

Super-Orthogonal Space-Time Turbo Transmit Diversity for CDMA

Daniël J. van Wyk

*RapidMobile (Pty)Ltd, Persequor Park, Pretoria 0020, South Africa
Email: danie@rapidm.co.za*

Louis P. Linde

*Department of Electrical, Electronic and Computer Engineering, University of Pretoria, Pretoria 0002, South Africa
Email: llinde@postino.up.ac.za*

Pieter G. W. van Rooyen

*Broadcom, 15435 Innovation Drive, San Diego, CA 92128, USA
Email: pieter@broadcom.com*

Received 8 September 2003; Revised 30 July 2004

Studies have shown that transmit and receive diversity employing a combination of multiple transmit-receive antennas (given ideal channel state information (CSI) and independent fading between antenna pairs) will potentially yield maximum achievable system capacity. In this paper, the concept of a layered super-orthogonal turbo transmit diversity (SOTTD) for downlink direct-sequence code-division multiple-access (CDMA) systems is explored. This open-loop transmit diversity technique improves the downlink performance by using a small number of antenna elements at the base station and a single antenna at the handset. In the proposed technique, low-rate super-orthogonal code-spread CDMA is married with code-division transmit diversity (CDTD). At the mobile receiver, space-time (ST) RAKE CDTD processing is combined with iterative turbo code-spread decoding to yield large ST gains. The performance of the SOTTD system is compared with single- and multiantenna turbo-coded (TC) CDTD systems evaluated over a frequency-selective Rayleigh fading channel. The evaluation is done both by means of analysis and computer simulations. The performance results illustrate the superior performance of SOTTD compared to TC CDTD systems over practically the complete useful capacity range of CDMA. It is shown that the performance degradation characteristic of TC CDTD at low system loads (due to the inherent TC error floor) is alleviated by the SOTTD system.

Keywords and phrases: transmitter diversity, space-time coding, code-division transmit diversity, layered super-orthogonal turbo transmit diversity, low-rate spreading and coding, CDMA wireless communications.

1. INTRODUCTION

Space-time (ST) processing techniques, such as receive diversity and antenna beamforming, can significantly improve the downlink and uplink capacity of cellular direct-sequence (DS) code-division multiple-access (CDMA) systems. Recent studies have explored the limits of multiple-antenna systems performance in frequency-selective multipath fading environments from an information-theoretic point of view [1, 2]. It has been shown that, with perfect receiver channel state information (CSI) and independent fading between pairs of transmit-receive antennas, maximum system

capacity may potentially be achieved. When multiple receive antennas are not available, multiple transmit antennas have been proven to be an alternative form of spatial diversity that may significantly improve spectral efficiency. Other forms of transmit diversity, such as antenna selection, frequency offset, phase sweeping, and delay diversity, have been studied extensively [3, 4, 5]. Recently, space-time (ST) coding was proposed as an alternative solution for high data rate transmission in wireless communication systems [6, 7, 8, 9, 10].

Depending on whether feedback information is utilized or not, transmit diversity schemes are usually categorized as being either closed- or open-loop methods. In closed-loop schemes, CSI estimated by the receiver is fed back to the transmitter, allowing for a number of different techniques to be considered. These techniques, such as beamforming,

adaptive antenna prefiltering, or antenna switching, are used to maximize the signal-to-noise ratio (SNR) at the receiver [11, 12]. When no feedback information is available, the temporal properties of the propagation environment and the transmission protocol can be used to improve the receiver's performance. Techniques utilizing these kinds of properties are commonly referred to as open-loop methods.

Foschini [2] has considered an open-loop layered space-time (ST) architecture with the potential to achieve a significant increase in capacity compared to single-channel systems. The spectrally efficient layered ST transmission process basically comprises the demultiplexing of a single primitive input data stream into n multiple equal-rate data streams. The n separately coded, chip-symbol-shaped and modulated data streams then individually drive separate multiple transmit antennas elements prior to radiation. A multiple-transmit multiple-receive ($M_T = n, M_R = n$)-antenna analysis (where M_T and M_R , respectively, denote the number of transmitter and receiver antenna elements) showed that the system capacity increased linearly with n , despite the random interference of the n received waves. With $n = 8$, an 1% outage probability, and 21 dB average SNR at each receiving antenna element, a spectral efficiency of 42 bps/Hz was shown to be achievable [2]. This implies a capacity increase of 40 times that of a $(M_T, M_R) = (1, 1)$ system at the same total radiated transmitter power and bandwidth. The layered ST concept basically relates to the exploitation of all available spatial and temporal dimensions provided by the layered combination of multielement transmit and/or receive antenna arrays and a vast range of available one-dimensional coding techniques to achieve maximum diversity gain through iterative processing at the receiver. For a detailed description and some illustrative examples of the layered ST architecture employing convolutional coding, as opposed to parallel concatenated iterative super-orthogonal turbo coding on each ST branch proposed in this paper, the interested reader is referred to references [2, 13].

This layered ST architecture forms the basis for the class of orthogonal decomposable coded ST codes presented in this paper. The Alamouti ST block codes are members of this class of codes [3, 6]. The condition of statistically independent (uncorrelated) fading, to maintain orthogonality, is seldom achieved in practice due to the scattering environment around the mobile and base station. However, decomposition or separation of the multiantenna channel into a number of nearly independent subchannels can be realized, provided that CSI is available at the receiver [2, 12]. Maximizing the free distance of the ST coded symbols transmitted over these nearly independent spatially separated channels, a spatial-temporal coding diversity gain can be achieved, referred to as space-time gain (STG).

DS-CDMA systems exhibit maximum capacity potential when combined with forward error correction (FEC) coding [14]. In CDMA, the positive tradeoff between greater distance properties of lower rate codes and increased cross-correlation effects (due to shorter sequence length) is funda-

mental to the success of coded CDMA. Most FEC systems, especially those with low code rates, expand bandwidth and can be viewed as spreading systems. It has been illustrated that the maximum theoretical CDMA capacity can only be achieved by employing very low-rate FEC codes utilizing the entire bandwidth, without further spreading by the multiple-access sequence [14, 15, 16]. These are known as code-spread CDMA systems.

Viterbi [17] has proposed the use of orthogonal convolutional codes as low-rate coding extensions for code-spread CDMA. Recently, two new classes of low-rate codes with improved performance have been proposed. Pehkonen and Komulainen [18, 19] proposed a coding scheme that combines super-orthogonal turbo codes (SOTC) with super-orthogonal convolutional codes (SOCC) [17]. A different approach was taken by Frenger et al. [15, 16], where a class of nested rate-compatible convolutional codes (RCCC), with maximum free distance (MFD), was derived and applied to code-spread CDMA.

