Space-Time Processing for Increased Capacity of Wireless CDMA

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Abstract— Spatial and temporal processing are combined to increase the capacity of CDMA-based wireless communications systems. Degrees of freedom provided by space-time processing are exploited to combat both fading and co-channel interference (the near-far effect). The following methods are formulated and studied: (1) space-time diversity, (2) cascade optimum spatial - diversity (RAKE) temporal, (3) cascade optimum spatial - optimum temporal, and (4) joint domain optimum processing. It is shown that, due to its interference cancellation capability, optimum combining provides significantly better performance than diversity processing. In particular, in a typical CDMA scenario with two antennas, the joint domain optimum combining system provides at least a 30% increase in capacity over diversity processing. Optimum combining may also be applied to compensate for imperfect power control of the signals received at the base station.

Introduction

Antenna arrays have been considered for increasing the capacity of CDMA-based wireless communication systems [1]. Early work in the application of adaptive antennas to mobile communication considered the flat fading channel model (narrowband communications) [2], [3]. As interest in CDMA systems grows, there is an increasing number of publications considering the benefits of antenna arrays to CDMA-based systems. A co-channel interference canceler based on null-steering is suggested in [4]. In [5] a space-time configuration that consists of an antenna array and a RAKE receiver is suggested. Further improvement is possible by considering optimum processing in both the space and time domains, either cascaded or jointly. We formulate and compare the following receiver configurations: (1) space-time maximum ratio combining (SMRC/TMRC), (2) cascade optimum space-MRC time (SOPT/TMRC) (optimum spatial processing cascaded with a RAKE receiver), (3) cascade optimum space-optimum time (SOPT/TOPT), and (4) joint domain optimum combining (JOPT). Configuration #2, SOPT/TMRC, is similar to the one studied in reference [5]. Thus our work generalizes the work in the reference and shows that the SOPT/TOPT and the JOPT structures provide better performance.

The sequence of this paper is organized as follows: the signal model is contained in section 2; we formulate the processing schemes and derive the various space-time filters weights in section 3; numerical results are given in section 4; near-far resistance is discussed in section 5, and the work is summarized in section 6.

Signal Model

Consider a cell site that serves $L+1$ users. The signal transmitted by the designated user is denoted $g_0(t)$, while signals transmitted by the other users are denoted $g_j(t)$, $j=1,\ldots,L$. The signals $g_j(t)$ are considered co-channel interference with respect to the designated user. The equivalent lowpass transmitted waveform for the ith user is

$$g_i(t) = A_i\, S_i\left(\lfloor \frac{t}{T_s} \rfloor\right) u_i(t), \tag{1}$$

where A_i denotes the amplitude of the signal. The amplitude of the signal transmitted by co-channel interference user j is $A_j = \psi_j\sqrt{P_j}$, where P_j is the power and ψ_j represents the voice activity modeled as a constant 3/8 for $j=1,\ldots,L$. In the case of perfect power control, all P_j are equal. Imperfect power control is characterized by the *power control error* (PCE) parameter. The notation $(\lfloor\cdot\rfloor)$ denotes the integer part, and $S_i(t)$ is the binary information data bit transmitted by the ith user with $E[|S_i(t)|^2]=1$. The users' signature waveforms, i.e., spreading waveforms, are represented by time functions $u_i(t)$, with period T_s, where T_s is the symbol interval. The spreading gain is given by $D=T_s/T_c$, with unit chip duration $T_c=1$, $D=\int_0^{T_s}|u_i(t)|^2dt$.

The typical CDMA channel is frequency-selective, i.e., the channel coherence bandwidth Δf_c is assumed smaller than the transmitted signal bandwidth W. This channel coherence bandwith is determined by the multipath spread T_m, with $\Delta f_c = 1/T_m$. Such a channel can be modeled with M resolvable paths [6], where $M=[T_mW]+1$. It is assumed that $T_s > T_m$.

