PERFORMANCE OF SPACE-TIME RECEIVER ARCHITECTURES FOR CDMA OVERLAY OF NARROWBAND WAVEFORMS FOR PERSONAL COMMUNICATION SYSTEMS

A. Shah and A.M. Haimovich
Department of Electrical and Computer Engineering
New Jersey Institute of Technology
Newark, NJ 07102

Abstract

This paper describes the performance analysis of a direct-sequence (DS) code-division multiple-access (CDMA) personal communication system (PCS) sharing a common spectrum with narrowband microwave radio links. Spatial and temporal processing are combined to improve the performance of the DS-CDMA system subject to narrowband interference. Degrees of freedom (DOF) provided by space-time (ST) optimum combining schemes are exploited to combat both fading and interference. The performance of several space-time receiver architectures is evaluated in terms of average bit error rate (BER) and it is investigated how this performance is influenced by various parameters.

1. Introduction

The concept of an overlay has been proposed for both the PCS band [1] and the cellular band [2]. In such scenarios, DS-CDMA signals are overlaid on existing narrowband user signals, thereby increasing overall spectrum efficiency [3]. A number of authors have studied the performance of DS-CDMA overlay system [4, 5], with a narrowband BPSK signal as an interference. In this paper we are concerned with the performance of the reverse link (mobile to base) of a BPSK DS-CDMA system that overlays a narrowband BPSK signal in a frequency-selective, slowly fading Rayleigh channel. The interference from adjacent cells is not considered.

Consider a cell site with L active users. The transmitted signal for the k-th DS-CDMA user is given by

$$S_k(t) = \sqrt{2P_k}\; d_k(t)\; u_k(t)\; \cos(2\pi f_o t + \theta_k), \qquad (1)$$

where P_k is the average power of the signal, $d_k(t) \in \{-1, 1\}$ is a random binary sequence representing data, $u_k(t) \in \{-1, 1\}$ is the signature sequence, f_o is the carrier frequency, and θ_k is a random phase uniformly distributed between $[0, 2\pi]$. The processing gain is $D = T_b/T_c$, where T_b is the bit duration and T_c is the chip duration. For a DS-CDMA signal with chip duration T_c, the spread spectrum bandwidth is approximately $B_s = 2/T_c$.

The time-variant frequency-selective Rayleigh fading channel is modeled as a tap-delay line with tap spacing T_c, and tap coefficients $\{c_m(t)\}$. The complex lowpass equivalent channel impulse response is given by [6]

$$h(\tau, t) = \sum_{m=0}^{M-1} c_m(t)\; \delta(t - \tau - mT_c), \qquad (2)$$

where $M = \lfloor \frac{T_m}{T_c} \rfloor + 1$ is the number of resolvable paths and $\delta(\cdot)$ denotes the Dirac impulse function. The notation $\lfloor \cdot \rfloor$ denotes the integer part. The total multipath delay spread T_m of the channel is assumed to be much smaller than the bit duration T_b. The channel coefficients $\{c_m(t)\}$ are modeled as zero-mean, complex-valued, stationary, mutually independent Gaussian random processes. The channel is characterized by slow fading such that $\{c_m(t)\} = \{c_m\}$ during the processing interval.

The narrowband interference is assumed to be a non-fading BPSK signal given by

$$J(t) = \sqrt{2J}\; b(t)\; \cos(2\pi(f_o + \nu)t + \vartheta), \qquad (3)$$

where ν is the offset of the interference carrier frequency from the DS-CDMA signal carrier frequency and parameters J and ϑ denote the received interference power and phase, respectively. The information sequence $b(t) \in \{-1, 1\}$ has bit rate $1/T_i$, where T_i is the bit duration. The interference bandwidth $B_i = 2/T_i$ and $B_i \ll B_s$. The ratio of the interference bandwidth to the DS-CDMA bandwidth is given by $p = \frac{B_i}{B_s} = \frac{T_c}{T_i}$. The ratio of the offset of the interference carrier frequency to half of the DS-CDMA bandwidth is defined by $q = \frac{2\nu}{B_s} = \nu T_c$.

