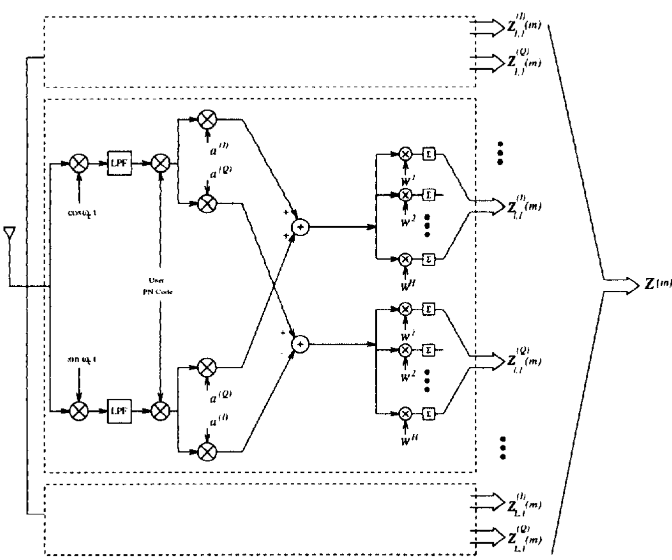

Figure 3: I and Q Hadamard-Walsh Correlators

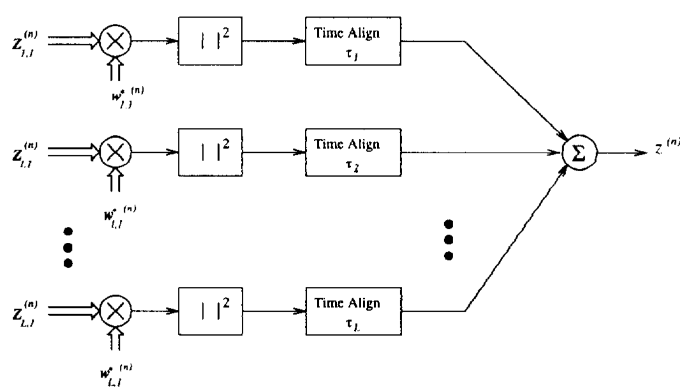

Figure 4: Optimum Beamforming and Incoherent RAKE Combining

vector and the array response vector of the new symbol will be quite different and losses will occur. In the next section, we will present some of our preliminary simulation results.

4. NUMERICAL AND SIMULATION RESULTS

We considered a base-station with three 120° sectors. In each sector, we considered a linear array of size 8λ. We considered the case when only thermal noise is present (*i.e.* no multiple access interference). We assumed that the number of multipath components $\mathcal{L}_1$ is 4. The array response vector was generated using the above channel model. We assumed that the noise power σ^2, the mobile power P_1, and the processing gain L are such that the average power in each multipath component $\frac{P_1 L}{\sigma^2} = 0$ dB. The results of a 100,000 runs with different values of $\Delta_{l,1}$ were used to estimate the distribution of $\frac{E_b}{N_o}$ at the output of the RAKE combiner. Figure 5 summerizes our results. This figure, shows the level above which $\frac{E_b}{N_o}$ is above 99% of the time as a function of the number of antennas in each sector's array. From this figure we can see the improvement in $\frac{E_b}{N_o}$. Also, note that with larger scattering angles, we get a slight additional improvement (order of 1dB) in $\frac{E_b}{N_o}$. The reason for this is that, with larger scattering angle, the correlation between the array response vector elements goes down according to the above model and hence an additional gain due to space diversity is obtained (since with low correlation a multipath component that is weak at one antenna is not necessarily so at another antenna). Note that our simulations did not include the effect of mobile speed or power control which will be included in a further study.

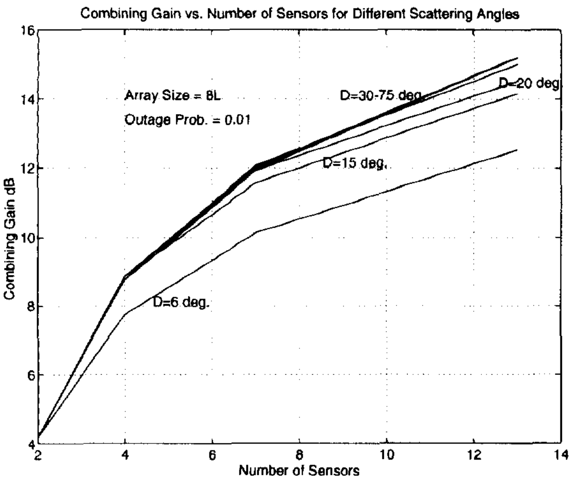

Figure 5: Beamforming gain for different Δ

5. CONCLUSIONS

In this paper, we presented an antenna array receiver structure for cellular DS/CDMA with M-ary orthogonal modulation. We developed the vector multipath channel and received signal models. We also derived an expression for the decision variables at the output of the RAKE combiner. We presented some of our preliminary simulations which show an improvement in the output $\frac{E_b}{N_o}$ due to the use of antenna arrays.

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