# BIT ERROR COMPUTATIONS FOR SPACE-TIME NARROWBAND INTERFERENCE SUPPRESSION IN CDMA COMMUNICATIONS

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

### Abstract

Spatial and temporal processing are combined to suppress narrowband interference in CDMA communications. Space-time (ST) processing provides degrees of freedom (DOF) for both interference cancellation and diversity combining. Two ST receiver architectures, cascade and joint, are studied. The main contributions of the paper are the development of new analytical expressions of (1) the asymptotic efficiency of each method, (2) the probability density function (PDF) of the signal-to-interference-plus-noise ratio (SINR) at the array output in a frequency selective Rayleigh fading environment, and (3) the average probability of bit error (BER) associated with each method.

### 1. Introduction

The need to suppress narrowband signals in CDMA systems arises in applications where narrowband signals are overlaid with wideband signals to increase spectral efficiency. This concept has been proposed for both the personal communication systems (PCS) band and the cellular band. The co-existence of these two different systems within the same frequency spectrum, will cause interference to both systems. In this work we are concerned with the interference caused by the narrowband signal to the DS-CDMA signal.

The conventional approach to rejecting narrowband interference has been to sample the received signal at the chip interval, and to exploit the high correlation between the interference samples prior to spread spectrum demodulation. This method essentially places a notch at the narrowband interference frequency. The notch, however, also removes a portion of the DS-CDMA signal. As the bandwidth of the interference increases, the notch widens and the DS-CDMA signal loss becomes more significant. A number of authors have explored the performance of DS-CDMA overlay system [1, 2] with a narrowband BPSK signal as an interference.

A different approach is the use of antenna arrays. In [3], the various ST receiver architectures considered were shown to combat co-channel interferences and fading in DS-CDMA based wireless communication systems. In this paper the performance of space-time receiver architectures, cascade and joint, is evaluated for suppressing a narrowband interference overlaid with a DS-CDMA signal in a frequency-selective slowly fading Rayleigh channel. Closed-form expressions are obtained for asymptotic efficiency, probability density function (PDF) of output SINR, and error probability associated with each architecture.

### 2. System Model

Consider the uplink of a coherent mobile communication system. The lowpass equivalent of the transmitted DS-CDMA signal is given by $s(t) = \sqrt{P}d(t)u(t)$, where $P$ is the signal power, $d(t) \in \{-1,1\}$ is the data bit with duration $T_b$, and $u(t)$ is the signature waveform with chip duration $T_c$. The signature sequence $u(t)$ is assumed such that

$$\int_0^{T_b} u(t-iT_c)u(t-jT_c)\,dt = \begin{cases} T_b & i=j \\ 0 & i \neq j, \end{cases} \qquad (1)$$

where $0 \leq i,j \leq (L-1)$ and $L = \frac{T_b}{T_c}$ is the processing gain. The time-variant frequency-selective Rayleigh fading channel is modeled as a tap-delay line with tap spacing $T_c$ and tap coefficients $\{c_m(t)\}$, where $0 \leq m \leq M-1$ and $M$ is the number of resolvable paths. Subsequent to slow fading assumption, we have $\{c_m(t)\} = \{c_m\}$ during the processing interval.

The narrowband interference is assumed to be a non-fading BPSK signal, and is defined by its equivalent lowpass representation as $J(t) = \sqrt{J}b(t)e^{j(2\pi\nu t+\vartheta)}$, where $\nu$ is the offset of the interference carrier frequency from the carrier frequency of DS-CDMA signal. The parameters $J$ and $\vartheta$ 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 ratio of the interference bandwidth to the DS-CDMA bandwidth is given by $p = \frac{B_i}{B_s} = \frac{T_c}{T_i}$ and $0 < p \leq 1$.