The base station uses an N-element uniform linear array with array elements assumed sufficiently separated such that spatial diversity (independent fading at each receive antenna) is achieved with respect to the DS-CDMA signals. The signal received from the desired user is denoted by the subscript '1', while other signals are considered co-channel interferences. The equivalent baseband received signal at the n-th antenna can be written:

$$x_n(t) \;=\; \sqrt{P} \sum_{m=0}^{M-1} c_{nm}^{(1)} d_1(t - mT_c - \tau_1) u_1(t - mT_c - \tau_1)$$

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$$+ \quad \xi_n(t) + v_n(t) + \sqrt{P}\sum_{k=2}^{L}\sum_{m=0}^{M-1} c_{nm}^{(k)} d_k(t - mT_c - \tau_k) u_k(t - mT_c - \tau_k), \qquad (4)$$

where $\{c_{nm}^{(k)}\}$, $k = 1, \ldots, L$, $m = 0, \ldots, (M-1)$, $n = 1, \ldots, N$ represent the complex-valued tap coefficients of the fading channels as seen by each of the DS-CDMA users and N is the number of antennas in the array. Samples of $\{c_{nm}^{(k)}\}$ are statistically independent between users k, between paths m, and between antennas n. The fact that all $\{P_k = P\}$ implies that the base station provides adaptive power control such that received signals from all L users have the same average power (to overcome the near-far problem). The received signals are assumed asynchronous, with τ_k denoting the delay of the k-th user uniformly distributed between $[0, T_b]$. Without loss of generality, we set $\tau_1 = 0$. The quantity $\xi_n(t)$ is the narrowband interference at the n-th antenna and is given by $\xi_n(t) = \sqrt{J}b(t)e^{j(2\pi\nu t + \vartheta)}e^{j\phi_n}$, where ϕ_n is the electrical angle of the interference at the n-th antenna. The additive noise $v_n(t)$ is modeled as complex white Gaussian with zero-mean and single-sided power spectral density of N_o. We assume perfect code synchronization.

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 spread spectrum demodulation with the desired user's signature $u_1(t)$. The general configuration at the base station is shown in Figure 1. The demodulator is shown in Figure 2. The output at the m-th tap correlator at the n-th antenna for the l-th symbol is given by

$$\begin{aligned} y_{nm}(l) &= \int_{lT_b}^{(l+1)T_b} x_n(t + mT_c)u_1(t)dt \\ &= \int_{lT_b}^{(l+1)T_b} \{\sqrt{P}\sum_{i=0}^{M-1} c_{ni}^{(1)} d_1(t + mT_c - iT_c) \\ &\quad u_1(t + mT_c - iT_c)\}u_1(t)dt + \int_{lT_b}^{(l+1)T_b} \\ &\quad \{\sqrt{P}\sum_{k=2}^{L}\sum_{i=0}^{M-1} c_{ni}^{(k)} d_k(t + mT_c - iT_c - \tau_k) \\ &\quad u_k(t + mT_c - iT_c - \tau_k)\}u_1(t)dt \\ &+ \int_{lT_b}^{(l+1)T_b} \xi_n(t + mT_c)u_1(t)dt \\ &+ \int_{lT_b}^{(l+1)T_b} v_n(t + mT_c)u_1(t)dt. \end{aligned} \qquad (5)$$

The quantity $y_{nm}(l)$ can be written explicitly in terms of the contributions of the desired user, co-channel interferences, narrowband interference and noise:

$$\begin{aligned} y_{nm}(l) &= \sqrt{P}d_1(l)B_{nm(0)}^{(1)} + \sqrt{P}d_1(l+1)B_{nm(1)}^{(1)} \\ &+ \sqrt{P}d_1(l-1)B_{nm(-1)}^{(1)} + \xi_{nm}(l) + \eta_{nm}(l) \\ &+ \sum_{k=2}^{L}\Big\{\sqrt{P}d_k(l)B_{nm(0)}^{(k)} + \sqrt{P}d_k(l+1)B_{nm(1)}^{(k)} \\ &+ \sqrt{P}d_k(l-1)B_{nm(-1)}^{(k)}\Big\}, \end{aligned} \qquad (6)$$