The general configuration at the base station is shown in Figure 1. The base station uses an N-element uniform array with antennas set sufficiently apart such that the signals can be assumed to reach the antennas via independent paths, and hence can be modeled as independent random processes. We further assume perfect code synchronization. The signal received at nth antenna can be written:

$$\begin{aligned} x_n(t) =& \sum_{m=0}^{M-1} C_{nm}^{(0)}(t)\, g_0(t-mT_d-\tau_0) \\ &+ \sum_{j=1}^{L}\sum_{m=0}^{M-1} C_{nm}^{(j)}(t) g_j(t-mT_d-\tau_j) + v_n(t) \end{aligned} \tag{2}$$

where $C_{nm}^{(j)}(t)$, $j=0,\ldots,L$, $m=0,\ldots,M-1$ are random processes characterizing the fading channel as seen by each of the users. Samples of $C_{nm}^{(j)}(t)$ are modeled as zero-mean, complex-valued, stationary Gaussian random processes—statistically independent between users j and between paths m. The received signals are assumed asynchronous, with τ_j denoting the delay of the jth user with respect to an arbitrary time origin. $T_d = 1/W$ represents the tap-delay in the channel model. The additive noise $v_n(t)$ is modeled white complex Gaussian with a variance of σ_v^2. We further assume that the channel is characterized by

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slow fading such that $C_{nm}^{(j)}(t) = C_{nm}^{(j)}$ during the processing period. Without loss of generality we set the time origin such that $\tau_0 = 0$.

A demodulator is used at each antenna element to collect the energy of the received signal from all independent paths and to despread the signals. The demodulator consists of an M-tap delay line and matched filters. The tap-delay line compensates for the delay propagation in the channel, providing the time alignment for demodulation with the user's signature $u_0(t)$. The received signal at the n-th antenna passes through the tap delay line and is fed into the correlators for spread spectrum demodulation. The demodulator is shown in Figure 2. The output at the m-th tap correlator and the k-th symbol is given by:

$$\begin{aligned} y_{nm}(k) &= \int_{kT_s}^{(k+1)T_s} x_n(t+mT_d)u_0^*(t)dt \\ &= \int_{kT_s}^{(k+1)T_s} \sum_{i=0}^{M-1} C_{ni}^{(0)} g_0(t+mT_d-iT_d)u_0^*(t)\,dt \\ &+ \int_{kT_s}^{(k+1)T_s} \sum_{j=1}^{L}\sum_{i=0}^{M-1} C_{ni}^{(j)} g_j(t+mT_d-iT_d-\tau_j)u_0^*(t)dt \\ &+ \int_{kT_s}^{(k+1)T_s} v_n(t+mT_d)\,u_0^*(t)\,dt. \end{aligned} \quad (3)$$

The quantity $y_{nm}(k)$ can be written explicitly in terms of the contributions of the designated user and the co-channel interference:

$$\begin{aligned} y_{nm}(k) &= A_0 S_0(k) B_{nm(0)}^{(0)} + A_0 S_0(k+1) B_{nm(1)}^{(0)} \\ &+ A_0 S_0(k-1) B_{nm(-1)}^{(0)} + \sum_{j=1}^{L} [A_j S_j(k) B_{nm(0)}^{(j)} \\ &+ A_j S_j(k+1) B_{nm(1)}^{(j)} + A_j S_j(k-1) B_{nm(-1)}^{(j)}] \\ &+ \eta_{nm}(k), \end{aligned} \quad (4)$$

where $B_{nm(\alpha)}^{(0)}$, $\alpha = 0, 1, -1$ represents the aggregate cross-correlation with the designated user waveform $u_0(t)$ of all paths and symbols S_0, as seen at the m-th tap delay in the n-th channel. $B_{nm(\alpha)}^{(0)}$ consists of contributions of the current symbol $S_0(k)$, as well as the previous and the next symbols. Similarly, $B_{nm(\alpha)}^{(j)}$ represent contributions of the co-channel interferences $S_j(k+\alpha)$, $j = 1, \ldots, L$ to the output of the cross-correlation with $u_0(t)$. The term $\eta_{nm}(k)$ is the noise at the output of the matched filter. Consistent with the slow fading model assumed, all user and co-channel interference factors B are fixed during the processing interval. In this paper we are concerned with the signal processing applied to the $y_{nm}(k)'s$, the demodulated outputs of the array/tap-delay structure. Various space-time processing schemes are formulated in the next section.

