



Increasing Data Rate over Wireless Channels

Space-Time Coding
and Signal Processing for
High Data Rate Wireless
Communications

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Recent developments in communications are driven by several things: 1) high data rate wireless communication between two portable terminals that may be located anywhere in the world, 2) the vision of a single phone that acts as a traditional cellular phone when used outdoors and as a conventional high quality phone when used indoors [3], and 3) the need for access media to the homes other than the conventional copper phone lines. The great popularity of cordless phones, cellular phones, radio paging, portable computing, and other personal communication services (PCS) demonstrates rising demand for these services. Rapid growth in mobile computing and other wireless data services is inspiring many proposals for high-speed data services in the range of 64-144 kb/s for micro cellular wide area and high mobility application and up to 2 Mb/s for indoor applications [4]. In addition to mobile applications, fixed wireless access (FWA) technologies offer the promise of bringing high quality telephony, high-speed Internet access, multimedia, and other broadband services to the home over wireless links [5], [6]. Research challenges in this area include the development of efficient coding and modulation, signal processing techniques to improve the quality and spectral efficiency of wireless communications, and better techniques for sharing the limited spectrum among different high capacity users.

The information capacity of wireless communication systems increases dramatically by employing multiple transmit and receive antennas [1], [2]. An effective approach to increasing data rate over wireless channels is to employ coding techniques appropriate to multiple transmit antennas, namely space-time coding. Space-time coding (STC) [31]-[43], [37] is a coding technique that is designed for use with multiple transmit antennas. ST codes introduce temporal and spatial correlation into signals transmitted from different antennas, in order to provide diversity at the receiver, and coding gain over an uncoded system without sacrificing the bandwidth. The spatial-temporal structure of these codes can be exploited to further increase the capacity of wireless systems with a relatively simple receiver structure [44]. In this article we will review STC and its associated signal processing framework.

The physical limitation of the wireless channel presents a fundamental technical challenge for reliable communications. The channel is subject to time-varying impairments such as noise, interference, and multipath [7]-[13]. Limitations on the power and size of the mobile terminal and of network terminating devices (NTDs) in a FWA application is a second major design consideration. Most personal communications and wireless services portables are meant to be carried in a briefcase and/or pocket and, therefore, must be small and lightweight, which translates to a low power requirement since small batteries must be used. Although an NTD in FWA appli-

cations may have more signal processing power than a mobile computing portable, power consumption and device and antenna size are still a concern. However, many of the signal processing techniques which may be used for reliable communications and efficient spectral utilization demand significant processing power, precluding the use of low power devices. Continuing advances in VLSI and application-specific integrated circuit (ASIC) technology for low power applications will provide a partial solution to this problem. Hence, placing more signal processing burden on fixed locations (base stations) with relatively larger power resources than the portable makes good engineering sense.

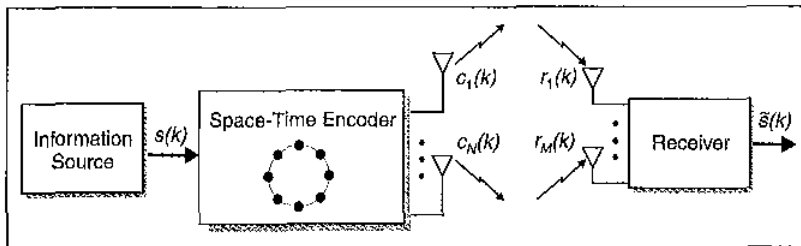
Several diversity techniques have been employed in wireless communication systems to improve the link margin. Diversity techniques which may be used include time, frequency, and space diversity.

▲ *Time Diversity*: Channel coding in combination with limited interleaving is used to provide time diversity. However, while channel coding is extremely effective in fast fading environments (high mobility), it offers very little protection under slow fading (low mobility and FWA) unless significant interleaving delays can be tolerated.

▲ *Frequency Diversity*: The fact that signals transmitted over different frequencies induce different multipath structure and independent fading is exploited to provide frequency diversity (sometimes referred to as path diversity). In time division multiple access (TDMA) systems, frequency diversity is obtained by the use of equalizers [14] when the multipath delay spread is a significant fraction of a symbol period. Global system for mobile communication (GSM) uses frequency hopping to provide frequency diversity. In direct sequence-code division multiple access (DS-CDMA) systems, RAKE receivers [15], [16] are used to obtain path diversity. When the multipath delay spread is small, compared to the symbol period, however, frequency or path diversity does not exist.

▲ *Space Diversity*: The receiver/transmitter uses multiple antennas that are separated and/or differently polarized for reception/transmission to create independent fading channels. Currently, multiple antennas at base stations are used for receive diversity at the base. It is difficult, however, to have more than one or two antennas at the portable unit due to the size limitations and cost of multiple chains of RF down conversion.

Both receive and polarization diversity have received much attention [11], [12], [17]. In fact, in current cellular applications, receive diversity is already used for improving reception from mobiles. Transmit diversity, on the other hand, has received comparatively little attention. The information theoretic aspects of transmit diversity were addressed in [1], [20], [21], and [2]. Previous work on transmit diversity can be classified into three broad categories: schemes using feedback, schemes with feedforward or training information but no feedback, and blind schemes. The first category uses feedback, either explicitly or implicitly, from the receiver to the transmitter to train the trans-



▲ 1. Space-time coding.

mitter. For instance, in time division duplex (TDD) systems [18], the same antenna weights are used for reception and transmission, so feedback is implicit in the exploitation of channel symmetry. These weights are chosen during reception to maximize the receive SNR, and during transmission to weight the amplitudes of the transmitted signals, and, therefore, will also maximize the SNR at the receiver. Explicit feedback includes switched diversity systems with feedback [19]. In practice, however, movement by either the transmitter or the receiver (or the surroundings such as cars) and interference dynamics causes a mismatch between the channel perceived by the transmitter and that perceived by the receiver.

Transmit diversity schemes mentioned in the second category use linear processing at the transmitter to spread the information across antennas. At the receiver, information is recovered by an optimal receiver. Feedforward information is required to estimate the channel from the transmitter to the receiver. These estimates are used to compensate for the channel response at the receiver. The first scheme of this type was proposed by Wittneben [22], and it includes the delay diversity scheme of [23] as a special case. The linear processing techniques were also studied in [24] and [25]. It was shown in [26] and [27] that delay diversity schemes are indeed optimal in providing diversity, in the sense that the diversity gain experienced at the receiver (which is assumed to be optimal) is equal to the diversity gain obtained with receive diversity. The linear filtering used (to create delay diversity) at the transmitter can be viewed as a channel code that takes binary or integer input and creates real valued out-

put. We show that there is a significant gain to be realized by viewing this problem from a coding perspective rather than from a purely signal processing point of view.

The third category does not require feedback or feedforward information. Instead, it uses multiple transmit antennas combined with channel coding to provide diversity. An example of

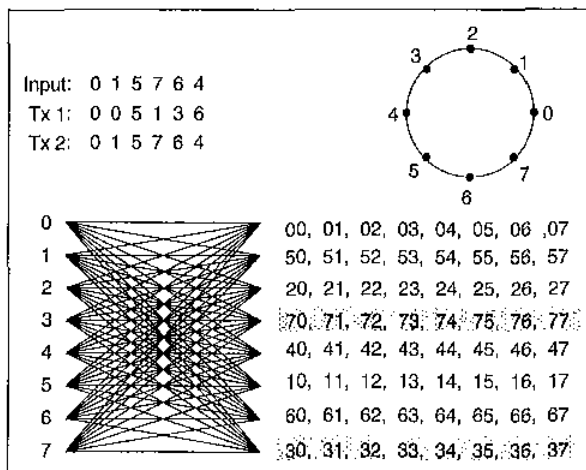
this approach is the use for channel coding along with phase sweeping [28] or for frequency offset [29] with multiple transmit antennas to simulate fast fading. An appropriately designed channel code/interleaver pair is used to provide the diversity benefit. Another approach in this category is to encode information by a channel code and transmit the code symbols using different antennas in an orthogonal manner. This can be done by either time multiplexing [28], or by using orthogonal spreading sequences for different antennas [30]. The disadvantage of these schemes as compared to the previous two categories is the loss in bandwidth efficiency due to the use of the channel code. Using appropriate coding, it is possible to relax the orthogonality requirement needed in these schemes and to obtain the diversity as well as a coding gain without sacrificing bandwidth. This will be possible to do if one views the whole system as a multiple input/multiple output system and uses channel codes that are designed with that view in mind.

Space-Time Coding

We will describe a basic model for a communication system that employs space time coding with N transmit antennas and M receive antennas. As shown in Fig. 1, the information symbol $s(l)$ at time l is encoded by the ST encoder as N code symbols $c_1(l), c_2(l), \dots, c_N(l)$. Each code symbol is transmitted, simultaneously, from a different antenna. The encoder chooses the N code symbols to transmit so that both the coding gain and diversity gain at the receiver are maximized.

Signals arriving at different receive antennas undergo independent fading. The signal at each receive antenna is a noisy superposition of the faded versions of the N transmitted signals. A flat fading channel is assumed. We assume that the signal constellation is scaled so that the average energy of the constellation points is one. Also, let us assume that E_s is the total energy transmitted (from all antennas) per input symbol. Therefore, the energy per input symbol transmitted from each transmit antenna is E_s / N . Let $r_j(l)$, $j = 1, \dots, M$ be the received signal at antenna j after matched filtering. Assuming ideal timing and frequency information, we have

$$r_j(l) = \sqrt{E_s / N} \cdot \sum_{i=1}^N h_{ij}(l) c_i(l) + \eta_j(l), \quad j = 1, \dots, M \quad (1)$$



▲ 2. Eight-PSK eight-state ST code with two transmit antennas.

where $\eta_i(l)$ are independent samples of a zero mean complex white Gaussian process with two-sided power spectral density $N_0/2$ per dimension. It is also assumed that $\eta_i(l)$ and $\eta_k(l)$ are independent for $j \neq k$, $1 \leq j, k \leq M$. The gain $h_{ij}(l)$ models the complex fading channel gain from transmit antenna i to receive antenna j . It is assumed that $h_{ij}(l)$ and $h_{jk}(l)$ are independent for $i \neq q$ or $j \neq k$, $1 \leq i, q \leq N$, $1 \leq j, k \leq M$. This condition is satisfied if the transmit antennas are well separated (by more than $\lambda/2$) or by using antennas with different polarization.

