Macroscopic Space-Time Coding: Motivation, Performance Criteria, and a Class of Orthogonal Designs

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Abstract — **Macroscopic space-time coding, where each single-antenna base station acts as one of the antennas in a space-time coding scheme, is an attractive method to improve the performance of wireless systems, particularly for applications requiring at least partial broadcast. However, because the users to whom the signals are broadcast can be spread across a wide geographical region, the time of arrival of a signal from one base station relative to that of a signal arriving from another base station varies greatly, thus complicating system design. In this paper, this novel macroscopic space-time coding problem is clearly motivated, and then a performance criterion, optimal receiver, and family of code designs are presented. A matched filter bound analysis with supporting numerical and simulation results demonstrates the improvement in bit error rate of the proposed code designs. Finally, the matched filter bound analysis is extended to demonstrate significant improvements in coverage for such a system over currently employed systems and standard space-time coding approaches applied across the same set of base stations.**

I. INTRODUCTION

Space-time coding offers a method to provide diversity against multipath fading on a downlink transmission from a multiple-antenna base station to a simple single-antenna mobile [1, 2, 3]. In standard space-time coding, the antennas at the transmitter are located in such close geographical proximity that the propagation delay from one transmit antenna to the receive antenna is similar to the propagation delay from any other transmit antenna to the receive antenna, in the sense that the difference in these propagation delays is negligible compared to the symbol interval. In contrast, interest here is on when the difference in such propagation delays is on the order of the symbol interval, which we will term the "macroscopic space-time coding" problem.

Consider public safety radio systems (e.g. those used by police officers), where the users are broken into "talk groups." In each talk group, one person speaks at a time and the rest of the group listens. Whereas this results in only a small amount of uplink traffic, there is significant identical downlink traffic that can span many cells in the commonly-employed cellular architecture. In this case, the most efficient resource allocation method (in terms of supporting the most talk groups) is often to broadcast on a single frequency across the network to a given talk group, hence not requiring frequency reuse for that talk group and yielding up to a sevenfold increase in talk group capacity if the members of each talk group are uniformly spread across the network.

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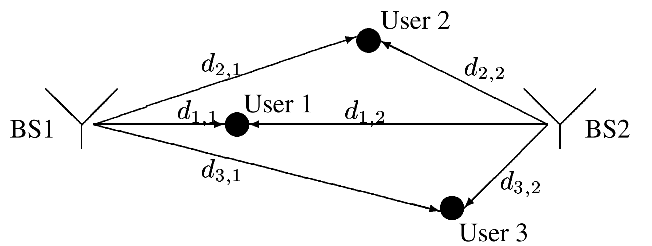

Figure 1: Two base stations broadcasting to three users in a given "talk group" on a single frequency channel through broadcast. Note, in particular, how the relative propagation delay of the transmissions from the two base stations varies widely among users.

In Figure 1, an example of a small region of such a system is shown. Because of the resource allocation arguments above, systems often employ "simulcasting" , where the same signal is sent from each of the base stations. However, in the same way that a standard space-time does not achieve diversity if the same thing is sent off each antenna at the same time, simulcasting does not provide diversity in the important center of the region on frequency-nonselective Rayleigh multipath fading channels.

The logical first solution to this problem is to employ one of the standard space-time solutions *across* the multiple base stations. However, these have severe limitations. In public safety radio system applications, the distance between base stations can be on the order of 20 to 40 miles, which can cause the signals from different base stations to experience propagation delays that differ by more than a symbol interval if the user is not in the center of the cell (e.g. User 1). Although one signal will be significantly weaker than the other in general due to the disparate path-loss, this intersymbol interference is a major problem in simulcasting systems, and it will be clearly demonstrated in Section V for a system experiencing path-loss, shadowing, and multipath fading that this is indeed a critical factor. Timing the transmissions of the two transmitted signals so that the corresponding symbols arrive at User 1 at the correct time immediately comes to mind, but this is not attractive because the resource allocation scheme mandates broadcast - and thus we desire to support geographically disparate users on the same radio channel. This motivates the design of space-time codes that are robust to differences in the propagation delays for the signals from different transmit antennas; however, there has been almost no research work (only the analysis in [4]) on the analysis and/or design of a link where *multiple antennas that are not co-located are transmitting to a single receiver* in a space-time fashion.

