Transmit Diversity With More Than Two Antennas

R. Michael Buehrer, Robert A. Soni and Quinn Li
Bell Laboratories - Lucent Technologies
67 Whippany Rd. Room 3A-220
Whippany, NJ 07981
mbuehrer@bell-labs.com

Abstract Recently, a new form of transmit diversity has been developed and included for *cdma2000* , the third generation successor for IS-95 code division multiple access (CDMA) systems. This transmit diversity scheme was developed by Bell Laboratories and uses Space-Time coding techniques. This paper presents performance results of this new transmit diversity method, termed space-time spreading, and investigates "open-loop" improvements to the scheme which use more than two antennas. Additionally, we investigate the performance improvements possible with closed loop techniques.

1 Introduction

Recently, Phase II of the *cdma2000* standardization process has been completed where a review of forward link antenna techniques was completed. A few different schemes were proposed as possible enhancements for the system. The standard originally supported a method of transmit diversity known as orthogonal transmit diversity (OTD). This method offered significant performance gains for rate 1/4 convolutional codes at low speeds, but did not offer the same types of gains for rate 1/2 codes. The scheme is open-loop, and makes no use of user specific data such as location or condition of its channel, other than through user independent power control.

Through the efforts of Bell Laboratories [1, 2, 3], an additional open-loop scheme was developed which significantly improved performance of weaker convolutional codes or codes with higher rates. This scheme which we will term "Space-Time Spreading" or STS can offer significant performance gains over the existing form of open-loop transmit diversity. This scheme is similar in concept and performance to the schemes proposed for the UMTS system (W-CDMA) [4], and by Tarokh et al. [5] for TDMA applications. This paper discusses the performance of this scheme, and shows some of the performance results which were used to win approval of this scheme in the *cdma2000* standardization process. Additionally, we investigate methods of improving the performance of Space-Time Spreading through the use of more than two transmit antennas with and without feedback (*i.e.,* closed loop techniques).

2 An *cdma2000* System Model

For a system with K mobiles receiving signals from a common base station, the transmitted signal on a single antenna can be modeled as:

$$x(t) = \left(\sum_{i=1}^{K} \sqrt{P_i} s_i(t) w_i(t) + \sqrt{P_p} w_0(t) \right) p(t) \tag{1}$$

where P_i is the power transmitted to the i^{th} mobile, $s_i(t)$ and $w_i(t)$ are the data signal and unique Walsh function intended for the i^{th} mobile respectively, P_p is the power of the pilot signal which uses Walsh function 0, and $p(t)$ is the covering code for the base station of interest. Further, the Walsh functions are orthogonal and repeat every symbol time T_s, i.e.

$$\frac{1}{T_s} \int_0^{T_s} w_i(t) w_j(t) dt = \begin{cases} 1 & i = j \\ 0 & i \neq j \end{cases} \tag{2}$$

At the mobile the following signal is received on a single antenna

$$r(t) = \gamma(t)x(t) + n(t) \tag{3}$$

where $\gamma(t)$ is the complex multiplicative distortion caused by the wireless channel and $n(t)$ is thermal noise and all other interference. Mobile i correlates the received signal with the i^{th} Walsh function during the k^{th} symbol interval after uncovering to achieve the decision statistic $z_i[k]$:

$$\begin{aligned} z_i[k] &= \int_{(k-1)T_s}^{kT_s} r(t)p^*(t)w_i(t)dt \\ &= \sqrt{P_i}\gamma(t)s_i[k] + n[k] \end{aligned} \tag{4}$$

where $\gamma(t)$ represents the cumulative effect of the channel $\gamma(t)$ over the kth symbol interval, and $s_i[k]$ is the k^{th} transmitted symbol for the i^{th} mobile.

The transmitted symbol can be recovered by using an estimate of the channel distortion $\hat{\gamma}[k]$ obtainable from the pilot channel, i.e.,

$$\hat{s}_i[k] = f\left(z_i[k]\hat{\gamma}^*[k]\right) \tag{5}$$

where $f(\cdot)$ is an appropriate decision function. Alternately, in a coded system $z_i[k]\hat{\gamma}^*[k]$ may be used directly as a symbol metric. If the channel is a flat, slow Rayleigh faded channel, in the absence of fast, accurate power control, the resulting performance of the link will be rather poor due to the lack of diversity. As a result, it is desirable to have a second antenna at the receiver to allow diversity reception, improving performance considerably. However, mobile handsets do not easily allow a second antenna to be added.

3 Transmit Diversity Methods

One method of achieving diversity performance is to transmit the same signals on multiple carriers. However, this is wasteful of the one resource we cannot afford to waste in mobile communications, namely bandwidth. As an alternative, re-transmitting the same waveform with a chip-level delay, also known as delay diversity, can help performance in some instances, but it can also degrade performance in other instances as it increases the amount of self-interference which degrades the performance of a typical Rake receiver.

Orthogonal transmit diversity (OTD) which is available in *cdma2000* as an option transmits half of the bits via one antenna and half of the bits via a second antenna spaced approximately 10 wavelengths away. The received stream of coded bits will be

$$\{\gamma_{i_0}[0]s_i[0], \gamma_{i_1}[0]s_i[1], \gamma_{i_0}(1)s_i[2], \gamma_{i_1}(1)s_i[3], \ldots\} \tag{6}$$

Using a Viterbi decoder, the link-level performance of the forward link becomes a function of the quality of both channels. Transmissions via channels with slow fading conditions benefit greatly from this method. However, this method offers less performance gain as the speed increases[1] and the code rate decreases. Specifically, with high rate codes, the performance gains are reduced since OTD relies on the decoder to obtain the diversity. This problem is alleviated by a technique termed *Space-Time Spreading* or STS.

4 Space-Time Spreading

Based upon space-time block codes, attributed to Alamouti [6], a signal transmission scheme which utilized the multiple orthogonal code structure already available in the standard was developed for *cdma2000* [2, 3].

