A Suboptimal Receiver with Turbo Block Coding for Ultra-Wideband Communications

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Abstract—**In this paper, the performance of adaptive equalization and turbo product coding is investigated for pulse-based UWB communications in short-range indoor environments. The sensitivity of adaptive LMS linear and nonlinear (decision-feedback) equalizers with respect to the number of training symbols and number of taps is considered. To reduce the error performance variation with respect to changing channel conditions, a turbo product code (TPC) with two component $(31, 26, 3)$ Hamming codes is proposed. We report simulation results showing that channel coding not only improves error performance, but also reduces significantly the sensitivity of UWB systems in short-range indoor wireless communications.**

I. INTRODUCTION

In recent years, a significant research effort has been devoted to the study of ultra-wideband (UWB) communication systems. The basic principle behind ultra-wideband communications was first used in radar systems over thirty years ago [1]-[3]. More recently, the UWB concept was used to develop impulse radio, where baseband pulses are transmitted over the channel [4]. This technique greatly simplifies the transmitter and receiver designs; however, the transmitted bandwidth extends to the gigaHertz range.

Any application of UWB technology must conform to the regulations imposed on radio-frequency transmissions. In the United States, these regulations are established and enforced by the Federal Communications Commission (FCC). The FCC regulation [5] permits transmission of signals with -10 dB bandwidths that lie in the 3.1 to 10.6 GHz band, provided that the transmitted signals have an effective isotropic radiated power (EIRP) below -41.3 dBm in this band and a minimum bandwidth of 500 MHz. The proposed IEEE 802.15.3a standard defines the requirements of a wireless personal-area network (WPAN) communication system [6]. These requirements include a bit rate of at least 110 Mbps at a distance of up to 10 m and 200 Mbps at up to 4 m, with desired rates up to 480 Mbps. A pulse-based UWB approach is a good candidate for meeting the WPAN requirements.

In this paper, we investigate the error performance of adaptive equalization and turbo product coding with pulse-based BPSK modulation for UWB communications in a short-range indoor environment. Performance is measured using the UWB multipath channel models generated by the IEEE 802.15.3a standard group. We consider a communication system in which pilot symbols are used, either to train an adaptive equalizer or to estimate the channel and initialize the equalizer taps. The performance of adaptive LMS linear and nonlinear (decision-feedback) equalizers is studied. Also considered is the sensitivity to the number of training symbols and the number of taps for a given UWB channel type. The results confirm the superiority of decision feedback structures. To reduce the variation in the bit error rate with respect to changing multipath channel conditions, a turbo product code (TPC) with two component $(31, 26, 3)$ Hamming codes is proposed.

The rest of the paper is organized as follows: In section II, the UWB communication system model used in this study is presented. An equivalent symbol-spaced UWB channel model is obtained at the boundaries before the pulse shaping filter and after the matched filter. Section III considers the performance of suboptimal adaptive equalizers as low-complexity alternatives to the Viterbi equalizer. Performance is studied in terms of equalizer length and sensitivity to different channel realizations. Simulation results, presented in section IV, of combinations of adaptive equalizers and a turbo product code, illustrate the benefits of using channel coding as an effective way to improve performance and reduce sensitivity to channel variations.

II. UWB COMMUNICATION SYSTEM MODEL

Fig. 1 shows a block diagram of a binary pulse-based UWB system. The symbol rate of the BPSK modulator is 250 Mbps, i.e., a symbol period $T = 4$ ns. The turbo product code is constructed from two identical $(31, 26, 3)$ Hamming codes. This results in an effective information rate equal to 175.86 Mbps. The output of the modulator is a binary-valued sequence $\{s_n\}$, with $s_n \in \{-1, +1\}$. A wideband unit-energy real-valued pulse shape $p(t)$ is employed such that the output of the transmit filter is given by

$$s(t) = \sum_{n=-\infty}^{\infty} s_n p(t - nT). \tag{1}$$

The pulse shape $p(t)$ can be either a single UWB pulse or a sequence of UWB pulses with good autocorrelation properties.