II. System Model and Optimal Receiver

Per Figure 1, let $d_{i,j}, i = 1, \ldots, K, j = 1, \ldots, n_T$ be the distance of the i^{th} user from the j^{th} base station, and let $\tau_{i,j}$ be the proportional propagation delay. Using a single system channel (i.e. broadcast mode), it is desirable to design the signal transmitted from the antenna at each of the base stations and the receivers employed at the mobile units such that $\max_{\underline{d} \in \mathcal{D}} P_e(\underline{d})$ is minimized, where P_e is the metric of interest (say, frame error rate or bit error rate) of the system given $\underline{d}_i = (d_{i,1}, d_{i,2}, \ldots, d_{i,n_T})$, and $\mathcal{D}$ is the set of all possible distance vectors for a given system configuration.

Let $u_j(t)$ be the complex baseband representation of the signal transmitted from the j^{th} base station of the form

$$u_j(t) = \sum_{l=-\infty}^{\infty} \sum_{n=0}^{N-1} C_{j,n}^{(l)} p(t - (lN + n)T_s)$$

where N is the codeword length, $p(\cdot)$ is the pulse shaping function, T_s is the symbol period, and $C_{j,n}^{(l)}$ is the n^{th} symbol of the l^{th} codeword sent from base station j. In particular, the symbol $C_{j,n}^{(l)}$ will be the n^{th} element of the j^{th} row of the $n_T \times N$ macroscopic space-time codeword matrix $C^{(l)}$ selected as the l^{th} transmitted codeword from the collection $\mathcal{C}$ of macroscopic space-time codeword matrices. Then, assuming a frequency-nonselective Rayleigh fading channel from each base station to the mobile, the complex baseband representation of the received signal for the i^{th} user is given by

$$r_i(t) = \sum_{j=1}^{n_T} \sqrt{E_{i,j}} X_{i,j} u_j(t - \tau_{i,j}) + n(t) \quad (1)$$

$$= \sum_{j=1}^{n_T} \sqrt{E_{i,j}} X_{i,j} \sum_{l=-\infty}^{\infty} \sum_{n=0}^{N-1} C_{j,n}^{(l)} p(t - (lN + n)T_s - \tau_{i,j}) + n(t)$$

where $E_{i,j}$ is the average received signal energy at the i^{th} receiver from the j^{th} base station, which includes the effects of path-loss and shadowing, $X_{i,j}$ is a zero-mean complex Gaussian random variable representing the multipath fading from the j^{th} base station to the i^{th} mobile, and $n(t)$ is a stationary Gaussian random process with power spectral density $\frac{N_0}{2}$. It is important to note that (1) looks similar to a standard intersymbol interference (ISI) channel model in a single-antenna system, and that similarity will suggest avenues of research; however, it is also important to note that, since the transmitted signal from each of the two base stations is different, the received signal cannot be written as the convolution of a single signal and a linear filter, which makes system design challenging.

As seen from (1), the delay between the arrival of signals from different base stations causes both intersymbol and inter-codeword interference, thus greatly complicating receiver design. However, receiver analogs from multi-user detection and equalization can be applied, with the receiver type depending strongly on the type of pulse shaping assumed. If the pulse shape $p(t)$ only has support on $[0, T_s]$ (e.g. rectangular pulse shapes), and the maximum differential delay between two arriving signals is assumed to be T_s, the ideas found in [5] can be extended to this scenario. As shown in [6], the maximum likelihood sequence estimator (MLSE) receiver consists of a bank of n_T matched filters followed by a Viterbi algorithm operating on a 2^{n_T-1}-state trellis. In contrast to [5], each state transition corresponds to an entire codeword (as opposed to a single symbol), and thus the metric on each branch of the trellis consists of a correlation of the received signal with the "expected" codeword for that branch in addition to a periodic "fixed" component. A simplified version of this receiver that corresponds to differential delays in $\{-T_s, 0, T_s\}$ will be given in the next section to aid in code design and will help elucidate the key differences from [5].

For pulse shapes that are band-limited (hence, not time-limited) and for which the differential delays do *not* fall on multiples of the symbol interval, receiver design is greatly complicated. Receiver design and performance for this important case will be considered in Section IV.

III. Code Design

In this section, the problem of macroscopic space-time code design is considered. For ease of exposition, it will be assumed that there are only two base stations and that the difference in the propagation delays from the two transmitters to the receiver is less than a single symbol (i.e. in the interval $[-T_s, T_s]$). The interested reader is referred to [6] for generalizations to higher numbers of base stations and a larger interval for the possible relative signal delays.