[1] The interleaver helps compensate for loss of bits during bursty errors, because at high speeds, the duration of the errors tend to be shorter.

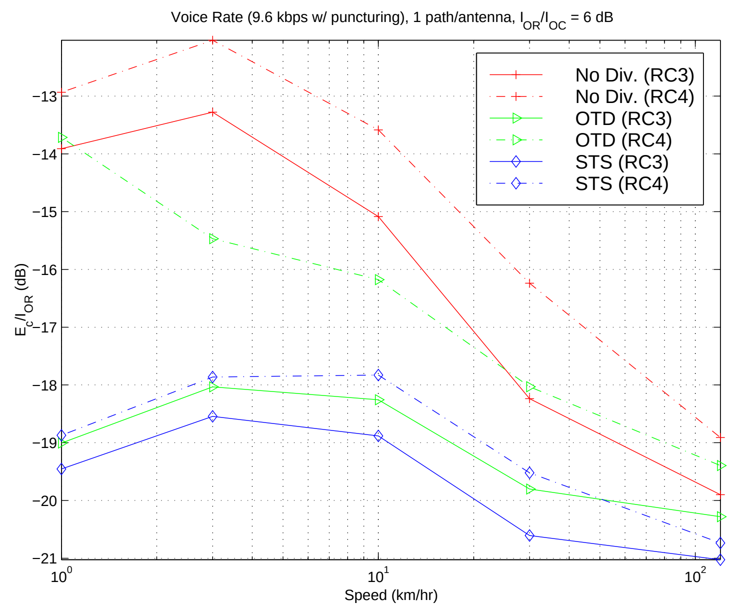

Figure 1: Fundamental Channel (9.6 kbps) Transmit Power, RC3 & RC4, One Path, $I_{OR}/I_{OC} = 6dB$ (See Table 1)

This method also uses two antennas spaced approximately 10 wavelengths apart. On the first antenna we transmit

$$x_1(t) = \left(\sqrt{\frac{P_1}{2}} s_e(t) w_1(t) - \sqrt{\frac{P_2}{2}} s_o^*(t) w_2(t) + \sqrt{P_{p1}} w_{p1}(t)\right) p(t) \tag{7}$$

and on the second antenna we transmit,

$$x_2(t) = \left(\sqrt{\frac{P_1}{2}} s_e^*(t) w_2(t) + \sqrt{\frac{P_2}{2}} s_o(t) w_1(t) + \sqrt{P_{p2}} w_{p2}(t)\right) p(t) \tag{8}$$

where $s_e(t)$ and $s_o(t)$ represent the even and odd streams of bits, respectively. Note that a separate Walsh code is required for each transmit antenna to support a pilot on each antenna. Since the data rate has been reduced by a factor of two by this scheme for each stream of bits, it is possible to use double length Walsh codes and not utilize additional bandwidth or Walsh resources. That is we can convert a single Walsh code into two double length Walsh codes using the repetition pattern

$$\begin{bmatrix} + & + \\ + & - \end{bmatrix} \tag{9}$$

At the receiver, we again uncover and correlate with the two Walsh codes. At the output of the two Walsh correlations we obtain (dropping the dependence on symbol interval)

$$z_1 = \sqrt{\frac{P_1}{2}} \gamma_1 s_e + \sqrt{\frac{P_2}{2}} \gamma_2 s_o + n_1 \tag{10}$$

$$z_2 = \sqrt{\frac{P_1}{2}} \gamma_2 s_e^* - \sqrt{\frac{P_2}{2}} \gamma_1 s_o^* + n_2 \tag{11}$$

where γ_1 and γ_2 are the effects of the complex channel. This obviously introduces interference terms in the decision statistics. However, if we have estimates of the channel distortions $\hat{\gamma}_1$ and $\hat{\gamma}_2$ from pilot signals 1 and 2, we can obtain a signal estimate for the even bits by

$$\begin{aligned} \hat{s}_e &= f\left\{\hat{\gamma}_1^* z_1 + \hat{\gamma}_2 z_2^*\right\} \\ &= f\{\hat{\gamma}_1^* (\sqrt{\frac{P_1}{2}} \gamma_1 s_e + \sqrt{\frac{P_1}{2}} \gamma_2 s_o + n_1) \end{aligned}$$

Table 1: Simulation Parameters for Figure 1

Base Station Antennas:	1 (No Div.) & 2 (STS/OTD)
Bit Rate:	9600bps
Chip Rate:	1.2288Mcps
Coding:	RC3 & RC4 (1/4 conv. & 1/2 conv.)
Frame Duration:	20ms
Frequency:	1.9GHz
Mobile Geometry:	I_{or}/I_{oc}= 6 dB
Pilot E_c/I_{or}:	-7, -13 dB
Max/Min power allocation:	-3 dB/-40 dB
Inner-loop PC rate:	800Hz
PC command error rate:	4%
Inner loop PC step:	±0.5dB
Outer loop PC:	1% FER target
Channel:	Rayleigh fading

$$
\begin{aligned}
&+\hat{\gamma}_2(\sqrt{\frac{P_1}{2}}\gamma_2 s_e^* - \sqrt{\frac{P_1}{2}}\gamma_1 s_o^* + n_2)^*\} \\
&= f\left\{\left(\sqrt{\frac{P_1}{2}}|\gamma_1|^2 + \sqrt{\frac{P_1}{2}}|\gamma_2|^2\right) s_e + \gamma_1^* n_1 + \gamma_2 n_2^*\right\}
\end{aligned} \tag{12}
$$

where we've assumed that the channel estimation is exact, $\hat{\gamma}_1 = \gamma_1$ and $\hat{\gamma}_2 = \gamma_2$. Similarly, we can estimate the data for the odd stream of bits as

$$
\begin{aligned}
\hat{s}_o &= f\{\hat{\gamma}_2^* z_1 - \hat{\gamma}_1 z_2^*\} \\
&= f\left\{\left(\sqrt{\frac{P_1}{2}}|\gamma_1|^2 + \sqrt{\frac{P_1}{2}}|\gamma_2|^2\right) s_o + \gamma_2^* n_1 - \gamma_1 n_2^*\right\}
\end{aligned} \tag{13}
$$

It can be easily shown that this is identical to the decision statistic for two-antenna diversity reception (without the 3dB aperture gain) [7].