Space-Time Combining Schemes

We consider two approaches to space-time processing: cascade space-time and joint domain processing. The cascade space-time processor consists of a temporal processor using the outputs of the spatial processor as shown in Figure 3. Processing could be maximum ratio combining or optimum in each domain. The joint processor is applied simultaneously to all the signals in the array/tap-delay line structure. In the following, first the spatial combiner is defined, then it is used to formulate the cascade space-time configurations. The section concludes with the presentation of joint domain processing.

A. Spatial Combiner

Spatial processing combines the signals following spread spectrum demodulation, i.e., it combines the signals y_{nm} for each m. Unlike conventional phased arrays where a single (line-of-sight) path is assumed between the source and the array, our model assumes that each sensor receives signals independent of the other sensors. This can be achieved by allowing sufficient separation between sensors. Define the array vector at the output of the m-th tap matched filter as $\mathbf{y}_m^T(k) = [y_{1m}(k), \ldots, y_{Nm}(k)]$. As can be seen from eq. (4), this output consists of the designated signal, interference and noise, and can be written:

$$\mathbf{y}_m(k) = A_0 S_0(k)\, \mathbf{B}_{m(0)}^{(0)} + \mathbf{I}_m + \mathbf{N}_m, \quad (5)$$

where $\mathbf{B}_{m(0)}^{(0)} = \left[B_{1m(0)}^{(0)}, \ldots, B_{Nm(0)}^{(0)}\right]^T$. This expression is similar in form to the output of a conventional phased array, with the random vector $\mathbf{B}_{m(0)}^{(0)}$ consisting of the transfer complex gain of the desired signal $S_0(k)$, from the source to each of the sensors. The interference vector can be expressed,

$$\mathbf{I}_m(k) = \sum_{j=0}^{L} \sum_{\substack{\alpha=-1 \\ \alpha,j\neq 0,0}}^{1} A_j S_j(k+i) \mathbf{B}_{m(\alpha)}^{(j)}. \quad (6)$$

Following spatial processing with the spatial weight vector $\mathbf{f}_m$, the array output at the m-th tap-delay is given by $z_m(k) = \mathbf{f}_m^H \mathbf{y}_m(k)$. Due to the independence between successive symbols of the same source, as well as the independence of the sources and the noise, the cross-correlation of the array vector and the desired signal is given by,

$$\begin{aligned} \mathbf{r}_m &= E[\mathbf{y}_m(k) S_o(k)] \\ &= A_0 \mathbf{B}_{m(0)}^{(0)}. \end{aligned} \quad (7)$$

A practical way to compute $\mathbf{r}_m$ using a training sequence or previous bit estimates $\widehat{S}_o(k)$:

$$\widehat{\mathbf{r}}_m = \frac{1}{K}\sum_{k=1}^{K} \mathbf{y}_m(k)\widehat{S}_o(k). \quad (8)$$

The SNIR is optimized by the optimum combining weight vector:

$$\mathbf{f}_m = \mathbf{R}_m^{-1}\mathbf{r}_m, \quad (9)$$

where $\mathbf{R}_m$ is the array interference and noise covariance matrix at the output of the m-th correlator and is defined by:

$$\mathbf{R}_m = E\left[(\mathbf{y}_m(k) - S_o(k)\mathbf{r}_m)(\mathbf{y}_m(k) - S_o(k)\mathbf{r}_m)^H\right]. \quad (10)$$

This covariance matrix can be estimated similar to the cross-correlation vector:

$$\widehat{\mathbf{R}}_m = \frac{1}{K}\sum_{k=1}^{K}\left(\mathbf{y}_m(k) - \widehat{S}_o(k)\widehat{\mathbf{r}}_m\right)\left(\mathbf{y}_m(k) - \widehat{S}_o(k)\widehat{\mathbf{r}}_m\right)^H. \quad (11)$$

