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Performance of multilevel-turbo codes with blind/non-blind equalization over WSSUS multipath channels

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SUMMARY

In this paper, in order to improve error performance, we introduce a new type of turbo codes, called 'multilevel-turbo codes (ML-TC)' and we evaluate their performance over wide-sense stationary uncorrelated scattering (WSSUS) multipath channels. The basic idea of ML-TC scheme is to partition a signal set into several levels and to encode each level separately by a proper component of the turbo encoder. In the considered structure, the parallel input data sequences are encoded by our multilevel scheme and mapped to any modulation type such as MPSK, MQAM, etc. Since WSSUS channels are very severe fading environments, it is needed to pass the received noisy signals through non-blind or blind equalizers before turbo decoders. In ML-TC schemes, noisy WSSUS corrupted signal sequence is first processed in equalizer block, then fed into the first level of turbo decoder and the first sequence is estimated from this first Turbo decoder. Subsequently, the other following input sequences of the frame are computed by using the estimated input bit streams of previous levels. Here, as a ML-TC example, 4PSK 2 level-turbo codes (2L-TC) is chosen and its error performance is evaluated in WSSUS channel modelled by COST 207 (Cooperation in the field of Science & Technology, Project #207). It is shown that 2L-TC signals with equalizer blocks exhibit considerable performance gains even at lower SNR values compared to 8PSK-turbo trellis coded modulation (TTCM). The simulation results of the proposed scheme have up to 5.5 dB coding gain compared to 8PSK-TTCM for all cases. It is interesting that after a constant SNR value, 2L-TC with blind equalizer has better error performance than non-blind filtered schemes. We conclude that our proposed scheme has promising results compared to classical schemes for all SNR values in WSSUS channels. Copyright © 2005 John Wiley & Sons, Ltd.

KEY WORDS: multilevel-turbo codes; WSSUS; EVA; COST207

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1. INTRODUCTION

The challenge to find practical decoders for large codes has not been considered until turbo codes, which was introduced by Berrou *et al.* in 1993 [1]. Turbo codes are a new class of error correction codes that were introduced along with a practical decoding algorithm. The importance of turbo codes is that they enable reliable communications with power efficiencies close to the theoretical limit predicted by Claude Shannon. Since turbo codes use convolutional codes as their constituent codes, a natural extension of the Turbo concept, which improves bandwidth efficiency, is its application to systems using TCM. The main principle of turbo codes, which is given in Reference [1], is conveyed to TCM and this new scheme is denoted as turbo trellis coded modulation (TTCM). Hence, the important properties and advantages of both structures are retained. Just as binary turbo codes, TTCM uses a parallel concatenation of two binary recursive convolutional encoders, two recursive TCM encoders are concatenated, and interleaving and puncturing are adapted as in References [2, 3].

In trellis-based structures, to improve the bit error probability, many scientists not only study the channel parameters as in References [4, 5] but as in References [6–9] they have also used multilevel coding as an important band and power efficient technique, since it provides significant amount of encoding gain and low coding complexity. Nowadays, there are also many attempts to improve the performance of turbo-based systems. For the same purpose, we construct a new type of turbo codes called as multilevel-turbo codes (ML-TC). Multilevel-turbo coding maps the outputs of more than one encoder into a related symbol each encoder is defined as a level. Since the number of encoders and decoders are the same, for each level of the multilevel encoder, there is a corresponded decoder, defined as a stage. Furthermore, except the first decoding stage, the information bits, which are estimated from the previous decoding stages, are used for the next stages.

Here, performance of ML-TC schemes are evaluated for mobile radio communication channels which are modelled as linear time varying multipath channels. The simplest non-degenerate class of processes, which exhibits uncorrelated dispersiveness in propagation delay and Doppler shift is known as the wide-sense stationary uncorrelated scattering (WSSUS) channel introduced by Bello [10]. In WSSUS, linear superposition of uncorrelated echoes and wide-sense stationary is assumed. Here, WSSUS channel is modelled as in COST 207 (Cooperation in the field of Science & Technology, Project #207) and typical urban (TU), bad urban (BU) and hilly terrain (HT) profiles of standard COST-207 are chosen. Since COST 207 is a severe fading channel, receivers employ various non-blind/blind equalizers such as least mean squares (LMS), recursive least squares (RLS), eigen vector algorithm (EVA).