The performance of STS was simulated for the *cdma2000* standard using a one path Rayleigh fading channel model for the fundamental channel [8]. The transmit power fractions (*i.e.,* the required fraction of the base station power), E_c/I_{OR}, for full-rate voice using radio configurations RC3 and RC4 [8] were derived from simulation. E_c represents the energy per chip, and I_{OR} represents the total transmit power spectral density. The quantity, I_{OR}/I_{OC} represents the ratio of the transmit power spectral density to the out of cell interference plus any additional thermal noise. It is commonly referred to as "mobile geometry" with low values associated with mobile locations near the edge of the cell and high values associated with mobiles close to the base station. The geometry is directly related to the signal to noise ratio of the decision statistics. The effects of power control, puncturing, and coding using the interleaver specified in the ballot version of the proposal [8] were included. The major simulation parameters are summarized in Table 1.

The simulation results for the fundamental channel are summarized in Figure 1. For both RC3 and RC4, it is clear that STS offers a significant performance advantage. Since RC3 uses 1/4 rate convolutional codes while RC4 uses 1/2 rate codes, RC3 uses length 64 Walsh codes while RC4 uses length 128 Walsh codes. Thus, RC4 is less likely to experience a capacity limit due to a Walsh code limitation and may be preferable in situations where the number of Walsh codes is a concern.

As shown in Figure 1 for a geometry of $I_{OR}/I_{OC} = 6$dB, STS offers up to 5 dB performance improvement over OTD at low speeds, and a minimum of 1.5 dB improvement at high speeds for RC4. For RC3, the gains of STS over OTD are smaller due to the stronger convolutional coding, with STS offering a minimum of 0.5 dB improvement over OTD and up to 1dB at low speeds. STS also provides significant improvement over no diversity achieving gains of 1-5dB. Most importantly transmit diversity helps where the system needs it most, at low speeds. This flattens out the performance curve versus speed and increases capacity.

5 Transmit Diversity with Four Antennas

We have seen that adding an additional antenna at the base station to provide transmit diversity is beneficial for *cdma2000* . The next question we must ask is "Can we improve upon this with additional antennas?" In this section, we discuss three options for extending this diversity scheme to four antennas. Extension to three antennas is also similar with one column of the transmission matrices being ignored. Note that this extension is meant to increase diversity performance, that is we can achieve higher orders of diversity by using more transmit antennas.

To allow four transmit antennas, we first extend the Walsh code for a particular user twice to obtain four Walsh codes with four times the length, where the extension pattern is

$$\begin{bmatrix} + & + & + & + \\ + & - & + & - \\ + & - & - & + \\ + & + & - & - \end{bmatrix} \tag{14}$$

This allows for no code sharing and can be compared to the STS case discussed previously.

To help describe the method of transmission, we define the concept of a transmission matrix. The transmission matrix simply describes the way symbols are transmitted. The rows of the matrix determine the Walsh codes used and the columns determine the antennas on which the symbols are transmitted. For example, we can see that from equations (7) and (8) in STS the transmission matrix is

$$\mathbf{T} = \begin{bmatrix} s_1 & -s_2^* \\ s_2 & s_1^* \end{bmatrix} \tag{15}$$

To obtain a transmission matrix for four transmit antennas, we require an orthogonal matrix with four columns and thus at least four rows [9]. While a 4×4 orthogonal matrix with four complex variables does not exist [5], we find that a 4×4 matrix with three complex variables does exist. One such transmission matrix is

$$\mathbf{T} = \begin{bmatrix} s_1 & -s_2 & 0 & s_3^* \\ s_2^* & s_1^* & -s_3^* & 0 \\ 0 & s_3 & s_1 & s_2^* \\ s_3 & 0 & s_2 & -s_1^* \end{bmatrix} \tag{16}$$

The received vector of Walsh outputs is then $\mathbf{r} = \mathbf{T}\gamma + \mathbf{n}$ where γ is the vector of complex channel distortions. This can be rearranged as $\mathbf{r} = \mathbf{\Gamma s} + \mathbf{n}$. To remove the self-interference we apply the channel matrix $\mathbf{\Gamma}^\dagger$ to $\mathbf{r}$. That is $\hat{\mathbf{s}} = \mathbf{\Gamma}^\dagger \mathbf{r}$ and $\mathbf{\Gamma}^\dagger \mathbf{\Gamma} = \left(|\gamma_1|^2 + |\gamma_2|^2 + |\gamma_3|^2 + |\gamma_4|^2\right) \mathbf{I}$. Thus, we can achieve four-fold diversity. However, in order to achieve this, we must reduce the data rate to 3/4 the original rate. This can be seen by noticing that while we use four codes (i.e., the rows of $\mathbf{T}$) we only transmit three symbols.

A second option for using four transmit antennas without reducing the data rate, is to use the transmission matrix

$$\mathbf{T} = \begin{bmatrix} s_1 & -s_2 & 0 & 0 \\ 0 & 0 & s_3 & -s_4^* \\ s_2^* & s_1^* & 0 & 0 \\ 0 & 0 & s_4^* & s_3^* \end{bmatrix} \tag{17}$$

Which guarantees orthogonality, but only achieves two-fold diversity before the decoder. However, if the interleaving is done correctly, we see that going into the decoder the metrics are:

$$s_1\left(|\gamma_1|^2+|\gamma_2|^2\right),\quad s_2\left(|\gamma_3|^2+|\gamma_4|^2\right),\quad s_3\left(|\gamma_1|^2+|\gamma_2|^2\right),\quad s_4\left(|\gamma_3|^2+|\gamma_4|^2\right),\quad \ldots \tag{18}$$

Thus, while two-fold diversity is achieved prior to decoding, the Viterbi decoder can see up to four-fold diversity in the path metrics. Thus, while we rely on the decoder to achieve the diversity gain from 2 to 4, we do not lose data rate.

