Soft-Decision Decoding of Systems with Tx/Rx Diversity

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***Abstract* — In this paper, we describe the concatenation of Turbo/Convolutional codes with transmit and receive diversity schemes by using Space-Time Block Code. It is shown that, by using two transmit antennas and one/or two receive antenna, large coding gain for the bit error rate is achieved over the system without diversity. Simulation results show that, by using systems with transmit and receive diversity, high gain can be achieved with very low complexity. It turns out that at $\mathrm{BER} = 10^{-4}$, the gain of 9 dB can be achieved for system using STTD transmit diversity only (without using any channel codes) and 2 dB gain can be achieved over channel coding systems using hard-decision decoding with much lower complexity. The most important conclusion is that, using soft-decision decoding systems enhanced with transmit diversity can provide very high coding gain; e.g., in convolutional coded system using soft-decision Viterbi decoder, the coding gain is 12 dB over uncoded system and 5 dB over hard-decision decoding in flat fading channel, while the coding gain is about 13 dB for turbo coded systems using soft-decision decoding based on SOVA algorithm with transmit diversity and the coding gain is 15 dB if the decoder is based on Log-MAP algorithm. In systems using transmit and receive diversity, the coding gain is much higher, e.g., for convolutional-coded systems, the coding gain is 20 dB, while for turbo-coded systems using SOVA and Log-MAP algorithms, the coding gain are a little more than 20 dB and 21 dB, respectively.**

***Index Terms* — Space-Time Block Code, Turbo Codes, SOVA, Transmit diversity, Receive diversity.**

I. INTRODUCTION

In future wireless communication systems, high data rates need to be reliably transmitted over time-varying bandlimited channels. The wireless channel mainly suffers from time-varying fading due to multipath propagation and destructive superposition of signal received over different paths, which make it hard for the receiver to reliably determine the transmitted signal unless some less attenuated replica of the signal are provided to the receiver. Transmitting the replica of the signal is called diversity. It is common for wireless systems to employ both diversity techniques and channel coding for error detection and/or correction. A widely applied technique to reduce the effects of multipath fading is antenna diversity. Usually, multiple antennas are used at the receiver with some kind of combining of the received signals, e.g., maximum ratio combining [1]. However, transmit and receive diversity techniques can be applied in the uplink and/or the downlink.

In digital mobile communication systems, powerful channel coding is paramount to combat the effects of fading, interference and noise to obtain sufficient reception quality. Turbo codes [2], can achieve remarkable error performance at a low signal-to-noise ratio (SNR) with moderate decoding complexity.

Recently, several schemes that combine space-time and Turbo codes were proposed. A scheme that combines a binary Turbo encoder with transmit antenna diversity was proposed in [3]. In [4], the outputs of parallel-concatenated TCM modules are routed to two separate antennas. This arrangement is a direct extension of [5], where the TCM modules are connected to a single antenna through a selector. A scheme that consists of two parallel-concatenated STCs was proposed in [6]. However, the scheme lacks a mechanism for puncturing the output resulting in Turbo codes with reduced data rates compared to the constituent codes. Schemes of serial concatenation of STCs and Turbo codes can be found in [7,8].

This paper is organized as follows. In section 2, the principles of space-time block codes are introduced. In section 3, soft-decision decoding of systems with transmit and receive diversity is described. The simulation results are presented in section 4, while we conclude in section 5.

II. PRINCIPLES OF SPACE-TIME BLOCK CODE

In this section, we review the theory of space-time block code by considering the classic MRC technique [9,10].