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 signal. The equivalent baseband received signal at the $n$-th antenna can be written:

$$x_n(t) = \sqrt{P}\sum_{m=0}^{M-1} c_{nm}d(t-mT_c)u(t-mT_c) + \xi_n(t) + v_n(t),$$

0-7803-3925-8/97 $10.00 

where $\{c_{nm}\}$, $n = 1, \ldots, N$, $m = 0, \ldots, M-1$, represent the complex-valued tap coefficients of the fading channels as seen by DS-CDMA user. Samples of $\{c_{nm}\}$ are statistically independent between paths $m$, and between antennas $n$. 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 $\phi_n$ 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 variance $\sigma^2$. 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 signal. The demodulator consists of an $M$ tap-delay line and matched filters. 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

$$\begin{aligned} y_{nm}(l) &= \int_{lT_b}^{(l+1)T_b} x_n(t+mT_c)\, u(t)\, dt \\ &= \sqrt{P}Ld(l)c_{nm} + \xi_{nm}(l) + \eta_{nm}(l), \end{aligned} \quad (2)$$

where $\xi_{nm}(l)$ and $\eta_{nm}(l)$ are narrowband interference and noise at the output of the matched filter. The last line in (2) follows from the assumption made in (1).

## 3. Spatial Processing

In this section we consider spatial processing only, i.e., all the terms except for $m = 0$ vanishes in (2). In other words, we consider flat Rayleigh fading at each antenna. Define an $N$ dimensional array vector $\mathbf{y}(l)$ for the $l$-th symbol at the output of the matched filter as

$$\mathbf{y}(l) = \sqrt{P}Ld(l)\mathbf{c} + \mathbf{\Upsilon}(l) + \mathbf{\Psi}(l), \quad (3)$$

where $\mathbf{c} = [c_1, \ldots, c_N]^T$, is the vector of channel coefficients. The terms $\mathbf{\Upsilon}(l) = [\xi_1(l), \ldots, \xi_N(l)]^T$ and $\mathbf{\Psi}(l) = [\eta_1(l), \ldots, \eta_N(l)]^T$ are interference and noise spatial vectors, respectively. The maximum SINR at the output of the array is [4]:

$$\mu = PL^2\mathbf{c}^H\mathbf{R}_{ni}^{-1}\mathbf{c}, \quad (4)$$

where $\mathbf{R}_{ni} = \mathbf{\Upsilon}\mathbf{\Upsilon}^H + (\sigma^2 L)\mathbf{I}_N$, is the interference-plus-noise covariance matrix at the output of the matched filters and $\mathbf{I}_N$ is the identity matrix.

Using unitary transformation in (4), the PDF of $\mu$ can be shown to be given by

$$f_\mu(\mu) = \frac{(JN+\sigma^2)\mu^{N-1}e^{\frac{-(JN+\sigma^2)\mu}{P_s}}\,{}_1F_1\left(N-1; N; \frac{JN\mu}{P_s}\right)}{\sigma^2\Gamma(N)h^N} \quad (5)$$

where $P_s = PL\mathrm{E}\left[|c_n|^2\right], n = 1, \ldots, N$, is the mean desired signal power per antenna element, $h = \frac{P_s}{\sigma^2}$ is the mean signal-to-noise ratio per antenna element and ${}_1F_1(\cdot)$ is the Kummer's confluent hypergeometric function.

The asymptotic efficiency is defined as the ratio $\zeta = \frac{\gamma_{\text{eff}}}{\gamma_0}$ in the region of low noise power, where $\gamma_{\text{eff}}$ is measured in the presence of the interference, and $\gamma_0$ is observed with the interference absent. Using (4), $\gamma_{\text{eff}} = \mathrm{E}[\mu] = \frac{P_s}{JN+\sigma^2} + \frac{(N-1)P_s}{\sigma^2}$. Without interference, $\gamma_0 = \frac{NP_s}{\sigma^2}$. Hence the asymptotic efficiency of the spatial processor is given by $\zeta = 1 - \frac{1}{N}$. This relation clearly illustrates the loss of one DOF incurred by the interference cancellation process.

## 4. Space-Time Processing

In this section two space-time processing schemes, cascade and joint, are formulated and their performance evaluated.