The last option is similar to Option 1 and uses the orthogonal design from [5]. The transmission matrix is

$$\mathbf{T}=\begin{bmatrix} s_1 & s_2 & \frac{s_3}{\sqrt{2}} & \frac{s_3}{\sqrt{2}} \\ -s_2^* & s_1^* & \frac{s_3}{\sqrt{2}} & -\frac{s_3}{\sqrt{2}} \\ \frac{s_3^*}{\sqrt{2}} & \frac{s_3^*}{\sqrt{2}} & \frac{-s_1-s_1^*+s_2-s_2^*}{2} & \frac{-s_2-s_2^*+s_1-s_1^*}{2} \\ \frac{s_3^*}{\sqrt{2}} & \frac{-s_3^*}{\sqrt{2}} & \frac{s_1-s_1^*+s_2-s_2^*}{2} & \frac{-s_2+s_2^*-s_1-s_1^*}{2} \end{bmatrix} \tag{19}$$

This option also achieves four-fold diversity prior to decoding, but also suffers from a 25% loss in data rate. The main difference between this option and Option 1, is that this allows all four codes to be used on all four antennas.

All three of these options essentially provide four-fold diversity performance although option 2 will suffer some degradation when puncturing is included on higher rate codes just as in OTD. However, we should note two things. First consider Figure 2 (a). This figure plots the theoretical performance of several options for four transmit antennas. Included are the theoretical performance of four branch diversity, two-branch diversity with two-branch aperture gain, four-branch aperture, and four-branch diversity and aperture gain. The theoretical performance is well known to be

$$P_e=\left[\frac{1}{2}(1-\mu)\right]^L\sum_{k=0}^{L-1}\binom{L-1+k}{k}\left[\frac{1}{2}(1+\mu)\right]^k \tag{20}$$

where L is the degree of diversity and $\mu=\sqrt{\frac{N\gamma_c}{1+N\gamma_c}}$, γ_c is the SNR per branch without any aperture gain, and N is the amount of aperture gain being achieved. Also shown in Figure 2 (a) for comparison purposes are no diversity and two-branch diversity. We can see that going from two-branch to four-branch diversity does not provide huge gains particularly at high BER's. Additionally, we see that two-branch diversity with two-branch aperture gain provides better performance in the region of 1% BER than four-branch diversity. As a second note, consider Figure 1. Notice that at high speeds the advantage of diversity is diminished compared to the gains at low speeds. This is due to the diversity obtained in the decoder due to fast fading and interleaving. Additionally, the benefits of additional diversity in frequency selective fading are much less than in flat fading. This suggests that increased diversity performance is not necessarily the best option. However, in order to implement the methods shown in Figure 2 (a) which are more than pure diversity, either feedback or uplink-based estimation is needed. We will discuss this next.

6 Methods which Require Feedback

The preceding discussion of higher order diversity was based on the premise that we wished to improve performance without requiring mobile feedback or uplink estimation. In this section we discuss the performance improvements when mobile feedback is allowed. There are essentially three options for performance improvement with four antennas when feedback is allowed. They are

- Four element fully adaptive transmit diversity
- Four element steered beam

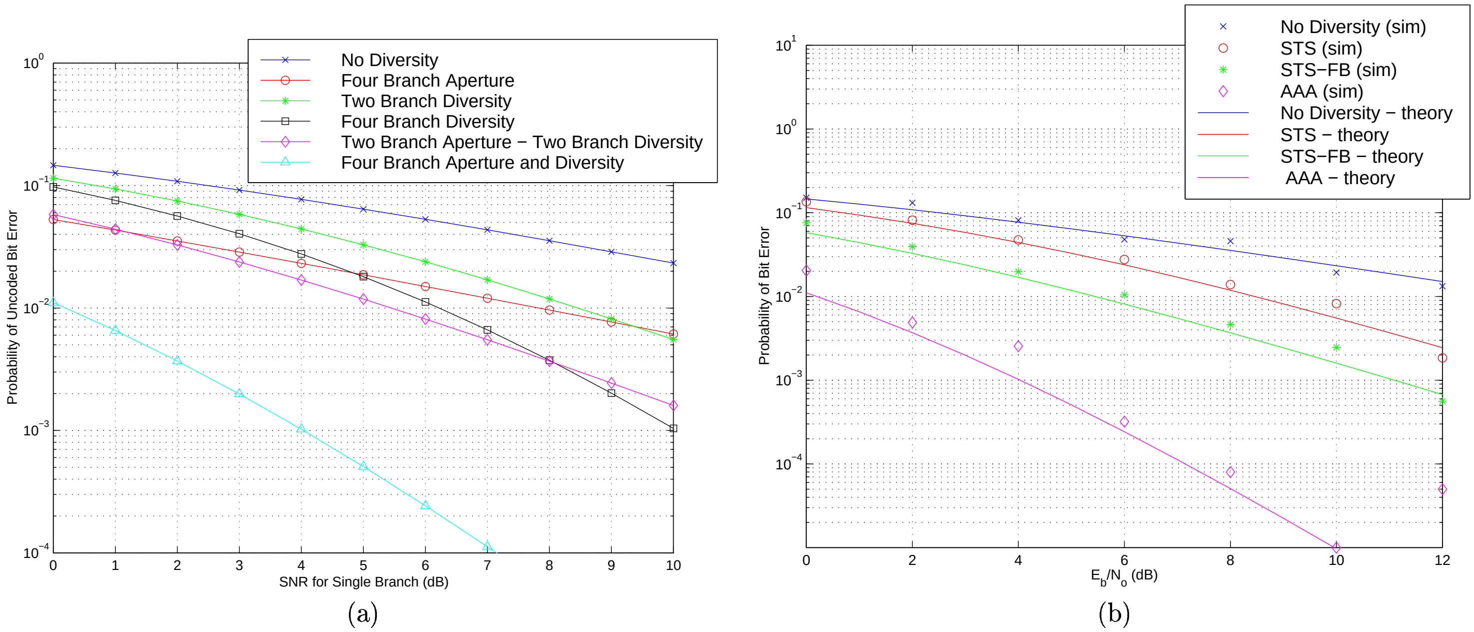

Figure 2: Theoretical Performance of Several Multiple Antenna Options (a) and Simulated Uncoded BER Performance of Feedback Schemes with Four Antennas with Perfect Feedback (b)[simulations assume 1.25ms pilot filter, pilot $E_s/N_o = 12E_b/N_o$]

