Adaptive Sequence Detection of Channel-Interleaved Trellis-Coded Modulation Signals over Multipath Fading ISI Channels

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Abstract— **A new tree-search receiver is proposed for the decoding of the channel-interleaved trellis-code transmitted over fast fading ISI channels. The feedforward channel estimation [1] and the adaptive diversity combining matched filter bank and whitening filter [2] are used, to combine diversity signals and shape the overall response to be causal, symbol-spaced taps. Then, the proposed receiver performs a T-algorithm joint tree-search over the state-machines of the code, the deinterleaver and the causal channel. We use the Ungerboeck's 8-PSK, 8-state trellis code with a modest block interleaving to show that the proposed receiver achieves the available time-diversity benefit of the code for the fast Rayleigh fading ISI channel, with a very moderate increase in the decoding complexity compared to the uncoded DFE.**

I. INTRODUCTION

Trellis coded modulation (TCM) is an efficient coding technique, which achieves the coding gain at no cost in bandwidth. This makes the use of TCM very attractive for land mobile radio communications where the spectrum and the battery power are limited resources.

TCM is originally designed for additive white Gaussian noise (AWGN) channels [4]. The design criterion is to increase the free Euclidean distance d_{free} of the coded sequence. One method of decoding is using the Viterbi algorithm (VA) to search the code trellis for the maximum likelihood sequence that has the minimum Euclidean metric. Considered for a static intersymbol interference (ISI) channel, the optimum decoding of TCM can be achieved by first forming a joint trellis which combines the state machines of the code and the ISI and then employing the Viterbi algorithm to search the joint trellis for the minimum Euclidean metric path. Suboptimal but reduced complexity search techniques such as the reduced state sequence estimation (RSSE) [8], the M-algorithm or the T-algorithm, can also be considered when the number of states of the joint trellis is large.

For fading channels such as Rayleigh or Rician flat-fading channels, the TCM design criterion is to obtain as much signal diversity as possible. Thus, first it is desirable to have the encoded symbols interleaved so as to provide independent fading on adjacent symbols. Then, the primary code design criterion is to increase the length of the shortest error event path; the secondary one is to increase the product of branch distances along that path, to achieve as large time-diversity as possible. The Viterbi decoder or other reduced search techniques can be used to search the deinterleaved sequence.

For fading ISI channels, such as the time-varying, frequency-selective, Rayleigh fading channels we are considering in this paper, the optimum decoder must again search the combined trellis of the encoder and the ISI. However, the use of the interleaver-deinterleaver forbids the formation of a joint trellis due to the prohibitive complexity. Provided an interleaver is not used, a joint trellis can be formed but little signal diversity can be achieved from the use of TCM. It was reported that TCM designed for flat fading channel may bring worse bit error rate (BER) performance than an uncoded modulation, where the receiver uses the VA to search the joint trellis without interleaving [6].

In this paper, we propose a new receiver scheme to decode TCM signals which are interleaved and transmitted over fast Rayleigh fading frequency-selective channels. The receiver employs the feedforward channel estimation techniques [1] and the pre-processing receiver developed in [2]. The pre-processor optimally combines diversity antenna signals and provides the symbol spaced, sufficient statistics and causal overall channel estimates to the post-processing receiver, a sequence estimator using the T-algorithm. The T-algorithm receiver then operates on the combined tree of the TCM encoder, the deinterleaver and the overall channel. In [2] it is shown that for uncoded signal transmission over the fading ISI channels the T-algorithm receiver brings a substantial SNR benefit over a decision feedback equalizer at a moderate increase in complexity. We show here that by the use of T-algorithm the joint decoding can be performed even for interleaved sequences and the efficiency of T-algorithm search is further enhanced while achieving the coding benefit.

This paper is organized as follows. Section II. describes the system in consideration. Section III. explains the receivers. Section IV. presents the simulation results. Section V. provides the conclusion of the paper.