B. Space-Time Combiner

Let $\mathbf{z}^T(k) = [z_1(k), \ldots, z_M(k)]$ be a vector that consists of the M outputs of the spatial combiners. The vector $z(k)$ is fed into the temporal combiner. It can be expressed:

$$\mathbf{z}(k) = A_0 S_0(k)\mathbf{H} + \mathbf{I}_t + \mathbf{N}_t, \qquad (12)$$

where $\mathbf{H}^T = \left[\mathbf{f}_0^H \mathbf{B}_{1(0)}^{(0)}, \ldots, \mathbf{f}_{M-1}^H \mathbf{B}_{M-1,(0)}^{(0)}\right]$, $\mathbf{I}_t^T = \left[\mathbf{f}_0^H \mathbf{I}_0, \ldots, \mathbf{f}_{M-1}^H \mathbf{I}_{M-1}\right]$, and $\mathbf{N}_t^T = \left[\mathbf{f}_0^H \mathbf{N}_0, \ldots, \mathbf{f}_{M-1}^H \mathbf{N}_{M-1}\right]$. The output of the space-time combiner is given by $\varsigma(k) = \mathbf{g}^H \mathbf{z}(k)$, where $\mathbf{g}$ is the temporal weight vector. The weight vector $\mathbf{g}$ is derived similar to the spatial weight vector $\mathbf{f}_m$. Several space-time cascade configurations are considered. (For acronyms, please refer to the introduction.) The SMRC/TMRC configuration is defined by

$$\mathbf{g} = \mathbf{H}, \qquad (13)$$

where in the definition of $\mathbf{H}$, the spatial MRC weight vectors $\mathbf{f}_m$ are given by $\mathbf{f}_m = \mathbf{r}_m$. Assuming independence between the various random variables, the cross-correlation between the space-time combiner output and the desired signal is given by

$$\begin{aligned} \mathbf{r}_t &= E[\mathbf{z}(k) S_o(k)] \\ &= A_0 \mathbf{H}. \end{aligned} \qquad (14)$$

Therefore the TMRC weight vector can be found from the estimate of the cross-correlation vector. With the SOPT/TMRC configuration, the SNIR at the output of the temporal processor is maximized by the optimum weight vector,

$$\mathbf{g} = \mathbf{R}_t^{-1} \mathbf{r}_t, \qquad (15)$$

where $\mathbf{R}_t$ is the array interference and noise covariance matrix at the input of the temporal combiner and is given by

$$\mathbf{R}_t = E\left[(\mathbf{z}(k) - S_o(k)\mathbf{r}_t)(\mathbf{z}(k) - S_o(k)\mathbf{r}_t)^H\right]. \qquad (16)$$

The table below summarizes the space-time receiver configurations.

Spatial Proc.	Temporal Proc.	Space-Time
$\mathbf{f}_m = \mathbf{r}_m$ (SMRC)	$\mathbf{g} = \mathbf{H}$ (TMRC)	SMRC/TMRC
$\mathbf{f}_m = \mathbf{R}_m^{-1} \mathbf{r}_m$ (SOPT)	$\mathbf{g} = \mathbf{H}$ (TMRC)	SOPT/TMRC
$\mathbf{f}_m = \mathbf{R}_m^{-1} \mathbf{r}_m$ (SOPT)	$\mathbf{g} = \mathbf{R}_t^{-1} \mathbf{r}_t$ (TOPT)	SOPT/TOPT