The UWB channel model employed in this work is compliant with the IEEE 802.15.3a model [7]. Here we use a simulation sampling time $\tau = 0.02$ ns. The UWB channel

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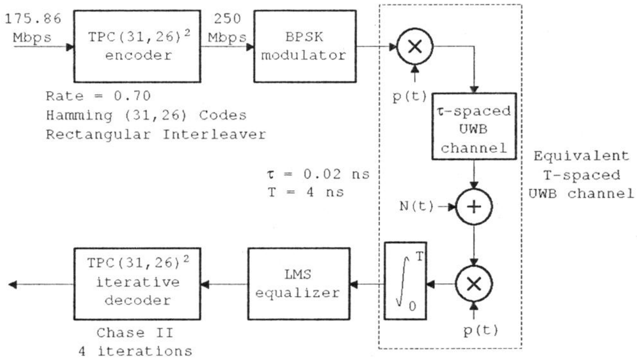

Fig. 1. A pulse-based UWB communication system.

impulse response (CIR) is given by

$$h(t) = \sum_{\ell=0}^{L-1} \alpha_\ell \delta(t - \tau_\ell), \tag{2}$$

where α_ℓ and τ_ℓ are the gain and delay of the ℓ-th channel path, for $\ell = 0, 1, \cdots, L-1$.

The path delays can be expressed as multiples of the sampling time: $\tau_\ell = m_\ell \tau$, where for $0 \leq \ell \leq L-1$, m_ℓ is a positive integer and $m_0 < m_1 < \cdots < m_{L-1}$. The delay spread of the channel is equal to τ_{L-1}. It is assumed that the noise process introduced at the receiver is denoted by $N(t)$ and modeled as AWGN with $\sigma^2 = N_0/2$. Consequently, the output of the matched filter at $t = mT$ is

$$\begin{aligned} Y_m &= \sum_{\ell=0}^{L-1} \alpha_\ell \int_{(m-1)T}^{mT} s(t-\tau_\ell)\, p(t)\, dt + W \\ &= \sum_{n=-\infty}^{\infty} s_n \sum_{\ell=0}^{L-1} \alpha_\ell \int_{(m-1)T}^{mT} p(t-\tau_\ell - nT)\, p(t)\, dt \\ &\qquad + W \\ &= \sum_{i=0}^{P-1} \beta_i s_{mM-i} + W, \end{aligned}$$

where $P = \lfloor T/\tau \rfloor$ is the number of sampling periods within a symbol period (200 in our case), and W is a Gaussian r.v. of zero mean and variance $\sigma^2 = N_0/2$. This results in an equivalent symbol-spaced UWB CIR which is given by

$$g(t) = \sum_{j=0}^{J-1} \beta_j\, \delta(t - jT), \tag{3}$$

where $J = \lfloor \tau_{L-1}/T \rfloor$ is the ratio of the channel delay spread to the symbol period. The path gains of the T-spaced CIR, β_j, depend on the autocorrelation function of $p(t)$ and on the path gains of the τ-spaced CIR α_ℓ, over the period $[(m-J+1+j)T, (m-J+1+j+1)T]$, for $0 \leq j \leq J-1$. Figure 2 shows the CIR of a sample realization of UWB channel type CM4. We assume that the channel remains static for a period of time in the order of thousands of symbols. This is a realistic assumption in an indoor environment.

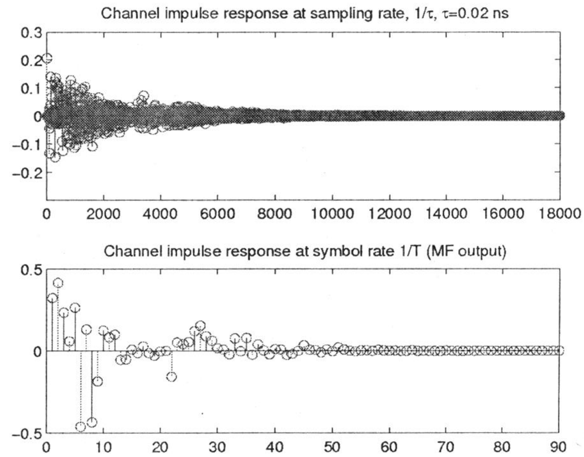

Fig. 2. UWB CIR at sampling rate and symbol rate for $P = 200$.

The output of the matched filter is input to an adaptive LMS equalizer. In this work, both linear and nonlinear (decision-feedback) structures are studied. The equalized output symbols are then delivered to a soft-input soft-output (SISO) iterative decoder for the TPC. After several iterations of SISO decoding, the estimated information bits are obtained.