Furthermore, in the code *design* process, two assumptions are made:

1. The average received energy from the two base stations is identical.
2. The relative propagation delay $\tau_{i,1} - \tau_{i,2}$ is a multiple of T_s, implying $\tau_{i,1} - \tau_{i,2} \in \{-T_s, 0, T_s\}$.

Note that both of these assumptions are simply to aid the design process - the derived codes will be tested on the general (1) below, so the effects of relative energy and any relative propagation delay in $[-T_s, T_s]$ *will* be considered for the performance characterization.

It is instructive to consider what the optimal receiver would look like for standard space-time coding solutions. Assume that the system is employing a pulse shape $p(t)$ that satisfies the zero ISI criterion. First, suppose that a simulcasting system is employed; that is, the system transmits the same signal from the two antennas. Then, the problem reduces to the standard ISI problem [7], and, under the assumptions of the previous paragraph with k known at the receiver, a symbol-spaced matched filter followed by the standard forward Viterbi algorithm yields the optimal receiver [8]. In fact, standard equalization results of all types are readily applied to the simulcast problem. However, per Section I, such a system does not provide diversity to the geographically-disadvantaged users directly between the two base stations and thus is not desirable (as will be Section V).

Next, consider the case when the transmissions from the two antennas are different; in particular, suppose an Alamouti scheme [2] is employed. To simplify the discussion (and, more importantly, the resulting diagrams), it is assumed that binary phase-shifting keying (BPSK) is employed, but the results generalize easily[6]. Under this assumption, the Alamouti scheme for transmitting two information bits $b_0 \in \{0, 1\}$ and $b_1 \in \{0, 1\}$ is defined by the transmitted codeword matrix:

$$C_{AL} = \begin{bmatrix} s_0 & s_1 \\ -s_1 & s_0 \end{bmatrix},$$

where the first row is transmitted from the first base station, the second row is transmitted from the second base station, and $s_i = (-1)^{b_i}, i = 0, 1$. Under a relative propagation delay of a single symbol, it can be shown that the optimal receiver consists of a matched filter followed by a symbol-spaced sampler and a forward Viterbi algorithm with modified path metrics. To observe such, consider the trellis shown in Figure 2 for representing the *effective* transmitted signal, which is defined as the transmitted signal *after* the relative delay is applied but *before* the fading and noise are added, in a system where

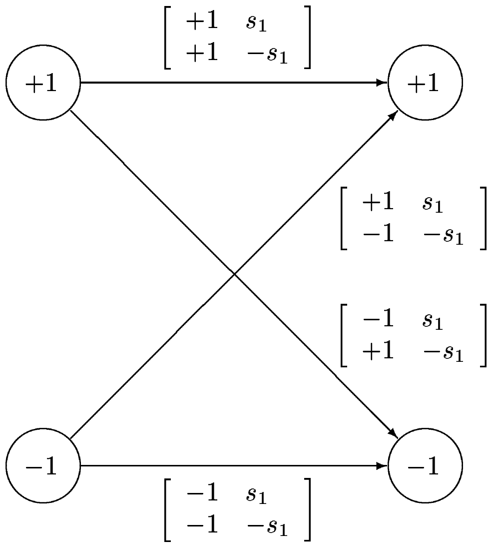

Figure 2: Trellis diagram for the effective transmitted signal in the Alamouti scheme when the signal of the second transmitter is delayed one symbol relative to that of the first transmitter, assuming BPSK modulation (the generalization to higher order modulation simply adds more states to the trellis diagrams [6]). The loss in diversity of the Alamouti scheme for this delay is immediately observed - any pair of parallel paths leads to a rank one result.