For nonoptimum multiuser receivers, such as the matched filter (MF) or RAKE, coding gain comes at the cost of an increased multiple-access interference (MAI) level. Note that as the spreading factor (SF) decreases, so does the potential number of users that can be accommodated, due to the smaller spreading sequence family size available. In such a case, the Gaussian approximation of the MAI does not apply, as the central limit theorem does not hold any more. However, when transmit diversity is considered, this situation (from a coding perspective) is improved due to the introduction of additional MAI as a result of the multiple transmission paths that are created through the application of the multiple transmit antenna diversity concept. Especially, when turbo coding is considered, the coding gain potential becomes significant. For a finite effective code rate (and hence a finite spreading ratio), the level of MAI, under additive white Gaussian noise (AWGN) and equal power conditions, is fixed. For a RAKE receiver with perfect CSI, the soft-input soft-output (SISO) turbo-based decoder will perform equally well in AWGN and fully interleaved multipath fading channels.

In this paper, a layered ST super-orthogonal turbo transmit diversity (SOTTD) architecture for a downlink DS-CDMA system, operating over a frequency-selective fading channel, is investigated. This open-loop transmit diversity technique is well-suited for code-spread CDMA systems where downlink performance is improved by using a small number of transmit antennas ($M_T = 3$) at the base station and a single antenna ($M_R = 1$) at the mobile handset receiver. In the proposed technique, low-rate super-orthogonal code-spread CDMA is married with code-division transmit diversity (CDTD), and at the mobile receiver, ST RAKE CDTD processing is combined with iterative turbo code-spread decoding. In Section 2, the description of the SOTTD code-spread CDMA system is presented. In Section 3, the performance of the SOTTD system is compared with single- and multiantenna turbo-coded (TC) CDTD systems. The evaluation is done by both analysis and computer simulations. Section 4 concludes the paper.

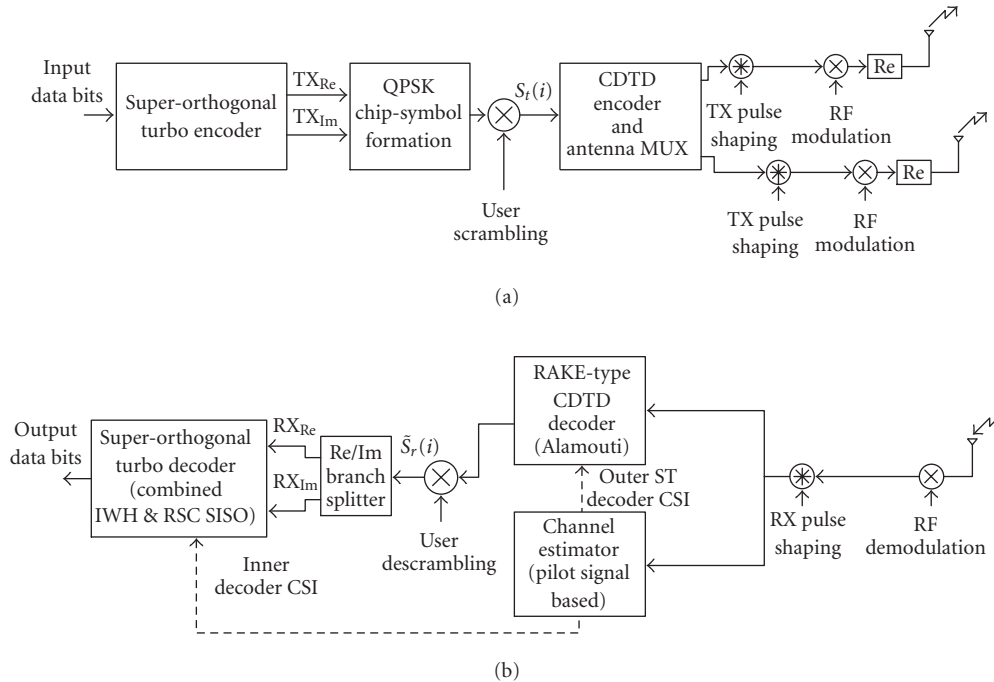


FIGURE 1: SOTTD system block diagrams: (a) transmitter, $M_T = 2$; (b) receiver, $M_R = 1$.

2. SYSTEM DESCRIPTION

2.1. Transmitter and receiver

A downlink (base station to mobile handset) dual transmit, $M_T = 2$, and single receive antenna, $M_R = 1$, multiuser DS-CDMA-based communication system with K simultaneous users is considered. The general structures of the DS-CDMA transmitter and receiver under investigation are illustrated in Figure 1.

With reference to Figure 1a, the transmitter consists of the following modules:

- (1) super-orthogonal turbo encoder producing complex code-spread sequences for CDMA;
- (2) Gray-coded quadrature-phase-shift-keyed (QPSK) chip-symbol formation;
- (3) user-specific scrambling of QPSK chip-symbols, for example, using a IS-95-like long pseudonoise (PN) scrambling sequence;
- (4) code-division transmit diversity (CDTD) encoder based on Alamouti ST block encoder and antenna multiplexer [3];
- (5) transmitter chains for M_T transmit antennas, each comprising chip-pulse shaping, RF modulation, and antenna transmission of RF-modulated (real-part only) signals for the M_T transmit antennas.

With reference to Figure 1b, the receiver consists of the following blocks:

- (1) receiver chain for $M_R = 1$ receive antenna, comprising RF demodulation and chip-pulse shaping;

- (2) channel estimation providing the fading coefficients for each of the transmitter antennas through the transmission of known pilot signals;
- (3) RAKE-type ST receiver based on the Alamouti ST block decoder and maximal-ratio combining (MRC);
- (4) user-specific descrambling of QPSK chip-symbols;
- (5) splitting of the spread-coded chip sequence into real and imaginary components;
- (6) super-orthogonal SISO-based turbo decoder.

In the following paragraphs, details concerning the super-orthogonal turbo encoder and decoder, as well as the ST RAKE CDTD decoder, will be given.