III. Adaptive Equalization

In this section, the performance of adaptive LMS equalizers for the binary pulse-based UWB communication system outlined in the previous section is considered. As shown in Fig. 2, the equivalent symbol-spaced UWB channel exhibits a severe amount of intersymbol interference (ISI). It is well known that the optimum receiver is a Viterbi equalizer (VE) [11]. However, the VE requires a complexity that grows exponentially with the length J of the symbol-spaced ISI channel. Although in some cases the complexity of the VE can be reduced through the use of a reduced-state trellis [12], it still remains an exponential function of J.

It is also well known that an alternative approach to reduce the complexity of the VE solution is the use of adaptive equalization techniques [8]-[10]. At the cost of a performance loss, the complexity of adaptive equalizers is a linear function of J. Consequently, it becomes of practical interest to study the performance of adaptive equalizers in UWB communications.

An adaptive equalizer needs to be trained, either by using a pilot sequence and estimating the channel to provide an initial setting of the coefficients [13], or by the use of a training sequence. For UWB applications, fast acquisition becomes important and channel estimation is the preferred method. In the simulation results presented below, we used a training sequence of 10000 symbols to initialize the equalizer. To provide a justification for this choice of training sequence length, Figs. 3 and 4 show the performance of 95-tap linear and nonlinear adaptive equalizers for an extreme multipath channel (CM4 type), respectively. Although the simulation results in this work are obtained using a training sequence,

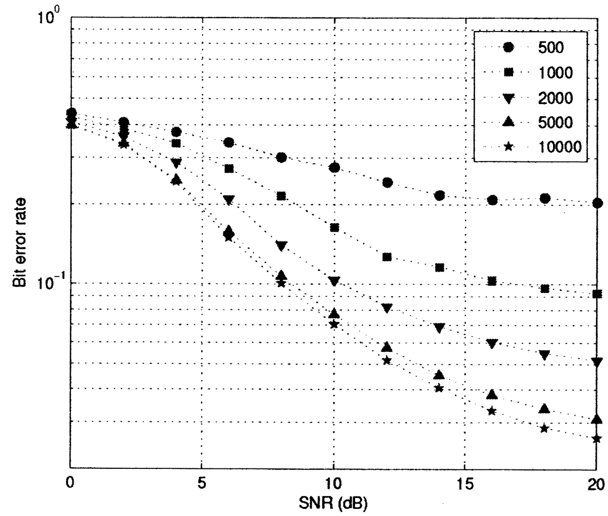

Fig. 3. Performance of a 95-tap linear equalizer with different number pilot symbols. CM4 channel.

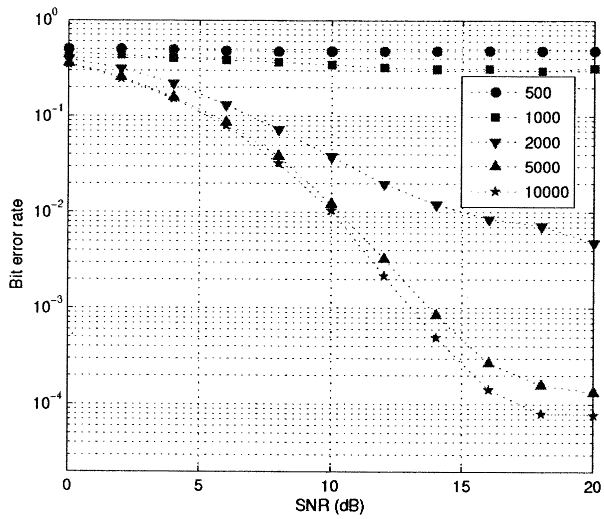

Fig. 4. Performance of a 95-tap nonlinear equalizer ($M = 63$, $M_f = 32$) with different number pilot symbols. CM4 channel.

the results and conclusions can be extended to the case of channel estimation using pilot sequences [13].