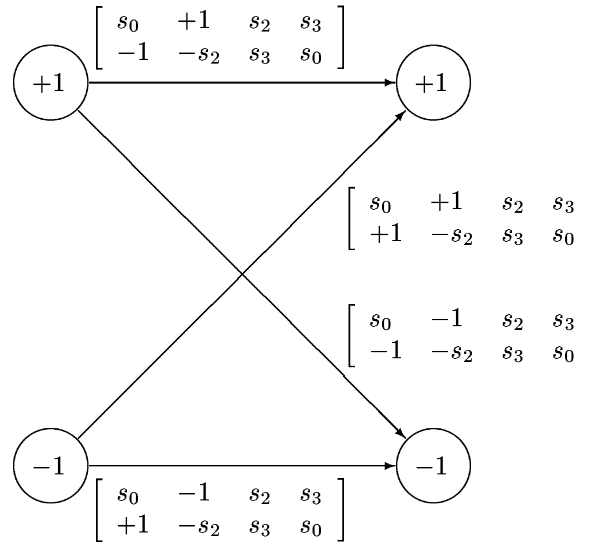

Figure 3: Trellis diagram for the effective transmitted signal in the proposed scheme, when the signal of the second transmitter is delayed one symbol relative to that of the first transmitter, assuming BPSK modulation (the generalization to higher order modulation simply adds more states to the trellis diagrams [6]). The proposed scheme retains rank two difference matrices for any two possible paths through the trellis.

the signal from base station two is delayed one symbol relative to the signal from base station one. As in standard coded modulation and equalization, any effective transmitted sequence is represented by a path through the trellis, and any path through the trellis has a corresponding effective transmitted sequence. Thus, the latter portion of the optimal receiver is easily realized as a two-state forward Viterbi algorithm, but, more importantly for the transmitter design results below, performance results can be derived for the system by considering paths that diverge at a given state and later remerge in the trellis.

Per Figure 2, the loss in rank of a standard Alamouti scheme is easily observed when the relative delay is equal to one symbol period. Not surprisingly, since this was not part of their design criterion, other schemes designed assuming no relative delay between the signals from the transmitted antennas often suffer a similar fate, since many have structural similarities to the Alamouti scheme at some level. For example, a simple visual inspection reveals that the three codes described in Figures 4 and 5 of [1] become rank one when the signal from the second antenna arrives one symbol period after that from the first antenna, because they have aspects of the Alamouti codes embedded in them.

Now suppose that it is desirable to design a code that works for all relative delays in the set $\{-T_s, 0, +T_s\}$. Although it might be anticipated that this more restrictive condition necessitates an inability to maintain full rate, it is possible to find BPSK orthogonal designs that are full-rate macroscopic space-time code mapped to BPSK that maintains full rank in such a scenario. An example of such a code for the relative delays in the set $\{-T_s, 0, T_s\}$ is given by:

$$C_{PROP} = \begin{bmatrix} s_0 & s_1 & s_2 & s_3 \\ -s_2 & s_3 & s_0 & -s_1 \end{bmatrix},$$

which is two interleaved Alamouti codes. The trellis for the effective transmitted signal for such a code is given in Figure 3. Other BPSK orthogonal designs of length $4 * \text{lcm}(2, 3, \ldots, L)$, where $\text{lcm}(\cdot)$ is the least common multiple of the elements in the set that form its argument, have been found for larger delay sets $\{-LT_s, (-L+1)T_s, \ldots, -T_s, 0, T_s, \ldots, LT_s\}$ [6]. The simulated performance of the Alamouti and proposed schemes is shown in Figure 4.

IV. Performance Characterization

To consider relative delays that are not equal to an integer multiple of symbol intervals, it is important to first consider the optimal receiver design in such a scenario. If the transmitted pulse shape is time-limited (hence, *not* bandlimited), it is straightforward to extend Verdu's MLSE receiver [5] targeted for code-division multiple-access (CDMA) to this case as discussed in Section II. However, since it is of more practical interest to consider bandlimited (hence, *not* time-limited) signals, the main goal is to consider here bandlimited signals with arbitrary relative delay τ between the arriving signals. In this case, the form of the optimal receiver becomes complicated; hence, the practical receiver will likely be a generalization of a fractionally-spaced equalizer. The latter is complicated in the macroscopic space-time formulation, since (1) does not allow the received signal to be expressed simply as the linear filtering of a single transmitted signal. Thus, rather than design and analyze these receivers, their performance is estimated through other techniques in this section.