Let $\mathbf{c}_l = [c_1(l), \dots, c_N(l)]^T$ be the $N \times 1$ code vector transmitted from the N antennas at time l , $\mathbf{h}_j(l) = [h_{1j}(l), \dots, h_{Nj}(l)]^T$ be the corresponding $N \times 1$ channel vector from the N transmit antennas to the j th receive antenna, and $\mathbf{r}(l) = [r_1(l), \dots, r_M(l)]^T$ be the $M \times 1$ received signal vector. Also, let $\eta(l) = [\eta_1(l), \dots, \eta_M(l)]^T$ be the $M \times 1$ noise vector at the receive antennas. Let us define the $M \times N$ channel matrix \mathcal{H}_l from the N transmit to the M receive antennas as $\mathcal{H}(l) = [\mathbf{h}_1(l), \dots, \mathbf{h}_M(l)]^T$. Equation (1) can be rewritten in a matrix form as

$$\mathbf{r}(l) = \sqrt{E_s/N} \cdot \mathcal{H}(l) \cdot \mathbf{c}_l + \eta(l). \quad (2)$$

We can easily see that the SNR per receive antenna is given by

$$\text{SNR} = \frac{E_s}{N_0}. \quad (3)$$

Space-Time Trellis Codes

Suppose that the code vector sequence

$$\mathcal{C} = \mathbf{c}_1, \mathbf{c}_2, \dots, \mathbf{c}_L$$

was transmitted. We consider the probability that the decoder decides erroneously in favor of the legitimate code vector sequence

$$\tilde{\mathcal{C}} = \tilde{\mathbf{c}}_1, \tilde{\mathbf{c}}_2, \dots, \tilde{\mathbf{c}}_L.$$

Consider a frame or block of data of length L and define the $N \times N$ error matrix \mathcal{A} as

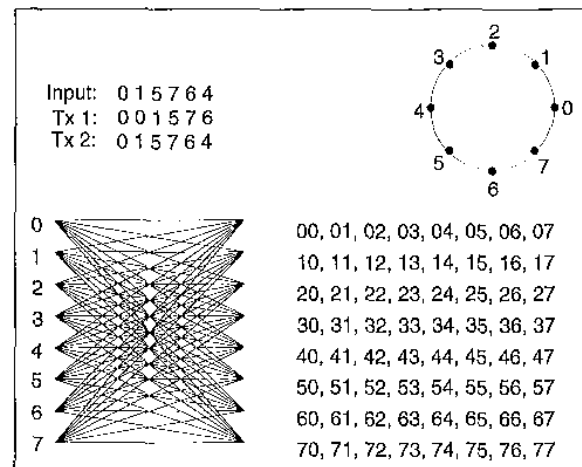
$$\mathcal{A}(\mathcal{C}, \tilde{\mathcal{C}}) = \sum_{l=1}^L (\mathbf{c}_l - \tilde{\mathbf{c}}_l)(\mathbf{c}_l - \tilde{\mathbf{c}}_l)^* \quad (4)$$

where $(\cdot)^*$ denotes the conjugate operation for scalars and the conjugate transpose for matrices and vectors. If ideal channel state information (CSI) $\mathcal{H}(l) = 1, \dots, L$ is available at the receiver, then it is straightforward to show that the probability of transmitting \mathcal{C} and deciding in favor of $\tilde{\mathcal{C}}$ is upper bounded by [45]

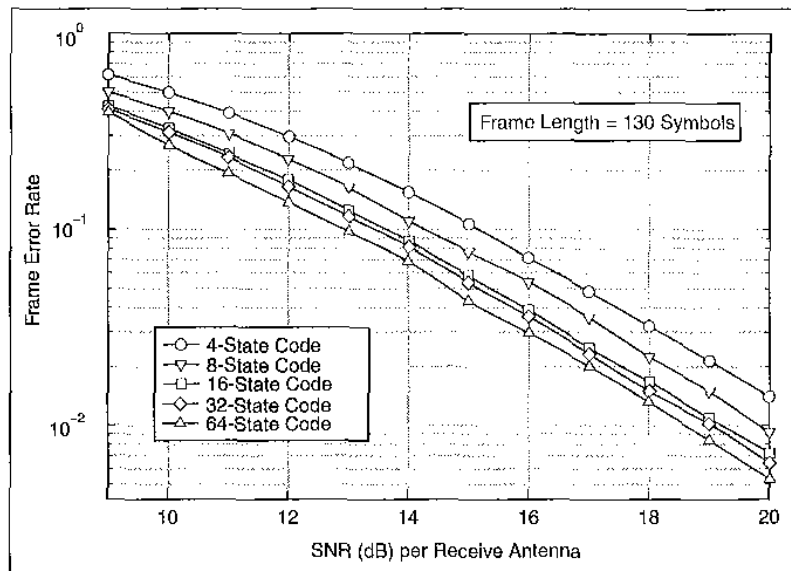
$$P(\mathcal{C} \rightarrow \tilde{\mathcal{C}}) \leq \left(\prod_{i=1}^r \lambda_i \right)^{-M} \cdot (E_s / 4N_0)^{-rM} \quad (5)$$

where r is the rank of the error matrix \mathcal{A} and $\lambda_i, i=1, \dots, r$ are the nonzero eigenvalues of the error matrix \mathcal{A} . We can easily see that the probability of error bound in (5) is similar to the probability of error bound for trellis coded modulation for fading channels. The first term $\mathcal{G}_r = \prod_{i=1}^r \lambda_i$ represents the coding gain achieved by the ST code, and the second term $(E_s / 4N_0)^{-rM}$ represents a diversity gain of rM . It is clear that in designing an ST trellis code, the rank of the error matrix r should be maximized (thereby maximizing the diversity gain) and at the same time \mathcal{G}_r should be also maximized (thereby maximizing the coding gain).

As an example for ST trellis codes, we provide an 8-PSK (phase-shift keyed) eight-state ST code designed for two transmit antennas. Figure 2 provides a labeling of the



▲ 3. Eight-PSK eight-state delay diversity as an ST code.



▲ 4. Performance of 4-PSK ST trellis codes with two transmit and one receive antennas.

One of the goals of the third- and fourth-generation wireless systems is to provide broadband access to both mobile and stationary users.

8-PSK constellation and the trellis description for this code. Each row in the matrix shown in Fig. 2 represents the edge labels for transitions from the corresponding state. The edge label $s_1 s_2$ indicates that symbol s_1 is transmitted over the first antenna and that symbol s_2 is transmitted over the second antenna. The input bit stream to the ST encoder is divided into groups of 3 bits, and each group is mapped into one of eight constellation points. This code has a bandwidth efficiency of 3 bits/channel use.

Figure 3 shows the STC representation of delay diversity. It is also interesting to note that the two trellis codes in Figs. 2 and 3 are similar. In fact, we can get the code in Fig. 2 by swapping the row that starts with a "1" with the row that starts with a "5" and the row that starts with a "3" with the row that starts with a "7." By looking at the constellation points in Fig. 2, we will easily realize that this ST code is delay-diversity except that the delayed symbol is multiplied by -1 if it is an odd symbol $\{1,3,5,7\}$ and by $+1$ if it is an even symbol $\{0,2,4,6\}$. This simple mapping of the delayed symbol gives a 2.5 dB of coding gain as compared to simple delay diversity.

For decoding ST codes, we assume that the channel information $\mathcal{H}(l), l=1, \dots, L$ is available at the receiver. Suppose that a code vector sequence $\mathcal{C} = \mathbf{c}_1, \mathbf{c}_2, \dots, \mathbf{c}_L$ has been transmitted, and $\mathcal{R} = \mathbf{r}_1, \mathbf{r}_2, \dots, \mathbf{r}_L$ has been received, where \mathbf{r}_l is given by (2). At the receiver, optimum decoding amounts to choosing a vector code sequence $\tilde{\mathcal{C}} = \tilde{\mathbf{c}}_1, \tilde{\mathbf{c}}_2, \dots, \tilde{\mathbf{c}}_L$ for which the *a posteriori* probability

$$\Pr(\tilde{\mathcal{C}} | \mathcal{R}, \mathcal{H}(l), l=1, \dots, L)$$

is maximized. Assuming that all codewords are equiprobable, then since the noise vector is assumed to be

a multivariate AWGN, it can be easily shown that the optimum decoder is [45]

$$\tilde{\mathcal{C}} = \arg \min_{\tilde{\mathcal{C}} = \tilde{\mathbf{c}}_1, \dots, \tilde{\mathbf{c}}_L} \sum_{l=1}^L \|\mathbf{r}(l) - \sqrt{E_s} \cdot \mathcal{H}(l) \cdot \tilde{\mathbf{c}}_l\|^2. \quad (6)$$

For the ST codes with trellis representations (as in the example in Fig. 2), it is obvious that the optimum decoder in (6) can be implemented using the Viterbi algorithm. Note that knowledge of the channel is required for decoding. The receiver, therefore, must estimate the channel either blindly or by using pilot/training symbols. Figure 4 shows the performance of 4-PSK ST trellis codes for two transmit and one receive antennas with different numbers of states.