II. THE SYSTEM DESCRIPTION

Fig. 1 (a) describes the baseband equivalent system used for the simulation. It is a part of the complete system, from A to B in Fig. 2. The complete system will be used to explain the operation of the decoding processes in which the detailed system (a) is replaced with the simplified tapped delay line

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(TDL) model (b). More explanation will be followed later in this section.

In Fig. 1, the modulated symbol sequence $\{I_k\}$ is transmitted using the transmit shaping filter (TX) with 35% excess bandwidth. Then, the transmitted signal is received through the L frequency-selective channels, assumed to be mutually independent by the use of L space-diversity antennas. Since the shaping filter employs an excess bandwidth, a half symbol period sampling of the received signal is assumed. Accordingly, the TX, diversity channels and matched filters (MF^l) are realized with half symbol-spaced finite impulse response (FIR) filters. MF^l at each diversity branch is matched to the cascade of TX and the l-th channel Ch^l. Then, the matched filtered signals are combined and symbol-rate sampled.

The mean-square whitening filter (MS-WF) is an anticausal, symbol-spaced FIR filter, which whitens the noise colored by the matched filtering at each diversity branch and provides the "quasi" minimum phase overall channel response between the input symbols $\{I_k\}$ and the output symbols $\{y_k\}$.

In [1], the same diversity combining structure of Fig. 1 is derived under the criterion of minimum mean squares error-DFE. In [2], the same structure is also shown to be the optimum (in MLSE sense) pre-processor to be used with the T-algorithm post-processor for uncoded use. In fact, by the use of a finite length MS-WF, instead of using an infinite length WF, the overall channel is generally not a minimum phase nor is the resulting noise perfectly whitened. The non-causal part of the response tends to vanish but not exactly zero-valued, even at the perfect knowledge of the channel. As it is reported in [2], however, the MS-WF provides a practical and stable solution suitable to be used in the presence of channel estimation error and for channels with a null (nulls) in the folded spectrum (i.e. severe ISI).

For the purpose of T-algorithm search, therefore, the non-causal part of the overall response $\mathbf{f}(k)$ is ignored, and the noise is assumed to be whitened. Then, the input/output relationship between $\{I_k\}$ and $\{y_k\}$is given by as depicted in Fig. 1 (b)

$$y_k = \sum_{i=0}^{N_h} f_i(k)\, I_{k-i} + \eta_k, \tag{1}$$

where N_h is the length of $\mathbf{f}(k)$ and η_k is assumed to be white Gaussian noise.

To update the MF, MS-WF and thus $\mathbf{f}(k)$, we use the channel estimation and tracking methods described in [1]. That is, we assume a contiguous transmission of frames, where a frame constitutes a training segment and an unknown data segment. A set of 4 channel estimates obtained during the training segments is interpolated to track the channel variation for the second data block. From the interpolated channel estimates, the MF^l and the MS-WF are obtained. Readers are directed to [2] for further details of the pre-processing receiver and the procedure to obtain the MF, MS-WF and $\hat{\mathbf{f}}(k)$ from the channel estimates. Main focus of this paper is on the post-processing receiver using the T-algorithm.

Fig. 2 provides the conceptual description of complete system where the detailed system given inside the box of Figure-1 is replaced with the overall channel estimate $\mathbf{f}(k)$ and the equivalent noise η_k. The equally-likely uncoded bits are mapped to the encoded symbol sequence and the modulated sequences are interleaved by the $(N_I \times N_J)$ interleaver before being transmitted. The training sequence of length N_t is inserted into each row of the interleaved sequence as described in Fig. 3, and transmitted row by row. This training symbols are used for the feedforward channel estimation as well as for the start and end of a decoding process (i.e., a sequence starts with a known state and ends to a known state.

III. THE RECEIVERS

In this section, we describe the proposed receiver and the other suboptimal receivers that are in comparison. All of the receivers use the Euclidean distance metric, obtained under the criterion of maximum likelihood sequence estimation (MLSE). Thus, we first begin with the derivation of metric.