To characterize the performance of well-designed practical receivers for the case of a band-limited pulse shape, attention is turned to the matched filter bound [9, 10, 11], which studies the "one-shot" performance of the system with perfect channel estimation and no ISI, and hence lower bounds system error probabilities. Consider the probability of choosing the space-time codeword corresponding to

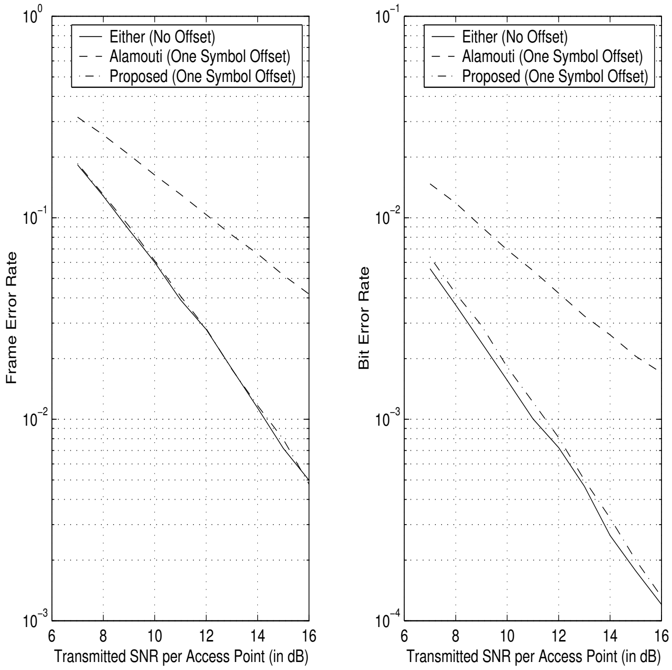

Figure 4: Simulated performance (200-bit frames) of the standard Alamouti scheme and the proposed scheme versus signal-to-noise ratio per transmitting base station. Each data point is the result of the simulation of 100,000 frames. The Alamouti scheme loses rank when the relative offset of the signals received from the two base stations is positive or negative one symbol period, whereas the proposed scheme displays only the slightest performance degradation.

the matrix V, where $v_{i,j}$ is the j^{th} symbol sent from the i^{th} base station, when the space-time codeword corresponding to the matrix C was sent. Because the relative delays considered will be in fractions of a symbol interval, simple space-time matrices must be abandoned in favor of waveforms. Thus, let the 2 by 1 matrix of waveforms, where the i^{th} row corresponds to the waveform sent from the i^{th} base station, for the transmitted codeword be given by

$$\begin{bmatrix} c_1(t) = \sum_{n=0}^{N-1} c_{1,n} p(t - nT_s) \\ c_2(t) = \sum_{n=0}^{N-1} c_{2,n} p(t - nT_s) \end{bmatrix}$$

and, for the codeword to which the error is made be

$$\begin{bmatrix} v_1(t) = \sum_{n=0}^{N-1} v_{1,n} p(t - nT_s) \\ v_2(t) = \sum_{n=0}^{N-1} v_{2,n} p(t - nT_s) \end{bmatrix}$$

where $p(\cdot)$ is the pulse-shaping waveform. Now, the matched filter bound to the pairwise error probability for a given τ is derived as:

$$\begin{aligned} P(C \to V) &\geq E_{X_1,X_2}\left[Q\left(\frac{d_{C,V,\tau}(X_1,X_2)}{\sqrt{2N_0}}\right)\right] \\ &= \frac{1}{\pi}\int_0^{2\pi} E_{X_1,X_2}\left[\exp\left(-\frac{d^2_{C,V,\tau}(X_1,X_2)}{4N_0\sin^2\theta}\right)\right] d\theta \\ &= \frac{1}{\pi}\int_0^{2\pi} \frac{1}{1+\frac{\lambda_1(C,V,\tau)}{4N_0\sin^2\theta}} \frac{1}{1+\frac{\lambda_2(C,V,\tau)}{4N_0\sin^2\theta}} d\theta \qquad (2) \end{aligned}$$

where $d_{C,V,\tau}(X_1, X_2)$ is the "received" Euclidean distance between the faded waveforms corresponding to C and V conditioned on the complex fading values X_1 and X_2, and the second line above exploits the alternate form of the $Q(\cdot)$ function [12]. The quantities $\lambda_1(C, V, \tau)$ and $\lambda_2(C, V, \tau)$ are the eigenvalues of the matrix A with entries:

$$\begin{aligned} A_{1,1} &= \int |v_1(t) - c_1(t)|^2 dt \\ A_{1,2} &= \int (v_1(t) - c_1(t))(v_2^*(t-\tau) - c_2^*(t-\tau)) dt \\ A_{2,1} &= \int (v_1^*(t) - c_1^*(t))(v_2(t-\tau) - c_2(t-\tau)) dt \\ A_{2,2} &= \int |v_2(t-\tau) - c_2(t-\tau)|^2 dt \end{aligned}$$

The above is the necessary waveform generalization of the discrete-time pairwise error probability derived in [1] for standard space-time systems.