The SNIR at the output of the space-time combiner is given by the expression:

$$\rho = \frac{S_b}{I+N} = \frac{A_0^2 E\left[\left|\mathbf{g}^H \mathbf{H}\right|^2\right]}{E\left[\left|\mathbf{g}^H \mathbf{I}_t\right|^2\right] + E\left[\left|\mathbf{g}^H \mathbf{N}_t\right|^2\right]}, \qquad (17)$$

where S_b is the signal power, I is the interference power and N is the noise power. The expectation is taken with respect to the noise over a time interval during which the channel is considered constant. Thus ρ is viewed as a random variable and is a function of the channel parameters $B_{nm(\alpha)}^{(j)}$, the sources' power P_j, and the voice activity ψ_j. The conditional bit error rate (BER) is given by

$$P_{e|\rho} = Q\left(\sqrt{2\rho}\right). \qquad (18)$$

It is common practice to evaluate the system performance in terms of the outage, defined as the probability of the BER exceeding a set level. To derive an expression for the outage, consider the various components of ρ. The signal power S_b can be set to 1, without affecting the SNIR, by suitable scaling of the weight vector $\mathbf{g}$. The interference and noise, $I+N$ term, is the aggregate of a large number of independent sources, symbols and noise, hence, according to the central limit theorem, it may be regarded as a normal random variable with some mean μ_I and variance σ_I^2. The outage can then be calculated from [7],

$$\Pr\left[\frac{S_b}{I+N} \leq \xi\right] = Q\left(\frac{1/\xi - \mu_I}{\sigma_I}\right). \qquad (19)$$

C. Joint Domain Combiner

With the joint domain combiner, processing occurs jointly in the space-time domains. To formulate the weight vectors of the joint domain combiner, define the NM-dimensional stacked vector after spread spectrum demodulation:

$$\begin{aligned} \mathbf{Y}(k) &= \left[\mathbf{y}_0^T(k), \ldots, \mathbf{y}_{M-1}^T(k)\right]^T \\ &= A_0 S_0(k)\mathbf{B}_0 + \mathbf{I}(k) + \mathbf{N}(k), \end{aligned} \qquad (20)$$

where $\mathbf{B}_o^T = \left[\mathbf{B}_{0(0)}^{(0)T}, \ldots, \mathbf{B}_{M-1,(0)}^{(0)T}\right]$, $\mathbf{I}^T(k) = \left[\mathbf{I}_0^T(k), \ldots, \mathbf{I}_{M-1}^T(k)\right]$, and $\mathbf{N}^T(k) = \left[\mathbf{N}_0^T(k), \ldots, \mathbf{N}_{M-1}^T(k)\right]$. The joint domain cross-correlation vector and the space-time covariance matrix are given by

$$\mathbf{r} = E[\mathbf{Y}(k) S_o(k)] = A_o \mathbf{B}_o \qquad (21)$$

and

$$\begin{aligned} \mathbf{R} &= E[(\mathbf{I}(k) + \mathbf{N}(k))(\mathbf{I}(k) + \mathbf{N}(k))^H] \\ &= E[(\mathbf{Y}(k) - S_o(k)\mathbf{r})(\mathbf{Y}(k) - S_o(k)\mathbf{r})^H], \end{aligned} \qquad (22)$$

respectively. The JMRC weight vector is given by $\mathbf{w} = \mathbf{r}$. It is readily shown that, with proper scaling, the SMRC/TMRC and JMRC configurations provide exactly the same outputs. The cascade and joint configurations, however, do not provide the same performance when optimum processing is applied. This is illustrated in the next section. The JOPT weight vector is given by

$$\mathbf{w} = \mathbf{R}^{-1}\mathbf{r}. \qquad (23)$$

The difference in performance between the cascade and the joint domain weight vectors is a consequence of the number of degrees of freedom available to each configuration. The cascade configuration has $(N + M - 2)$ degrees of freedom while the joint domain configuration has $(NM - 1)$ degrees of freedom. In a typical CDMA scenario, the number of interferences $L \gg NM$, hence each additional degree of freedom provides increased performance. The outage for joint domain processing is evaluated using the following SNIR expression:

$$\rho = \frac{S_b}{I+N} = \frac{A_0^2 E\left[\left|\mathbf{w}^H \mathbf{B}_0\right|^2\right]}{E\left[\left|\mathbf{w}^H \mathbf{I}\right|^2\right] + E\left[\left|\mathbf{w}^H \mathbf{N}\right|^2\right]}. \qquad (24)$$