A. The Euclidean Distance Metric from MLSE

The maximum likelihood bit-sequence can be determined from

$$\hat{\mathbf{b}} = \arg\max_{\mathbf{b} \in \mathrm{B}} \Pr\{\mathbf{y} \mid \mathbf{b}\}, \tag{2}$$

where $\mathbf{b}$ is the sequence of independent, equally likely bits and B is the set of all possible bit sequences. The code trellis provides a one-to-one mapping from a sequence of bits $\mathbf{b}$ to a sequence of trellis coded modulation symbols $\mathbf{I}(\mathbf{b})$. Therefore, (2) can be rewritten as

$$\hat{\mathbf{b}} = \arg\max_{\dot{\mathbf{I}} \in \{\mathbf{I}(\mathbf{b}) : \mathbf{b} \in \mathrm{B}\}} \Pr\{\mathbf{y} \mid \dot{\mathbf{I}}\}. \tag{3}$$

Then, from the relationship given by (1) we have

$$\hat{\mathbf{b}} = \arg\min_{\dot{\mathbf{I}} \in \{\mathbf{I}(\mathbf{b}) : \mathbf{b} \in \mathrm{B}\}} \{|\mathbf{y} - \dot{\mathbf{y}}|^2\} = \arg\min_{\dot{\mathbf{I}}} \left\{\sum_{k=0}^{N-1} |y_k - \dot{y}_k|\right\} \tag{4}$$

where we used $\dot{\mathbf{y}} := (\dot{y}_{N-1} \dots \dot{y}_0)^t$ and $\dot{y}_k := \sum_{i=0}^{N_h} f_i(k)\, \dot{I}_{k-i}$. The equalities follow from the assumption that η_k is white Gaussian.

B. Joint Decoder without the Use of Interleaver

Now consider a situation where no interleaver is used, i.e. $N_I = 1$. Then, the ISI trellis with M^{N_s} states and the encoder trellis with S states can be readily combined to form a super-trellis, and the complete search of (4) can be performed by the use of VA which searches over the joint-trellis with $O(S \times M^{N_s})$ states. When the number of trellis state becomes too large for the VA to be of any practical use, reduced search techniques can be considered. In [5], the M-algorithm, the T-algorithm and RSSE are applied to decode trellis coded signals transmitted over a static ISI channel, and it is reported that the T-algorithm which operates on the joint trellis achieves the performance of RSSE at much less average computational cost.

In the design of trellis-coded signals for fading channels the primary objective is not to obtain a large free Euclidean distance but to achieve as large a diversity order as possible. Then the potential diversity gain of the code can be achieved fully for an ideal system operating on an independent fading channel, but partially for systems which use the interleaver to implement independent signal fading. When applied to fading channels, therefore, the joint trellis decoder may not provide any coding benefit [9] at all since the interleaver is not used.

In this paper, the joint decoder without an interleaver uses the same T-algorithm with $N_I = 1$ described in section D. This joint decoder employs the least mean squares (LMS) per-survivor channel tracking scheme developed in [2] to enhance the channel estimate and thus improve the decision. The intermittent training symbol sequence enforces the sequence to start from a known state and end to a known state within a frame, and thus the decoding is performed on a per frame basis. We refer to this receiver as 'no interleaver joint T-algorithm' (NI Joint T-alg).