A. Classical Maximum Ratio Combining

Figure 1 shows the baseband representation of the classic MRC technique in conjunction with two receivers. At a particular instant, a symbol x is transmitted. As we can see from the figure, the transmitted symbol x propagates through two different channels, namely, h_1 and h_2. For simplicity, all channels are assumed to be constituted of a single nondispersive or flat-fading propagation path and can be modeled as complex multiplicative distortion, which consists of a magnitude and phase response given by [9,10]

$$h_1 = |h_1| e^{j\theta 1} \tag{1}$$

$$h_2 = |h_2| e^{j\theta 2} \qquad (2)$$

where $|h_1|$, $|h_2|$ are the fading magnitudes and θ_1, θ_2 are the phase values. Each receiver, as shown in Figure 1, adds noise. Hence, the resulting received baseband signals are

$$y_1 = h_1 x + n_1 \qquad (3)$$

$$y_2 = h_2 x + n_2 \qquad (4)$$

where n_1 and n_2 are complex noise samples. In matrix form, this can be written as

$$\begin{pmatrix} y_1 \\ y_2 \end{pmatrix} = x \begin{pmatrix} h_1 \\ h_2 \end{pmatrix} + \begin{pmatrix} n_1 \\ n_2 \end{pmatrix} \qquad (5)$$

Assuming that perfect channel information is available, the received signals y_1 and y_2 can be multiplied by the conjugate of the complex channel transfer functions $\bar{h}_1$ and $\bar{h}_2$, respectively, in order to remove the channel's effects. Then, the corresponding signals are combined at the input of the maximum likelihood detector of Figure 1 according to

$$\tilde{x} = \bar{h}_1 y_1 + \bar{h}_2 y_2 = \left(|h_1|^2 + |h_2|^2\right) x + \bar{h}_1 n_1 + \bar{h}_2 n \qquad (6)$$

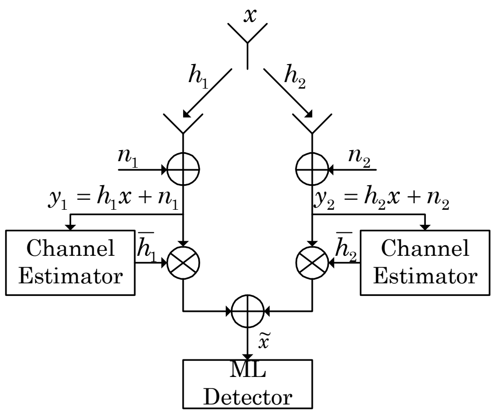

Figure 1: Baseband representation of the MRC technique using two receivers.

The combined signal $\tilde{x}$ is then passed to the maximum likelihood detector, as shown in Figure 1. The most likely transmitted symbol is determined by the maximum likelihood detector based on the Euclidean distances between the combined signal $\tilde{x}$ and all possible transmitted symbols. The simplified decision rule is based on choosing x_i if and only if

$$dist(\tilde{x}, x_i) \le dist(\tilde{x}, x_j), \quad \forall i \ne j \qquad (7)$$

where $dist(\alpha, \beta)$ is the Euclidean distance between signals α and β and the index j spans all possible transmitted signals. From (7), we can see that maximum likelihood transmitted symbol is the one having the minimum Euclidean distance from the combined signal $\tilde{x}$.

B. Transmission Model and Diversity Criterion

In analogy to the MRC matrix formula of (5), a STB code describing the relationship between the original transmitted signal x and the signal replicas artificially created at the transmitter for transmission over various diversity channels is defined by an $n \times p$ dimensional transmission matrix. The entries of the matrix are constituted of linear combinations of the k-ary input symbols $x_1, x_2, \ldots, x_k$ and their conjugates. The k-ary input symbols x_i, $i = 1, \cdots, k$ are used to represent the information-bearing binary bits to be transmitted over the transmit diversity channels. In a signal constellation having 2^b constellation points, a number b of binary bits are used to represent a symbol x_i. Hence, a block of $k \times b$ binary bits is entered into the STB encoder at a time and it is, therefore, referred to as a STB code. The number of transmitter antennas is p and n represents the number of time slots used to transmit k input symbols. Hence, a general form of the transmission matrix of a STB code is written as

$$\begin{pmatrix} g_{11} & \cdots & g_{1p} \\ \vdots & \ddots & \vdots \\ g_{n1} & \cdots & g_{np} \end{pmatrix} \qquad (8)$$