Space-Time Block Codes

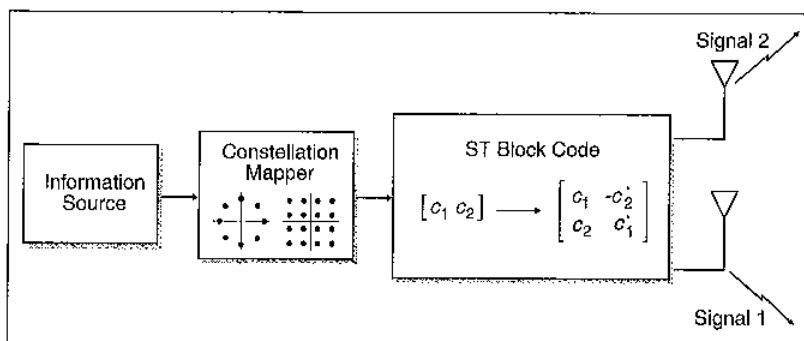
When the number of antennas is fixed, the decoding complexity of ST trellis coding (measured by the number of trellis states at the decoder) increases exponentially as a function of the diversity level and transmission rate [35]. In addressing the issue of decoding complexity, Alamouti [37] discovered a remarkable ST block coding scheme for transmission with two antennas. This scheme supports maximum-likelihood detection based only on linear processing at the receiver. This scheme was later generalized in [38] to an arbitrary number of antennas and is able to achieve the full diversity promised by the number of transmit and receive antennas. Here, we will briefly review the basics of ST block codes (STBCs). Figure 5 shows the baseband representation for ST block coding with two antennas at the transmitter. The input symbols to the ST block encoder are divided into groups of two symbols each. At a given symbol period, the two symbols in each group $\{c_1, c_2\}$ are transmitted simultaneously from the two antennas. The signal transmitted from antenna 1 is c_1 , and the signal transmitted from antenna 2 is c_2 . In the next symbol period, the signal $-c_2^*$ is transmitted from antenna 1 and the signal c_1^* is transmitted from antenna 2. Let h_1 and h_2 be the channels from the first and second transmit antennas to the receive antenna, respectively. The major assumption here is that h_1 and h_2 are constant over two consecutive symbol periods, that is

$$h_i(nT) = h_i((n+1)T), \quad i=1,2.$$

We assume a receiver with a single receive antenna and denote the received signals over two consecutive symbol periods as r_1 and r_2 . The received signals can be written as

$$r_1 = h_1 c_1 + h_2 c_2 + \eta_1 \quad (7)$$

$$r_2 = -h_1 c_2^* + h_2 c_1^* + \eta_2 \quad (8)$$



▲ 5. Transmitter diversity with ST block coding.

where η_1 and η_2 represent the AWGN and are modeled as i.i.d. complex Gaussian random variables with zero mean and power spectral density $N_0/2$ per dimension. We define the received signal vector $\mathbf{r} = [r_1 \ r_2]^T$, the code symbol vector $\mathbf{c} = [c_1 \ c_2]^T$, and the noise vector $\boldsymbol{\eta} = [\eta_1 \ \eta_2]^T$. Equations (7) and (8) can be rewritten in a matrix form as

$$\mathbf{r} = \mathbf{H} \cdot \mathbf{c} + \boldsymbol{\eta} \quad (9)$$

where the channel matrix \mathbf{H} is defined as

$$\mathbf{H} = \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix} \quad (10)$$

The vector $\boldsymbol{\eta}$ is a complex Gaussian random vector with zero mean and covariance $N_0 \cdot \mathbf{I}$. Let us define \mathcal{C} as the set of all possible symbol pairs $\mathbf{c} = \{c_1, c_2\}$. Assuming that all symbol pairs are equiprobable and since the noise vector $\boldsymbol{\eta}$ is assumed to be a multivariate AWGN, we can easily see that the optimum maximum-likelihood decoder is

$$\hat{\mathbf{c}} = \arg \min_{\mathbf{c} \in \mathcal{C}} \|\mathbf{r} - \mathbf{H} \cdot \mathbf{c}\|^2 \quad (11)$$

The ML decoding rule in (11) can be further simplified by realizing that the channel matrix \mathbf{H} is orthogonal and, hence, $\mathbf{H}^T \mathbf{H} = \rho \cdot \mathbf{I}$ where $\rho = |h_1|^2 + |h_2|^2$. Consider the modified signal vector $\tilde{\mathbf{r}}$ given by

$$\tilde{\mathbf{r}} = \mathbf{H}^T \cdot \mathbf{r} = \rho \cdot \mathbf{c} + \tilde{\boldsymbol{\eta}} \quad (12)$$

where $\tilde{\boldsymbol{\eta}} = \mathbf{H}^T \cdot \boldsymbol{\eta}$. In this case the decoding rule becomes

$$\hat{\mathbf{c}} = \arg \min_{\mathbf{c} \in \mathcal{C}} \|\tilde{\mathbf{r}} - \rho \cdot \mathbf{c}\|^2 \quad (13)$$

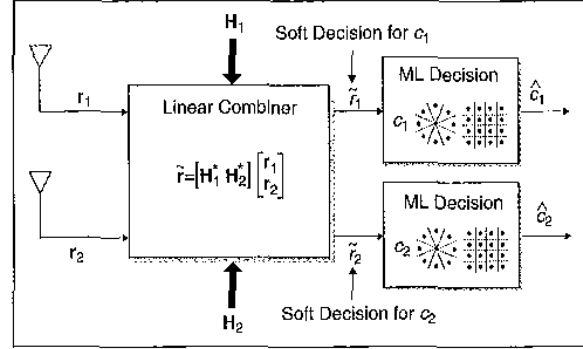
Since \mathbf{H} is orthogonal, we can easily verify that the noise vector $\tilde{\boldsymbol{\eta}}$ will have a zero mean and covariance $\rho N_0 \cdot \mathbf{I}$, i.e., the elements of $\tilde{\boldsymbol{\eta}}$ are independent and identically distributed. Hence, it follows immediately that by using this simple linear combining, the decoding rule in (13) reduces to two separate, and much simpler, decoding rules for c_1 and c_2 , as established in [37]. In fact, for the above 2×2 STBC, only two complex multiplications and one complex addition per symbol are required for decoding. Also, assuming that we are using a signaling constellation with 2^b constellation points, this linear combining reduces the number of decoding metrics that has to be computed for ML decoding from 2^{2b} to 2×2^b . It is also straight forward to verify that the SNR for c_1 and c_2 will be

$$\text{SNR} = \frac{\rho \cdot E_s}{N_0} \quad (14)$$

and hence a two-branch diversity performance (i.e., a diversity gain of order two) is obtained at the receiver.

When the receiver uses M receive antennas, the received signal vector \mathbf{r}_m at receive antenna m is

$$\mathbf{r}_m = \mathbf{H}_m \cdot \mathbf{c} + \boldsymbol{\eta}_m \quad (15)$$



▲ 6. Receiver for ST block coding.

where $\boldsymbol{\eta}_m$ is the noise vector and \mathbf{H}_m is the channel matrix from the two transmit antennas to the m th receive antenna. In this case the optimum ML decoding rule is

$$\hat{\mathbf{c}} = \arg \min_{\mathbf{c} \in \mathcal{C}} \sum_{m=1}^M \|\mathbf{r}_m - \mathbf{H}_m \cdot \mathbf{c}\|^2 \quad (16)$$

As before, in the case of M receive antennas, the decoding rule can be further simplified by premultiplying the received signal vector \mathbf{r}_m by \mathbf{H}_m^T . In this case, the diversity order provided by this scheme is $2M$. Figure 6 shows a simplified block diagram for the receiver with two receive antennas. Note that the decision rule in (13) and (16) amounts to performing a hard decision on $\tilde{\mathbf{r}}$ and $\tilde{\mathbf{r}}_M = \sum_{m=1}^M \mathbf{H}_m^T \mathbf{r}_m$, respectively. Therefore, as shown in Fig. 6, the received vector after linear combining, $\tilde{\mathbf{r}}_M$, can be considered as a soft decision for c_1 and c_2 . When the STBC is concatenated with an outer conventional channel code, like a convolutional code, these soft decisions can be fed to the outer channel decoder to yield a better performance. Note also that for the above 2×2 STBC, the transmission rate is one while achieving the maximum diversity gain possible with two transmit antennas.

The extension of the above STBC was studied in [38]. There, a general technique for constructing STBCs for $N > 2$ that provide the maximum diversity promised by the number of transmit and receive antennas was developed. These codes retain the simple ML decoding algorithm based on only linear processing at the receiver [37]. It was also shown that for real signal constellations (PAM constellation), STBCs with transmission rate 1 can be constructed [38]. However, for a general complex constellations like M-QAM or M-PSK, it is not known whether an STBC with transmission rate 1 and simple linear processing that will give the maximum diversity gain with $N > 2$ transmit antennas does exist or not. Moreover, it was also shown that such code where the number of transmit antennas N equals the number of equals both the number of information symbols transmitted and the number of time slots need to transmit the code block does not exist. However for rates < 1 , such codes can be found. For example, assuming that the transmitter unit uses four transmit antennas, a rate $4/8$ (i.e., it is a rate $1/2$) STBC is given by

STC is a new coding/signal processing framework for wireless communication systems with multiple transmit and multiple receive antennas. This new framework has the potential of dramatically improving the capacity and data rates.

$$\begin{bmatrix} c_1 \\ c_2 \\ c_3 \\ c_4 \end{bmatrix} \rightarrow \begin{bmatrix} c_1 & -c_2 & -c_3 & -c_4 & c_1^* & -c_2^* & -c_3^* & -c_4^* \\ c_2 & c_1 & c_4 & -c_3 & c_2^* & c_1^* & c_4^* & -c_3^* \\ c_3 & -c_4 & c_1 & c_2 & c_3^* & -c_4^* & c_1^* & c_2^* \\ c_4 & c_3 & -c_2 & c_1 & c_4^* & c_3^* & -c_2^* & c_1^* \end{bmatrix} \quad (17)$$

In this case, at time $t=1$, c_1, c_2, c_3, c_4 are transmitted from antennas 1 through 4, respectively. At time $t=2$, $-c_2, c_1, -c_4, c_3$ are transmitted from antenna 1 through 4, respectively, and so on. For this example, let r_1, r_2, \dots, r_8 be the received signals at time $t=1, 2, \dots, 8$, respectively. Define the new received signal vector $\mathbf{r} = [r_1, r_2, r_3, r_4, r_5, r_6, r_7, r_8]^T$. In this case we can write the received signal vector \mathbf{r} at the receive antenna as