C. Separate Equalization and Decoding

In this system the interleaver-deinterleaver is employed to achieve the potential diversity benefit provided by the TCM. The sequence of symbols $(I_0 I_1 I_2 \ldots)$ are interleaved and the interleaved sequences $\left(I_0 I_{N_I} I_{2N_I} \cdots\right)$ are transmitted over the multipath fading channel as shown in Fig. 2. The equalization and the decoding steps are separated by the deinterleaver as shown inside the receiver box in Fig. 2. The T-algorithm equalization [2] is first performed on the received signals $\{y_k\}$, which are assumed to be corrupted by the overall channel estimate $\mathbf{f}(k)$. The T-algorithm then searches on the interleaved sequence without exploiting the sequence-constraint imposed by the code trellis. The sequence estimator works on a per frame basis using the training symbols at both ends. The T-algorithm equalization provides hard decisions $\left(\hat{I}_0 \hat{I}_{N_I} \hat{I}_{2N_I} \cdots\right)$ on the transmitted symbols in the expanded signal set. These hard decisions are then used to cancel the ISI and generate the sequence of soft equalized outputs $\left(\tilde{I}_0 \tilde{I}_{N_I} \tilde{I}_{2N_I} \cdots\right)$ which if the hard decisions were correct can be described by

$$\tilde{I}_{kN_I+i} = \frac{y_{k+iN_J} - \sum_{p=1}^{N_h} f_p(k+iN_J)\, \hat{I}_{(k-p)N_I+i}}{f_0(k+iN_J)} = I_{kN_I+i} + noise \quad (5)$$

for $k = 0, 1, \ldots, N_J - 1$ and $i = 0, 1, \ldots, N_I - 1$. The LMS per-survivor channel tracking [2] is again used to reduce the number of survivors and improve on the hard decisions.

The deinterleaved soft output sequence $\left(\tilde{I}_0 \tilde{I}_1 \tilde{I}_2 \ldots\right)$ is fed to the Viterbi decoder which searches the code trellis to decide the minimum metric path. We refer this receiver as the 'T-alg. & VA' receiver.

D. The Proposed Joint Tree-Searching T-algorithm Decoding

We now describe the proposed T-algorithm receiver which performs jointly the decoding, deinterleaving and equalization. Since it is a joint search of the maximum likelihood path, there is no information loss due to early decisions (definitely there is some information loss due to the use of the suboptimal T-algorithm search, instead of complete search) nor is "turbo" like iteration required between the ISI trellis and the code trellis. It would be the case of the separate equalization and decoding scheme where a turbo-iteration would be beneficial, the two state machines are separated by the deinterleaver. The result of the iteration would converges to that of the joint-search.

To be elaborated more in the sequel, the T-algorithm follows the code tree, while cancelling the contribution of post-cursor ISI by the use of the tentative decision symbols stored in the survivor-sequence, or by the use of the decided symbols when the depth of the T-algorithm is shorter than $N_I N_h$. The decided symbols are quite reliable since they are results of sequential search, not a symbol by symbol decision as in the case of DFE. That is, the ISI cancelling is carried out in each of the survivors.

We now describe the proposed T-algorithm receiver. Following parameters are pre-selected by the receiver designer:

- P_{max} denotes the maximum number of survivors allowed.
- ζ denotes the threshold value and ζ_b denotes the reduction value.
- N_g, less than or equal to N_h, determines the depth of the tree N_D, i.e., $N_D = N_g \cdot N_{row}$. When $N_g < N_h$, we use the decided symbols to cancel the corresponding ISI.

For the description of the algorithm we use the following notation.

- i denotes the survivor index, i.e., $i = 0, 1, 2, \ldots, P_{max} - 1$.
- j denotes the contender index, i.e., $j = 0, 1, 2, \ldots, MP_{max} - 1$.
- $\mathbf{I}^i_k$ denotes the i-th survivor, $i = 0, 1, 2, \ldots, P_{max} - 1$, a $(N_D \times 1)$ vector which stores a history of hypothetical encoded symbols.
- $\mathbf{S}^i_k$ denotes the history of encoder-states of the i-th survivor.
- $B_{met}(i, q)$ denotes the metric of the branch which is the q-th transition from the state of i-th survivor, and is computed by

$$B_{met}(i, q) = \left| y_k - \sum_{p=1}^{N_h} f_p(k)\, \dot{I}^i_{k-pN_I} - \left(\dot{I}_k(i, q) \cdot f_0(k) \right) \right|, \quad (6)$$

where $\dot{I}_k(i, q)$ represents the modulation symbol defined in the encoder-trellis for the transition.