where the entries g_{ij} represent linear combinations of the symbols $x_1, x_2, \ldots, x_k$ and their conjugates. More specifically, the entries g_{ij}, where $i = 1, \cdots, p$ are transmitted simultaneously from transmit antennas $1, \ldots, p$ in each time slot $j = 1, \ldots, n$. For example, in time slot $j = 2$, signals $g_{12}, g_{22}, \ldots, g_{p2}$ are transmitted simultaneously from transmit antennas $Tx1, Tx2, \ldots, Txp$. We can see in the transmission matrix defined in (8) that encoding is carried out in both space and time; hence, the term space–time coding. The $n \times p$ transmission matrix in (8) (which defines the STB code) is based on a complex generalized orthogonal design, as defined in [11]. Since there are k symbols transmitted over n time slots, the code rate of the STB code is given by

$$R = k/n\,. \qquad (9)$$

At the receiving end, one can have an arbitrary number of q receivers. It was shown in [6] that the associated diversity order is $p \times q$. A combining technique similar to MRC can be applied at the receiving end, which may be generalized to q receivers.

C. Decoding Algorithm of Space–Time Block Code

A simple transmit diversity scheme for two transmit antennas was introduced by Alamouti in [9]. The transmission matrix is

$$\mathbf{G}_2 = \begin{pmatrix} x_1 & x_2 \\ -\bar{x}_2 & \bar{x}_1 \end{pmatrix}. \qquad (10)$$

We can see in the transmission matrix $\mathbf{G_2}$ that there are $p=2$ (number of columns in the matrix $\mathbf{G_2}$) transmitters, $k=2$ possible input symbols, namely, x_1, x_2, and the code spans over $n=2$ (number of rows in the matrix $\mathbf{G_2}$) time slots. Since $k=2$ and $n=2$, the code rate given by (9) is unity.

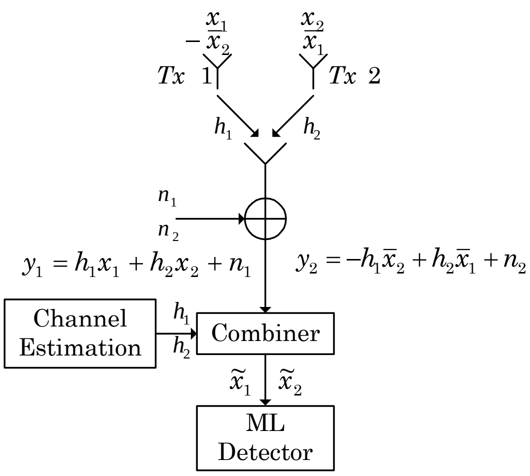

Figure 2: Simple two-transmitter STB code $\mathbf{G_2}$ using one receiver.

Figure 2 shows the baseband representation of a simple two-transmitter STB code, namely, that of the $\mathbf{G_2}$ code seen in (10) using one receiver. We can see from the figure that there are two transmitters, namely, $Tx1$ as well as $Tx2$ and they transmit two signals simultaneously. As mentioned earlier, the complex fading envelope is assumed to be constant across the corresponding two consecutive time slots. Therefore, one can write

$$h_1 = h_1(T=1) = h_1(T=2) \tag{11}$$

$$h_2 = h_2(T=1) = h_2(T=2) \tag{12}$$

At the receiver, independent noise samples are added in each time slot; hence the signals received over nondispersive or narrow-band channels can be expressed with the aid of (10) as

$$y_1 = h_1 x_1 + h_2 x_2 + n_1 \tag{13}$$

$$y_2 = -h_1 \bar{x}_2 + h_2 \bar{x}_1 + n_2 \tag{14}$$

where y_1 is the first received signal and y_2 is the second. Note that the received signal y_1 consists of the transmitted signals x_1 and x_2, while y_2 consists of their conjugates. In order to determine the transmitted symbols, we have to extract the signals x_1 and x_2 from the received signals y_1 and y_2. Therefore, both signals y_1 and y_2 are passed to the combiner, as shown in Figure 2. In the combiner-aided by the channel estimator, which provides perfect estimation of the diversity channels in this example-simple signal processing is performed in order to separate the signals x_1 and x_2. Specifically, the maximum likelihood detection minimizes the decision metric