- Combined transmit diversity and steering (Steered STS)

The first option has the largest potential gains and constitutes using a four element array with large element spacing. Relying on mobile feedback the transmitter would dynamically adjust the phase and gains of each antenna to ideally coherently combine all four transmit antennas. This would ideally provide both diversity and aperture gain but obviously requires fast accurate mobile feedback. Such feedback would be limited by inherent delays and would need to be added to the current standard. The second option does not necessarily require a standards change but does not achieve any diversity gain and might be susceptible to deep fades. A steered beam solution achieves approximately a 6dB aperture gain, although the gains could be less in a rich scattering environment. This would require either mobile feedback (*i.e.,* a standards change) or uplink direction-of-arrival (DOA) estimation. The third option also does not necessarily require feedback and thus would not require a standards change. It provides two branch diversity as well as two times aperture gain.

To provide some feel for the performance of the above options, simulations were run without power control or coding and a 1.25ms average pilot filter providing a pilot $E_s/N_o = 12E_b/N_o$. The DOA for the S-STS case was estimated over one frame at the mobile and fed back to the base. Figure 2 (b) plots the results of these simulations for no diversity, STS, combined STS and beam-steering (called Steered STS or STS-Feedback), and a fully adaptive array with four diversity branches. The STS and Steered STS results assume two independently faded Rayleigh channels, while the fully adaptive array assumes four independently faded Rayleigh channels. Feedback is perfect, *i.e.,* full precision, no delay and no feedback error. The results match well with theory. STS provides two-branch diversity gain over the no diversity case (4-5dB at 1% BER). Steered STS provides a 3dB gain over STS and the fully adaptive array achieves both a 3dB aperture improvement and a diversity improvement over Steered STS.

The preceding simulations assumed that feedback was perfect. In Figure 3 (a) we remove this constraint and allow the feedback to be quantized by 4 bits and bit errors to occur in the feedback process. Further we examine the performance as the feedback error rate increases from zero to 20%. Of course the performance of no diversity and STS remains unchanged. Surprisingly, the performance of S-STS also remains unchanged

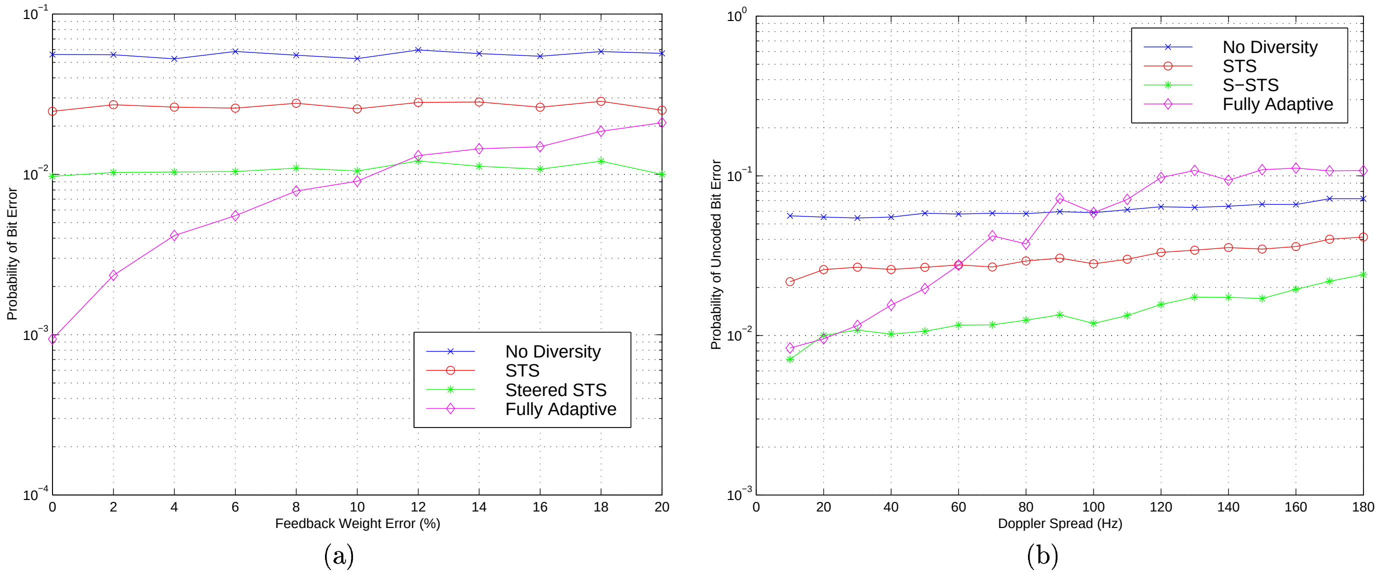

Figure 3: Simulated Uncoded BER Performance of Feedback Schemes with Four Antennas with 4 Bit Weight Quantization and (a) Zero Feedback Delay and (b) 2 PCG Feedback Delay (4% Weight Error) - [all assume 1.25ms pilot filter, pilot $E_s/N_o = 12E_b/N_o$]

over the error rate of interest. However, we see that the performance of the fully adaptive array degrades significantly as the feedback error increases. In fact the fully adaptive approach loses all of its advantage over S-STS if the feedback error is 10% or higher. Note that in order to keep the delay to a reasonable level the feedback will have to be uncoded (*i.e.,* we cannot wait for the Viterbi decoding of the entire frame to obtain feedback bits). Thus, the error rate on the bits could be fairly high.