- $J_{cum}(i)$ the cumulative metric of the i-th survivor.
- $J_{cont}(j)$ denotes the cumulative metric of the j-th contender, i.e.,

$$J_{cont}(j = iN_b + q) = J_{cum}(i) + B_{met}(i, q), \quad (7)$$

where N_b is the number of branches out of a state.

- $\mathbf{D}_I$ denotes the decided symbol sequence.
- $\mathbf{D}_E$ denotes the decided encoder-state sequence.
- P denotes the length of survivor list that is updated at each epoch.

Then, the joint-tree searching T-algorithm can be described as:

- (Step-1) Start from the state-0 of the encoder, and thus set $J_{cum}(0) = 0.0$, $\hat{S}^0_{N_D} = 0$ and the length of the survivor $P = 1$.

Then for each $k = 0, 1, \ldots, N_I \cdot N_J - 1$ the following steps are taken:

- (Step-2) For $i = 0, \ldots, P-1$, extend the i-th survivor into N_b contenders. At each extension step, $J_{cont}(j)$ is computed by (7), the minimum metric J_{min} and the best survivor index i_{min} are updated by a binary comparison ($J_{min} = J_{cont}(0)$), and the survivor-path index i, i.e. $P_{id}(j) = i$, are recorded.
- (Step-3) Mark and count the contenders which pass the threshold test

$$J_{cont}(j) - J_{min} < \zeta \tag{8}$$

and possess the same path-history symbol as the one in the best metric path. If the counter p reaches P_{max} before j reaches PN_b, stop and lower the threshold by ζ_b, and then mark and count again. From the marked paths, generate a survivor list which records the contender's index $S_{id}(p) = j$. P is the size of the survivor list.

- (Step-4) For $p = 0, 1, \ldots, P-1$ obtain the index of the survivors using S_{id} and P_{id}, i.e. $r = P_{id}(S_{id}(p))$, and form the new survivors $\dot{\mathbf{I}}^p_{k+1}$ and $\dot{\mathbf{S}}^p_{k+1}$ by concatenating the new symbol and the new encoder-state which are obtained from the trellis to $\dot{\mathbf{I}}^r_k$ and $\dot{\mathbf{S}}^r_k$ respectively.
- (Step-5) For $k \geq N_D$, release the symbol and the encoder-state of the best metric path to $\mathbf{D}_I$ and $\mathbf{D}_E$.

IV. SIMULATION RESULTS AND DISCUSSION

In this section we study the performance of the proposed receiver via computer simulations using the similar method described in [1]. We use the half-symbol spaced system for the system in Fig. 1. The transmit filter (TX) uses a 9 tap square root raised cosine filter with 35% roll-off, which is a 5-symbol period truncation. Each diversity channel is a 3 tap filter, and each independent Rayleigh fading tap is realized with the sum of nine-sinusoids method as explained in [1][2]. The average powers of the three fading taps are (0.6652,0.2447,0.0901), for which the rms delay spread is about 1/3 the symbol period. During the reception of signal the channel taps are continuously varied according to the given fading rate f_{dm}. As a worst case scenario of 128 km/hr vehicle speed, the fading rate reaches 100 Hz. This requires the frequency of training to be at least every 120 symbols for the purpose of channel interpolation. In this paper we insert $N_t = 11$ training symbols for every $N_I = 69$ symbols, so that a frame consists of 80 symbols ($B = 80$).

A Monte Carlo method with 2,000-50,000 independent trials was used to obtain an averaged performance over the randomly varied channel. To evaluate the adaptation on continuously transmitted frames, each trial consisted of 5-16 frames.