2.2. Super-orthogonal turbo encoder description

The detailed structure of the super-orthogonal turbo encoder is shown in Figure 2. The heart of the encoding scheme is formed by the $Z = 2$ rate-(1/16) constituent encoders, consisting of the combination of a rate-(1/4) recursive systematic convolutional (RSC) encoder, a rate-(4/16) Walsh-Hadamard (WH) encoder, parallel-to-serial (P/S) converter, and puncturing modules. A definition and description of the iterative generation of WH codes, together with their correlation properties, are given in Proakis [20, Chapter 8, pages 424–425]. The combined encoder is referred to as the super-orthogonal RSC&WH encoder. These encoders are concatenated in parallel. A binary data sequence of length N is fed into the encoder. The first encoder processes the original data sequence, whereas before passing through the second encoder, the data sequence is permuted by a pseudorandom interleaver of length N . The outputs of the rate-(1/4)

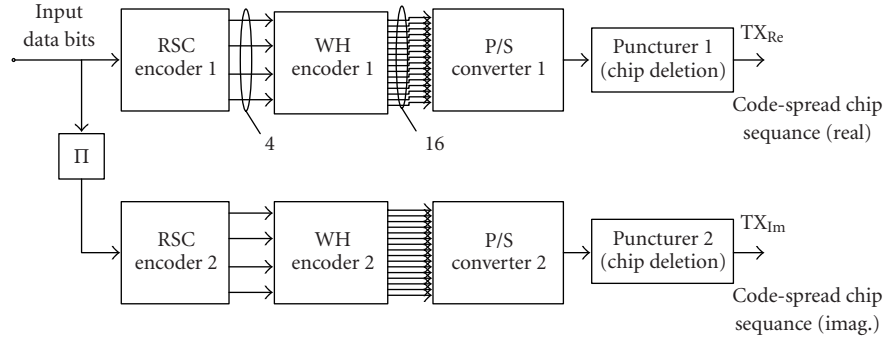


FIGURE 2: Super-orthogonal turbo encoder for $Z = 2$ constituent RSC&WH encoders.

RSC encoder is fed to the rate-(4/16) WH encoder, producing a sequence of length $L_{WH} = 16$ from a set of 16 sequences. By combining the constituent encoder outputs, the code rate from the turbo encoder before puncturing is $R_c = 1/(ZL_{WH}) = 1/32$.

Figure 3 depicts the rate-(1/4), 8-state RSC encoder block diagram and associated trellis diagram. The trellis diagram is important in the evaluation of code distance properties and for Viterbi decoding.

As a last stage of encoding, after P/S conversion, the outputs of the two constituent RSC&WH encoders are punctured to produce the code-spread chip sequences (TX_{RE} , TX_{IM}). The puncturing (chip deletion) operation can be seen as a form of rate matching to provide a wide range of spread-code rates. Note that the final code rate of the super-orthogonal turbo encoder determines the code-spread factor, G , where $G \leq 1/R_c$, in general. In the case of no puncturing, $G = 1/R_c = 2L_{WH} = 32$.

The complex chip output sequences of the super-orthogonal turbo encoder is Gray-mapped into a QPSK symbol constellation. The in-phase (I) and quadrature (Q) QPSK chip-symbol sequences are complex-scrambled with a user-specific IS-95-like long complex pseudonoise (PN) scrambling sequence. The complex result of this complex scrambling process, $S_t(i)$, is fed to a code-division transmit diversity (CDTD) block encoder based on the Alamouti ST block encoder and antenna multiplexer [3, 6].

The CDTD encoder in Figure 1a maps two symbols into an orthogonalising (2×2) code matrix according to

$$D_{M_T} = \begin{bmatrix} s_t(2i-1) & s_t(2i) \\ -s_t^*(2i) & s_t^*(2i-1) \end{bmatrix}, \quad (1)$$

where $i = 1, 2, \dots$. The symbol $s_t(n)$ denotes the transmitted QPSK chip-symbol for time instant n .

Finally, the real part of the complex transmit pulse-shaped and RF-modulated outputs of the CDTD encoder are radiated from multiple transmit antennas, as shown in Figure 1a. In this way, low-rate super-orthogonal code-spread CDMA and an open-loop code-division transmit diversity (CDTD) technique have been combined to potentially facilitate significantly improved downlink performance through appropriate iterative ST receiver and decoder processing.

2.3. Space-time RAKE CDTD receiver/decoder description

Figure 4 shows the general architecture of the RAKE-type CDTD ST receiver for the SOTTD system.

It has been shown that the conditions where ST decoding yields significant diversity gains are independent of those conditions that are favorable for a RAKE-type receiver [21]. In other words, ST diversity is not adversely affected by sub-optimal multipath diversity gain. Under the best conceivable conditions, the multipath components have equal expected power and arrive such that the delayed spreading codes are perfectly orthogonal. Then an L_R -finger RAKE receiver, where $L_R = J$ denotes the number of resolvable paths, would be equivalent to having J receiver antennas, and both the diversity and the expected SNR would theoretically be increased by the factor J [12, 21].

In order to maximize multipath diversity gain, the following assumptions are made.

- (1) The J paths from antenna m experience independent Rayleigh fading, expressed through the channel coefficients, h_{jm} , $j = 1, 2, \dots, J$ and $m = 1, 2$.
- (2) Each pair of paths from the two transmitter antennas arrives with the same set of delays at the receiver antenna. (This assumption is justified by the fact that in the cellular personal communication frequency bands, the propagation delay between the two transmitter antenna elements is measured in nanoseconds, while the multipath delays are measured in microseconds [12]).
- (3) Path delays are approximately a few chips in duration and small compared with the symbol period so that intersymbol interference can be neglected.

Multipath RAKE and ST decoding is performed on knowledge of the multipath delays and fading coefficients for each of the M_T transmitter antennas and J possible multipaths. This information is provided to the mobile receiver by the channel estimator block.

The channel estimator operates on the principle of pilot signals transmission. Increasing the number of transmitter antennas tends to give greater diversity gains, but if the total pilot power is fixed, the individual estimates for the fading coefficients deteriorate, and crosstalk increases among