$$\left|y_1 - h_1 x_1 - h_2 x_2\right|^2 + \left|y_2 + h_1 \bar{x}_2 - h_2 \bar{x}_1\right|^2 \tag{15}$$

over all possible values of x_1 and x_2. We expand the above metric and delete the terms that are independent of the codewords and observe that the minimization is equivalent to minimizing

$$\begin{aligned} &-[y_1\bar{h}_1\bar{x}_1 + \bar{y}_1 h_1 x_1 + y_1\bar{h}_2\bar{x}_2 + \bar{y}_1 h_2 x_2 - y_2\bar{h}_1 x_2 - \bar{y}_2 h_1 \bar{x}_2 + \\ &y_2\bar{h}_2 x_1 + \bar{y}_2 h_2 \bar{x}_1] + (|x_1|^2 + |x_2|^2)(|h_1|^2 + |h_2|^2). \end{aligned} \tag{16}$$

The above metric decomposes into two parts, one of which

$$-[y_1\bar{h}_1\bar{x}_1 + \bar{y}_1 h_1 x_1 + y_2\bar{h}_2 x_1 + \bar{y}_2 h_2 \bar{x}_1] + |x_1|^2\ (|h_1|^2 + |h_2|^2) \tag{17}$$

is only a function of x_1, and the other one

$$-[y_1\bar{h}_2\bar{x}_2 + \bar{y}_1 h_2 x_2 - y_2\bar{h}_1 x_2 - \bar{y}_2 h_1 \bar{x}_2] + |x_2|^2\ (|h_1|^2 + |h_2|^2) \tag{18}$$

is only a function of x_2. Thus the minimization is equivalent to minimizing these two parts separately. This in turn is equivalent to minimizing the decision metric

$$\left|\left(y_1\bar{h}_1 + \bar{y}_2 h_2\right) - x_1\right|^2 + \left(-1 + |h_1|^2 + |h_2|^2\right)|x_1|^2 \tag{19}$$

for detecting x_1 and the decision metric

$$\left|\left(y_1\bar{h}_2 - \bar{y}_2 h_1\right) - x_2\right|^2 + \left(-1 + |h_1|^2 + |h_2|^2\right)|x_2|^2 \tag{20}$$

for decoding x_2. From equations (19) and (20), we can obtain the "optimum" soft values of the codewords x_1 and x_2 as follows:

$$\hat{x}_1 = \left(\bar{h}_1 y_1 + h_2 \bar{y}_2\right) \tag{21}$$

$$\hat{x}_2 = \left(\bar{h}_2 y_1 - h_1 \bar{y}_2\right) \tag{22}$$

Clearly, from (21) and (22), we can see that we have separated the signals x_1 and x_2 by simple multiplications and additions. Due to the orthogonality of the STB code $\mathbf{G_2}$ in (10) [11], the unwanted signal x_2 is canceled out in (21) and vice versa, signal x_1 is removed from (22). In hard-decision decoding, both signals $\hat{x}_1$ and $\hat{x}_2$ are passed to the maximum likelihood detector, which applies (7) to determine the most likely transmitted symbols.

There may be applications where a higher order of diversity is needed and multiple receive antennas at the remote units are feasible. In such cases, it is possible to provide a diversity order of 4 with two transmit and two receive antennas. The encoding and transmission sequence of the information symbols for this configuration is identical to the case of a single receiver. Then, the received signals at the second receive antenna will be:

$$y_3 = h_3 x_1 + h_4 x_2 + n_3 \tag{23}$$

$$y_4 = -h_3 \bar{x}_2 + h_4 \bar{x}_1 + n_4 \tag{24}$$

where n_3 and n_4 are complex random variables representing receiver thermal noise and interference. The combiner at the receiver builds the following two soft values of the codewords x_1 and x_2 as follows [9]:

$$\hat{x}_1 = \left(\bar{h}_1 y_1 + h_2 \bar{y}_2 + \bar{h}_3 y_3 + h_4 \bar{y}_4\right) \qquad (25)$$

$$\hat{x}_2 = \left(\bar{h}_2 y_1 - h_1 \bar{y}_2 + \bar{h}_4 y_3 - h_3 \bar{y}_4\right) \qquad (26)$$

As before, in hard-decision decoding, both signals $\hat{x}_1$ and $\hat{x}_2$ are passed to the maximum likelihood detector, which applies (7) to determine the most likely transmitted symbols.