### 4.1. Cascade Space-Time Processing

The cascade ST receiver consists of a temporal processor using the outputs of the spatial processor. Using (2), define the $N$-dimensional array vector at the output of the $m$-th tap matched filter for the $l$-th symbol as $\mathbf{y}_m^T(l) = [y_{1m}(l), \ldots, y_{Nm}(l)]$. The optimum weight vector, which maximizes SINR, is given by $\mathbf{f}_m = \mathbf{R}_m^{-1}\mathbf{r}_m$, where $\mathbf{r}_m = \mathrm{E}[\mathbf{y}_m(l)d(l)] = \sqrt{P}L\mathbf{c}_m$ is the cross-correlation vector, $\mathbf{R}_m = \mathrm{E}\left[\{\mathbf{y}_m(l) - d(l)\mathbf{r}_m\}\{\mathbf{y}_m(l) - d(l)\mathbf{r}_m\}^H\right]$ is the interference-plus-noise covariance matrix at the output of the $m$-th correlator and $\mathbf{c}_m^T = [c_{1m}, \ldots, c_{Nm}]$ is the vector of channel coefficients at the $m$-th tap delay. Following spatial processing with the spatial weight vector $\mathbf{f}_m$, the spatial output at the $m$-th tap-delay line is $z_m(l) = \mathbf{f}_m^H\mathbf{y}_m(l)$. 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. Define the output of the ST combiner as $\rho(l) = \mathbf{g}^H\mathbf{z}(l)$, where $\mathbf{g}$ is the temporal weight vector. We consider two ways to combine the elements of $\mathbf{z}(l)$: (1) straight combiner ($\mathbf{g} = \mathbf{1}$) and (2) optimum combiner in which $\mathbf{g}$ is derived in a similar fashion as the optimum spatial weight vector $\mathbf{f}_m$.

#### Straight Temporal Combiner ($\mathbf{g} = \mathbf{1}$)

Due to mutual independence of the spatial outputs $\{z_m(l)\}$, the SINR at the output of the cascade ST processor is $\mu_{st1} = \sum_{m=0}^{M-1}\mu_m$, where $\mu_m$ is the SINR associated with each of the spatial combiner. The PDF of $\mu_{st1}$ can be shown to be given by

$$\begin{aligned} f_{\mu_{st1}}(\mu) &= \frac{(JN+\sigma^2)^M\mu^{NM-1}e^{\frac{-(JN+\sigma^2)\mu}{P_s}}}{(\sigma^2)^M\Gamma(NM)h^{NM}} \\ &\quad {}_1F_1\left((N-1)M; NM; \frac{JN\mu}{P_s}\right), \end{aligned} \quad (6)$$

where $P_s = PL\mathrm{E}\left[|c_{nm}|^2\right], n = 1, \ldots, N, m = 0, \ldots, M-1$.

Following procedure similar to that in section 3, the asymptotic efficiency of the cascade optimum space-straight temporal (COSST) processor can be calculated to be $\zeta_{st1} = 1 - \frac{1}{N}$. This relation clearly demonstrates a loss of $M$ DOF out of a total of $NM$ due to presence of interference.

#### Optimum Temporal Combiner

The vector $\mathbf{z}(l)$, fed into the temporal combiner, can be expressed as $\mathbf{z}(l) = \sqrt{P}Ld(l)\mathbf{B} + \mathbf{\Upsilon}_t + \mathbf{\Psi}_t$, where $\mathbf{B}^T = \left[\mathbf{f}_0^H\mathbf{c}_0, \ldots, \mathbf{f}_{M-1}^H\mathbf{c}_{M-1}\right]$, $\mathbf{\Upsilon}_t^T = \left[\mathbf{f}_0^H\mathbf{\Upsilon}_0, \ldots, \mathbf{f}_{M-1}^H\mathbf{\Upsilon}_{M-1}\right]$ is the interference vector and $\mathbf{\Psi}_t^T = \left[\mathbf{f}_0^H\mathbf{\Psi}_0, \dots, \mathbf{f}_{M-1}^H\mathbf{\Psi}_{M-1}\right]$ is the noise vector. The optimum temporal weight vector $\mathbf{g}$, which maximizes SINR, is given by $\mathbf{g} = \mathbf{R}_t^{-1}\mathbf{r}_t$, where $\mathbf{r}_t = \mathrm{E}[\mathbf{z}(l)d(l)] = \sqrt{P}L\mathbf{B}$ is the cross-correlation vector, $\mathbf{R}_t = \mathrm{E}\left[\{\mathbf{z}(l) - d(l)\mathbf{r}_t\}\{\mathbf{z}(l) - d(l)\mathbf{r}_t\}^H\right]$ is the interference-plus-noise covariance matrix at the input of the temporal combiner. Hence, the SINR at the output of cascade optimum space-optimum temporal (COSOT) processor is given by $\mu_{st2} = \frac{PL^2\mathrm{E}\left[|\mathbf{g}^H\mathbf{B}|^2\right]}{\mathrm{E}\left[|\mathbf{g}^H\mathbf{\Upsilon}_t|^2\right] + \mathrm{E}\left[|\mathbf{g}^H\mathbf{\Psi}_t|^2\right]}$. The expectation is taken with respect to the noise over a time interval during which the channel is considered constant (due to slow fading assumption, channel coefficients are constant for several bit intervals). Thus, $\mu_{st2}$ is as a random variable parameterized by the channel coefficients $\{c_{nm}\}$. Unfortunately, the PDF of $\mu_{st2}$ is not known.