A. Linear equalization

An adaptive linear equalizer (LE) is an FIR filter in which M tap coefficients $\{w_m\}$ are updated in order to optimize a given cost function. Here, we use the mean square error (MSE) as the cost function. The coefficients are modified in order to minimize the MSE using the least mean square (LMS) algorithm, as follows. The output of the LE at time $t = nT$ is

$$c_n = \sum_{m=0}^{M-1} w_m Y_m. \qquad (4)$$

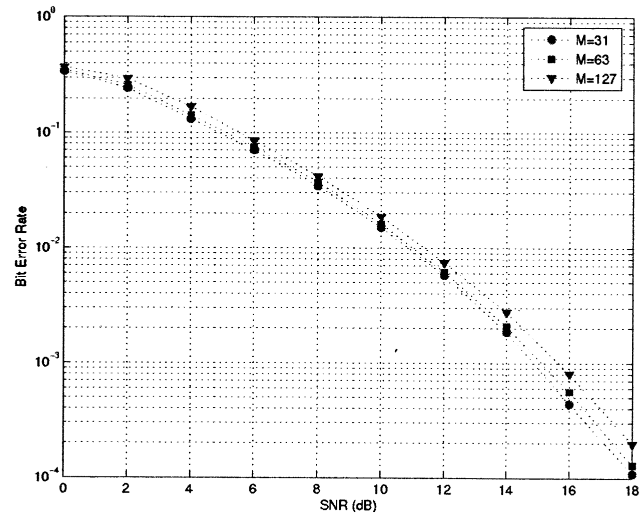

Fig. 5. Linear equalizer with different number of taps. CM1 channel.

Let the error sample be defined as $e_n = \hat{c}_n - c_n$, where $\hat{c}_n = \mathrm{sgn}(c_n)$ is the output of the BPSK slicer in decision-directed mode and a pilot symbol in training mode. Then the coefficients are updated via

$$w_m(n+1) = \mu\, w_m(n) + \Delta\, Y_n\, e_n, \qquad (5)$$

where Δ is the step size and μ is the forgetting factor, $0 < \mu \leq 1$. Simulation results, not reported here, show that $\Delta = 0.00085$ and $\mu = 0.75$ are good choices.

Fig. 5 shows the performance of an LE with different numbers of taps M over a UWB channel type CM1. From these results, it evident that the performance variation is relatively small, provided that the equalizer length M is larger than the ISI length J of the UWB channel. However, note that over an extreme multipath density channel (type CM4), Fig. 6 shows that performance may degrade considerably, not only in comparison to the CM1 channel, but also increasing as a function of M.

B. Nonlinear decision-feedback equalization

One way to improve the performance of a linear equalizer is by feeding back previous decisions. The resulting structure is nonlinear and known as a decision-feedback equalizer (DFE). Nonlinear equalizers for wideband communications have been studied extensively [13]-[20]. The output of the DFE at time $t = nT$ is now

$$c_n = \sum_{m=0}^{M-1} w_m Y_m + \sum_{\ell=1}^{M_f} v_m \hat{c}_{n-\ell}, \qquad (6)$$

where, as before, $e_n = \hat{c}_n - c_n$. Using the LMS algorithm, the feedforward and feedback coefficients are updated via

$$w_m(n+1) = \mu\, w_m(n) + \Delta\, Y_n\, e_n \qquad (7)$$
$$v_m(n+1) = \mu\, v_m(n) + \Delta\, Y_n\, e_n. \qquad (8)$$

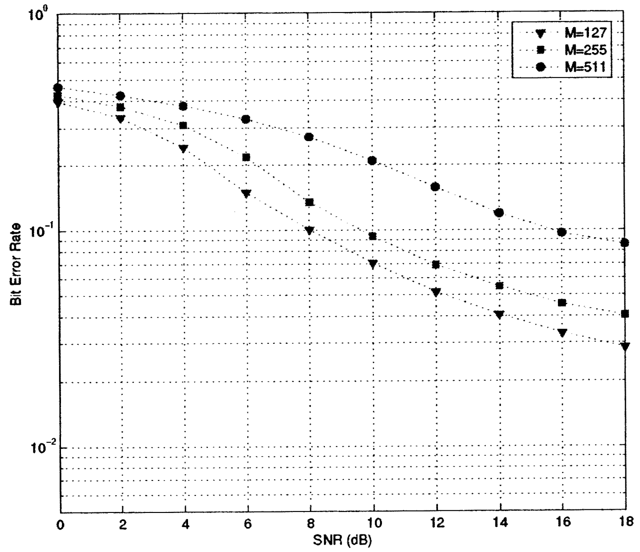

Fig. 6. Linear equalizer with different number of taps. CM4 channel.