To illustrate the performance of the joint T-algorithm receiver over the 3 path Rayleigh fading channel, we used the 8-PSK, 8 state code [4]. The length of the shortest error event path in the trellis is 2 and the square product Euclidean distance is equal to 8.0. Thus, it provides a potential diversity gain of order 2. The code has 3.6 dB asymptotic coding gain over AWGN.

We simulate all the receiver schemes in section III. for the purpose of comparison with the proposed receiver. All the following receivers use the same number of taps for the preprocessing filters. In particular, the matched filter at each diversity branch uses 12 half symbol-spaced taps. The symbol-spaced MS-WF uses 6 symbol-spaced taps.

- The 'NT-DFE' represents the non-Toeplitz DFE [1] which is used to decode Gray-mapped 4-QAM signals. The feedback filter uses 6-taps.
- The 'NI-Joint T-alg.' stands for the joint T-algorithm receiver described in section III.B. The T-algorithm parameters are $(P_{max}, \zeta, N_D, \Delta) = (1000, 3.5, 50, 0.005)$, where Δ is the stepsize of the least mean squares (LMS) algorithm [2]. We use larger numbers for P_{max} and ζ to approximate a Viterbi algorithm joint decoding.
- The 'T-alg. & VA' implies the receiver scheme described in section III.C. in which the T-algorithm [2] is employed to obtain the equalized, soft-output sequence $\{y'_k\}$, then deinterleaved and fed to the VA decoder which searches the 8-state trellis. That is, the equalization and the decoding are separated by the use of deinterleaver. The T-algorithm parameters are $(P_{max}, \zeta, N_D, \Delta) = (100, 2.5, 50, 0.005)$.
- The 'Joint T-alg.' represents the proposed receiver in section III.D. where the equalization and decoding are jointly performed using the joint T-algorithm. The T-algorithm parameters are $(P_{max}, \zeta, N_D) = (100, 2.5, 50)$.
- The 'Ideal Joint T-alg.' is the Joint T-alg. operating with perfect knowledge of the fading channel. Doppler fading of 200 Hz is used to simulate an ideal interleaving. The T-algorithm parameters are $(P_{max}, \zeta, N_D) = (1000, 2.5, 50)$.

Fig. 4 shows the average BER performance of the receivers at the fading rate $f_{dm} = 100$ Hz. First we note that the NI-Joint T-alg. receiver achieves no coding benefit at all, showing only a slight performance advantage over the NT-DFE at high SNR. On the other hand, the receivers with the use of interleaver/deinterleaver show drastic performance improvement. The 'T-alg. & VA' and 'Joint T-alg.' receivers shows a substantial SNR benefit, which is about 5 - 6 dB for $L = 1$ and 3 - 4 dB for $L = 2$ at the average BER 10^{-4} over the NT-DFE. Comparing the 'Joint T-alg.' receiver with the 'Ideal' receiver the SNR loss due to channel estimation error can be estimated, which is about 5 dB SNR loss for $L = 1$ and 4 dB for $L = 2$.

Comparing the 'T-alg. & VA' and the proposed 'Joint T-alg', the SNR difference is not as dramatic. The 'Joint T-alg' receiver provides SNR benefit less than 1.0 dB compared to the T-alg. & VA receiver. The BER advantage of the Joint T-algorithm, however, is obtained by keeping a far smaller number of survivors on average. Fig. 5 indicates the average number of survivors required for the T-algorithm employed in different receivers. The 'T-alg. & VA' receiver requires more than 60 survivors for equalization alone. Of course, an additional computation is required for the VA decoding. In addition, the overflow percentage of this receiver reaches 100%, suggesting the need to lower the threshold value at the expense of bit error rate increase. On the other hand, the Joint T-alg receiver shows very low average number of survivors, requiring about 10 average survivors in the SNR region where BER is acceptable. The overflow percentage is less than 0.1%.