where $B_{nm(\alpha)}^{(1)}$, $\alpha = 0, 1, -1$, represents the aggregate cross-correlation with the desired user signature waveform $u_1(t)$ of all paths and symbols d_1, as seen at the m-th tap delay at the n-th antenna. $B_{nm(\alpha)}^{(1)}$ consists of contributions of the current symbol $d_1(l)$, as well as the previous and the next symbols. Similarly, $B_{nm(\alpha)}^{(k)}$ represent contributions of the co-channel interferences $d_k(l+\alpha)$, $k = 2, \ldots, L$, to the output of the cross-correlation with $u_1(t)$. The terms $\xi_{nm}(l)$ and $\eta_{nm}(l)$ are narrowband interference and noise at the output of the matched filter, respectively. Subsequent to the slow fading model assumed, coefficients $\{B_{nm(\alpha)}^{(k)}\}$ are constant during the processing interval.

2. Space-Time Optimum Combining

We consider two approaches to space-time (ST) optimum combining: cascade ST and joint ST. The cascade ST processor consists of a temporal processor using the outputs of the spatial processor as shown in Figure 3. The joint ST processor is applied simultaneously to all the signals in the array/tap-delay line structure.

Spatial Combiner

Spatial processing combines the signals following spread spectrum demodulation, i.e., it combines the signals y_{nm} for each m. Define the N-dimensional array vector at the output of the m-th tap matched filter for the l-th symbol as $\mathbf{y}_m^T(l) = [y_{1m}(l), \ldots, y_{Nm}(l)]$. As can be seen from 6, this output consists of the desired user signal, interference and noise, and can be written as

$$\mathbf{y}_m(l) = \sqrt{P}d_1(l)\mathbf{B}_{m(0)}^{(1)} + \boldsymbol{\Upsilon}_m(l) + \boldsymbol{\Psi}_m(l), \qquad (7)$$

where $\mathbf{B}_{m(0)}^{(1)} = \left[B_{1m(0)}^{(1)}, \ldots, B_{Nm(0)}^{(1)}\right]^T$. The term $\boldsymbol{\Psi}_m(l) = [\eta_{1m}(l), \ldots, \eta_{Nm}(l)]^T$ is the noise array vector. The interference array vector, which consists of self-interference, co-channel interferences, and narrowband interference, can be expressed as

$$\boldsymbol{\Upsilon}_m(l) = \sum_{k=1}^{L}\sum_{\substack{\alpha=-1 \\ \alpha\neq 0 \text{ if } k=1}}^{1} \sqrt{P}d_k(l+\alpha)\mathbf{B}_{m(\alpha)}^{(k)} + \boldsymbol{\Xi}_m(l), \qquad (8)$$

where $\boldsymbol{\Xi}_m(l) = [\xi_{1m}(l), \ldots, \xi_{Nm}(l)]^T$ is the narrowband interference vector. The reason α is not equal to 0 when k equals 1 is because it is the desired term. Following spatial processing with the spatial weight vector $\mathbf{f}_m$, the spatial output at the m-th tap-delay line is $z_m(l) = \mathbf{f}_m^H\mathbf{y}_m(l)$. Due to the independence between successive symbols of the same user, as well as mutual independence between DS-CDMA users, the narrowband interference, and the noise, the cross-correlation of the array vector and the desired signal at the m-th tap-delay is given by

$$\mathbf{r}_m = \mathrm{E}\left[\mathbf{y}_m(l)d_1(l)\right] = \sqrt{P}\mathbf{B}^{(1)}_{m(0)}. \quad (9)$$