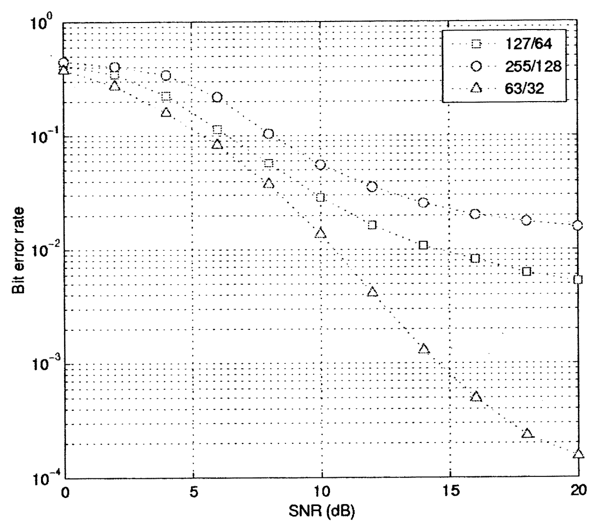

Fig. 7. DFE with various tap configurations. CM4 channel

The DFE parameters were set to the same values as the linear equalizer, i.e., $\Delta = 0.00085$ and $\mu = 0.75$.

Extensive simulations were performed while changing the number of taps in the feedforward (M) and feedback (M_f) filters. In Fig. 7, results obtained with particular choices of M and M_f over a CM4 channel are shown. A particular selection of M and M_f is labeled as M/M_f in the figure. These results suggest that best performance is achieved when the number of taps is kept to a minimum, while at the same time satisfying the condition $M + M_f > J$. With $M + M_f = 95$ taps, the DFE structure has various choices of M and M_f. Fig. 8 shows simulation results of the average BER over 10 cahnnel realizations, supporting the selection $M = 63$ and $M_f = 32$.

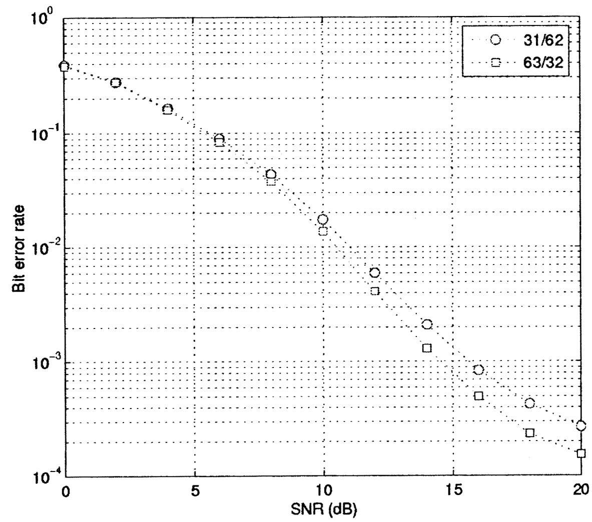

Fig. 8. Two taps distributions for a 95-tap nonlinear equalizer. CM4 channel.

The performance of this DFE is compared with that of a 95-tap LE in the following section.

Another interesting result is the sensitivity of the performance of an adaptive DFE to different channel realizations for the same UWB channel type. This is illustrated for the CM4 channel type in Fig. 9.

Two important observations can be made based on this result: An irreducible error floor may appear due to the fact that the DFE cannot completely remove the ISI. Also, the SNR value required to achieve a particular target bit error rate (BER) is expected to vary at least 3 dB. Similar studies were performed for the other UWB channel types, from the mildly dense multipath channel (CM1 type) to the dense multipath channel (CM3 type). It was found that the SNR variation grows with the multipath density or maximum delay spread, i.e., least variation for CM1 channels and most variation for CM4 channels.

IV. Turbo Product Coding

To improve upon the error performance of adaptive linear and nonlinear equalizers, a turbo product code (TPC) is applied [22]. The selected TPC is constructed from two identical $(31, 26, 3)$ Hamming codes and has a coding rate $R = (26/31)^2 = 0.7034$. Henceforth, we refer to this code as TPC $(31, 26)^2$. Iterative SISO decoding with the Chase type-II algorithm [23] and four decoding iterations was employed.

This channel coding approach is attractive from a practical perspective because of its very low complexity compared to other types of codes and decoding algorithms, while at the same time exhibiting turbo-like error performance. It is interesting to note that a similar TPC scheme, using an expurgated Hamming $(31, 25, 4)$ code, has been adopted as an optional mode in the IEEE 802.16 2004 standard for fixed broadband wireless communications [24].