### 4.2. Joint Space-Time Processing

With the joint ST combiner, processing is carried out simultaneously in the space-time domains. Define the $NM$-dimensional stacked vector for the $l$-th symbol after the spread spectrum demodulation as

$$\mathbf{Y}(l) == \sqrt{P}Ld(l)\mathbf{C} + \mathbf{\Upsilon}(l) + \mathbf{\Psi}(l), \tag{7}$$

where $\mathbf{C}^T = \left[\mathbf{c}_0^T, \dots, \mathbf{c}_{M-1}^T\right]$ is the vector of channel coefficients, $\mathbf{\Upsilon}^T(l) = \left[\mathbf{\Upsilon}_0^T(l), \dots, \mathbf{\Upsilon}_{M-1}^T(l)\right]$ is the interference vector and $\mathbf{\Psi}^T(l) = \left[\mathbf{\Psi}_0^T(l), \dots, \mathbf{\Psi}_{M-1}^T(l)\right]$ is the noise vector. Using procedure of section 3, the maximum SINR at the output of the joint ST combiner can be shown to be given by

$$\mu_{jd} = PL^2 \sum_{l=1}^{NM} \frac{|s_l|^2}{\lambda_l}, \tag{8}$$

where $\mathbf{s}^T = [s_1, \dots, s_{NM}]$, $\{\lambda_l\}$ are the eigenvalues of $\mathbf{R} = \mathbf{\Upsilon}\mathbf{\Upsilon}^H + (\sigma^2 L)\mathbf{I}_{NM}$ and $\mathbf{I}_{NM}$ is the identity matrix. Each of $\{|s_l|^2\}$ is a chi-square random variable with two degrees of freedom. The eigenvalues $\{\lambda_l\}$ of $\mathbf{R}$ are

$$\lambda_l = \begin{cases} \lambda_l & l = 1, \dots, r \\ \sigma^2 L & l = r+1, \dots, NM, \end{cases}$$

where $1 \leq r \leq M$. The value of $r$ depends on the bandwidth (BW) of the narrowband interference, which in turn determines the number of principal eigenvalues (eigenvalues containing most of the interference power) of $\mathbf{R}$. The number of principal eigenvalues can be predicted by the Landau-Pollak theorem and is $r \approx p(M-1) + 1$, where $p = \frac{T_c}{T_i}$ and $M$ is the number of taps in the temporal processor. Clearly, when the interference BW is very small compared to DS-CDMA signal BW, $r = 1$, whereas when the interference BW is same as DS-CDMA BW, $r = M$. Also, when $M = 1$, i.e., spatial processing only, $r = 1$ regardless of the value of $p$ demonstrating the robustness of spatial processor with respect to interference BW.

The PDF of $\mu_{jd}$ can be shown to be given by

$$f_{\mu_{jd}}(\mu) = \frac{\gamma_o{}^{-(NM-r)}\mu^{(NM-r)}}{\Gamma(NM-r+1)} \sum_{l=1}^{r} \frac{\pi_l e^{\frac{-\mu}{\gamma_l}}}{\gamma_l} \, {}_1F_1(NM-r; NM-r+1; (\frac{1}{\gamma_l} - \frac{1}{\gamma_o})\mu) \tag{9}$$

where $\pi_l = \prod_{k=1,k\neq l}^{r} \frac{\gamma_l}{\gamma_l - \gamma_k}$, $\gamma_l = \frac{LP_s}{\lambda_l}$ and $\gamma_o = \frac{P_s}{\sigma^2}$. Here, it was assumed that $\{\lambda_l, l = 1, \dots, r\}$ are distinct.

The asymptotic efficiency of the joint space-time optimum combiner can be shown to be given by $\zeta_{jd} = 1 - \frac{r}{NM}$ where $1 \leq r \leq M$. Interestingly, when the interference BW equals DS-CDMA BW, i.e., $r = M$, the asymptotic efficiencies of the COSST combiner and joint domain optimum combiner are identical.