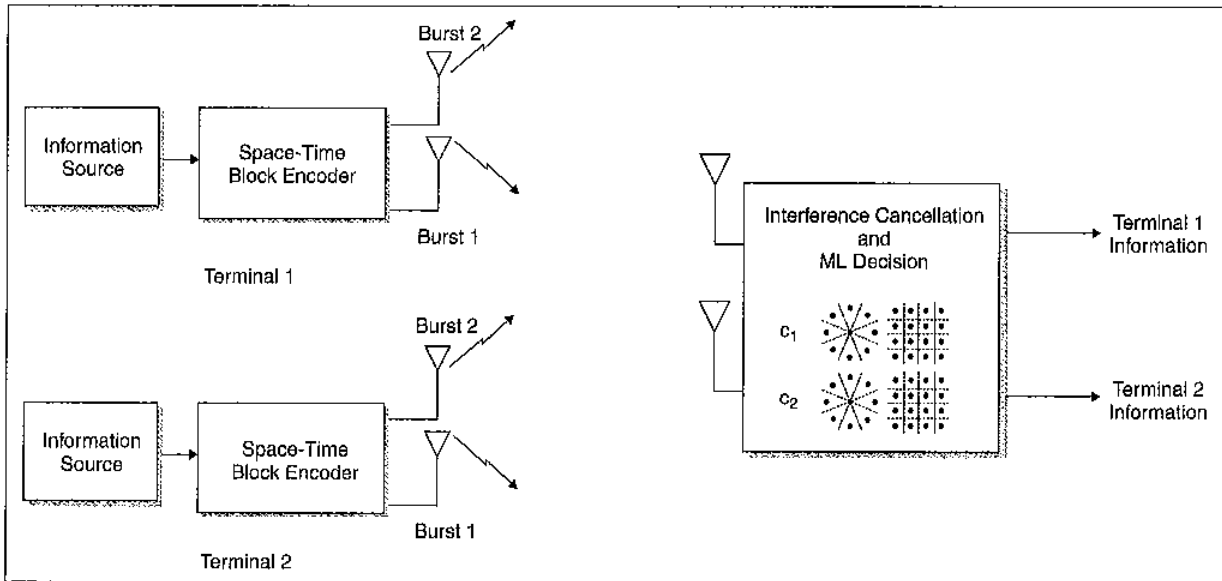
$$\mathbf{r} = \mathbf{H} \cdot \mathbf{c} + \boldsymbol{\eta} \quad (18)$$

where $\boldsymbol{\eta}$ is an 8×1 AWGN noise vector and \mathbf{H} is 8×4 channel matrix given by

$$\mathbf{H} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2 & -h_1 & h_4 & -h_3 \\ h_3 & -h_4 & -h_1 & h_2 \\ h_4 & h_3 & -h_2 & -h_1 \\ h_1^* & h_2^* & h_3^* & h_4^* \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & -h_4^* & -h_1^* & h_2^* \\ h_4^* & h_3^* & -h_2^* & -h_1^* \end{bmatrix} \quad (19)$$

We can immediately see that \mathbf{H} is orthogonal, that is $\mathbf{H}^T \mathbf{H} = \rho_4 \cdot \mathbf{I}$, where $\rho_4 = 2 \cdot \sum_{i=1}^4 |h_i|^2$. Therefore, the same procedure used for decoding the simple 2×2 STBC can be used for this code too. In this case, the SNR for c_1, \dots, c_4 is $\rho_4 E_s / N_0$, providing a 3 dB coding gain in addition to four-branch diversity performance. The 3 dB coding gain comes from the (intuitive) fact that eight time slots are used to transmit four information symbols.

Note that the decoding of STBCs requires knowledge of the channel at the receiver. The channel state information can be obtained at the receiver by sending training or pilot symbols or sequences to estimate the channel from each of the transmit antennas to the receive antenna [46]–[53]. For one transmit antenna, there exist differential detection schemes, such as DPSK, that neither require knowledge of the channel nor employ pilot or training symbol transmission. These differential decoding schemes are used, for example, in the IS-54 cellular standard ($\pi/4$ -DPSK). This motivates the generalization of differential detection schemes for the case of multiple transmit antennas. A partial solution to this problem was proposed in [41] for the 2×2 code, where it was assumed that the channel is not known at the receiver. In this scheme, the detected pair of symbols at time $t-1$ are used to estimate the channel at the receiver and these channel estimates are used for detecting the pair of symbols at

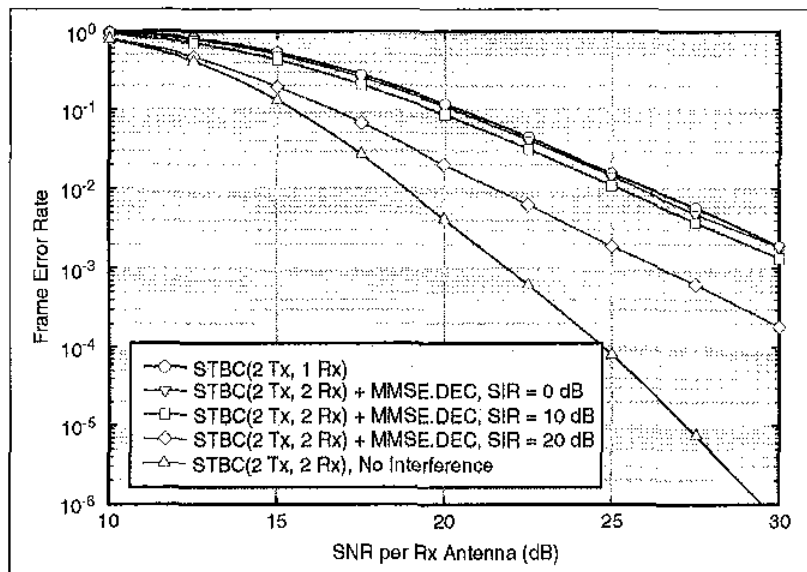


▲ 7. Interference cancellation with STBCs.

time t . The scheme in [41], however, requires the transmission of known pilot symbols at the beginning and hence is not fully differential. The scheme in [41] can be thought as joint data/channel estimation approach which can lead to error propagation. In [40], a true differential detection scheme for the 2×2 code was constructed. This scheme shares many of the desirable properties of DPSK: it can be demodulated with or without CSI at the receiver, achieves full diversity gain in both cases, and there exists a simple noncoherent receiver that performs within 3 dB of the coherent receiver. However, this scheme has some limitations. First, the encoding scheme expands the signal constellation for nonbinary signals. Second, it is limited only to the $N=2$ STBC for complex constellations and to the case $N \leq 8$ for real constellation. This is based on the results in [38] that the 2×2 STBC is an orthogonal design and complex orthogonal designs do not exist for $N > 2$. In [54], another approach for differential modulation with transmit diversity based on group codes was proposed. This approach can be applied to any number of antennas and to any constellation. The group structure of these codes greatly simplifies the analysis of these schemes and may also yield simpler and more transparent modulation and demodulation procedures. A different no-differential approach to transmit diversity when the channel is not known at the receiver is reported in [55] and [56] but this approach requires exponential encoding and decoding complexities.

Interference Suppression with Space-Time Block Codes

The properties of the ST block coding scheme in [37] and its extension in [38] can be further exploited to develop efficient interference suppression techniques that can be used to increase system capacity or increase throughput for individual users. In general, we consider a multiuser environment with K synchronous cochannel users where each user is equipped with N transmit antennas and uses an STBC with N transmit antenna. In general, in this scenario, there will be $K \times N$ interfering signals arriving at the receiver. Therefore, classical interference suppression techniques [57] with multiple receive antennas will require $N \times (K-1) + 1$ antennas at the receiver to suppress signals from the $K-1$ cochannel ST users and achieve a diversity order of N for the desired terminal. By exploiting the temporal and spatial structure of STBCs, it can be shown [42], [43], and [58] that only K antennas are required to suppress the interference from the $K-1$ cochannel users while maintaining the diversity order of



▲ 8. MMSE interference cancellation with STBCs.

Table 1. Zero-Forcing IC and ML Decoding Algorithm for Space-Time Block Codes.

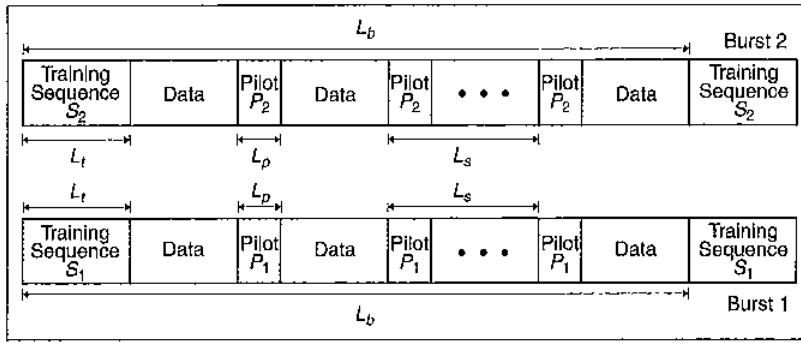
$(\hat{c}, \Delta) = \text{ZF_DEC}(r_1, r_2, H_1, H_2, G_1, G_2)$

$$\begin{cases} \tilde{r} = r_1 - G_1 G_2^{-1} r_2 \\ \tilde{H} = H_1 - G_1 G_2^{-1} H_2 \\ \sigma^2 = 1 + \text{Tr}\{G_1 G_2^{-1}\} / \text{Tr}\{G_2 G_2^H\} \\ \hat{c} = \arg \min_{c \in C} \|\sigma^{-1}(\tilde{r} - \tilde{H} \hat{c})\|^2 \\ \Delta = \|\sigma^{-1}(\tilde{r} - \tilde{H} \hat{c})\|^2 \end{cases}$$

Table 2. MMSE IC Decoding Algorithm for Space-Time Block Codes.