Numerical Results

This section presents numerical results on the performance of the configurations studied in the previous section. Both perfect and imperfect power control are considered, and the effect of the number of antennas on the BER and outage probability is studied. Based on a multipath coherence bandwidth of 300 kHz and a CDMA signal bandwidth of 1.25 MHz., the channel was modeled with four taps. The information symbols were modulated by Gold sequences of length 127. The SNR before spread spectrum demodulation was set to 0.8. For the case of perfect power control, all sources were assumed to have the same power as the desired signal. Covariance matrices and cross-correlation vectors were estimated from blocks of 50 symbols using relations similar to those in eqs. (11) and (8), respectively. Curves shown are averages of 100 Monte Carlo runs.

Figure 4 shows the outage as a function of the capacity for the different space-time configurations and for $N = 1, 2$ antennas. The power control was assumed perfect. According to the figure, the JOPT capacity, measured at an outage of 10^{-3}, is increased from 20 to 30 for a single antenna, and from 45 to 65 for two antennas. The configuration SOPT/TMRC provides only little improvement. Thus both JOPT and SOPT/TOPT are superior to SOPT/TMRC. This is not surprising since for two antennas there is only a single degree of freedom to exploit. The capacity of the SOPT/TOPT configuration is about half-way between the capacities of the SMRC/TMRC and JOPT configurations. The effect of imperfect power control on capacity can be assessed from Figure 5. It is observed that a 2 dB PCE generally causes a significant drop in capacity for each configuration. However, JOPT with power error is better than SMRC/TMRC without perfect power control. Thus, optimum combining can be viewed as providing a trade-off to power control.

Near-Far Resistance

The output of the demodulator has the following approximation

$$\begin{aligned} \mathbf{Y}(k) &= A_o S_o(k)\mathbf{B}_o + \mathbf{I}(k) + \mathbf{N}(k) \\ &\approx A_o D S_o(k)\mathbf{C} + \mathbf{I}(k) + \mathbf{N}(k), \end{aligned} \tag{25}$$

assuming inter-path contributions are small. $\mathbf{C} = [C_{10}^{(0)}, \ldots, C_{N(M-1)}^{(0)}]$ is the channel coefficient vector of the desired user. To simplify the analysis, we define the normalized channel vector $\mathbf{C} = \mathbf{c}/|\mathbf{c}|$ such that the array processing gain $|\mathbf{C}| = 1$. The space-time processor's output is given by

$$\zeta = \mathbf{w}^H\mathbf{Y} = A_o D S_o(k)\mathbf{w}^H\mathbf{C} + \mathbf{w}^H\mathbf{I}(k) + \mathbf{w}^H\mathbf{N}(k). \tag{26}$$

The noise covariance matrix is given by $E[\mathbf{N}(k)\mathbf{N}(k)^H] = D\sigma_v I$, where I is the identity matrix. The worst case is when all interferences combine to reduce the output due to the desired signal. We have the minimum SNR at the output

$$(SNR)_{min} = E\left[\frac{(A_o D|\mathbf{w}^H\mathbf{C}| - |\mathbf{w}^H\mathbf{I}(k)|)^2}{D\sigma_v^2|\mathbf{w}|^2}\right]. \tag{27}$$

The $(SNR)_{min}$ determines the probability of error P_e for the desired user. To achieve this error probability P_e, the SNR at the input of space-time processor or the output of the demodulator is required as $(SNR)_{in} = \frac{A_o^2 D^2}{D\sigma_v^2} = \frac{A_o^2 D}{\sigma_v^2}$, The *asymptotic efficiency* is defined as the ratio $(SNR)_{min}/(SNR)_{in}$ in the region of low noise power which can be written as follows:

$$\eta = \frac{(SNR)_{min}}{(SNR)_{in}} = E\left\{\frac{\left[\max\left(0, |\mathbf{w}^H\mathbf{C}| - \frac{|\mathbf{w}^H\mathbf{I}(k)|}{A_o D}\right)\right]^2}{|\mathbf{w}|^2}\right\}. \tag{28}$$

Notice that in the absence of interferences ($\mathbf{I}(k) = 0$) and since $|\mathbf{C}| = 1$, the asymptotic efficiency equals to 1 ($\eta = 1$).