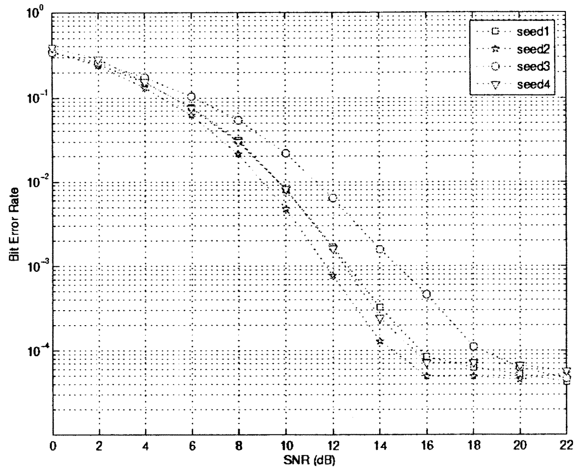

Fig. 9. DFE with four realizations of a CM4 channel.

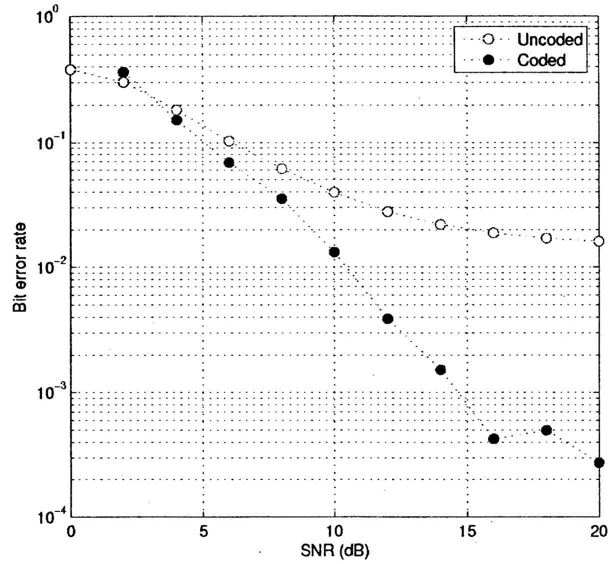

Fig. 10. Performance of TPC $(31,26)^2$ and LE for a CM1 channel.

A. Iterative decoding with Chase type-II algorithm

In the Chase type-II decoding algorithm, the equalizer outputs c_n are scored by their reliability values $|c_n|$. A bit position n is said to be reliable if the value of $|c_n|$ is high. Error patterns $\bar{e}$ are constructed for those code positions with low reliability values. For each error pattern $\bar{e}$, a noisy test vector $\bar{r} = \bar{z} + \bar{e}$ is generated, where $z_n = \text{sgn}(c_n)$ is the n-th component of the hard-decision received vector. The closest codeword $\bar{v}$ to the test vector $\bar{r}$ is determined via a hard-decision decoder. For a $(31,26,3)$ Hamming code, hard-decision decoding is extremely simple, using a combinatorial circuit to implement a syndrome look-up table.

At each decoding iteration with the Chase type-II algorithm, soft-outputs are generated using the two closest codewords, $\bar{v}_1$ and $\bar{v}_2$, to $\bar{z}$. In the event that these codewords are identical, we use the procedure suggested in [25]. In the simulations reported below, four iterations of decoding are performed as suggested by Pyndhia [22]. Increasing weights are used to modify the reliability correction factors when feeding back the extrinsic information in the iterative SISO decoder. Also, for each UWB channel type, 10 channel realizations and corresponding BER values as a function of the signal-to-noise ratio (SNR) were generated and the average BER evaluated. Both linear and nonlinear equalization schemes were examined for each UWB channel type.

B. Simulation results

The pulse-based BPSK modulation UWB system with a 95-tap adaptive LE and the proposed TPC $(31,26)^2$ scheme performed poorly, even over the relatively mild channel types CM1 and CM2, as shown in Figs. 10 and 11, respectively. As the multipath density and delay spread increase, the error performance of the linear equalizer worsens and the TPC is unable to provide any additional improvement. For the other channel types, CM3 and CM4, the performance of the TPC scheme becomes worse than that of the linear equalizer alone because the BER is not low enough to produce the familiar waterfall performance of a turbo code.