The reason that S-STS is relatively immune to the feedback error is that the range of phases required to steer antennas over $\pm 60^o$ is fairly small and thus any error perturbs the beam only a small amount. On the other hand with the fully adaptive approach the antennas require a much larger range of phase adjustments. This makes the range of phase error larger as well as the resulting degradation. Note that the fully adaptive approach also requires three times the feedback.

A second major degradation in a feedback system is due to delay. In *cdma2000* there is a minimum 2.5ms delay (equivalent to two power control groups). This will obviously be a problem for the fully adaptive approach as the fading rate increases. Figure 3 (b) presents the performance of the schemes with 4% feedback error as the Doppler rate increases from 10Hz to 180Hz. We can see that when feedback error and delay are considered, the fully adaptive approach suffers dramatically. Note that the performance of the fully adaptive approach degrades beyond that of the no diversity approach due to the feedback errors combined with improper weighting. The receiver relies on correct knowledge of the feedback bits in order to correctly phase the pilot. Thus, feedback error can cripple performance. Note that all suffer degradation at high speeds due to the pilot filter length.

7 Steered Space-Time Spreading

Results from the previous section lead us to consider the use of STS with beam-steering, *i.e.,* combined diversity and aperture gain. In this section we describe this idea more fully. Consider a linear antenna arrangement as shown in Figure 4. The base station transmits on M antennas divided into two groups. Group

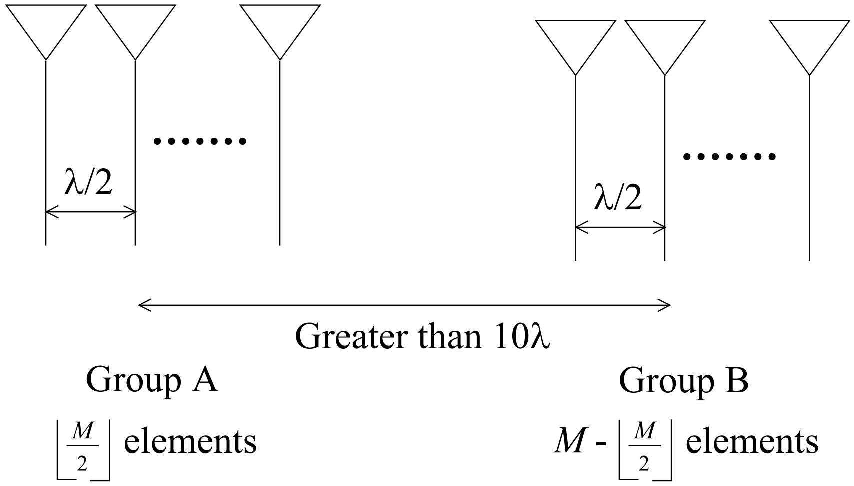

Figure 4: Antenna Configuration for Employing STS with Two or More Antennas

A has $\lfloor \frac{M}{2} \rfloor$ antennas[2] with inter-element spacing of approximately $\frac{\lambda}{2}$ where λ is the carrier wavelength.[3] Group B is separated from Group A by a distance large enough to insure that the two groups experience uncorrelated fading and contains $M - \lfloor \frac{M}{2} \rfloor$ elements.

On antennas within Group A the transmitted signal (ignoring other users) is

$$x_i(t) = \left\{ v_i \sqrt{P_i} \left(s_e(t) w_1(t) - s_o^*(t) w_2(t) \right) + \sqrt{P_p} w_{p_i}(t) \right\} p(t) \tag{21}$$

where i represents the antenna index which is $1 \leq i \leq \lfloor \frac{M}{2} \rfloor$ for Group A, P_i represents the transmit power for the data from antenna i, s_e and s_o are the data stream divided into even and odd streams respectively, P_p is the power in the dedicated pilot on each antenna, w_{p_i} is the Walsh code used for the pilot on that antenna[4], v_i is a complex weight to be described later, $w_1(t)$ and $w_2(t)$ are extended Walsh codes as described earlier, and $p(t)$ is a pseudo-random sector-specific covering code.

On antennas within group B the transmitted signal (ignoring other users) is

$$x_i(t) = \sqrt{P_i} \left\{ v_i \left(s_o(t) w_1(t) + s_e^*(t) w_2(t) \right) + \sqrt{P_p} w_{p_i}(t) \right\} p(t) \tag{22}$$

where $\lfloor \frac{M}{2} \rfloor + 1 \leq i \leq M$. Note that for $M = 2$ we have Space-Time Spreading [3]. At the receiver we have a single signal

$$r(t) = \sum_{i=1}^{\lfloor \frac{M}{2} \rfloor} x_i(t) \gamma_1(t) e^{j\theta_i} + \sum_{i=\lfloor \frac{M}{2} \rfloor + 1}^{M} x_i(t) \gamma_2(t) e^{j\theta_i} + n(t) \tag{23}$$

where $\gamma_1(t)$ and $\gamma_2(t)$ are the time-varying multiplicative distortion due to Rayleigh fading seen from groups A and B respectively, $\theta_i = \frac{2\pi d_i}{\lambda} \cos(\phi)$, d_i is the distance of the ith antenna from an arbitrary reference, ϕ is the angle formed between a line drawn from the base to the mobile and the array baseline, and $n(t)$ is temporally and spatially white complex Gaussian noise. By correlating the received signal (after removing the long code) with $w_1(t)$ and $w_2(t)$ and assuming that the channel is static over the integration period, we obtain the following correlation outputs:

[2] $\lfloor x \rfloor$ is defined as the largest integer less than or equal to x.

[3] Note that the exact element spacing is not crucial. However, it must be small enough so that the signals transmitted from all elements experience highly correlated fading. Additionally, a spacing of much greater than $\frac{\lambda}{2}$ will introduce grating lobes which is generally undesirable.