Let us consider one interference synchronized case. There, the interference term can be written as $\mathbf{I}(k) \approx A_I S_I(k)\mathbf{C}_I$, where $\mathbf{C}_I$ is the normalized channel coefficient vector of the interference user with $|\mathbf{C}_I| = 1$. The MRC method yields the weight vector $\mathbf{w}_m = \mathbf{C}$, so that the asymptotic efficiency is given by:

$$\begin{aligned} \eta_{mrc} &= E\left\{\left[\max\left(0, 1 - \frac{A_I}{A_o D}|\mathbf{C}^H\mathbf{C}_I|\right)\right]^2\right\} \\ &\approx \left[\max\left(0, 1 - \sqrt{\frac{\pi}{NM}} \cdot \frac{A_I}{2A_o D}\right)\right]^2. \end{aligned} \tag{29}$$

When the interference power increases, the asymptotic efficiency decreases until it vanishes. Hence, the MRC is not near-far resistant.

With optimum combining the weight vector is given by $\mathbf{w}_o = \mathbf{R}_n^{-1}\mathbf{C}$, where $\mathbf{R}_n = A_I^2\mathbf{C}_I\mathbf{C}_I^H + D\sigma_v^2\mathbf{I}$. Using the matrix inversion lemma, $\mathbf{R}_n^{-1} = \left(\mathbf{I} - \alpha\mathbf{C}_I\mathbf{C}_I^H\right)/(D\sigma_v^2)$, where $\alpha = A_I^2/(D\sigma_v^2 + A_I^2)$. The asymptotic efficiency of optimum combining can be written as follows:

$$\eta_{opt} = E\left[\frac{\left(1 - \alpha|\mathbf{C}^H\mathbf{C}_I|^2 - \frac{\alpha\sigma_v^2}{A_o A_I}|\mathbf{C}^H\mathbf{C}_I|\right)^2}{1 - \alpha(2-\alpha)|\mathbf{C}^H\mathbf{C}_I|^2}\right]. \tag{30}$$

In the region of low noise power, we have $\alpha \to 1$ and $\alpha\sigma_v^2/(A_o A_I) \to 0$ so that the asymptotic efficiency of optimum combining can be further simplified as follows:

$$\begin{aligned} \eta_{opt} &= E\left[1 - |\mathbf{C}^H\mathbf{C}_I|^2\right] \\ &= 1 - \frac{1}{NM}. \end{aligned} \tag{31}$$

It means that space-time processing with the optimum combining is near-far resistant in the one interference case.

Figure 6 shows the asymptotic efficiency of an adaptive array for two synchronized users with spread gain $D = 127$, system dimension $NM = 8$. Both simulation results and theory calculations show that the optimum combining is near-far resistant while MRC is not. The simulation curves were obtained by using eq. (28) and averaging over 1000 runs while the theory curves for the MRC and the optimum combining were obtained by using eq. (29) and eq. (31), respectively.

When the number of the interferences, (which include delayed version of multipaths), is smaller than the system dimension, optimum combining has sufficient degrees of freedom to null all interferences making the system near-far resistant. When the number of the interferences is larger than the system dimension, optimum combining cannot null all the interferences and the system ceases to be near-far resistant.