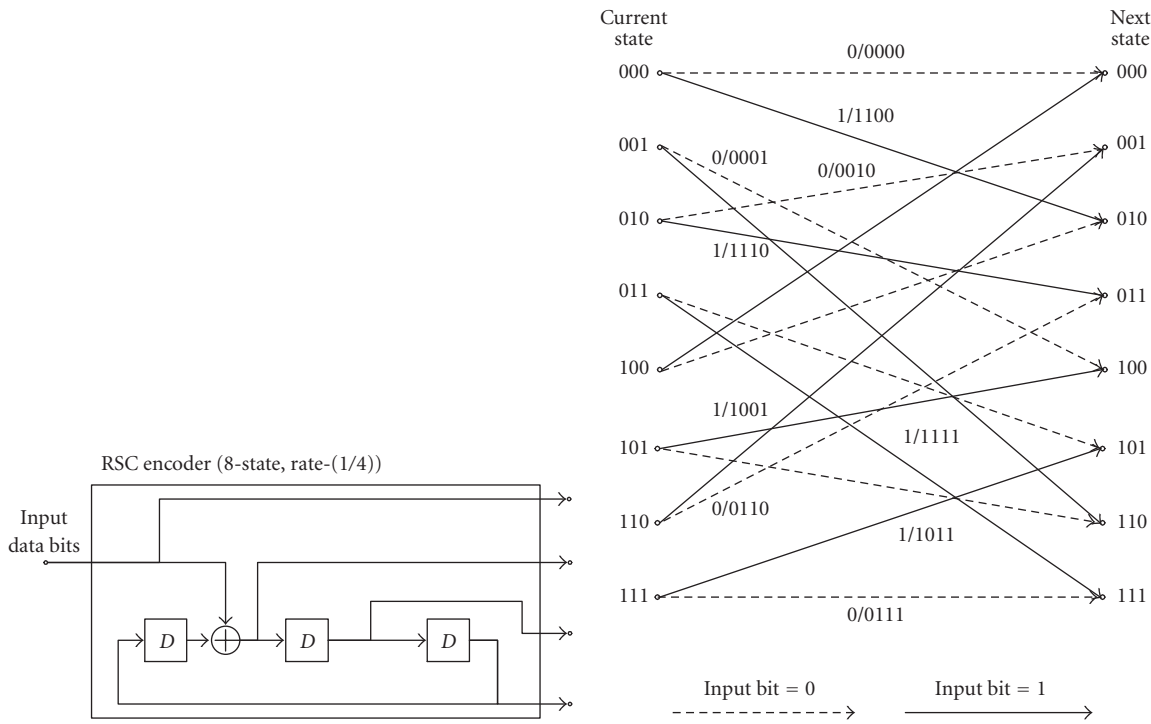


FIGURE 3: Constituent RSC encoder: (a) encoder's block diagram; (b) encoder's trellis diagram.

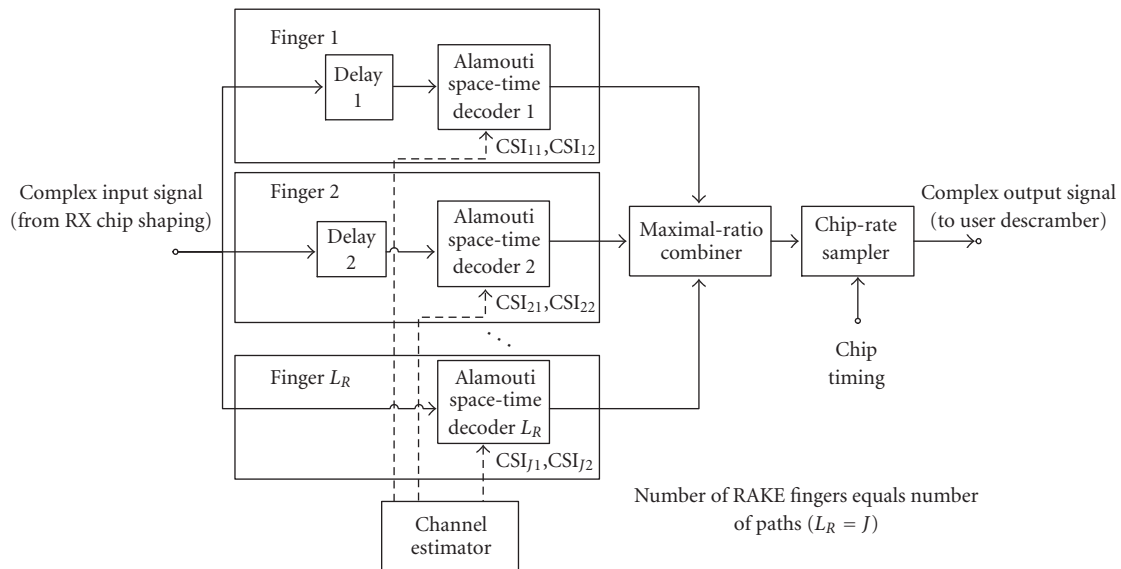


FIGURE 4: RAKE-type CDTD space-time (ST) receiver based on Alamouti ST block decoder.

the subchannels. Adding extra antennas requires the incorporation of additional pilot signals to enable the mobiles to accurately estimate the multiple-antenna propagation coefficients. As a rule of thumb, the individual powers of these pilot signals should be inversely proportional to the number of transmit antennas. In this paper, perfect CSI is assumed, and the channel estimation error-related RAKE ST receiver problems are not treated here.

From Figure 4, it can be seen that paths $j = 1, 2, \dots, J - 1$ are delayed before ST decoding is attempted. This path delay should be equal to the time-of-arrival difference between path j and the last path J , and is done to synchronize individual path powers for maximal-ratio combining (MRC).

The ST decoder shown in Figure 4 is based on the Alamouti ST length-two block encoder and decoder [3, 6]. Recall that the encoder mapped two symbols into a (2×2) code

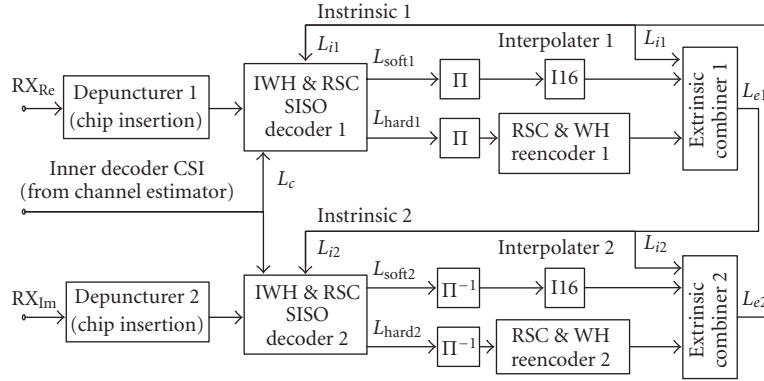


FIGURE 5: Super-orthogonal turbo decoder.

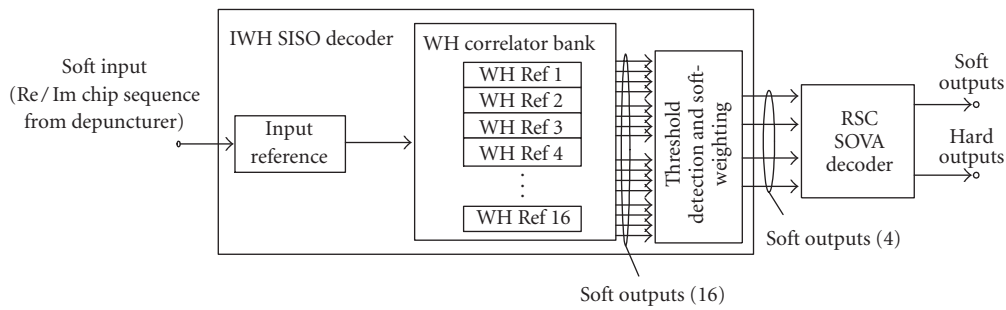


FIGURE 6: Combined inverse Walsh-Hadamard (IWH) and recursive systematic convolutional (RSC) soft-input soft-output (SISO) decoder.

matrix according to (1). Since the symbols are also orthogonal across antennas, the soft-input block decoder simply calculates:

$$\begin{aligned} s_r(2i-1) &= h_{j_1}^* r(2i-1) + h_{j_2} r^*(2i), \\ s_r(2i) &= h_{j_2}^* r(2i-1) + h_{j_1} r^*(2i). \end{aligned} \quad (2)$$

In (2), $r(n)$ denotes the received chip symbols. $s_r(n)$ is the block-decoder soft output for time instant n that determines to which quadrant in the QPSK constellation the chip symbols most likely belong. The likelihood (or confidence level) of this determination is the soft output passed on to the channel decoder after MRC.