The cross-correlation vector $\mathbf{r}_m$ can be estimated using a training sequence or previous bit estimates $\widehat{d}_1(l)$:

$$\widehat{\mathbf{r}}_m = \frac{1}{K}\sum_{i=1}^{K}\mathbf{y}_m(i)\widehat{d}_1(i), \quad (10)$$

where K is the number of samples used in the estimation. The optimum weight vector, which maximizes the signal-to-interference-plus-noise ratio (SINR) at the output of the spatial combiner is given by

$$\mathbf{f}_m = \mathbf{R}_m^{-1}\mathbf{r}_m, \quad (11)$$

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

$$\mathbf{R}_m = \mathrm{E}\left[\{\mathbf{y}_m(l) - d_1(l)\mathbf{r}_m\}\{\mathbf{y}_m(l) - d_1(l)\mathbf{r}_m\}^H\right]. \quad (12)$$

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

$$\widehat{\mathbf{R}}_m = \frac{1}{K}\sum_{i=1}^{K}\left[\mathbf{y}_m(i) - \widehat{d}_1(i)\widehat{\mathbf{r}_m}\right]\left[\mathbf{y}_m(i) - \widehat{d}_1(i)\widehat{\mathbf{r}_m}\right]^H. \quad (13)$$

Space-Time Combiner

Let $\mathbf{z}(l) = [z_0(l), \ldots, z_{M-1}(l)]$ be a vector that consists of the M outputs of the spatial combiners. The vector $\mathbf{z}(l)$ is fed into the temporal combiner. It can be expressed as

$$\mathbf{z}(l) = \sqrt{P}d_1(l)\mathbf{U} + \mathbf{\Upsilon}_t + \mathbf{\Psi}_t, \quad (14)$$

where $\mathbf{U}^T = \left[\mathbf{f}_0^H\mathbf{B}^{(1)}_{0(0)}, \ldots, \mathbf{f}_{M-1}^H\mathbf{B}^{(1)}_{(M-1)(0)}\right]$, the interference vector $\mathbf{\Upsilon}_t^T = \left[\mathbf{f}_0^H\mathbf{\Upsilon}_0, \ldots, \mathbf{f}_{M-1}^H\mathbf{\Upsilon}_{(M-1)}\right]$, and the noise vector $\mathbf{\Psi}_t^T = \left[\mathbf{f}_0^H\mathbf{\Psi}_0, \ldots, \mathbf{f}_{M-1}^H\mathbf{\Psi}_{(M-1)}\right]$. The output of the cascade ST combiner is $\rho(l) = \mathbf{g}^H\mathbf{z}(l)$, where g is the temporal weight vector. Assuming mutual independence of the terms in (14), the cascade ST cross-correlation vector is given by

$$\mathbf{r}_t = \mathrm{E}\left[\mathbf{z}(l)d_1(l)\right] = \sqrt{P}\mathbf{U}. \quad (15)$$

The optimum weight vector, which maximizes SINR at the output of the temporal processor is given by

$$\mathbf{g} = \mathbf{R}_t^{-1}\mathbf{r}_t, \quad (16)$$

where $\mathbf{R}_t$ is the interference-plus-noise covariance matrix at the input of the temporal combiner and is defined as:

$$\mathbf{R}_t = \mathrm{E}\left[\{\mathbf{z}(l) - d_1(l)\mathbf{r}_t\}\{\mathbf{z}(l) - d_1(l)\mathbf{r}_t\}^H\right]. \quad (17)$$

The SINR at the output of the cascade ST combiner is given by

$$\mu = \frac{P\mathrm{E}\left[\left|\mathbf{g}^H\mathbf{U}\right|^2\right]}{\mathrm{E}\left[|\mathbf{g}^H\mathbf{\Upsilon}_t|^2\right] + \mathrm{E}\left[|\mathbf{g}^H\mathbf{\Psi}_t|^2\right]}, \quad (18)$$

where P is the desired signal power and E denotes expectation. The expectation is taken with respect to the noise over a time interval during which the channel is considered constant. Thus, μ is a random variable parameterized by the channel $B^{(k)}_{nm(\alpha)}$. When the number of active users L is large and power control is incorporated, then according to the central limit theorem, the sum of the residual interference (self, co-channel, and narrowband) and noise can be approximated by a Gaussian random variable. Thus, BER conditioned on μ (a specific set of fading channels, under the above Gaussian assumption) is given by

$$P_{e|\mu} = \mathrm{Q}(\sqrt{2\mu}), \quad (19)$$

where $\mathrm{Q}(\cdot)$ is the area under the tail of the gaussian probability density function. The unconditional BER, i.e., the one averaged over all fading channels, is then given by

$$P_e = \int_0^\infty \mathrm{Q}(\sqrt{2\mu})\, f_\mu(\mu)\, d\mu, \quad (20)$$

where $f_\mu(\mu)$ is the probability density function of μ.