## 5. Probability of Error Calculations

In this section we calculate the average probability of bit error (BER) for the spatial optimum combiner, COSST combiner and joint ST optimum combiner. The sum of residual interference and noise at the output of the ST combiner (array) is assumed Gaussian. The BER is simply

$$P_e = \int_0^\infty \frac{1}{2}\mathrm{erfc}(\sqrt{\mu}) f_\mu(\mu) d\mu \tag{10}$$

where $\mathrm{erfc}(\cdot)$ is the complementary error function, and $f_\mu(\mu)$ is the PDF of $\mu$.

Using (5), (10) and [5, 6], the BER of the spatial optimum combiner can be calculated and is

$$\begin{aligned} P_e &= \frac{1}{2} - \frac{(JN+\sigma^2)\Gamma(N+\frac{1}{2})P_s^{N+\frac{1}{2}}}{\sqrt{\pi}\sigma^2\Gamma(N)h^N(JN+\sigma^2+P_s)^{N+\frac{1}{2}}} \\ &\quad F_2\left(N+\frac{1}{2}, 1, N-1; \frac{3}{2}, N;\right. \\ &\quad \left.\frac{P_s}{JN+\sigma^2+P_s}, \frac{JN}{JN+\sigma^2+P_s}\right), \end{aligned} \tag{11}$$

where $F_2(\cdot)$ is Appell's hypergeometric function of two variables.

Using (6), (10) and [5, 6], the BER of the COSST combiner can be calculated and is

$$\begin{aligned} P_e &= \frac{1}{2} - \frac{(JN+\sigma^2)^M\Gamma(NM+\frac{1}{2})}{\sqrt{\pi}(\sigma^2)^M\Gamma(NM)h^{NM}(\frac{JN+\sigma^2+P_s}{P_s})^{NM+\frac{1}{2}}} \\ &\quad F_2\left(NM+\frac{1}{2}, 1, (N-1)M; \frac{3}{2}, NM;\right. \\ &\quad \left.\frac{P_s}{JN+\sigma^2+P_s}, \frac{JN}{JN+\sigma^2+P_s}\right) \end{aligned} \tag{12}$$

Using (9), (10) and [5, 6], the BER of the joint ST optimum combiner can be calculated and is

$$\begin{aligned} P_e &= \frac{1}{2} - \sum_{l=1}^{r} \frac{\Gamma(NM-r+\frac{3}{2})\pi_l(1+\frac{1}{\gamma_l})^{-NM+r-\frac{3}{2}}}{\sqrt{\pi}\Gamma(NM-r+1)\gamma_o{}^{(NM-r)}\gamma_l} \\ &\quad F_2\left(NM-r+\frac{3}{2}, 1, NM-r; \frac{3}{2}, NM-r+1;\right. \\ &\quad \left.\frac{\gamma_l}{\gamma_l+1}, \frac{1-\frac{\gamma_l}{\gamma_o}}{\gamma_l+1}\right) \end{aligned} \tag{13}$$

Equations (11), (12) and (13), though exact, do not offer any meaningful insight. Hence, meaningful upper bounds of these BERs are derived.

### 5.1. Upper Bounds for the BER

Using the bound $\text{erfc}(\sqrt{\mu}) < \frac{e^{-u}}{\sqrt{\pi}}$ in (10), the upper bound for the BER of the spatial optimum combiner is

$$P_e < \frac{(JN+\sigma^2)}{2\sqrt{\pi}(JN+\sigma^2+P_s)(1+\frac{P_s}{\sigma^2})^{N-1}}. \quad (14)$$

This expression provides meaningful insight into a few special cases. When there is no interference, i.e., $J = 0$, $P_e < \frac{1}{2\sqrt{\pi}(1+h)^N}$. This is the upper bound for the BER of an $N$-order space diversity receiver employing maximal ratio combining (MRC) and without interference, as expected. When the interference power is infinite, i.e., $J \to \infty$, the expression becomes $P_e < \frac{1}{2\sqrt{\pi}(1+h)^{N-1}}$. This is the upper bound for the BER of an $(N-1)$-order diversity MRC receiver without interference. This implies that presence of interference with infinite power in the channel results in the loss of one diversity path.