Given perfect multipath and diversity gain, the RAKE-type ST decoder has a combined multipath and ST diversity gain of $L_R M_T$, where $L_R = J$ denotes the number of received signal paths (which are here assumed to be equal to the number of fingers employed in the RAKE receiver structure) and M_T denotes the number of transmit antennas.

2.4. Super-orthogonal turbo decoder description

Figure 5 shows the general architecture for the super-orthogonal iterative turbo decoding strategy.

Before the actual decoding takes place, for those chips that were punctured (deleted), zero values are inserted. Therefore, the decoder regards the punctured chips as erasures. The iterative decoding of the turbo coding scheme re-

quires two component decoders using soft code-spread chip inputs and providing soft outputs. Two SISO decoders are employed in the component decoders as shown in Figure 6. The first is a SISO inverse Walsh-Hadamard (IWH) decoder and the second a SISO RSC decoder, based on the soft-output Viterbi algorithm (SOVA). Details concerning the actual decoding process will now be given, with reference to Figure 5.

Let RX_{Re} and RX_{Im} be the associated received and demodulated branch code-spread chip sequences with L_c the corresponding reliability values of the CSI. The decoder accepts a priori values $L_i(b)$ for all the information bit sequences and soft-channel outputs $L_c \cdot RX_{Re}$ and $L_c \cdot RX_{Im}$. In the IWH SISO decoder, the branch metric calculation is performed very efficiently by using soft-outputs based on the IWH transformation, which basically correlates the received soft chip-spread sequences, RX_{Re} and RX_{Im} , with the branch WH sequences. The soft outputs from the IWH SISO decoder are passed to the RSC SOVA, which produces hard (L_{hard}) and soft (L_{soft}) outputs. Without loss of generality, the indices, $z = 1, 2$, denoting the constituent component decoders (shown in Figure 5), have been omitted in this discussion.

The IWH&RSC SISO component decoders delivers a posteriori soft outputs $L(\hat{b})$ for all the information bits and extrinsic information $L_e(b)$. The latter is only determined for the current bit by its surrounding bits and the code constraints. It is therefore independent of the intrinsic information and the soft output values of the current bit.

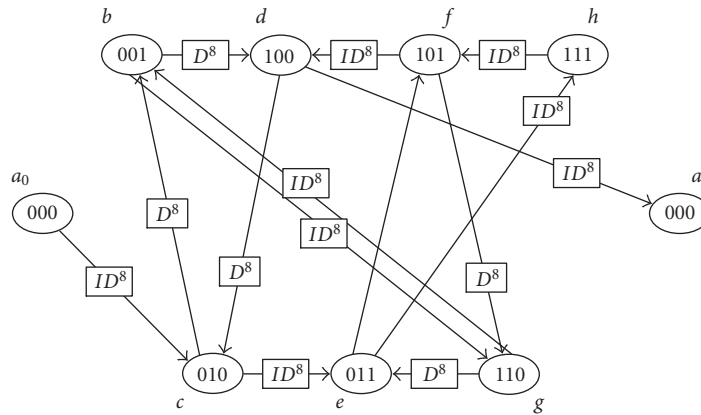


FIGURE 7: State diagram of combined RSC&WH constituent encoder. (Note that the state transitions are determined by RSC encoder (shown in Figure 3), while output-word Hamming distances are determined by the WH encoder.)

The extrinsic information is given by

$$L(\hat{b}) = [L_{\text{hard}} \otimes f(L_{\text{hard}}) + g(L_{\text{soft}})] - L_i(b), \quad (3)$$

where the first term, $L_{\text{hard}} \otimes f(L_{\text{hard}})$, is the reencoded chip code-spread sequence, with $f(L_{\text{hard}})$ being the function denoting the combined RSC and WH reencoding process. The symbol “ \otimes ” denotes convolution. The second term, $g(L_{\text{soft}})$, represents the interpolated soft outputs from the component SISO decoders, with interpolation factor $L_{\text{WH}} = 16$. It is important to note that all the above-mentioned sequences are vectors of length $L_{\text{WH}} = 16$.

The log-likelihood ratio (LLR) soft output of the decoder for the information bit b is written as

$$L(\hat{b}) = [L_c(\text{RX}_{\text{Re}} + \text{RX}_{\text{Im}}) + L_i(b)] + L_e(\hat{b}) \quad (4)$$

implying that there are three independent estimates that determine the LLR of the information bits, namely, the a priori values, $L_i(b)$, the soft-channel outputs of the received sequences, $L_c \cdot \text{RX}_{\text{Re}}$ and $L_c \cdot \text{RX}_{\text{Im}}$, and the extrinsic LLR's $L_e(\hat{b})$.

At the commencement of the iterative decoding process, there usually are no a priori values $L_i(b)$; hence the only available inputs to the first decoder are the soft-channel outputs obtained during the actual decoding process. After the first decoding process, the intrinsic information on b is used as independent a priori information at the second decoder. The second decoder delivers a posteriori information, which is an output produced by the first decoder too. Note that initially the LLRs are statistically independent. However, since the decoders directly use the same information, the improvement through the iterative process becomes marginal, as the LLRs become progressively more correlated.

It is important to note that the constituent RSC&WH encoders may produce similar WH codewords. Since these codewords are transmitted over different antennas, the full-rank characteristic of the system is still guaranteed. Under multipath fading scenarios, some of the orthogonality will be destroyed. The latter is not a function of the specific WH codeword transmitted at the different antennas, but rather

dependent on the delay spread of the channel. Transmitting the same WH codewords over different antennas will have an effect on the channel estimation and initial synchronisation procedures.

3. PERFORMANCE EVALUATION

3.1. Union-bound BEP derivation of combined RSC and WH code

One of the objectives of this section is to shed some light on the contribution of the parallel-concatenated WH codes to the overall SOTTD systems performance. Towards this end, an upper bound is derived for the average bit error probability (BEP) performance of parallel-concatenated WH codes, stemming from the characteristics of the combined RSC&WH code.

The performance of the SOTTD system depends not on the distance properties of the WH code, but actually on the distance properties of the combined RSC&WH code. In this context, the most important single measure of the code's ability to combat interference is d_{min} . Figure 7 depicts the modified state diagram of the RSC&WH constituent code under consideration. The state diagram provides an effective tool for determining the transfer function, $T(L, I, D)$, and consequently, d_{min} of the code. The exponent of D on a branch describes the Hamming weight of the encoder corresponding to that branch. The exponent of I describes the Hamming weight of the corresponding input word. L denotes the length of the specific path.