$(\hat{c}, \Delta) = \text{MMSE_DEC}(r_1, r_2, H_1, H_2, G_1, G_2, \Gamma)$

$$\begin{cases} \tilde{r} = [r_1^T r_2^T]^T \\ \tilde{H} = \begin{bmatrix} H_1 & G_1 \\ H_2 & G_2 \end{bmatrix} \\ M = \tilde{H} \tilde{H}^H + \frac{1}{\Gamma} I_4 \\ h_1 = [b_{11} \ b_{21} \ b_{12} \ b_{22}]^T \\ \quad = \text{first column of } \tilde{H} \\ h_2 = [b_{21} - b_{11}^* \ b_{22} - b_{12}^*]^T \\ \quad = \text{second column of } \tilde{H} \\ w_1 = M^{-1} h_1 \\ w_2 = M^{-1} h_2 \\ \hat{c} = \arg \min_{c \in C} \{\|w_1^H r - \hat{c}_1\|^2 + \|w_2^H r - \hat{c}_2\|^2\} \\ \Delta = \|w_1^H r - \hat{c}_1\|^2 + \|w_2^H r - \hat{c}_2\|^2 \end{cases}$$



▲ 9. Downlink slot structure for STCM-based modem.

N provided by the STBC. Given the assumption that the receiver is equipped with $M \geq K$ antennas, zero forcing (ZF) and minimum mean-squared error (MMSE) interference suppression techniques that exploit the structure of the STBC are developed in [42] and [58]. These techniques will perfectly suppress the interference from the $K-1$ cochannel users and provide a diversity order of $N \times (M - K + 1)$ while maintaining the simple linear processing feature of the STBCs.

We outline these interference cancellation schemes for the 2×2 case here. For a more detailed treatment the reader is referred to [42]. Figure 7 shows a simple scenario for two synchronous cochannel ST users (each employs the 2×2 STBC) and a receiver with two receiver antennas. Using the signal model developed above, the received signal vector at antenna 1 and 2 are

$$\mathbf{r}_1 = \mathbf{H}_1 \cdot \mathbf{c} + \mathbf{G}_1 \cdot \mathbf{s} + \boldsymbol{\eta}_1 \quad (20)$$

$$\mathbf{r}_2 = \mathbf{H}_2 \cdot \mathbf{c} + \mathbf{G}_2 \cdot \mathbf{s} + \boldsymbol{\eta}_2 \quad (21)$$

where \mathbf{r}_1 is the received vector at antenna 1, \mathbf{r}_2 is the received vector at antenna 2, \mathbf{c} is the vector of code symbols from first user, and \mathbf{s} is the vector of code symbols from second user. The matrices \mathbf{H}_1 and \mathbf{H}_2 are the channel matrices from the first ST user to the first and second receive antennas, respectively, and are defined similar to (10). Similarly, the matrices \mathbf{G}_1 and \mathbf{G}_2 are the channel matrices from the second ST user to the first and second receive antennas, respectively. The last two equations can be rewritten as

$$\mathbf{r} = \begin{bmatrix} \mathbf{r}_1 \\ \mathbf{r}_2 \end{bmatrix} = \mathbf{H} \cdot \tilde{\mathbf{c}} + \boldsymbol{\eta} \\ = \begin{bmatrix} \mathbf{H}_1 & \mathbf{G}_1 \\ \mathbf{H}_2 & \mathbf{G}_2 \end{bmatrix} \begin{bmatrix} \mathbf{c} \\ \mathbf{s} \end{bmatrix} + \begin{bmatrix} \boldsymbol{\eta}_1 \\ \boldsymbol{\eta}_2 \end{bmatrix} \quad (22)$$

In the zero-forcing solution, the interference between the two ST cochannel users is removed, without any regard to noise enhancement, by using a matrix linear combiner \mathbf{W} such that

$$\mathbf{W} \cdot \mathbf{r} = \begin{bmatrix} \tilde{\mathbf{r}}_1 \\ \tilde{\mathbf{r}}_2 \end{bmatrix} = \begin{bmatrix} \tilde{\mathbf{H}} & \mathbf{0} \\ \mathbf{0} & \tilde{\mathbf{G}} \end{bmatrix} \begin{bmatrix} \mathbf{c} \\ \mathbf{s} \end{bmatrix} + \begin{bmatrix} \tilde{\boldsymbol{\eta}}_1 \\ \tilde{\boldsymbol{\eta}}_2 \end{bmatrix} \quad (23)$$

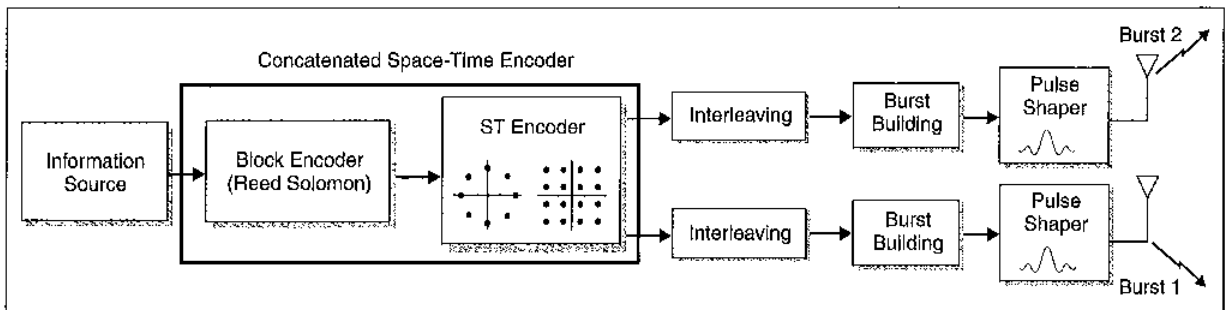
In this case, the modified received signal vector $\tilde{\mathbf{r}}_1$ depends only on signals from first terminal and the modified received signal vector $\tilde{\mathbf{r}}_2$ depends only on signals from second terminal. It was shown in [42] that a solution for \mathbf{W} is given by

$$\mathbf{W} = \begin{bmatrix} \mathbf{I}_2 & -\mathbf{G}_1 \mathbf{G}_2^{-1} \\ -\mathbf{H}_2 \mathbf{H}_1^{-1} & \mathbf{I}_2 \end{bmatrix} \quad (24)$$

It is interesting to note that by using this matrix linear combiner \mathbf{W} , the matrices $\tilde{\mathbf{H}}$ and $\tilde{\mathbf{G}}$ will have the same structure as that of the channel matrix \mathbf{H} in (10). Hence, using the matrix linear combiner in (24) will reduce the problem of detecting the two cochannel ST users into two separate problems that have a much simpler solution as pointed out before. Table 1 shows the algorithm for the zero-forcing interference cancellation and maximum-likelihood decoding of STBC.

In the MMSE interference suppression technique, let us assume, for example, that we are interested in decoding signals from the first ST user. In this case, the receiver selects two linear combiners \mathbf{w}_1 and \mathbf{w}_2 such that

$$J_1(\mathbf{w}_1) = \|\mathbf{w}_1' \mathbf{r} - c_1\|^2 \\ \text{and} \\ J_2(\mathbf{w}_2) = \|\mathbf{w}_2' \mathbf{r} - c_2\|^2 \quad (25)$$



▲ 10. Base station transmitter with STCM and two transmit antennas.

are minimized. It was shown in [42] that the optimum solution is given by

$$\begin{aligned} \mathbf{w}_1 &= \mathbf{M}^{-1} \mathbf{h}_1 \\ \text{and} \\ \mathbf{w}_2 &= \mathbf{M}^{-1} \mathbf{h}_2 \end{aligned} \quad (26)$$

where $\mathbf{M} = \mathbf{H}\mathbf{H}^H + 1/\Gamma\mathbf{I}$, $\Gamma = E_s/N_0$ is the SNR, $\mathbf{h}_1 = [h_{11} \ h_{21} \ h_{12} \ h_{22}]^T$ is the first column of \mathbf{H} , and $\mathbf{h}_2 = [h_{21} \ h_{11} \ h_{22} \ h_{12}]^T$ is the second column of \mathbf{H} . It was shown in [42] that \mathbf{w}_1 and \mathbf{w}_2 are orthogonal, and hence, errors in decoding c_1 do not affect decoding c_2 and visa versa, thereby maintaining the separate detection feature for STBC decoding. Note that the MMSE solution will reduce to the ZF solution outlined earlier as $\Gamma \rightarrow \infty$. Table 2 outlines the algorithm description for MMSE interference suppression and decoding of STBC. For a more detailed treatment of both the ZF and MMSE solutions the reader is referred to [42]. Figure 8 shows the performance of the MMSE interference cancellation scheme as a function of SNR and signal-to-interference ratio (SIR) for two cochannel ST users each using the 2×2 STBC and a receiver with two receive antennas. Note that the performance of the ZF interference cancellation will always be the same as that of a single ST user with one receiver antenna.

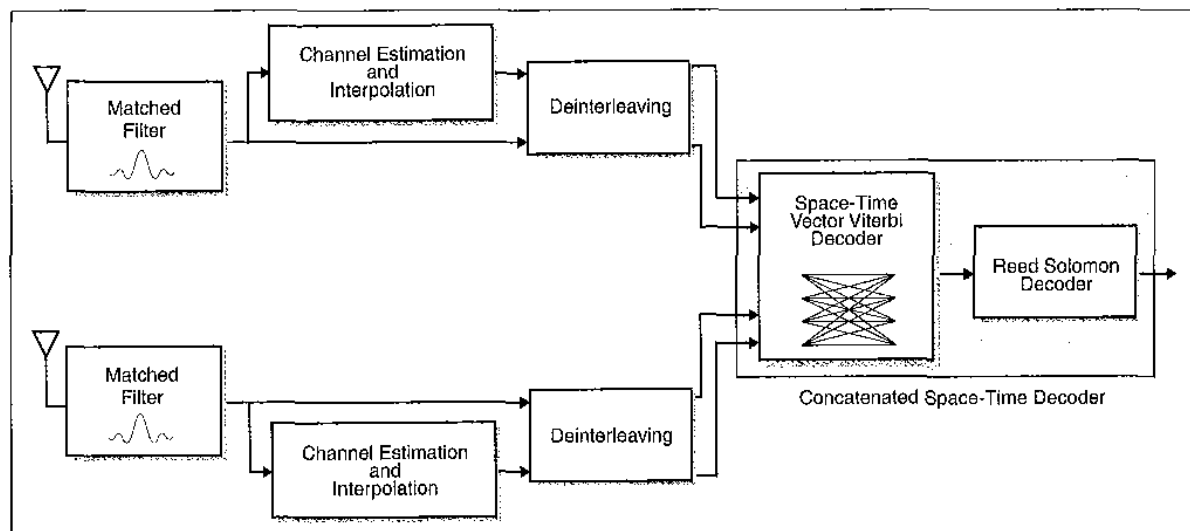
Applications of Space-Time Coding to Wireless