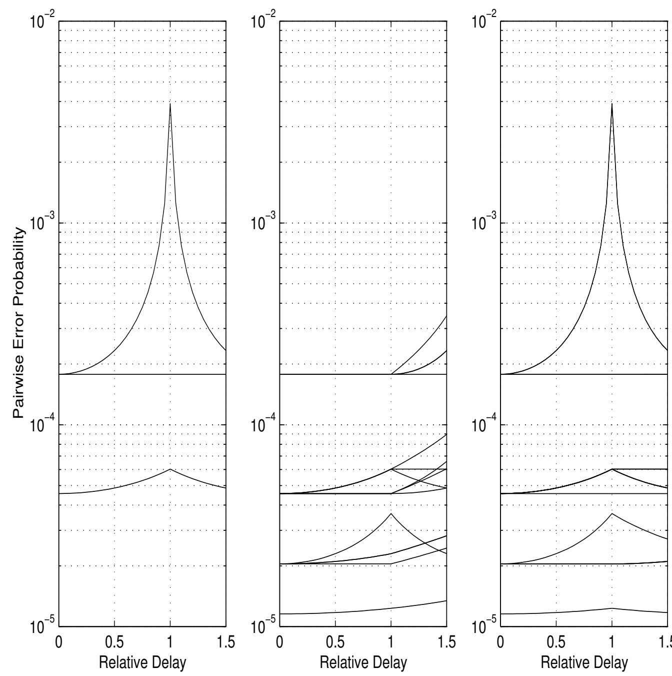

Figure 5: The matched filter bound to the pairwise error probability of various codeword pairs for (from left to right): (a) the Alamouti scheme, (b) the proposed scheme, and (c) a scheme which concatenates two Alamouti codewords together. The average received signal-to-noise ratio (SNR) is 15 dB per base station, and rectangular pulse shaping is employed. Note that for each delay, the effective code in each of the three cases is geometrically uniform, and thus it has been assumed that the codeword corresponding to information bits that are all zeros has been sent.

Although interest here is mainly in bandlimited pulse shapes, the matched filter bound is plotted in Figure 5 for rectangular pulse shapes for verification below. In Figure 6, the matched filter bound is plotted for raised cosine pulse shaping. From Figures 5 and 6, it is clear that the loss in rank of the Alamouti scheme at a relative delay of one symbol is symptomatic of weakened performance over a much broader range. The robustness over the delays in the

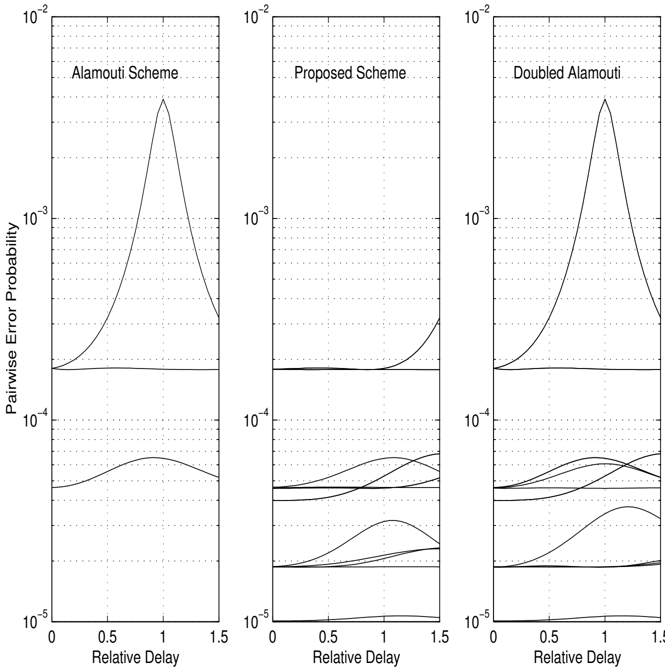

Figure 6: The matched filter bound to the pairwise error probability of various codeword pairs for (from left to right): (a) the Alamouti scheme, (b) the proposed scheme, and (c) a scheme which concatenates two Alamouti codewords together. The average received signal-to-noise ratio (SNR) is 15 dB per base station, and raised cosine pulse shaping with 50% excess bandwidth is employed. Note that for each delay, the effective code in each of the three cases is geometrically uniform, and thus it has been assumed that the codeword corresponding to information bits that are all zeros has been sent.