Through visual inspection, the minimum distance path, of length $L = 4$, can be identified as $a_0 \rightarrow c \rightarrow b \rightarrow d \rightarrow a_1$. This path has a minimum distance of $d_{\text{min}} = 4 \times 8 = 32$ from the all-zero path, and differs from the all-zero path in 2 bit inputs.

Given an $(32, 1)$ RSC&WH constituent code, its input-redundancy weight enumerating function (IRWEF) is used to characterize the complete encoder [22]. The IRWEF makes implicit in each term of the normal weight enumerating function the separate contributions of the information and

the parity-check bits to the total Hamming weight of the codewords. When the contributions of the information and redundant bits to the total codeword weight are separated, the IRWEF for the constituent RSC&WH code is obtained as

$$A(I, D) = 1 + 4ID^7 + 6I^2D^2 + 4I^3D^5 + I^4D^4. \quad (5)$$

When employing a turbo interleaver of length N , the IRWEF of the new constituent $(n, k) = (32N, N)$ code is given by $A^N(I, D) = [A(I, D)]^N$, for all Z constituent codes (see [22, page 157, equation (5)] for a similar approach), where n denotes the code length and k the number of encoded data symbols in the code word.

To compute an upper bound to the BEP, the IRWEF can be used with the union bound assuming maximum-likelihood (ML) soft decoding. The BEP, including the fading statistics (assumed to be slowly fading), is of the form shown in (6) [14, 21], where σ_{oc} denotes the effective SNR, and S denotes the power of the received signal:

$$P_{b|S} \leq \frac{1}{k} Q\left(\sqrt{d_{\min} \sigma_{oc} S}\right) \cdot e^{d_{\min} \sigma_{oc} S} \cdot \left. \frac{\partial A^N(I, D)}{\partial I} \right|_{I=D=e^{-\sigma_{oc} S}}. \quad (6)$$

On an AWGN channel, the total effective output SNR term used in (6) is $\sigma_{oc} = R_c E_b / N_0$. Assuming that the cellular system is employing omnidirectional antennas, the total output SNR term used in (6) can be determined as in (7) [11, 21]:

$$\sigma_{oc} = \left(\frac{1}{R_c} \frac{N_0}{2E_b} + \frac{(K \cdot M_T - 1)}{3G} \right)^{-1}. \quad (7)$$

Recall that K denotes the number of simultaneous users, G is the code-spread ratio, and E_b/N_0 is the energy-per-bit-to-noise spectral density ratio. The CDMA normalized system load is given as K/G . Also, if it is assumed that the M_T transmitters have equal power, with constant correlation between the branches, and transmitted over a Rayleigh fading channel, the components of the received power vector \mathbf{S} are identically distributed, with probability density function (pdf) given by (8), with $\zeta = 1 - \rho + \rho M_T L_R$ (see [21, Sections 6.3.2 to 6.3.4, pages 93–98]):

$$\begin{aligned} p_S(S) &= \frac{1}{\Omega^2 \Gamma(M_T L_R)} \left(\frac{S}{\Omega^2} \right)^{M_T L_R - 1} \\ &\times \frac{\exp(S/(1-\rho)\Omega^2) \cdot {}_1F_1(1, M_T L_R, \rho M_T L_R S / \zeta(1-\rho)\Omega^2)}{\zeta(1-\rho)^{(M_T L_R - 1)}}. \end{aligned} \quad (8)$$

In the above equation, ${}_1F_1(\cdot)$ denotes the confluent hypergeometric function, Ω^2 is the average received path strength, ρ the correlation between transmit or receive branches, and L_R the number of RAKE receiver fingers.

Finally, the BEP is computed using (6) and (7), by averaging (6) over the fading statistics defined in (8).

TABLE 1: System parameters for analytical and simulation BEP performance analysis.

Parameter	Simulation value
Spreading ratio	$G = 32$
Operating environment	2-path frequency-selective fading
Number of users	$K = 1, 2, \dots, G$
Number of RAKE fingers	$L_R = J = 2$
Transmit diversity technique	CDTD and SOTTD
Transmit diversity elements	$M_T = 1, 2$ ($\rho = 0$)
Interleaver length	$N = 256$

3.2. Numerical analysis of CDTD and SOTTD CDMA systems

The performance of the proposed super-orthogonal transmit diversity (SOTTD) CDMA system is compared to that of an uncoded, as well as convolutional- and turbo-coded code-division transmit diversity (CC and TC CDTD) CDMA systems. In order to calculate the BEP of the coded CDTD and SOTTD systems, the output SNR should include the transmit diversity interference term as shown in (7).

Using the system parameters outlined in Table 1, the BEP performance of a cellular CDMA system employing the different techniques has been determined numerically. The performance of single and $M_T = 2, 3$ transmit diversity systems are shown in Figure 8. From the curves, it is clear that the superior performance predicted for TC CDTD may be achieved with the SOTTD system over the complete CDMA capacity range. Also of importance is the fact that the performance degradation of TC CDTD at low system loads (due to inherent TC error floor) is alleviated by the SOTC system—hence the superior performance of SOTTD. This is explained in terms of the higher minimum free distance offered by the rate-(1/16) constituent encoders, as opposed to the use of rate-(1/2) constituent encoders in TC systems.

3.3. Simulation results

Monte-Carlo simulations were conducted to verify the BEP bounds presented above. In the computer simulations, a root-raised cosine (RRC) chip-pulse shaping with roll-off factor of $\alpha = 0.22$ was used. The length of the pulse-shaping filter was set to 8 chips, and 4 samples per chip were taken. A single receiver antenna and $J = 2$ resolvable Rayleigh fading multipaths with equal average power were assumed.

For the simulation, perfect synchronization, coherent detection, and perfect channel state information (CSI) estimation are assumed. The simulated fading channel assumed a flat Doppler power spectrum. A mobile velocity of 3 km/h was selected (corresponding to slow fading), producing nearly static fading over the frame (and interleaver) length of $N = 256$ information bits used in the simulations. The individual path gains are assumed nearly constant (quasistatic) during one frame and change independently from one another. The multipath spread was randomized and evenly distributed with a minimum resolution of