The upper bound for the BER of the COSST combiner is

$$P_e < \frac{(JN+\sigma^2)^M}{2\sqrt{\pi}(JN+\sigma^2+P_s)^M(1+\frac{P_s}{\sigma^2})^{(N-1)M}}. \quad (15)$$

Clearly, the COSST combiner, without interference, performs as an $NM$-order diversity receiver, whereas presence of an interference with infinite power results in a loss of $M$ DOF.

Similarly, the upper bound for the BER of the joint ST optimum combiner is

$$P_e < \frac{1}{2\sqrt{\pi}}\sum_{l=1}^{r}\frac{\pi_l}{(1+\gamma_l)(1+\gamma_o)^{NM-r}}. \quad (16)$$

These expressions for the upper bounds clearly demonstrate that the interference cancellation entails a loss of DOF, with a corresponding loss in the diversity performance.

## 6. Numerical Results

This section presents both analytical and simulations results on the performance of the space-time combining schemes studied in the previous sections. The channel was modeled with $M = 4$ taps. The data symbols were modulated by Gold sequence of length $L = 31$ and $T_b = T_i = LT_c = 31$. The interference-to-signal ratio prior to spread spectrum demodulation (at the input of the correlators) is $J/S = 25$ dB, where the signal power $S = P\beta$, and $\beta = \text{E}\{\mid c_{nm} \mid^2\}$. The number of antennas $N = 2$ and the offset of the interference carrier frequency $\nu = \frac{1}{LT_c} = \frac{1}{31}$.

In Figure 3, the BER is plotted as a function of the average total SNR $= NM\frac{P_s}{\sigma_2}$. Covariance matrices and the cross-correlation vectors used for optimal combining were estimated from blocks of 50 samples. The simulations results shown are averages of 4000 Monte Carlo runs. The COSST simulations provide a good match to the theoretical BER curve. In this case the ratio of the number of samples used to estimate the covariance matrix and the signal dimension is $\frac{50}{2}$ (optimum combining is carried out only in the spatial domain with two antennas). Notice that the COSST and the COSOT show similar performance. For the joint ST case the simulation curve indicates slightly higher BER's than predicted by theory. This is explained by the covariance matrix estimation errors. In this case the ratio of the number of samples used to estimate the covariance matrix and the signal dimension is $\frac{50}{8}$ $(NM = 8)$. The effect of the number of samples used to estimate the covariance matrix is illustrated in Figure 4 where the joint ST simulations were generated using 50 and 100 samples estimates. A closer match between theory and simulations is clearly evident when the number of samples used to estimate the covariance matrix is increased.

In Figure 5, the BER of the overlay system is shown as a function of the ratio $p$ of the interference BW to the DS-CDMA signal BW. The performance of the COSST and the COSOT combiners is not affected by the interference BW as mentioned in section 4. The performance of the joint ST combiner approaches that of the COSST when the $p = 1$, as indicated by the asymptotic efficiency expressions derived in section 4.

## 7. Conclusions

In this paper we developed new analytical expressions of the asymptotic efficiency, the PDF of the SINR at the array output in a frequency selective Rayleigh fading environment, and the BER associated with several space-time receiver architectures. The simulations results corroborated the analytical results.

# References


[1] J. Wang and L. B. Milstein, "CDMA overlay situations for microcellular mobile communications," *IEEE Transactions on Communications*, vol. 43, pp. 603–614, February, March, April 1995.

[2] P. Wei et al., "Adaptive interference suppression for CDMA based PCN systems," *IEEE International Conference on Communications, Geneva, Switzerland*, pp. 1890–1894, May 1993.

[3] X. Wu and A. Haimovich, "Space-Time Optimum Combining for CDMA Communications," *Wireless Personal Communications*, no. 3, pp. 73–89, 1996.

[4] R. A. Monzingo and T. W. Miller, *Introduction to Adaptive Arrays.* New York, USA: John Wiley & Sons, Inc., 1980.

[5] L. C. Andrews, *Special Functions for Engineers and Applied Mathematicians.* New York, USA: Macmillan Publishing Company, 1985.

[6] H. Exton, *Handbook of Hypergeometric Integrals: Theory, Applications, Tables, Computer Programs.* New York, USA: John Wiley & Sons, Inc., 1978.