Conclusions

This paper studied space-time processing for CDMA-based wireless communications. Specifically, the following configurations were suggested: (1) space-time diversity, (2) cascade optimum spatial-diversity temporal (RAKE), (3) cascade optimum spatial-optimum temporal, and (4) joint domain optimum processing. We considered a typical CDMA wireless communications scenario modeled by a four tap-delay line and showed that a two-antenna system provides double the capacity of a single antenna system. Joint domain optimum combining provides an additional 25% increase in capacity over diversity processing

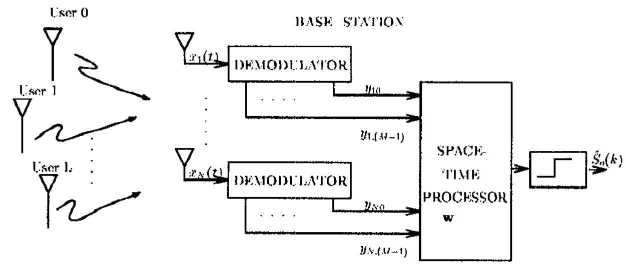

Fig. 1. General configuration of the space-time CDMA receiver.

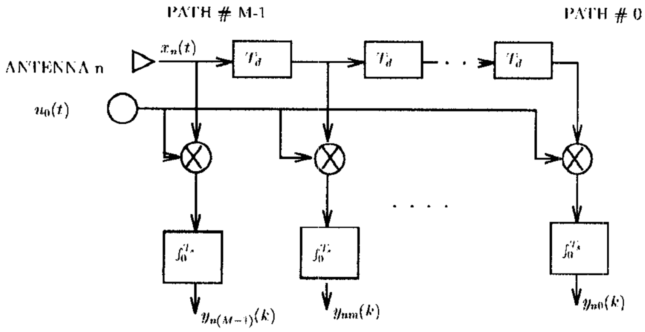

Fig. 2. Configuration of the demodulator

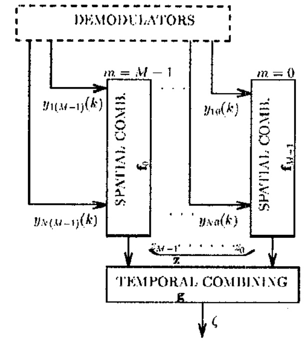

Fig. 3. Configuration of cascade space-time processing

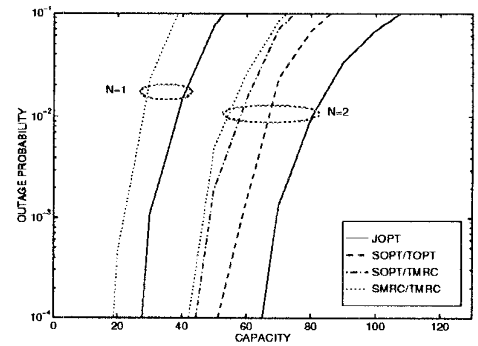

Fig. 4. Outage probability vs. the capacity with perfect power control.

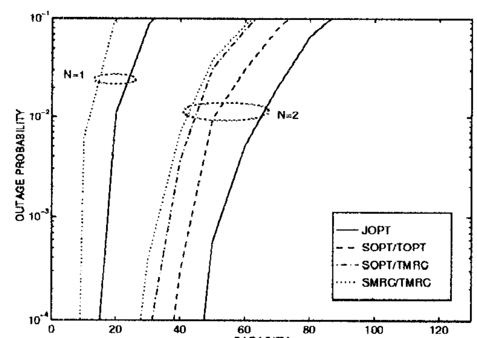

Fig. 5. Outage probability vs. the capacity with PCE=2dB.

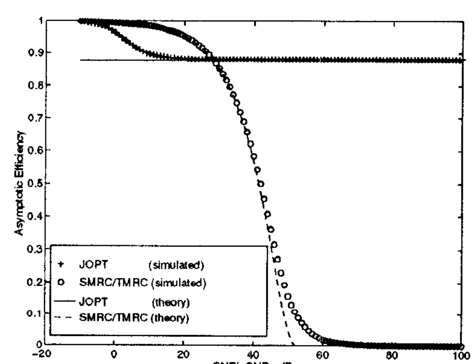

Fig. 6. Asymptotic efficiency of an adaptive array for two users with NM=8

only. Optimum combining may be also used to compensate for power control errors. These improvements come at only modest cost due to the few degrees of freedom that need to be optimized.

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