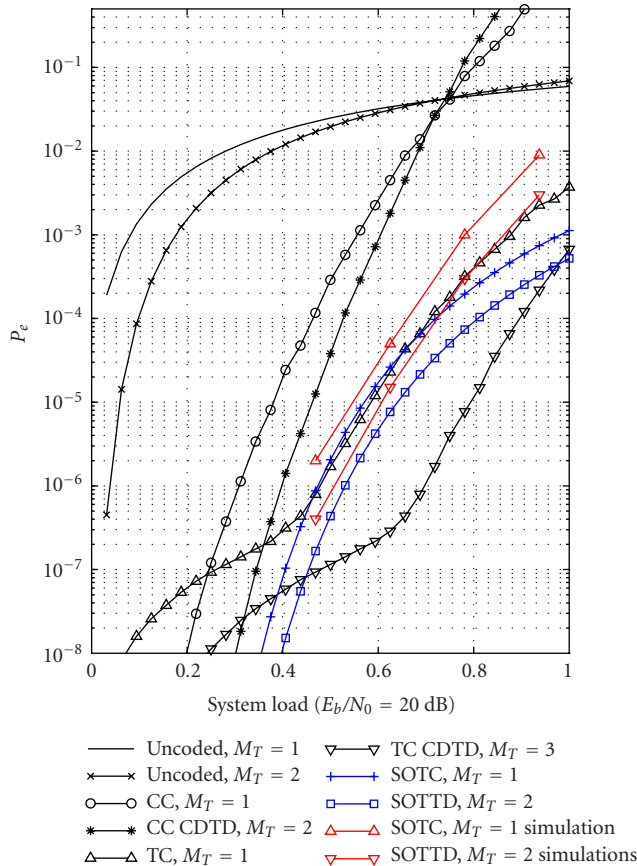


FIGURE 8: Bit error probability as a function of the load (number of users/total spreading = K/G), with the operating point at $E_b/N_0 = 20$ dB.

one sample. In addition, the turbo decoding configuration for $Z = 2$ constituent codes operates in serial mode, that is, “SISO decoder 1” processes data before “SISO decoder 2” starts its operation, and so on (refer to Figure 5).

Using the system parameters outlined in Table 1, the BER performance of a SOTTDD CDMA system has been determined by means of simulation. Figure 8 compares the simulated SOTTDD performance with the theoretical performance bounds of convolutional and turbo-coded CDTD. $E_b/N_0 = 20$ dB and $G = 32$, unless otherwise stated.

Concentrating on the BER curves of the SOTTDD system, slight disparities between the simulation results and performance bounds can be identified for target BER values of 10^{-6} or worse. As can be seen from the graphs, the simulation curves are very close to the simulation bounds, for normalized user loads (K/G) of less than 0.75. For the conditions of low load ($P_b < 10^{-6}$), the performance of the simulated system is dominated by the performance of the suboptimal (non-ML) decoder and the practical choice of a random interleaver.

For the higher load conditions, the simulation results are also worse than the bounding performance, since the performance is limited in frequency-selective channels due to increased interference.

4. SUMMARY AND CONCLUSION

In this paper, a new concept of layered super-orthogonal turbo transmit diversity (SOTTDD) has been presented for application in code-division multiple-access (CDMA) communication systems. The techniques of low-rate spreading and coding have been combined with orthogonal code-division transmit diversity (CDTD) and iterative “turbo” processing at the receiver. In contrast to layered ST turbo-coded (TC) CDTD, where a turbo encoder (and its associated iterative decoder) is required for every transmit diversity branch available, SOTTDD requires a single turbo encoder-decoder pair, making it particularly attractive for CDMA wireless applications, the only requirement being that the number of constituent encoders Z be greater or equal to the transmit diversity order M_T .

From the performance results presented, it may be deduced that the proposed SOTTDD system provides a very powerful and practical extension to the TC CDTD schemes, and yields superior performance compared to TC CDTD over the practically complete capacity range of CDMA. Another significant observation is the fact that the performance degradation of TC CDTD at low system loads (due to inherent TC error floor) is alleviated by the SOTTDD system. This is explained in terms of the higher minimum free distance offered by the low rate-(1/16) constituent encoders, as opposed to the use of rate-(1/2) (256-state) constituent encoders in TC systems.

In conclusion, the interpretation of the performance bounds presented in this paper should be done within the confidence limits imposed by the use of the union bound, as well as the restrictions set by practical considerations, as such bounds are only valid for the case of ML decoding, and they may diverge significantly from the true performance at low values of E_b/N_0 . Also, in the simulation, a suboptimal non-ML decoding algorithm was employed, as well as a pseudo-random interleaver. Furthermore, the performance of practical systems is strongly influenced by the availability of reliable CSI, which also plays a major role in the correct operation of virtually all adaptive receiver subsystems, including channel estimation, multipath decomposition and RAKE MRC, Doppler tracking, equalization, and several others. Clearly, the absence of reliable CSI will produce a noticeable degradation in the system performance. However, despite the restrictions and limitations, the results presented are close to the theoretical bounds for most of the normal CDMA operational range and thus provide useful design and comparative performance guidelines for SOTTDD CDMA application scenarios.

REFERENCES

- [1] N. Chiurtu, B. Rimoldi, and I. E. Telatar, “On the capacity of multi-antenna Gaussian channels,” in *Proc. IEEE International Symposium on Information Theory (ISIT '01)*, p. 53, Washington, DC, USA, June 2001.
- [2] G. J. Foschini, “Layered space-time architecture for wireless communication in a fading environment when using multi-element antennas,” *Bell Labs Technical Journal*, vol. 1, no. 2, pp. 41–59, 1996.