This paper is organized as follows: multilevel turbo encoders are described in Section 2. In Section 3, Equalization of WSSUS multipath channels are considered. In Section 4, multilevel turbo decoding process is explained. As an example, 2L-TC system is investigated in Section 5. The error performance of the proposed system in WSSUS channel is discussed in Section 6.

2. MULTILEVEL TURBO ENCODER

Multilevel-turbo encoder and decoder consist of many parallel turbo encoder/decoder levels as in Figure 1. At the receiver, there is one binary turbo encoder at every level of multilevel turbo encoder. Firstly, input bit stream is converted from serial to parallel. Each turbo encoder has

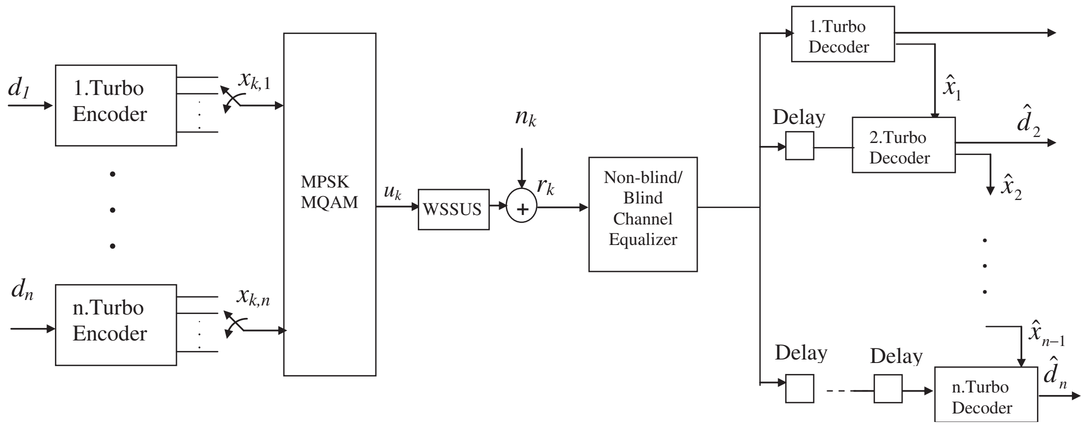

Figure 1. ML-TC structure block diagram.

only one input and encoding is processed simultaneously. The outputs of these encoders can be punctured and then are mapped to any modulated scheme using group set partitioning technique as in Figure 2. Here, if the d_k is the input of multilevel encoder at time k, the encoder output x_k is equal to

$$x_k = f(d_k) \tag{1}$$

where $f(.)$ corresponds to the encoder structure. In partitioning, for the first level encoder, $x_{k,1}$ is the output bit of the first level turbo encoder where signal set is divided into two subsets. If $x_{k,1} = 0$ then the first subset is chosen, if $x_{k,1} = 1$ then the second subset is chosen. The $x_{k,2}$ bit is the output bit of the second level turbo encoder and divides the subsets into two same as before. This partitioning process continues until the last partitioning level is occurred. To clarify the partitioning in multilevel-turbo coding, the signal set for 8-PSK with three level Turbo encoder is chosen as in Figure 2. Here, if the output bit of the first level turbo encoder is $x_{k,1} = 0$, then u_0^1 set, if it is $x_{k,1} = 1$, then u_1^1 set is chosen. The output bits of the second level turbo encoder determines whether u_1^2 or u_0^2 subsets to be chosen. Here, the rule is to transmit u_0^3 signal if the output bit of the third level turbo encoder is 0, and to transmit u_1^3 signal if the output bit is 1.

In Figure 3, as an example, we design multilevel-turbo coding system for 4PSK 2 level-turbo codes (2L-TC). In each level, we consider a 1/3 recursive systematic convolutional (RSC) encoder with memory size $M = 2$. Firstly, bit sequence is converted from serial to parallel. Then, for each level, bit sequences are encoded by the turbo encoders. At turbo encoder outputs, the encoded bit streams to be mapped to M-PSK signals are determined after a puncturing procedure. The first bit is taken from the first level turbo encoder output, the second bit is taken from second level encoder output and the other bits are obtained in similar way. Following this process, the bits at the output of the Turbo encoders are mapped to 4-PSK signals by using group set partitioning technique. In our example, the remainder w_k of the encoder can be found using feedback polynomial $g^{(0)}$ and feedforward polynomial is $g^{(1)}$. The feedback variable is