On the other hand, when the proposed TPC $(31,26)^2$ scheme is combined with a 95-tap adaptive DFE ($M = 63$ and $M_f = 32$), significant coding gains are observed consistently across all UWB channel types, as shown in Fig. 12. On the average, the variation in the SNR value required to achieve a given target BER is reduced dramatically to about 1 dB for all simulated UWB channel conditions.

It should be noted that TPC $(31,26)^2$ has minimum distance $d_{min} = 9$ and is expected to have better performance than a rate-3/4 punctured convolutional code obtained from a standard 64-state rate-1/2 convolutional code with $d_{min} = 5$ [26], not only because the distance increases, but also because of the use of iterative SISO decoding in TPC $(31,26)^2$.

These results suggest that a powerful capacity-achieving (or "turbo-like") channel coding scheme is useful, not only to provide diversity and error performance improvements, but also to reduce the sensitivity of a UWB communication system to different channel conditions in an indoor environment.

V. Conclusions

In this paper, the performance of adaptive equalization and turbo product coding for UWB communication systems in indoor environments has been studied. It was shown that the performance of an adaptive equalizer is strongly dependent on the number of taps. The number of taps should be selected as small as possible to give good performance, while at the same time larger than the CIR length. The length of a pilot sequence to train or initialize the taps of the equalizer should be large enough to result in good performance.

Our results also indicate that the performance of an adaptive DFE equalizer is very sensitive to channel conditions in an indoor environment. We have shown that this sensitivity

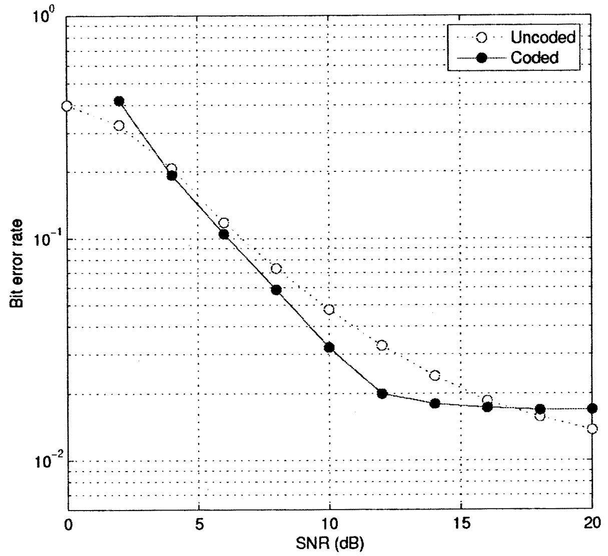

Fig. 11. Performance of TPC $(31,26)^2$ and LE for a CM2 channel.

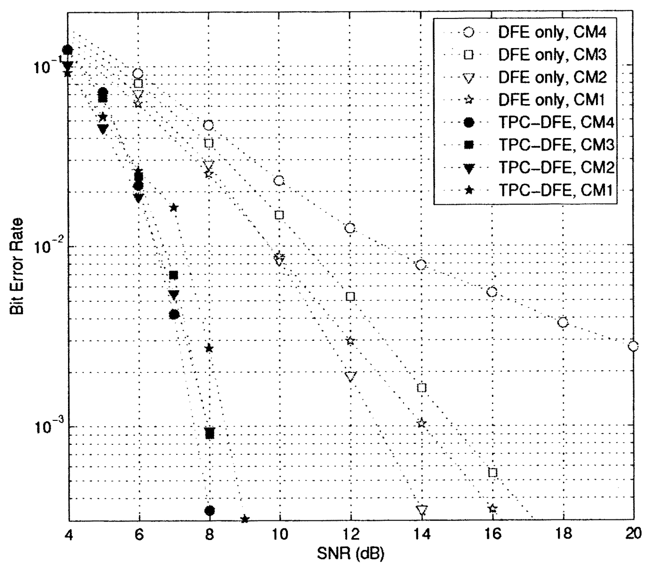

Fig. 12. Performance of TPC $(31,26)^2$ and an adaptive DFE.

can be drastically reduced through the use of turbo coding schemes, such as the TPC $(31,26)^2$ presented in this paper. Better performance will be achieved with the use of more powerful coding schemes, such as parallel and serial concatenated convolutional codes and low-density parity-check (LDPC) codes [26]. The main contribution of this work is understanding the potential increase in the robustness of a UWB communication system, resulting from the combination of an adaptive DFE and a powerful channel coding scheme.

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