- [3] S. M. Alamouti, "A simple transmit diversity technique for wireless communications," *IEEE J. Select. Areas Commun.*, vol. 16, no. 8, pp. 1451–1458, 1998.
- [4] A. Hiroike, F. Adachi, and N. Nakajima, "Combined effects of phase sweeping transmitter diversity and channel coding," *IEEE Trans. Veh. Technol.*, vol. 41, no. 2, pp. 170–176, 1992.
- [5] W.-Y. Kuo and M. P. Fitz, "Design and analysis of transmitter diversity using intentional frequency offset for wireless communications," *IEEE Trans. Veh. Technol.*, vol. 46, no. 4, pp. 871–881, 1997.
- [6] N. Seshadri and J. H. Winters, "Two signaling schemes for improving the error performance of frequency division duplex (fdd) transmission systems using transmitter antenna diversity," *International Journal of Wireless Information Networks*, vol. 1, no. 1, pp. 49–60, 1994.
- [7] V. Tarokh, A. F. Naguib, N. Seshadri, and A. R. Calderbank, "Low-rate multi-dimensional space-time codes for both slow and rapid fading channels," in *8th IEEE International Symposium on Personal, Indoor and Mobile Radio Communications (PIMRC '97)*, pp. 1206–1210, Helsinki, Finland, September 1997.
- [8] N. Seshadri, V. Tarokh, and A. R. Calderbank, "Space-time codes for wireless communication: code construction," in *IEEE 47th Vehicular Technology Conference (VTC '97)*, pp. 637–641, Phoenix, Ariz, USA, May 1997.
- [9] V. Tarokh, N. Seshadri, and A. R. Calderbank, "Space-time codes for high data rate wireless communication: performance criterion and code construction," *IEEE Trans. Inform. Theory*, vol. 44, no. 2, pp. 744–765, 1998.
- [10] A. F. Naguib, V. Tarokh, N. Seshadri, and A. R. Calderbank, "A space-time coding modem for high-data-rate wireless communications," *IEEE J. Select. Areas Commun.*, vol. 16, no. 8, pp. 1459–1478, 1998.
- [11] M. P. Lötter, *Numerical analysis of spatial/temporal cellular CDMA systems*, Ph.D. thesis, University of Pretoria, Pretoria, South Africa, 1999.
- [12] P. G. W. van Rooyen, M. P. Lötter, and D. J. van Wyk, *Space-Time Processing for CDMA Mobile Communications*, Kluwer Academic Publishers, Boston, Mass, USA, 2000.
- [13] G. J. Foschini and M. J. Gans, "Capacity when using diversity at transmit and receive sites and the Rayleigh-faded matrix channel is unknown at the transmitter," in *Proc. 6th WINLAB Workshop on 3rd Generation Wireless Information Networks*, New Brunswick, NJ, USA, March 1996.
- [14] A. J. Viterbi, *CDMA: Principles of Spread Spectrum Communications*, Addison-Wesley Publishing, Reading, Mass, USA, 1995.
- [15] P. Frenger, P. Orten, and T. Ottosson, "Combined coding and spreading in CDMA systems using maximum free distance convolutional codes," in *48th IEEE Vehicular Technology Conference (VTC '98)*, pp. 2497–2501, Ottawa, Ontario, Canada, May 1998.
- [16] P. Frenger, P. Orten, and T. Ottosson, "Code-spread CDMA using low-rate convolutional codes," in *Proc. IEEE 5th International Symposium on Spread Spectrum Techniques and Applications (ISSSTA '98)*, pp. 374–378, Sun City, South Africa, September 1998.
- [17] A. J. Viterbi, "Very low rate convolution codes for maximum theoretical performance of spread-spectrum multiple-access channels," *IEEE J. Select. Areas Commun.*, vol. 8, no. 4, pp. 641–649, 1990.
- [18] K. Pehkonen and P. Komulainen, "A superorthogonal turbo-code for CDMA applications," in *Proc. IEEE 4th International Symposium on Spread Spectrum Techniques and Applications (ISSSTA '96)*, pp. 580–584, Mainz, Germany, September 1996.
- [19] P. Komulainen and K. Pehkonen, "Performance evaluation of superorthogonal turbo codes in AWGN and flat Rayleigh fading channels," *IEEE J. Select. Areas Commun.*, vol. 16, no. 2, pp. 195–205, 1998.
- [20] J. G. Proakis, *Digital Communications*, McGraw-Hill, New York, NY, USA, 4th edition, 2001.
- [21] D. J. van Wyk, *Space-time turbo coding for CDMA mobile communications*, Ph.D. thesis, University of Pretoria, Pretoria, South Africa, 2000.
- [22] S. Benedetto and G. Montorsi, "Average performance of parallel concatenated block codes," *Electronics Letters*, vol. 31, no. 3, pp. 156–158, 1995.

Daniël J. van Wyk received the B.Eng. and M.Eng. degrees, both cum laude, from the University of Pretoria in 1993 and 1996, respectively. During 1998–2000, he completed a Ph.D. thesis at the same university in the area of space-time turbo-coded processing. From 1995 till 1998, he worked at the Laboratory for Advanced Engineering (LGI), the University of Pretoria as development engineer. From 1998 to 2000, he was employed as a systems engineer at CSIR Defencetek where he lead research teams in electronic warfare system design. In August 2000, he joined Zyray Wireless in San Diego where he was the Lead System Designer for the Spinner WCDMA communication system. Currently, he is employed as a Senior DSP Specialist at RapidM in South Africa, focusing on the development of data modems for military and commercial markets. In 1997, he received, with L.P. Linde, a Design Institute Award from the South African Bureau of Standards (SABS) for a self-synchronizing BER analyzer product. He has cowritten the book "Space-time processing for CDMA mobile communications" (Kluwer, 2000). He is the author and coauthor of 5 patents in digital communications and has published a number of articles in international journals and at international conferences. Danie is a Member of the IEEE.



Louis P. Linde received the integrated B.Eng., with honors, degree from the University of Stellenbosch in 1973, and the M.Eng. (cum laude) and D.Eng. degrees from the University of Pretoria (UP), South Africa, in 1980 and 1984, respectively. He is presently a Professor and Group Head of the Signal Processing and Telecommunications Group in the Department of Electrical, Electronic and Computer Engineering, UP, and the Director of the Centre for Radio and Digital Communication (CRDC), where he directs a group of researchers in the fields of IP-based wireless multiple-access systems, MIMO channel estimation and modelling, and space-time coding. He is also the codirector of DiGiMod (Pty)Ltd, a private enterprise active in the development of innovative wireless communication products for industry. Examples include a novel multidimensional quasisynchronous orthogonal code-division multiple-access transceiver employing complex chirp-like spreading sequences, as well as a high-speed microwave/satellite modem and long-range power efficient broadband DSSS telemetry transceiver, jointly developed with Tellumat (Pty)Ltd, South Africa Professor Linde is the Editor of Telecommunications, Signal Processing, and Information Theory of the Transactions of the SAIEE. He is the author and coauthor of more than 60 conference presentations and journal papers and holds four patents. He is a registered Professional Engineer since 1976 and a Senior Member of the IEEE.



Pieter G. W. van Rooyen is presently a Chief Architect of Broadcom's Mobile and Wireless BU and was the founder and the Chief Technology Officer (CTO) of Zyray Wireless Inc. that has been acquired by Broadcom in 2004 for almost a 100 mil dollar. At Zyray Wireless, he was responsible for new technology development and for defining the overall technology strategy of the company.



He has focused on new technology development in the areas of smart antennas and space-time processing to further enhance Zyray's growing product family. Previously, van Rooyen founded and served as Director of the Alcatel Research Unit for Wireless Access (ARUWA) at the University of Pretoria, South Africa, conducting research into mobile communications systems with a particular emphasis on WCDMA/smart antenna cellular technology. He has also worked at Sony Advanced Telecommunications Laboratory (Tokyo, Japan), where he conducted research and product development on software-defined radio and space-time processing techniques for next-generation wireless communications. Prior to that, he spent two years at Alcatel Altech Telecoms and has served as a Professor in the Department of Electrical, Electronic and Computer Engineering at the University of Pretoria, South Africa. He has published numerous technical papers, holds a number of technical patents in the area of digital communications and is the coauthor of two books related to WCDMA/smart antenna mobile systems. Dr. van Rooyen holds a Ph.D. degree in engineering from the Rand Afrikaans University, Johannesburg, South Africa, in the area of CDMA and smart antenna techniques.