As pointed out earlier, one of the goals of the third- and fourth-generation wireless systems is to provide broadband access to both mobile and stationary users. Real-time multi-media services (such as video conferencing) would require data rates two to three orders of magnitude larger than what is offered by current wireless technologies. A higher spectral efficiency can be achieved by using multiple transmit and/or receive anten-

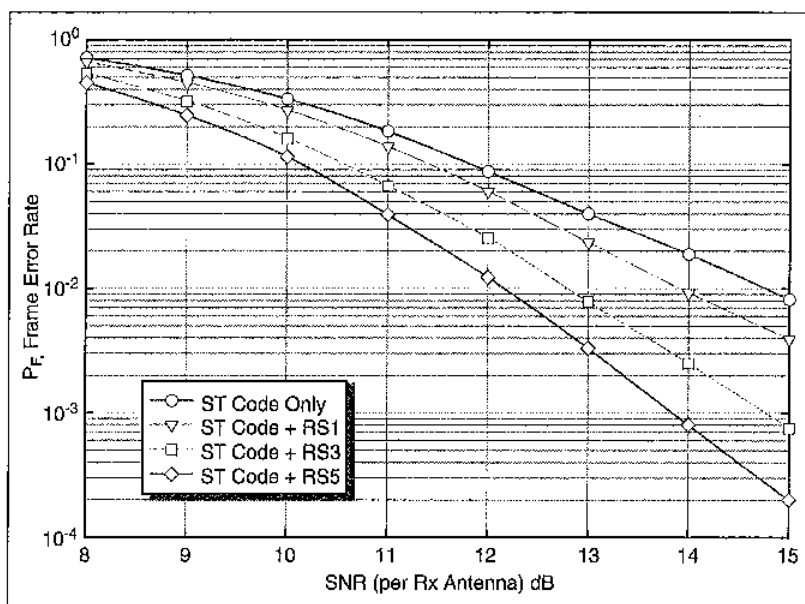
nas [1], [20], [59]. STC techniques with multiple transmit antennas offer the best possible trade-off between power consumption and spectral efficiency in multipath radio channels. STC and signal processing techniques with multiple transmit antennas have been recently adopted in third-generation cellular standard (e.g., CDMA-2000 [60] and W-CDMA [61]) and also have been proposed for wireless local loop applications (Lucent's BLAST project [6]) and wide-area packet data access (AT&T's Advanced Cellular Internet Service [5]). In this section we will outline several examples of application of STC to different wireless applications.

Application to Narrow Band TDMA Cellular

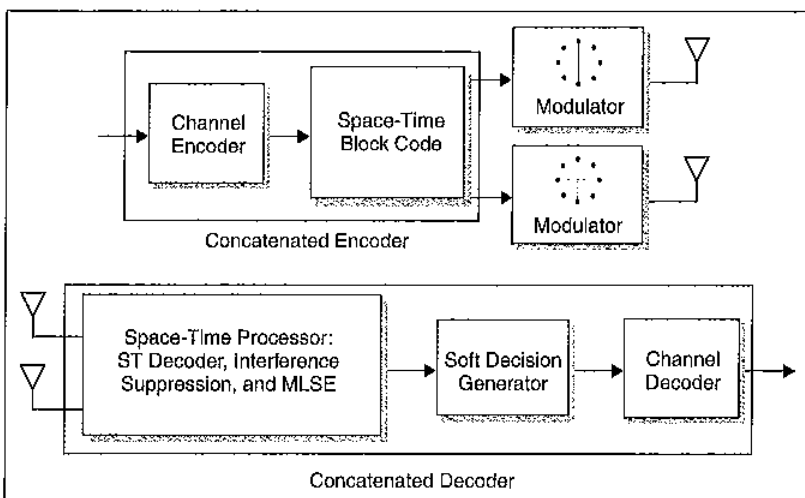
In this section, we will present a general architecture for a narrow band TDMA modem with STC and two transmit antennas [46]. For brevity, we will present the modem architecture for the downlink only. The uplink modem will have a similar architecture, except that the framing and timing structure will be different and must allow for a guard time between different asynchronous (due to difference in propagation delay) bursts from different users. The system architecture that we propose is similar, but not identical, to that of the IS-136 US cellular standard. Figure 9 shows the basic TDMA time slot structure, employing a signaling format which interleaves training and synchronization sequences, pilot sequences, and data is used. In each TDMA slot, two bursts are transmitted, one from each antenna. The training sequences \mathbf{S}_1 and \mathbf{S}_2 will be used for timing and frequency synchronization at the receiver. In addition, the transmitter inserts periodic and orthogonal pilot sequences \mathbf{P}_1 and \mathbf{P}_2 which are used, along with the training sequences \mathbf{S}_1 and \mathbf{S}_2 , at the receiver to estimate the channel from each of the transmit antennas to the corresponding receive antenna. Figure 10 shows a block diagram for the transmitter, where in addition to the ST



▲ 11. Mobile receiver with STC and two receive antennas.



▲ 12. FER Performance of 8-PSK 32-state ST code with two transmit and two receive antennas at 180 Hz Doppler with different coding rates.



▲ 13. Concatenation of STBC and conventional channel coding schemes for increasing capacity.

encoder, a high rate Reed-Solomon (RS) block encoder is used as an outer code. The RS outer code is used to correct the few symbol errors at the output of the ST decoder. The output of the RS encoder is then encoded by an ST channel encoder and the output of the ST encoder is split into two streams of encoded modulation symbols. Each stream of encoded symbols is then independently interleaved using a block symbol-by-symbol interleaver. The transmitter inserts the corresponding training and periodic pilot sequences in each of the two bursts. Each burst is then pulse shaped and transmitted from the corresponding antenna. The signal transmitted from the i th antenna, $i = 1, 2$, can be written as

$$s_i(t) = \sqrt{E_s} \cdot \sum_l c_i(l) p(t - lT_s) \quad (27)$$

where $T_s = 1/R_s$ is the symbol period and $p(t)$ is the transmit filter pulse. Figure 11 shows the corresponding block diagram of a mobile receiver equipped with two receive antennas. After down conversion to baseband, the received signal at each antenna element is filtered using a receive filter with impulse response $\bar{p}(t)$ that is matched to the transmit pulse shape $p(t)$. The output of the matched filters is oversampled at a rate that is an integer multiple of the symbol rate. Received samples corresponding to the training sequences S_1 and S_2 are used for timing and frequency synchronization. The received samples at the optimum sampling instant are then split into two streams. The first one contains the received samples corresponding to the pilot and training symbols. These are used to estimate the corresponding CSI $\hat{H}(l)$ at the pilot and training sequence symbols. The receiver then uses an appropriately designed interpolation filter to interpolate those trained CSI estimates and obtain accurate interpolated CSI estimates for the whole burst. The second stream contains the received samples corresponding to the superimposed information symbols. The interpolated CSI estimates along with the received samples corresponding to the information symbols are then deinterleaved using a block symbol-by-symbol deinterleaver and passed to a vector maximum-likelihood sequence decoder followed by a RS decoder.

Figure 12 shows the performance of the above modem architecture with a 32-state 8-PSK ST code with two transmit and two receive antennas and with different coding rate options (see [46] for details). At 10 Hz Doppler, this modem architecture with the 32-state 8-PSK STC would be able to deliver almost 56 kb/s (over a 30 kHz bandwidth) with 10% frame error rate at 18 dB and 11 dB SNR for 1 and 2 receive antennas, respectively. At 180 Hz Doppler, the required SNR would be 20 dB and 12 dB respectively. These results assume basic IS-136 channelization and framing structure. As pointed out in [46], this architecture has the potential of almost doubling the current data rates supported by the IS-136 cellular standard.

Applications to Increasing Capacity/Throughput of Wireless Systems

First, we consider a scenario where K synchronized terminal units each with two transmit antennas communicate with a base station having $M \geq K$ receive antennas. Increased system capacity (in terms of the number of cochannel terminals that can simultaneously communicate with the base station) can be attained while providing diversity benefits to each terminal by using a concatenated coding scheme where the inner code is an STBC and the outer code is a conventional channel error correcting code (a TCM, a convolutional code, or a RS code, for example), as shown in Fig. 13. More specifically, information symbols are first encoded using a conventional channel code. The output of this channel code is then encoded using an ST block encoder with two transmit antennas (N transmit antennas in general can be used with the appropriate STBC). At the receiver, the inner STBC is used to suppress interference from the other cochannel terminals using, for example, the MMSE interference suppression technique described above. In the above technique, a hard decision is applied on the output of the interference canceler to produce an estimate for the transmitted information symbols. That is, given the two IC weight vectors w_{i1} and w_{i2} corresponding to some terminal i , the receiver forms the two decision variables

$$\xi_{i1} = w_{i1}^H r \quad (28)$$

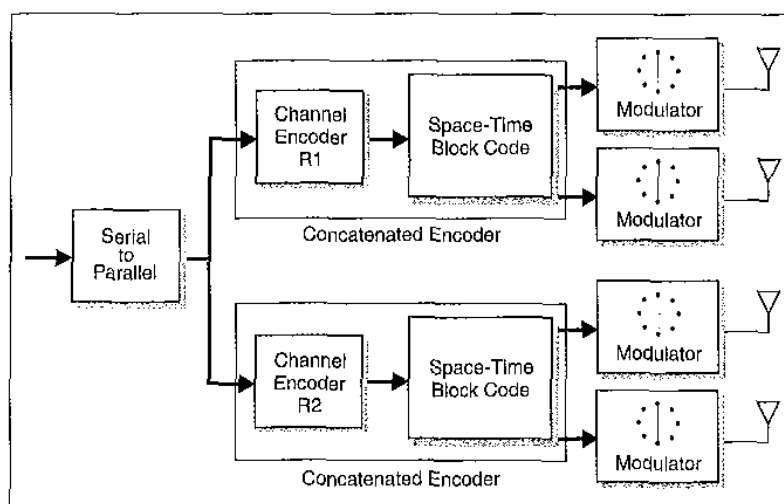
$$\xi_{i2} = w_{i2}^H r. \quad (29)$$

A hard decision is then applied on these decision variables to decode the two transmitted symbols corresponding to the i th terminal. In the case when the space-time code is concatenated with an outer conventional channel code, however, the decision variables ξ_1 and ξ_3 are used as soft decisions for the transmitted information symbols and then fed to the conventional channel decoder. This will improve the error rate performance of the conventional channel code as compared to using hard decision. Thus, in this scheme, we are using the structure of the inner ST code for interference suppression and we are able to support many cochannel terminals while providing diversity benefit to those terminals. At the same time, the inner ST decoder provides soft decision output for the outer con-