[4] As we will discuss later, a pilot per antenna is only necessary if mobile feedback is being used to steer the array. If uplink estimation is used to steer the array, only one pilot per group is necessary.

$$W_1 = \sum_{i=1}^{\lfloor \frac{M}{2} \rfloor} \sqrt{P_i} v_i e^{j\theta_i} \gamma_1 s_e + \sum_{i=\lfloor \frac{M}{2} \rfloor+1}^{M} \sqrt{P_i} v_i e^{j\theta_i} \gamma_2 s_o^* + N \tag{24}$$

$$W_2 = \sum_{i=1}^{\lfloor \frac{M}{2} \rfloor} -\sqrt{P_i} v_i e^{j\theta_i} \gamma_1 s_o^* + \sum_{i=\lfloor \frac{M}{2} \rfloor+1}^{M} \sqrt{P_i} v_i e^{j\theta_i} \gamma_2 s_e + N' \tag{25}$$

$$\tag{26}$$

Now, using the pilots we can obtain estimates for γ_1 and γ_2. We can then use the estimates to create decision statistics for the even and odd streams as

$$\hat{s_e} = f\left\{\hat{\gamma_1}^* W_1 + \hat{\gamma_2} W_2^*\right\} \tag{27}$$

$$\hat{s_o} = f\left\{\hat{\gamma_2}^* W_1 - \hat{\gamma_1} W_2^*\right\} \tag{28}$$

$$\tag{29}$$

where $f\{\cdot\}$ is an appropriate decision function. Expanding the first equation and assuming perfect channel knowledge for simplicity's sake results in

$$\hat{s_e} = f\left\{ s_e \left(\sum_{i=1}^{\lfloor \frac{M}{2} \rfloor} \sqrt{P_i} v_i e^{j\theta_i} |\gamma_1|^2 + \sum_{i=\lfloor \frac{M}{2} \rfloor+1}^{M} \sqrt{P_i} v_i e^{j\theta_i} |\gamma_2|^2 \right) + \ldots \right.$$
$$\left. \left(\sum_{i=1}^{\lfloor \frac{M}{2} \rfloor} \sqrt{P_i} v_i e^{j\theta_i} - \sum_{i=\lfloor \frac{M}{2} \rfloor+1}^{M} \sqrt{P_i} v_i e^{j\theta_i} \right) \gamma_1^* \gamma_2 s_o + \gamma_1^* N + \gamma_2 N'^* \right\} \tag{30}$$

From this equation we can see two things: (1) we wish $v_i = e^{-j\theta_i}$ and (2) we must set the transmit powers such that $\sum_{i=1}^{\lfloor \frac{M}{2} \rfloor} \sqrt{P_i} = \sum_{i=\lfloor \frac{M}{2} \rfloor+1}^{M} \sqrt{P_i}$. For an even number of antennas, the second condition is satisfied by giving all antennas equal power. The first condition however, must be accomplished by either (a) using information from the uplink to estimate θ_i or (b) using mobile feedback. We will discuss the options for mobile feedback in a moment.

Assuming that the two above conditions are met, the decision statistic for s_e is

$$z_{s_e} = s_e \left(\sum_{i=1}^{\lfloor \frac{M}{2} \rfloor} \sqrt{P_i} |\gamma_1|^2 + \sum_{i=\lfloor \frac{M}{2} \rfloor+1}^{M} \sqrt{P_i} |\gamma_2|^2 \right) + \gamma_1^* N + \gamma_2 N'^* \tag{31}$$

$$\tag{32}$$

Defining the SNR as $\frac{\mathrm{E}\{z_{s_e}\}^2}{\mathrm{var}\{z_{s_e}\}}$ and for the moment assuming that M is even (i.e., $P_i = P/M$), we can see that the SNR is a Chi-Square random variable with 4 degrees of freedom (i.e., two-fold diversity) and an expected value of

$$\overline{\mathrm{SNR}} = \frac{\frac{M^2}{4} \mathrm{E}\left(|\gamma_1|^2 + |\gamma_2|^2 \right)^2 \frac{P}{M}}{\sigma_n^2 \mathrm{E}\left(|\gamma_1|^2 + |\gamma_2|^2 \right)}$$
$$= \frac{P}{\sigma^2} \frac{M}{2} \tag{33}$$

where we have assumed that $\mathrm{E}\left\{|\gamma_i|^2\right\}$=1. Thus, we have an improvement of $\frac{M}{2}$ in SNR when compared to the case of standard STS which sees no aperture gain but merely a diversity gain.

7.1 Calculating Weights

7.1.1 Using an Uplink Array

A key to the scheme is the set of weights $\{v_i\}$. To maximize SNR we must set $v_i = e^{-j\theta_i}$. One method of setting the weights is to attempt to estimate θ_i from uplink information. In the presence of this uplink array we can estimate ϕ by measuring θ_i^{up} and using the relation $\theta_i^{up} = \frac{2\pi d_i}{\lambda^{up}}\cos(\phi)$. After estimating ϕ the weights are set to

$$\begin{aligned} v_i &= e^{-j(\theta_i^{dn})} \\ &= e^{-j\frac{2\pi d_i}{\lambda^{dn}}\cos(\hat{\phi})} \end{aligned} \tag{34}$$

However, this assumes that the distance between elements is known, the elements are phase matched, and there is symmetry between the uplink angle-of-arrival and downlink angle-of-arrival. All of these are either reasonable to assume or could be obtained through calibration. In this case, if the main pilot is put on the first antenna of each group and the other elements are phased with respect to it, the transmit signals per group will arrive at the mobile in phase and thus only one pilot per group is necessary. The mobile station in such a system would not need to know that beam-steering was being used.

7.1.2 Feedback Options

A second means of calculating the set of weights $\{v_i\}$ is to rely on mobile feedback. Since the weights depend ultimately on the angle-of-arrival, ϕ, they must only be updated at the rate at which ϕ changes which is likely very slow compared to channel fading rate. There are several possible methods of employing feedback.