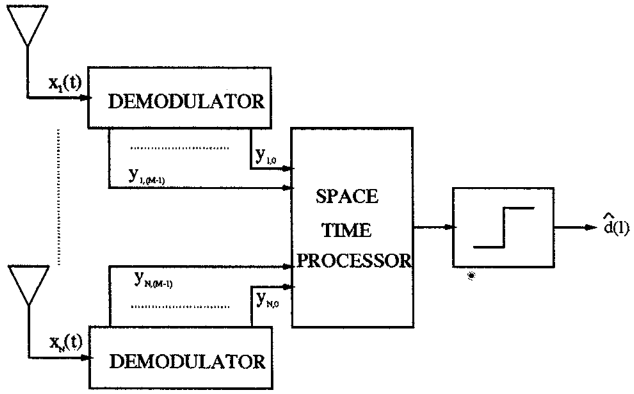


Figure 1: General configuration of space-time receiver.

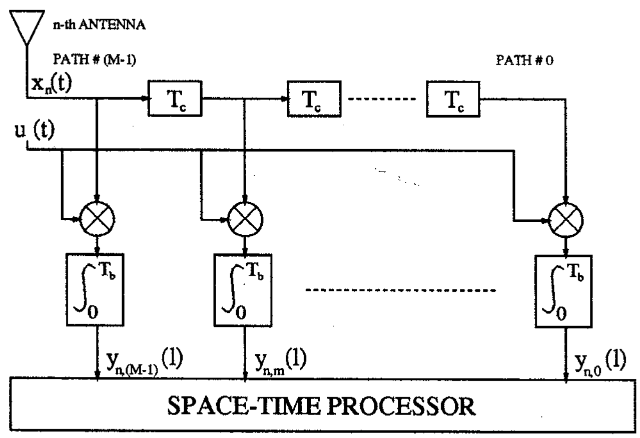


Figure 2: Demodulator.

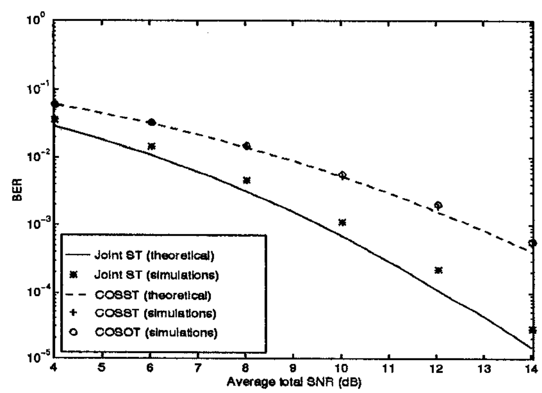


Figure 3: BER of space-time receiver for DS-CDMA overlay system.

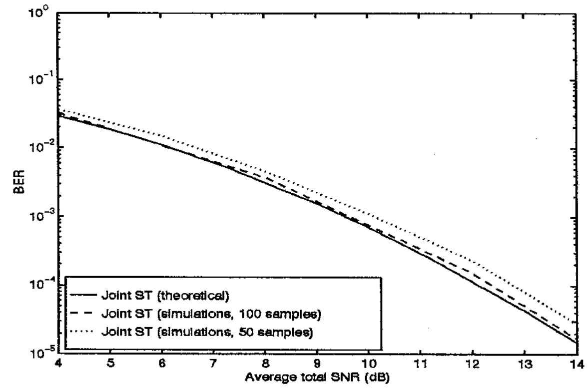


Figure 4: BER of the joint space-time optimum combiner.

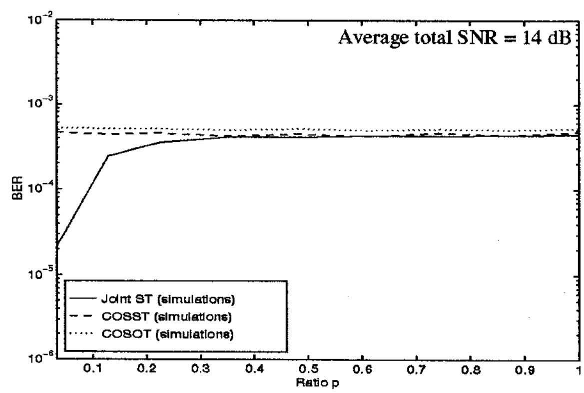


Figure 5: Performance of the space-time receiver as a function of ratio p.