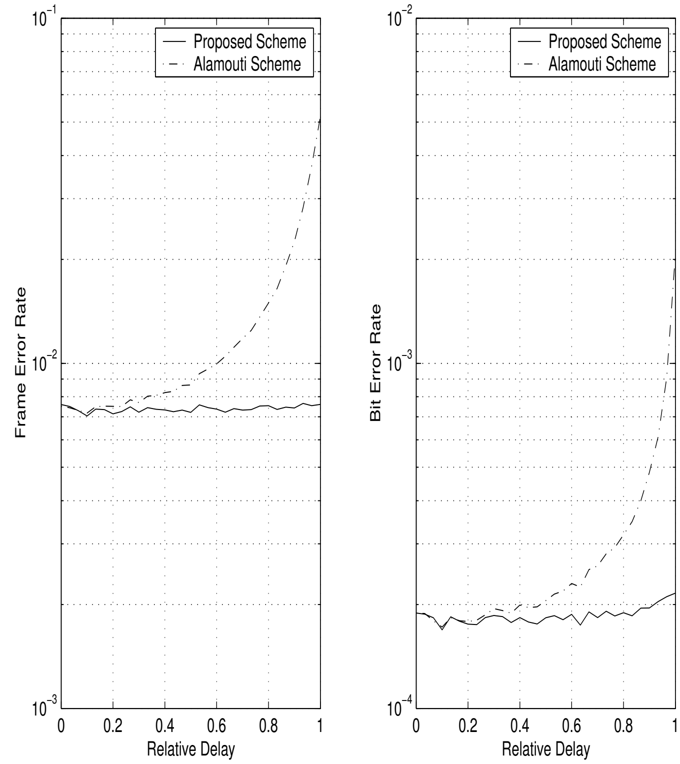

Figure 7: Simulated performance (200-bit frames) of the standard Alamouti scheme and the proposed scheme versus relative delay between the arrival times of the signals from the two base stations. The received SNR is 15 dB per per base station, and each data point is the result of the simulation of 100,000 frames. Rectangular pulse shaping is employed; thus, the optimal receiver derived in [6] and described in Section 1 is simulated. Note the close agreement with what the matched filter bound of Figure 5 predicts.

set $\{-T_s, 0, +T_s\}$ for the proposed scheme does in fact guarantee robustness over the interval $[-T_s, T_s]$ (results for only half of the interval are displayed, but everything is symmetric). Also, the performance gain of the proposed scheme is higher for the raised cosine pulse shape than for the rectangular pulse shape, which is promising since band-limited pulse shapes are used in practice. The increase in the matched filter bound for relative delays greater than T_s for the proposed scheme is expected, since this is beyond its designed range. In each figure, the matched filter bound results for a scheme which concatenates two Alamouti codes together; this is to show that the displayed broad gain in robustness of the proposed scheme over the Alamouti scheme is not an artifact of applying the matched filter bound technique to two codes of different lengths.

Next, the matched filter bound results of Figures 5 and 6 are verified through simulation. For each transmitted codeword in the Alamouti scheme, there is one other codeword for which the pairwise error corresponds to the poor-performing curve in Figure 5 or 6. Thus, it is anticipated that the simulated bit error rate performance of the Alamouti scheme should be one-half of the worst pairwise error probability curve. This is indeed what is observed through simulation as shown in Figure 7 for the case of rectangular pulses. By a similar argument, it is anticipated that the bit error probability of the proposed scheme should match that of the top curve in Figure 5 closely, which is also observed. Note that a simulation result for the case of raised cosine pulse shaping is not shown. Since the raised cosine pulse shape is not time-limited, the receiver described in Section 1 cannot be employed, and, as discussed above, the optimal receiver is very complex.

V. Coverage Results

Next, consider coverage results with parameters drawn from public safety radio - base station separation of 30 miles and symbol period of 0.05 milliseconds. Consider again Figure 1. Normalizing the distance between the base stations to unity, we will assume that the two base stations transmit together to mobiles over the one-by-one square with corners at (X, Y)-coordinates $(0.0, 0.0)$ and $(1.0, 1.0)$, with the base stations at the locations $(0.0, 0.5)$ and $(1.0, 0.5)$. Consider coverage in the rectangle with corners $(0.30, 0.0)$ and $(0.70, 1.0)$, which is the most difficult coverage region in such systems.