$$w_k = d_k + \sum_{j=1}^{K} w_{k-j} g_j^{(0)} \tag{2}$$

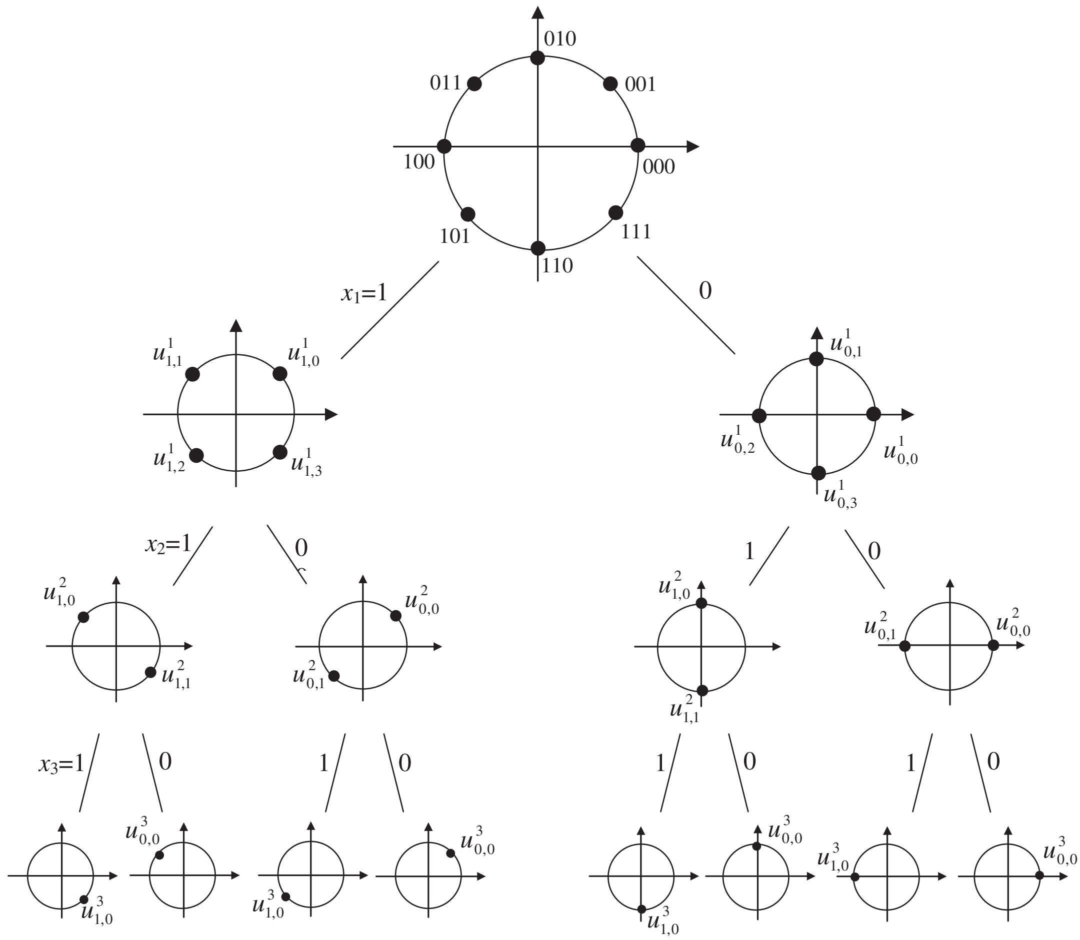

Figure 2. Set partitioning of 3L-TC.

and RSC encoders outputs $x_k^{1,2}$ called parity data are

$$x_k^{1,2} = \sum_{j=0}^{K} w_{k-j} g_j^{(1)} \tag{3}$$

In our case, RSC encoder has feedback polynomial $g^{(0)} = 7$ and feedforward polynomial $g^{(1)} = 5$, and it has a generator matrix

$$G(D) = \left[1 \quad \frac{1 + D + D^2}{1 + D^2}\right] \tag{4}$$

where D is memory unit.