Joint Domain Combiner

With the joint domain combiner, processing is carried out simultaneously in the space-time domains. To formulate the joint space-time optimal weight vector, define the NM-dimensional stacked vector for the l-th symbol after the spread spectrum demodulation as

$$\begin{aligned}\mathbf{Y}(l) &= \left[\mathbf{y}_0^T(l), \ldots, \mathbf{y}_{(M-1)}^T(l)\right]^T \\ &= \sqrt{P}d_1(l)\mathbf{B}_0 + \mathbf{\Upsilon}(l) + \mathbf{\Psi}(l),\end{aligned} \quad (21)$$

where $\mathbf{B}_0^T = \left[\mathbf{B}^{(1)T}_{0(0)}, \ldots, \mathbf{B}^{(1)T}_{(M-1),(0)}\right]$, the interference vector $\mathbf{\Upsilon}^T(l) = \left[\mathbf{\Upsilon}_0^T(l), \ldots, \mathbf{\Upsilon}_{(M-1)}^T(l)\right]$, and the noise vector $\mathbf{\Psi}^T(l) = \left[\mathbf{\Psi}_0^T(l), \ldots, \mathbf{\Psi}_{(M-1)}^T(l)\right]$. The joint domain cross-correlation vector and interference-plus-noise covariance matrix are given by

$$\mathbf{r} = \mathrm{E}[\mathbf{Y}(l)d_1(l)] = \sqrt{P}\mathbf{B}_0 \quad (22)$$

and

$$\mathbf{R} = \mathrm{E}\left[\{\mathbf{Y}(l) - d_1(l)\mathbf{r}\}\{\mathbf{Y}(l) - d_1(l)\mathbf{r}\}^H\right], \quad (23)$$

respectively. The joint domain weight vector, which maximizes the SINR, is given by

$$\mathbf{w} = \mathbf{R}^{-1}\mathbf{r}. \quad (24)$$

The SINR at the output of the joint ST combiner is given by

$$\mu = \frac{P\mathrm{E}\left[\left|\mathbf{w}^H\mathbf{B}_0\right|^2\right]}{\mathrm{E}\left[|\mathbf{w}^H\mathbf{\Upsilon}|^2\right] + \mathrm{E}\left[|\mathbf{w}^H\mathbf{\Psi}|^2\right]}. \quad (25)$$

3. Numerical Results

This section presents numerical results on the performance of the space-time optimum combining schemes studied in the previous section. The channel was modeled with $M = 4$ taps. The data symbols were modulated by Gold sequences of length 127. The interference-to-signal ratio at the input of the correlators is $J/S = 25$ dB, where the signal power $S = P\beta$, and $\beta = \mathrm{E}[| c_{nm}^{(k)} |^2]$. Covariance matrices and the cross-correlation vectors were estimated from blocks of 31 symbols using relations similar to those in (13) and (10), respectively. Curves shown are averages of 2000 Monte Carlo runs.

Figure 4 shows the BER performance of the DS-CDMA system as a function of the average total signal-to-noise ratio $\bar{\gamma}_b$ for $N = 2$. The average total signal-to-noise ratio is $\bar{\gamma}_b = NM\bar{\gamma}_c$, where $\bar{\gamma}_c$ is the average signal-to-noise ratio per path and is given by $\bar{\gamma}_c = \frac{E_b}{N_o}\beta$, and $E_b = PT_b$ is the bit energy. Clearly, the joint ST processor outperforms the cascade version. The difference in performance between the cascade and the joint processors is a consequence of the number of DOF available to each configurations. The cascade configuration has $(N + M - 2)$ DOF, with respect to interference cancellation, while the joint domain configuration has $(NM - 1)$. Since the number of co-channel interferences $L \gg NM$, each additional DOF provides increased performance. Note that as $\bar{\gamma}_b$ increases, the performance of both configurations becomes interference limited.