V. CONCLUSION

We propose a new receiver scheme which may be used to decode the coded symbol sequences transmitted over fast fading frequency-selective diversity channels. It constitutes the pre-processing receiver [2] and the post-processor using the T-algorithm.

The pre-processing receiver optimally combines the diversity channel outputs and provides the symbol-spaced sufficient statistic to the post-processing receiver. As the result of the pre-processing receiver, the overall channel response can be approximated to be a quasi minimum-phase ISI with the additive-noise term whitened (approximately). This is desirable for the T-algorithm processor. The efficiency of T-algorithms depends on the channel response. It makes early decisions to purge based on the metric difference. If the channel has a larger energy taps early in the response, then the metric difference will be larger between the correct and incorrect paths, and a more reliable early decision can be made.

Given the sufficient statistic sequence and the estimate of the minimum phase overall response at our disposal, this paper proposes the way to jointly and computationally-efficient decoding method for trellis-codes which are channel-interleaved and transmitted over the fast time-varying multipath fading ISI channels. A joint-trellis decoding over the code and the minimum phase ISI-trellises is almost impossible due to the use of interleaver-deinterleaver. However, the use of interleaver is necessary to realize the potential diversity benefit of the trellis-code. The separate decoding and equalization scheme is suboptimal due to early decision made at the equalization step.

On the other hand, the proposed T-algorithm is a tree-searching receiver, thus the formation of a joint-trellis is not required and the ISI cancellation and deinterleaving can be performed on a per survivor basis. The decoder follows the code tree and at each survivor cancels the contribution of the minimum phase ISI using the history symbols stored in the survivor. This differentiates the proposed receiver from the separate equalization and decoding receiver, where the ISI cancellation is performed using the hard decision symbols in the extended signal set and the decisions are made without the knowledge of the sequence constraint of the code. Our simulation results show that the proposed joint tree-search T-algorithm brings out the available coding benefit at a very moderate complexity increase, in terms of the average number of survivors.

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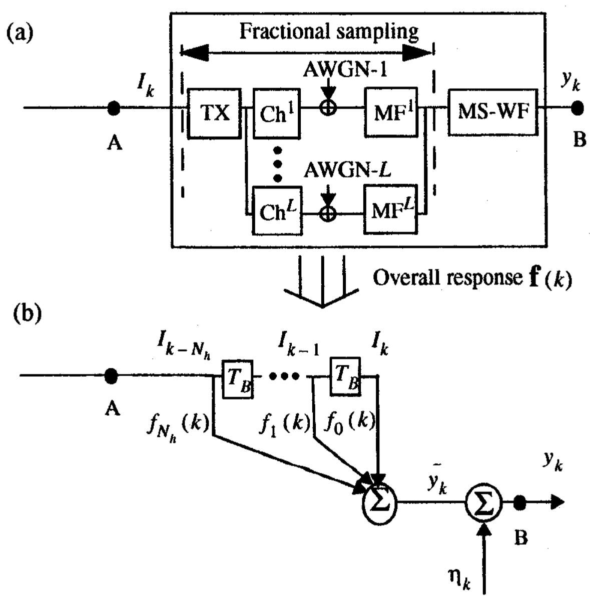

Fig. 1 (a) Baseband system description of the system from A to B in Fig. 2. (b) The symbol-spaced TDL is the model representing the system from A to B for the operation of T-algorithm receiver.