III. SOFT-DECISION DECODING OF SYSTEMS WITH TRANSMIT AND RECEIVE DIVERSITY

A. Serially Concatenated Coded Systems

We can concatenate a secondary channel encoder before the STBC transmitter and concatenate its decoder to the STBC receiver. The outer encoder is a classical non-recursive non-systematic convolutional code. Between outer encoder and inner encoder, we can insert an interleaver to decorrelate the burst errors produced by correlated fading. A block diagram of such system is shown in Figure 3.

If the output of the STBC decoder is hard, then hard-decision decoding is made in the outer decoder. If the output of the STBC decoder is soft, then soft-decision decoding is made in the outer decoder. If the outer code is convolutional encoder and its decoder use soft-decisions Viterbi decoder, then the STBC decoder must be soft output. Since STBC combats the fading by antenna diversity, the outer code combats the AWGN to achieve additional coding gain. To complete the decoding, the soft values must be deinterleaved and sent to the soft-decision Viterbi decoder.

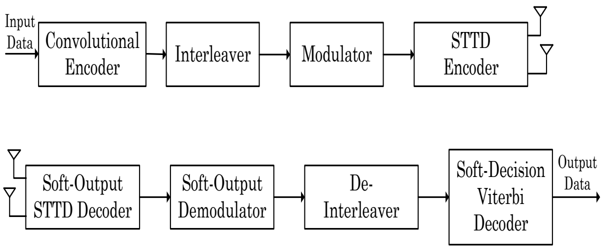

Figure 3: A block diagram of the concatenated system.

B. Iterative (Turbo) Decoding Strategy

We consider the channel code and the MIMO channel as a serially concatenated scheme [12], with an outer channel encoder (typically a convolutional or turbo code), interleaver, and inner space-time block encoder. The iterative decoding method is a well established method in decoding bits encoded by turbo encoders. The main idea is based on two individually optimal steps that can be iteratively repeated. Therefore, we are often content to solve the problems of having the MIMO detector incorporate soft reliability information provided by the channel decoder, and the channel decoder incorporate soft information provided by the MIMO detector. Information between the detector and decoder is then exchanged in an iterative fashion until desired performance is achieved. While this iterative process is not strictly optimal, it has been shown that the "turbo principle" is very effective and computationally efficient in other joint detection/decoding problems [13].

In the first step, from the channel samples we compute the a posteriori log-likelihood ratios of the coded symbols for each transmit antenna.

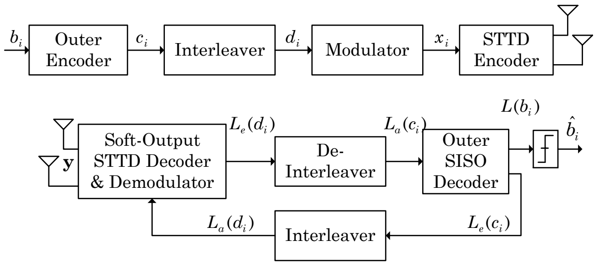

Figure 4: Iterative detection and decoding for a system with transmit diversity.