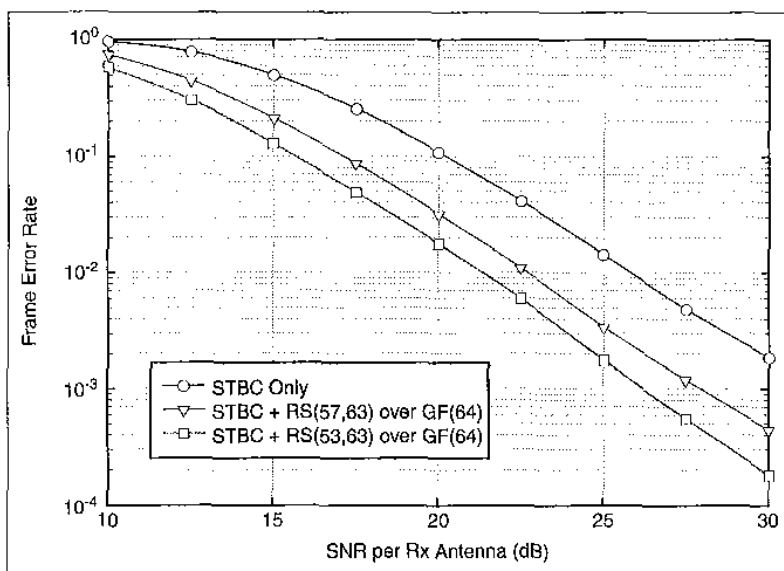
The spatial-temporal structure of space-time codes can be exploited to increase the capacity and/or data rate of wireless systems with a relatively simple receiver structure.

ventional channel code which will provide protection against channel errors.

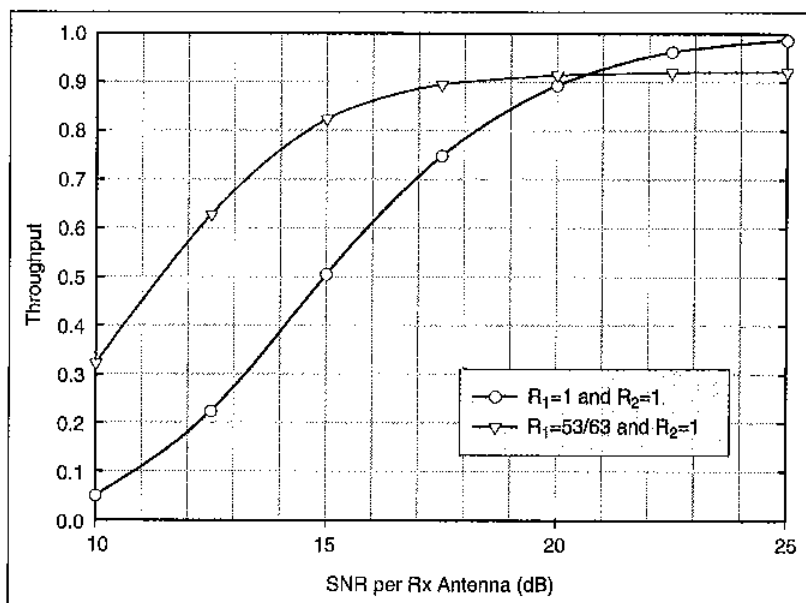
The above ST block coding and MMSE interference suppression technique can also be used in situations



▲ 14. Parallel transmission with ST block coding for increased system throughput.



▲ 15. FER Performance of four cochannel ST users with a concatenated coding scheme over a flat fading channel.



▲ 16. Throughput performance of parallel transmission with ST block coding with unequal coding.

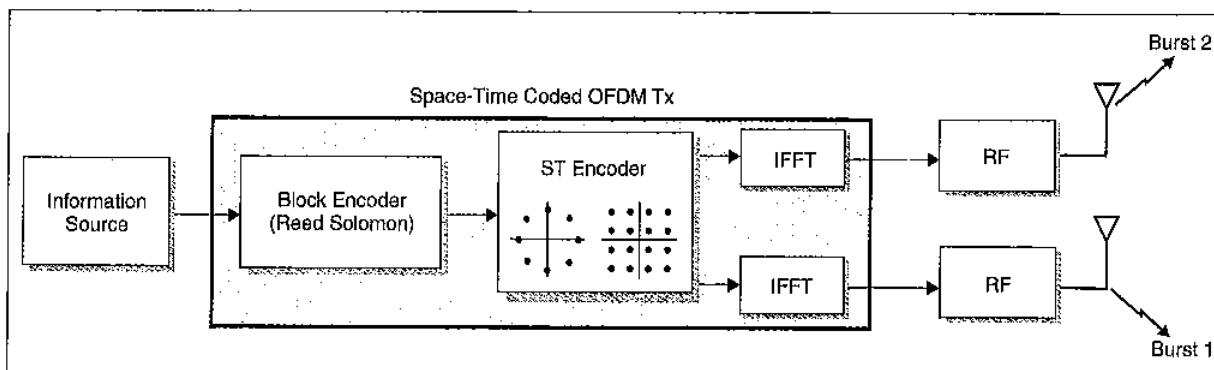
where increasing the data rate or the data throughput is of interest. In this case information symbols from a transmitting terminal are split into L parallel streams. Stream l is then encoded using a conventional channel code with rate R_l and then encoded with an ST block encoder with two transmitting antennas (as before, N transmit antennas in general can be used with the appropriate STBCs). The coding rates for each of the L parallel streams are chosen such that $R_1 < R_2 < \dots < R_L$. In this case, symbols transmitted in stream l will have better immunity against channel errors than to symbols transmitted in stream u where $u > l$. The base station receiver is assumed to be equipped with L receive antennas. The base station receiver treats each stream as a different user and uses the above MMSE interference suppression technique to generate soft decisions ξ_{11} and ξ_{12} for the data in the first stream. These soft decisions are then fed into the decoder corresponding to the first channel code. The output information symbols are then encoded with the same channel code for the first stream. Since the first stream has the

smallest coding rate R_1 , it will have the best immunity against channel errors and most likely it will be error free. The resulting symbols are then used to subtract the contributions of the first stream in the received signal while decoding the remaining $L-1$ streams. In decoding the remaining $L-1$ streams, the decoder will decode signals from the second stream first since it will have the best immunity against channel errors among the remaining $L-1$ streams. Then the receiver cancels out the contribution of the second stream in the received signal. This process is repeated until all streams are decoded. In this case, we define the system throughput as

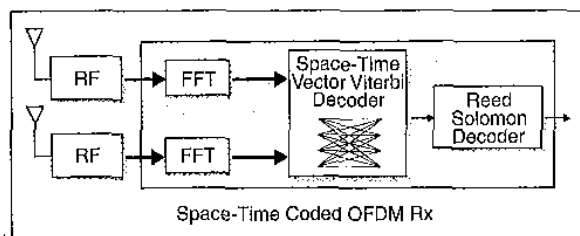
$$\rho = \frac{1}{L} \sum_{l=1}^L R_l \cdot (1 - \text{FER}_l) \quad (30)$$

where FER_l is the frame error rate of stream l . As we will see from the simulation, this will increase the system throughput at low SNRs. Figure 14 shows a block diagram for a terminal that uses four transmit antennas. In this case, the input information stream is split into two parallel streams i.e., ($L=2$).

Figure 15 shows the performance of the system in Fig. 13 where a concatenated coding scheme is used. The figure shows the FER of any of the four users with different coding rates. There was four cochannel users and each uses the 2×2 STBC and the receiver had four antennas. The above MMSE-IC scheme was used to separate the four users. This scheme is suitable for fixed wireless access applications. Figure 16 shows the throughput performance of the system in Fig. 14. Combined MMSE interference cancellation and decoding of the STBC was used to separate the two different data streams. Using this parallel transmission and making use of the STBC properties to separate the two streams will allow for doubling the data rate. Also if one of the two data streams is coded heavier than the other one, increased throughput can be obtained especially at low SNR. See [44] for more details.



▲ 17. Transmitter for ST coded OFDM for broadband applications.



▲ 18. Receiver for ST coded OFDM for broadband applications.