In Figure 5, the asymptotic ($N_o \to 0$) BER of the system is shown as a function of the ratio (p) of the interference bandwidth to the DS-CDMA signal bandwidth. The performance of cascade ST processor is invariant with respect to interference bandwidth and serves as the upper bound for the joint ST processor when the interference bandwidth approaches DS-CDMA signal bandwidth.

Figure 6 demonstrates the asymptotic BER performance of the ST configurations as a function of the ratio (q) of the offset of the interference carrier frequency to the half DS-CDMA signal bandwidth. The performance of the ST processors is very robust to the change in q. When $q = 1$ the narrowband interference is outside the DS-CDMA frequency band.

In Figure 7, the asymptotic BER of the DS-CDMA system as a function of the number of active users L is shown for different values of J/S. It is seen that for large J/S and for a given BER, the joint ST receiver can support many more users than can the cascade ST receiver. When $J/S \leq 10$ dB and the number of active users L is large, both configurations show similar performance.

4. Conclusions

In this paper we studied the performance of two space-time receiver configurations for DS-CDMA based PCS in an environment with narrowband interference. The performance of cascade ST processor is very robust with respect to the bandwidth of the narrowband signal and serves as the upper bound for the joint ST processor when the interference bandwidth approaches DS-CDMA signal bandwidth. It was shown that the performance of both ST receivers is robust with respect to change in interference carrier frequency. Although the space-time optimum combining schemes require matrix inversion, the matrix dimension is not large. We conclude that the space-time optimum combining schemes can be implemented to enhance both the performance and the capacity of a DS-CDMA system overlaying narrowband waveforms.

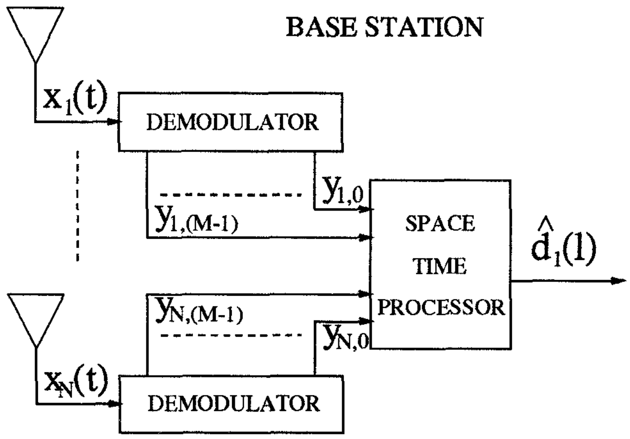

Figure 1: General configuration of space-time receiver.

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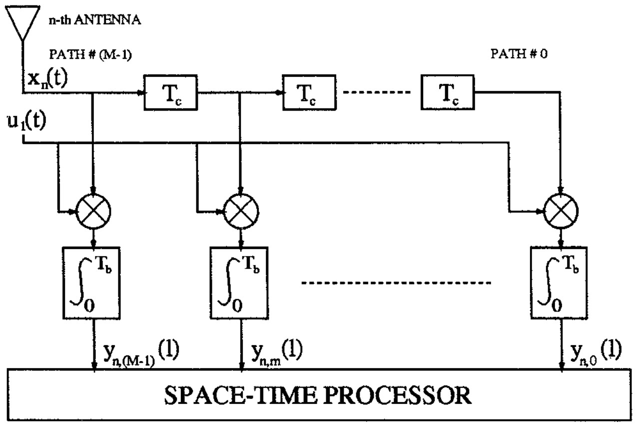

Figure 2: Demodulator.