Since we have samples from the receive antennas, and we want to compute soft values for the bits that are transmitted on the MIMO channel, this module can be interpreted as a multiple soft-in soft-out a posteriori probabilities estimator. In the second step, the a posteriori log-likelihood ratios of the coded bits are deinterleaved and fed to the outer SISO decoder [12]. The SISO decoder provides both the log-likelihood ratios of the information bits $L(b_i)$, and new/improved log-likelihood ratios of the coded bits $L_e(c_i)$. Following the iterative decoding principle [13], extrinsic log-likelihood ratios of the coded bits are computed by subtracting the decoder inputs from the decoder outputs, $L_e(c_i) = L(c_i) - L_a(c_i)$. This is to minimize the correlation with previously computed soft values. The extrinsic values are interleaved and fed back to the space-time decoder/demodulator where they are used in a new iteration as an estimate of the a priori log-likelihood ratios of the coded bits $L_a(d_i)$. Extrinsic information is also computed at the space-time decoder/demodulator output, $L_e(d_i) = L(d_i) - L_a(d_i)$. By repeating several times the above procedure, the performance of the system is greatly improved. In the final iteration, the decoded sequence of information bits is obtained by making hard decisions on $L(b_i)$.

IV. SIMULATION RESULTS

Simulations were carried out for frames sizes of 384 bits and transmission of 1 bit/s/Hz using QPSK modulation combined with rate $1/2$ Turbo/or convolutional encoder. We used two transmit antennas and one/or two receive antenna applying the space-time block code $\mathbf{G}_2$ in quasi-static fading, i.e. the channel is constant during transmission of one coded modulation block and changes from one block to the next independently.

A. Simulations for uncoded systems with transmit and receive diversity

- The first experiment is for a system using transmit diversity only (2T-1R), at $\mathrm{BER} = 10^{-4}$, the diversity gain is 9 dB over uncoded system.
- The second experiment is for a system using transmit and receive diversity only (2T-2R), at $\mathrm{BER} = 10^{-4}$, the diversity gain is 18 dB over uncoded system and 9 dB over the 2T-1R system.

B. Simulations for convolutional-coded systems with diversity

- The third experiment is for a system using convolutional encoder and hard-decision decoding without diversity, at $\mathrm{BER} = 10^{-4}$, the coding gain is 7 dB over uncoded system.
- The fourth experiment is for coded systems using convolutional encoder and soft-decision Viterbi decoder with transmit diversity (coded 2T-1R), at $\mathrm{BER} = 10^{-4}$, the coding gain is 12 dB over uncoded system and 5 dB over hard-decision decoding.
- The fifth experiment is for coded systems using convolutional encoder and soft-decision Viterbi decoder with transmit and receive diversity (coded 2T-2R), at $\mathrm{BER} = 10^{-4}$, the coding gain is 20 dB over uncoded system, 8 dB over convolutional coded system with transmit diversity only (2T-1R).

These results are shown in Figure 5. At $\mathrm{BER} = 10^{-5}$, we can see that the coding gain of the system using convolutional code with transmit and receive diversity over the coded system with transmit diversity only is 10 dB, while the coding gain is 14 dB over the system with transmit diversity only.

C. Simulations for turbo-coded systems with diversity

- The sixth experiment is for coded systems using turbo encoder and SOVA decoder with transmit diversity (coded 2T-1R), at $\mathrm{BER} = 10^{-4}$, the coding gain is 13.5 dB over uncoded system and 7 dB over hard-decision decoding.
- The seventh experiment is for coded systems using turbo encoder and SOVA decoder with transmit and receive diversity (coded 2T-2R), at $\mathrm{BER} = 10^{-4}$, the coding gain is more than 20 dB over uncoded system, 7 dB over turbo coded system with transmit diversity only (2T-1R).

These results are shown in Figure 6. At $\mathrm{BER} = 10^{-5}$, we can see that the coding gain of the system using turbo code with transmit and receive diversity over the coded system with transmit diversity only is 9 dB, while the coding gain is 14.5 dB over the system with transmit diversity only.

- The eighth experiment is for coded systems using turbo encoder and Log-MAP decoder with transmit diversity (coded 2T-1R), at $\mathrm{BER} = 10^{-4}$, the coding gain is 15 dB over uncoded system and 8 dB over hard-decision decoding.
- The ninth experiment is for coded systems using turbo encoder and Log-MAP decoder with transmit and receive diversity (coded 2T-2R), at $\mathrm{BER} = 10^{-4}$, the coding gain is more than 21 dB over uncoded system, 6 dB over turbo coded system with transmit diversity only (2T-1R).