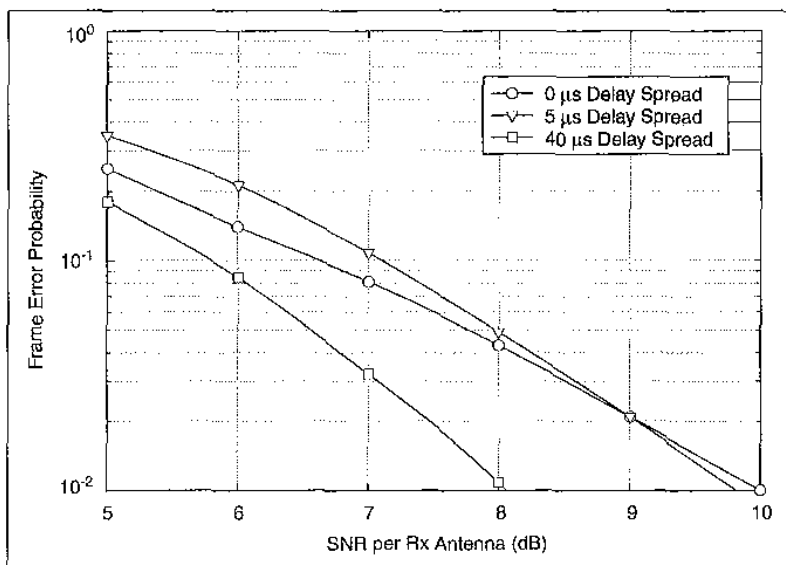
Application to Broadband Wireless

Figures 18 and 17 show simplified block diagrams for the transmitter and receiver, respectively, for an OFDM modem with a concatenated STC scheme. This architecture [62] is suitable for broadband wireless communications applications (similar work, but based on STBCs, can be found in [63] and [64]). The input information symbols are first encoded by an outer conventional channel code. The output of the outer code is then ST encoded. Each stream of the ST code output streams is then OFDM modulated and sent over the corresponding antenna. At the receiver, the signal at each receive antenna is OFDM demodulated. The demodulated signals from antennas are then fed into the ST decoder followed by the outer decoder. Figure 19 shows the simulation results for the above OFDM ST coded modem (STCM). In this simulation, the available bandwidth is 1 MHz and the maximum Doppler frequency is 200 Hz. The number of OFDM tones used for modulation is 256. These correspond to a subcarrier separation of 3.9 KHz and OFDM frame duration of 256 μ s. A cyclic prefix of 40 μ s duration is added to each frame. Each tones modulates a 4-PSK constellation, although higher order M-PSK or M-QAM may be used. We used a

16-state 4-PSK ST code [35] with two transmit and two receive antennas together with an outer (72,64,9) RS code over $GF(2^8)$. We plot the frame error probability as function of SNR for different delay spread. From this plot, we can see that an E_b/N_0 between 2.7-4 dB (depending on the delay spread) is needed to achieve a data rate of 1.5 Mb/s. This technique can be used also with the combined ST block coding and interference suppression scheme, as shown in Fig. 20, to yield even higher data rates (multiples of Mb/s/1 MHz) over a wireless channel.

Conclusions

STC is a new coding/signal processing framework for wireless communication systems with multiple transmit and multiple receive antennas. This new framework has the potential of dramatically improving the capacity and data rates. In addition, this framework presents the best trade-off between spectral efficiency and power consumption. ST codes (designed so far) come in two different types. ST trellis codes offer the maximum possible diversity gain and a coding gain without any sacrifice in the transmission bandwidth. The decoding of these codes, however, would require the use of a vector form of the Viterbi decoder. STBCs offer a much simpler way of obtaining transmit diversity without any sacrifice in bandwidth and without requiring huge decoding complexity. In fact, the structure of the STBCs is such that it allows for very simple signal processing (linear combining) for encoding/decoding, differential encoding/detection, and interference cancellation. This new signal processing framework offered by ST codes can be used to enhance the data rate and/or capacity in various wireless applications. That is the reason many of these STC ideas have already found their way to some of the current third-generation wireless systems standards.



▲ 19. FER of concatenated ST coded OFDM with 4-PSK 16-state STC with 2Tx and 2Rx antennas.

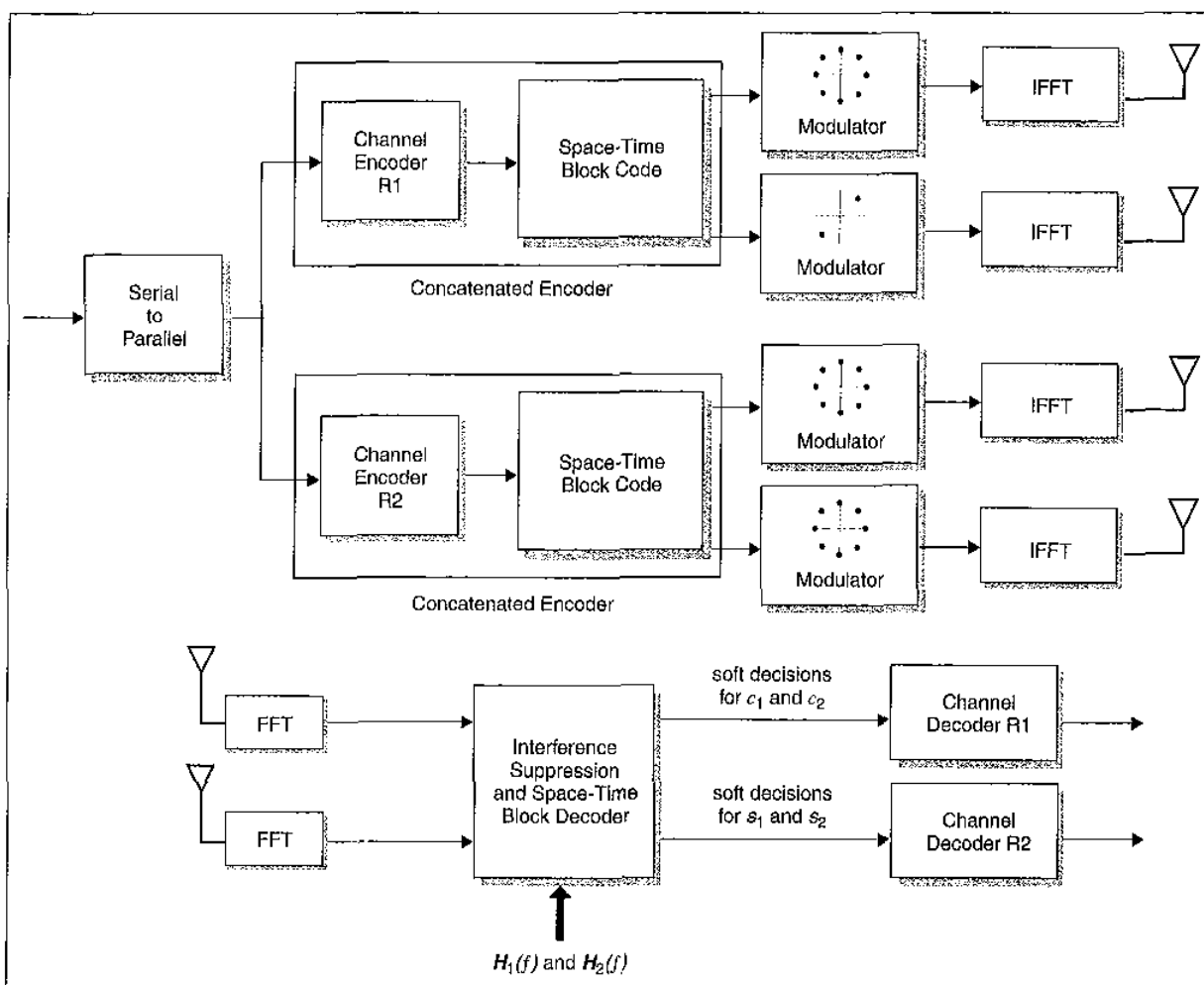
Ayman Naguib (Member) received the B.Sc. degree (with honors) and the M.S. degree in electrical engineering from Cairo University, Cairo, Egypt, in 1987 and 1990, respectively, and the M.S. degree in statistics and the Ph.D. degree in electrical engineering from Stanford University, Stanford, CA, in 1993 and 1996, respectively. From 1987 to 1989, he spent his military service at the Signal Processing Laboratory, The Military Technical College, Cairo, Egypt. From 1989 to 1990, he was employed with Cairo University as a research and teaching assistant in the Communication Theory Group, Department of Electrical Engineering. From 1990 to 1995, he was a Research and Teaching Assistant in the Information Systems Labo-

ratory, Stanford University. In 1996, he joined AT&T Labs, Florham Park, NJ, where he is now a Principal Member of Technical Staff. His current research interests include antenna arrays, signal processing, modulation, and coding for high data rate wireless and digital communications and modem design for broadband systems.

Nambi Seshadri (Fellow) is currently Technical Director for Communication Systems R&D at Broadcom Corporation, Irvine, CA. Prior joining Broadcom, and from 1996-1999, he was the Division Manager for Communications Research Department at AT&T Labs, Research, Florham Park, NJ. There his team was responsible for several innovations in communications and signal compression such as ST codes, incremental redundancy based radio link protocols for wireless data, and multiple description signal compression for video. From 1986-1995 he was with AT&T Bell Laboratories, Murray Hill, NJ. He received his undergraduate education from Regional Engineering College, Tiruchirapalli, India, in 1982, and his graduate education from Rensselaer Polytechnic Institute, Troy, NY, from where he earned his M.S. and

Ph.D. degrees. Along with Vahid Tarokh and Rob Calderbank, he is a recipient of the Best Paper award from the Information Theory Society for the year 2000. He was the Associate Editor for Coding Techniques for the Information Theory Society for the years 1996-1998.

A. Robert Calderbank (Fellow) received the B.S. degree in 1975 from Warwick University, U.K.; the M.S. degree in 1976 from Oxford University, U.K.; and the Ph.D. degree in 1980 from California Institute of Technology, Pasadena, all in mathematics. He joined AT&T Bell Laboratories in 1980, and prior to the split of AT&T and Lucent, he was a Department Head in the Mathematical Sciences Research Center at Murray Hill. He is now Director of the Information Sciences Research Center at AT&T Labs-Research in Florham Park, NJ. His research interests range from algebraic coding theory to wireless data transmission to quantum computing. At the University of Michigan and at Princeton University, he has developed and taught an innovative course on bandwidth-efficient communication. From 1986 to



▲ 20. Parallel transmission with ST block coding for increased system throughput over delay spread channels.

1989, Dr. Calderbank was Associate Editor for Coding Techniques for the *IEEE Transactions on Information Theory*. From 1996 to 1999, he was the Editor-in-Chief of the *IEEE Transactions on Information Theory*. He was also Guest Editor for the Special Issue on of the *IEEE Transactions on Information Theory* dedicated to coding for storage devices. He served on the Board of Governors of the IEEE Information Theory Society from 1990 to 1996. Dr. Calderbank received the 1995 Prize Paper Award from the Information Theory Society for his work on the Z4 linearity of the Kerdock and Preparata codes (jointly with A.R. Hammons, Jr., P.V. Kumar, N.J.A. Sloane and P. Sole). He also received the 1999 Information Theory Society Best Paper Award (jointly with V. Tarokh and N. Seshadri).

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