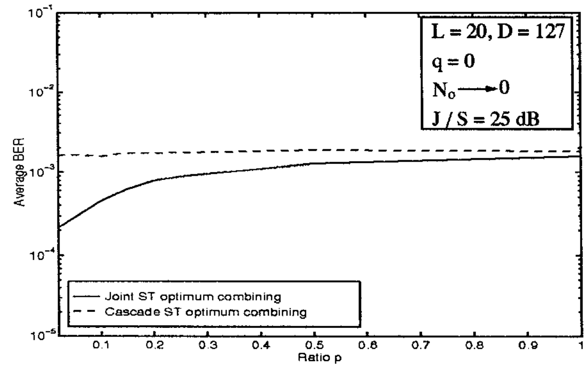

Figure 5: Performance of ST receiver as a function of ratio p.

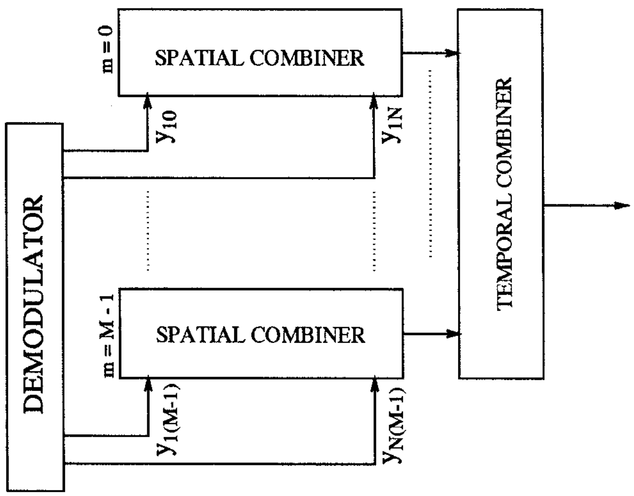

Figure 3: Configuration of cascade space-time processing.

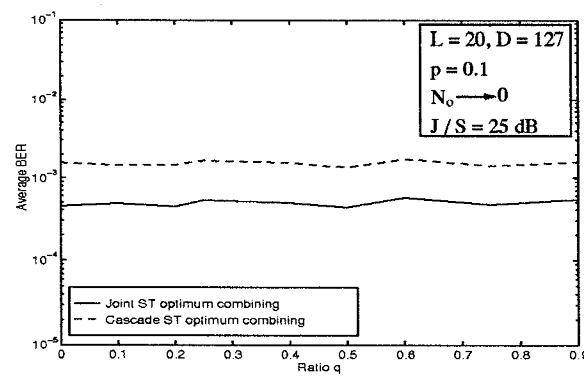

Figure 6: Performance of ST receiver as a function of ratio q.

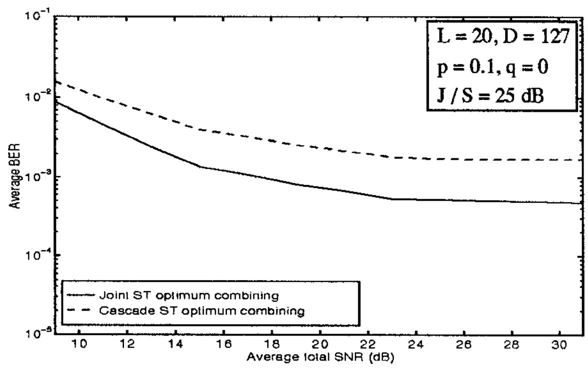

Figure 4: Performance of ST receiver for DS-CDMA overlay system.

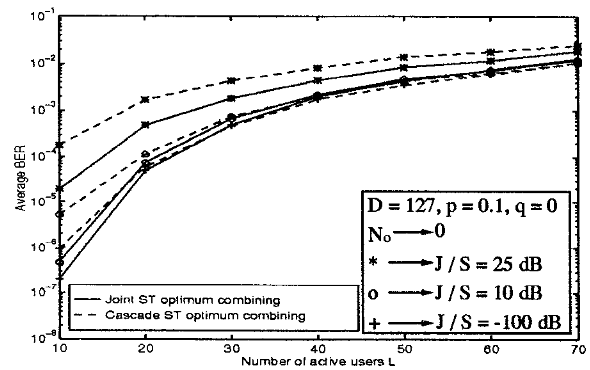

Figure 7: Performance of ST receiver as a function of number of active users L.