These results are shown in Figure 7. At $\mathrm{BER} = 10^{-5}$, we can see that the coding gain of the system using turbo code with transmit and receive diversity over the coded system with transmit diversity only is 10 dB, while the coding gain is 15 dB over the system with transmit diversity only.

V. CONCLUSIONS

In this paper, we have described the concatenation of Turbo/Convolutional codes with transmit and receive diversity schemes by using Space-Time Block Code. It is shown that, by using two transmit antennas and one/or two receive antenna, large coding gain for the bit error rate and frame error rate are achieved over the system without diversity. This scheme is expected to provide diversity improvement at all the remote units in the wireless communication system.

From the simulation results, we conclude that, using systems (coded or uncoded) with transmit and receive diversity, high gain can be achieved with very low complexity. It turns out that at $\mathrm{BER} = 10^{-4}$, the gain of 9 dB can be achieved for system using STBC transmit diversity only (without using any channel codes) and 2 dB gain can be achieved over channel coding systems using hard-decision decoding with much lower complexity. The most important conclusion is that, using soft-decision decoding systems enhanced with transmit diversity can provide very high coding gain; e.g., in convolutional coded system using soft-decision Viterbi decoder, the coding gain is 12 dB over uncoded system and 5 dB over hard-decision decoding in flat fading channel, while the coding gain is about 13 dB for turbo coded systems using soft-decision decoding based on SOVA algorithm with transmit diversity and the coding gain is 15 dB if the decoder is based on Log-MAP algorithm. In systems using transmit and receive diversity, the coding gain is much higher, e.g., for convolutional-coded systems, the coding gain is 20 dB, while for turbo-coded systems using SOVA and Log-

MAP algorithms, the coding gain are a little more than 20 dB and 21 dB, respectively.

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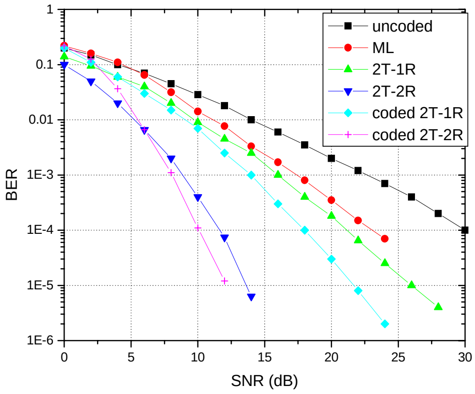

Figure 5: BER curves of convolutional-coded system with transmit and receive diversity.

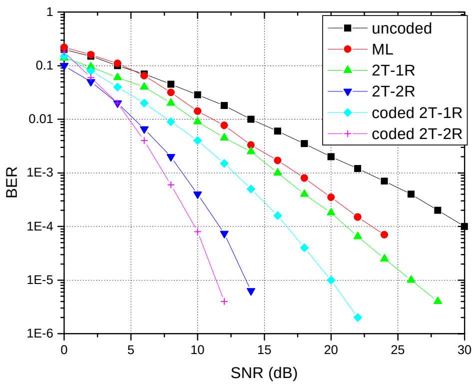

Figure 6: BER of turbo-coded system with transmit and receive diversity using SOVA.

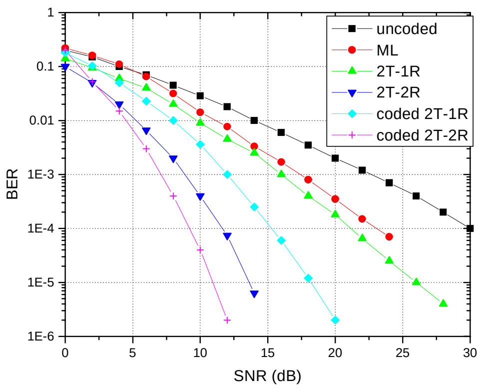

Figure 7: BER of turbo-coded system with transmit and receive diversity using Log-MAP.