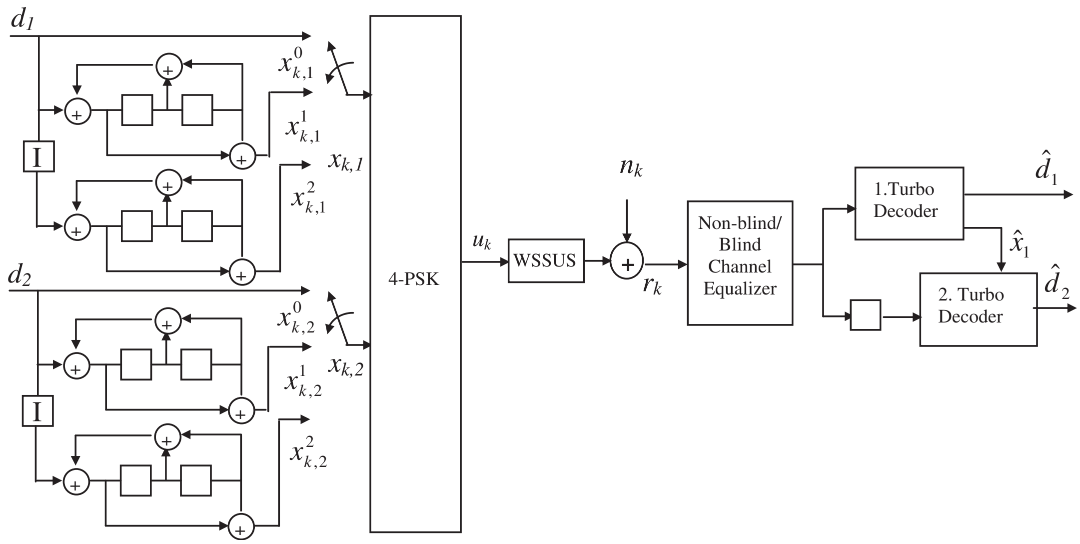

Figure 3. 2L-TC structure for WSSUS channel.

3. EQUALIZATION OF WIDE-SENSE STATIONARY UNCORRELATED SCATTERING CHANNELS

The main effects that cause and influence multipath propagation, are reflection, scattering, shadowing, diffraction and path loss. The multipath propagation results in frequency and time selective fading of the received signal at the receiver. In digital communication systems require the knowledge of the transmission channel's impulse response. Since this knowledge usually is not available, the problem of channel estimation arises. From the point of view of systems theory, channel estimation is a particular form of (linear) system identification which, in our case, is complicated by three main properties of the radio channel: (i) it consists of multiple propagation paths and is therefore frequency selective, (ii) its discrete-time equivalent baseband impulse response may be mixed phase and (iii) in a mobile environment, it is time-variant [11]. Because of the movement of the mobile station and, hence, the changing channel characteristics, the channel has to be treated as a time-variant, multipath fading channel. This can be characterized by statistical channel models such as WSSUS models. WSSUS models require only two sets of parameters to characterize fading and multipath effects: the power delay profiles and Doppler power spectra. Therefore, these can be efficiently be used within software (and hardware) simulation systems. Both parameter sets can be described by a single function, which is called scattering function. In a mobile scenario, the physical multipath radio channel is time-variant with a baseband impulse response depending on the time difference τ between the observation and excitation instants as well as the (absolute) observation time t. We adopt the stochastic zero mean Gaussian stationary uncorrelated scattering (GSUS) model leading to the following impulse response of the composite channel [12]

$$h^{\mathrm{c}}(\tau, t) = \frac{1}{\sqrt{N_{\mathrm{e}}}} \sum_{v=1}^{N_{\mathrm{e}}} \mathrm{e}^{\mathrm{j}(2\pi f_{d,v} t + \Theta_v)} \cdot g_{\mathrm{TR}}(\tau - \tau_v) \tag{5}$$

where N_e is the number of elementary echo paths, $g_{TR}(\tau)$ denotes the combined transmit/receiver filter impulse response, and the subscript in $h^c(\cdot)$ suggests its continuous-time property. Three-dimensional sample impulse responses can easily be determined from (5) by Doppler frequencies $f_{d,v}$ initial phases Θ_v and echo delay times τ_v from random variables with Jakes, uniform, and piecewise exponential probability density functions, respectively.

In this study, the performance of multilevel-turbo codes is investigated and as an example, 4PSK 2L-TC signals are transmitted over WSSUS multi path channel and corrupted by AWGN. During transmission, the signals are generally corrupted due to the severe transmission conditions of WSSUS channels. So it is needed to minimize these effects and receive the signals with minimum errors by using equalizers and decoders together. In order to improve performance of ML-TC signals; non-blind equalizers such as LMS, RLS and blind equalizer as eigen vector algorithm (EVA) are used. More information on equalizers is given below.