The four systems that will be considered are given as follows: (1) *Selection Macrodiversity:* The mobile associates only with the base station with which it achieves the highest *average* signal-to-noise ratio. (2) *Simulcast:* The base stations transmit exactly the same signal at exactly the same time. (3) *Alamouti:* The base stations transmit using the Alamouti scheme, except that one row of the space-time code matrix is sent from a given single-antenna base staion. (4) *Proposed:* The base stations transmit with the scheme proposed above, where one row of the space-time code matrix is sent from a given single-antenna base staion. Coverage results obtained through predictions from the matched filter bound technique are shown in Table 1, where duplicate locations arising from symmetry have been removed. Note the robust performance of the proposed scheme.

VI. Conclusions

In this paper, the macroscopic space-time coding problem has been introduced. In particular, the relative difference of the arrival times of the signals from the multiple base stations is a critical feature that cannot be ignored, particularly in systems employing partial broadcast. A family of space-time codes, where each code consists of codewords with orthogonal rows, has been introduced, and it has been demonstrated that such schemes greatly outperform previous solutions: standard selection macro-diversity, simulcasting, and standard space-time coding. This is due to the fact that standard space-time codes often demonstrate reduced (often no) diversity when employed in such an environment.

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X	Y	Slct	Simul	Alam	Prop
0.30	0.00	0.85	0.21	0.41	0.20
0.30	0.10	0.78	0.15	0.44	0.15
0.30	0.20	0.69	0.10	0.49	0.10
0.30	0.30	0.60	0.06	0.22	0.06
0.30	0.40	0.53	0.04	0.11	0.04
0.30	0.50	0.47	0.04	0.09	0.04
0.35	0.00	0.85	0.26	0.29	0.22
0.35	0.10	0.79	0.17	0.23	0.15
0.35	0.20	0.72	0.11	0.19	0.10
0.35	0.30	0.66	0.07	0.18	0.06
0.35	0.40	0.59	0.04	0.17	0.04
0.35	0.50	0.58	0.04	0.18	0.04
0.40	0.00	0.86	0.30	0.23	0.20
0.40	0.10	0.81	0.22	0.18	0.15
0.40	0.20	0.74	0.14	0.13	0.10
0.40	0.30	0.69	0.11	0.11	0.08
0.40	0.40	0.64	0.07	0.07	0.05
0.40	0.50	0.62	0.06	0.07	0.05
0.45	0.00	0.87	0.47	0.21	0.20
0.45	0.10	0.81	0.36	0.15	0.15
0.45	0.20	0.76	0.26	0.11	0.11
0.45	0.30	0.71	0.20	0.09	0.08
0.45	0.40	0.67	0.14	0.06	0.06
0.45	0.50	0.65	0.13	0.06	0.05
0.50	0.00	0.87	0.84	0.21	0.21
0.50	0.10	0.82	0.79	0.14	0.14
0.50	0.20	0.76	0.74	0.11	0.11
0.50	0.30	0.72	0.67	0.07	0.07
0.50	0.40	0.68	0.64	0.06	0.06
0.50	0.50	0.66	0.62	0.06	0.06

Table 1: Probability of outage at a given point, where outage is defined as the bit error rate being above threshold (which, per Figure 4, roughly translates to frame error rates that are equal for the compared systems) for each of the four considered schemes for various locations (X, Y) in the normalized region. The table goes roughly from locations further from the center of the cell at the top to locations roughly between the two base stations at the bottom. The path loss exponent is 3.0, and the standard deviation of the log-normal shadowing is $\sigma_Y = 8$ dB. Each point is the result of 5000 trials, where each trial requires an independent generation of the log-normal shadowing for that point. The average received SNR for a unit in the middle of the region per base station is 22.65 dB and the bit error rate threshold is 2×10^{-4}. As expected, selection macrodiversity performs the worst of the systems. Simulcast performs poorly in the center of the cell, but improves as it moves away from the center. The Alamouti scheme is excellent in the middle of the cell despite the low received average SNR, but it degrades away from there as expected from Figure 6. The proposed scheme shows robust performance across the region.