The most straightforward method of feedback is to transmit a dedicated pilot on each antenna and feed back the phase of the received pilots. One pilot per group could be used as a reference and the phase of the other pilots with respect to the reference pilots are then fed back. This requires $M-2$ phase values be fed back per update. For q bit quantization and F Hz feedback rate, this method requires $qF(M-2)$bps feedback. This method makes no assumptions about the array spacing and is thus robust to imperfect knowledge of the inter-element spacing.

Another method which requires less feedback is to feedback a single value for the entire array. If the inter-element spacing within each group of elements is the same, the elements should differ by a constant phase $\Delta\theta = \frac{2\pi d}{\lambda}\cos(\phi)$. While the method is simpler and requires less feedback, it is more sensitive to non-ideal element spacing. The feedback rate would be qFbps. For small array sizes (eg., M=4) this may not be a significant savings.

7.2 Performance

To investigate the detailed performance of Steered STS in *cdma2000* , simulations were run using the *cdma2000* standard. The simulation assumptions are given in Table 1 and with the exception that S-STS uses two groups of two antennas, /ie $M = 4$, only RC3 is simulated and $I_{or}/I_{oc} = 0dB$. Note that the geometry assumed is much worse (*i.e.,* lower E_b/N_o) than in Figure 1 thus the higher required power fraction than the previous case. The array assumed had four antennas (*i.e.,* two pairs).

We can see that Steered-STS provides significant performance improvement over both STS (approximately 3dB) and no diversity. An important point about S-STS is that even at high mobile speeds, the scheme still achieves significant gains over the baseline, which is not true of transmit diversity in general. That is the gains are not speed dependent. However, we see that the performance of S-STS is fairly flat with respect to speed.

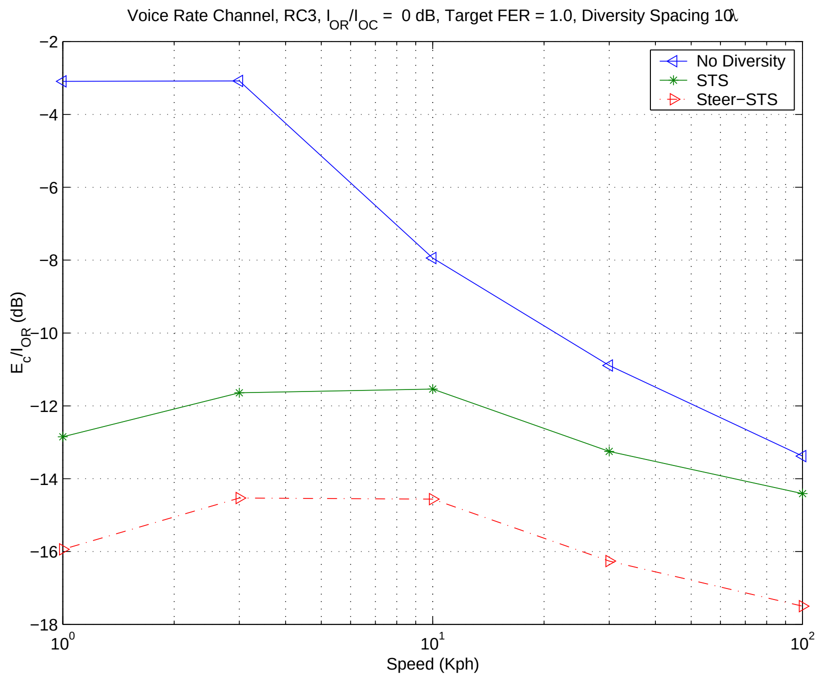

Figure 5: Simulated Performance of No Diversity, STS and Steered-STS in *cdma2000* (parameters listed in Table 1 except: S-STS uses 4 antennas, $I_{or}/I_{oc} = 0dB$, RC3 only)

8 Conclusion

In this paper we have investigated several possible extensions of transmit diversity to four antennas for the Third Generation standard *cdma2000* . It is clear that there are potential benefits from increasing the number of antennas at the mobile station. These benefits to the *cdma2000* downlink will be significant as data becomes more predominant.

Acknowledgments

The authors wish to express their appreciation to their colleagues at Lucent-Bell Labs who have contributed to the work in this paper including Roger Benning, Steve Nicoloso, Nallepilli Ramesh, Steve Allpress, Constantinos Papadias, Bert Hochwald and Tom Marzetta.

References

[1] C. Papadias, B. Hochwald, T. Marzetta, R.M. Buehrer, and R. Soni , "Space-time spreading for CDMA systems," *Stanford Sixth Workshop on Smart Antennas for Mobile Communications*, July 22-23 1999.

[2] Lucent Technologies, "Performance of Space Time Spreading (STS) for IS-2000," *Contribution 3GPP2-C30-19990914-013*, September 1999.

[3] R. Soni, R. Buehrer, and J.-A. Tsai, "Open-loop transmit diversity methods in IS-2000 systems," in *Proceedings of the Asilomar Conference on Signals, Systems and Computers*, October 1999.

[4] Texas Instruments, "Space time block coding improvements," *ETSI - UMTS Contribution*, December 1998.

[5] V. Tarokh, H. Jafarkhani, and A. Calderbank, "Space-time block codes from orthogonal designs," *IEEE Transactions on Information Theory*, vol. 45, pp. 1456–1467, July 1999.

[6] S. M. Alamouti, "A simple transmitter diversity scheme for wireless communications," *IEEE J. Select. Areas Commun.*, vol. 16, October 1998.

[7] J. Proakis, *Digital Communications.* New York, NY: McGraw-Hill, third ed., 1995.

[8] TR45.5, *Physical Layer Standard for cdma2000 Spread Spectrum Systems.* TIA/EIA/IS-2000.2, 1999. (Ballot Version).

[9] B. Hochwald and T. Marzetta, "Unitary space-time modulation for multiple-antenna communications in rayleigh flat fading," *IEEE Transactions on Information Theory*, vol. 46, pp. 543–564, March 2000.