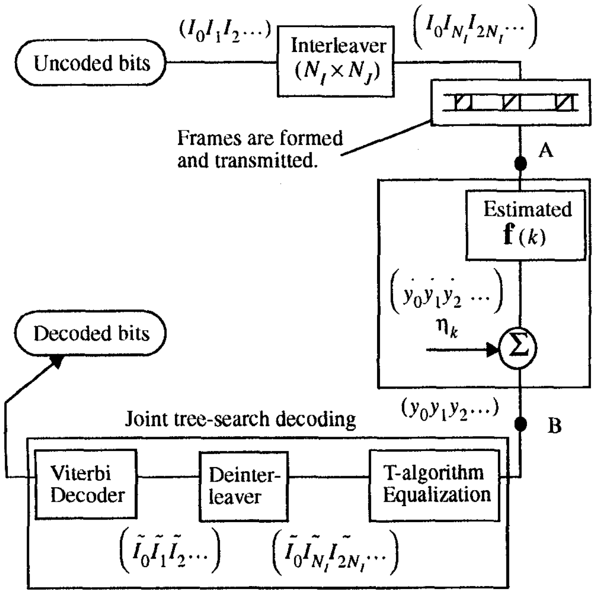

Fig. 2 The block diagram of the overall system: the details of the system from A to B are described in Fig. 1 (a).

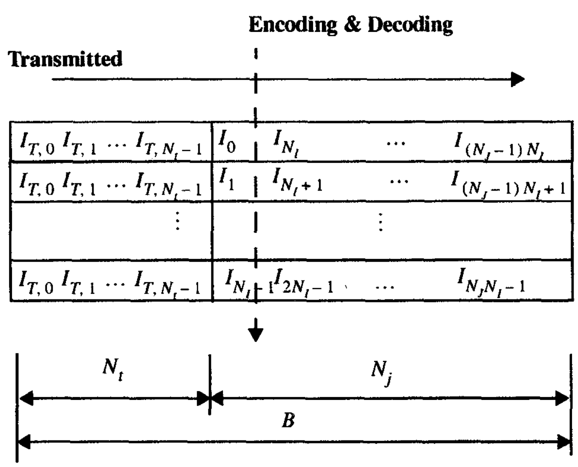

Fig. 3 $N_I \times B$ transmitted symbols. Each row is B symbols, N_t training symbols and N_j unknown symbols.

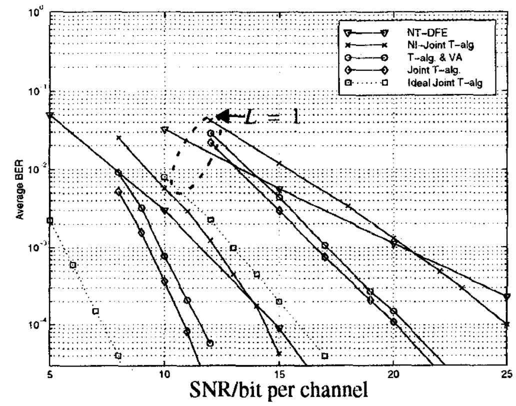

Fig. 4 Average BER performances vs. average SNR/bits/channel

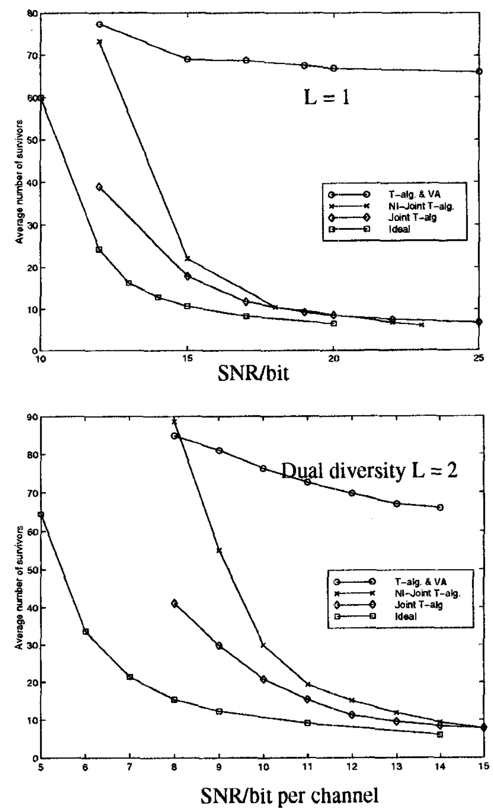

Fig. 5 Average number of survivors.