3.1. Non-blind equalization

If both the received sequence $r(k)$ and some transmitted data $u(k)$ (training sequence) are given the minimum mean square error (MMSE-(ℓ, k_0)) equalizer coefficients can be calculated using the well-known normal equation [11]

$$\mathbf{e}_{\mathrm{MMSE}(k_0)} = \mathbf{R}_{rr}^{-1}\mathbf{r}_{ru} \tag{6}$$

with

$$\mathbf{r}_{ru} \cong E\{\mathbf{r}_k u(k-k_0)\} \tag{7}$$

$$\mathbf{R}_{uu} \cong E\{u_k u_k^*\} \tag{8}$$

where $\mathbf{r}_{ru}$ and $\mathbf{R}_{uu}$ denote the cross-correlation vector and the non-singular $(l+1)\times(l+1)$ Hermitian Toeplitz autocorrelation matrix, respectively, and the vectors r_k and conjugate transpose form, r_k^* are defined as

$$\mathbf{r}_k \cong [r^*(k), r^*(k-1), \ldots, r^*(k-l)]^{\mathrm{T}} \tag{9}$$

$$\mathbf{r}_k^* \cong [r(k), r(k-1), \ldots, r(k-l)] \tag{10}$$

Then the MMSE-(ℓ, k_0) equalizer is as

$$\mathbf{e}_{\mathrm{MMSE}(k_0)} \cong [e_{\mathrm{MMSE}}(o), \ldots, e_{\mathrm{MMSE}}(\ell)]^{\mathrm{T}} \tag{11}$$

In the noiseless case, it approximates the channel's inverse system (deconvolution, zero forcing) in order to minimize intersymbol interference (ISI). If additive noise is present, however, its coefficients adjusted differently so as to minimize the total mean squared error (MSE) in the equalized sequence due to ISI and noise.

3.2. Blind equalization

The fundamental idea of blind channel estimation is to derive the channel characteristics from the received signal only, i.e. without access to the channel input signal by means of training sequences. Depending on the different ways to extract information from the received signal, some classes of algorithms such as EVA can be used [11]. Figure 4 shows block diagram of EVA

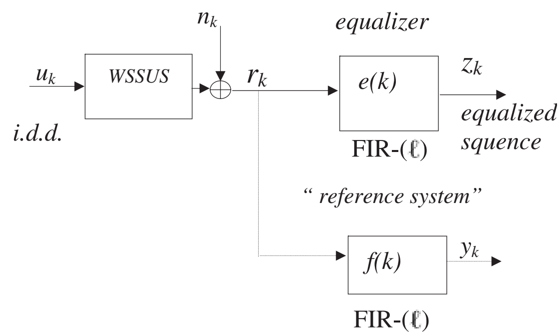

Figure 4. Block diagram of blind equalizer (EVA).

equalizer. The transmitted data u_k are an independent, identically distributed (i.i.d.) sequence of random variables with zero mean, variance σ_u^2, skewness γ_3^u, and kurtosis γ_4^u [11]. Each symbol period T, u_k takes a (possibly complex) value from a finite set. For this reason, the channel input random process clearly is non-Gaussian with a non-zero kurtosis ($\gamma_4^u \neq 0$), while its skewness vanishes ($\gamma_3^u = 0$) due to the even probability density function of typical digital modulation signals such as phase shift keying (PSK), quadrature amplitude modulation (QAM), etc. The objective is to determine the MMSE-(ℓ, k_0) equalizer coefficients without access to the transmitted data, i.e. from the received sequence r_k only. Similar to Shalvi and Weinstein's maximum kurtosis criterion [13,14], the EVA solution to blind equalization is based on a maximum 'cross-kurtosis' quality function [15]

$$\begin{aligned} c_4^{zy}(0,0,0) &= E\{|z_k|^2|y_k|^2\} - E\{|z_k|^2\}E\{|y_k|^2\} \\ &\quad - |E\{z_k^* y_k\}|^2 - |E\{z_k y_k\}|^2 \end{aligned} \tag{12}$$

i.e. the cross-kurtosis between u_k and r_k, can be used as a measure for equalization quality:

$$\max|c_4^{zy}(0,0,0)| = \begin{cases} \text{subject to} \\ r_{zz}(0) = \sigma_u^2 \end{cases} \tag{13}$$

The equalizer output z_k, can easily be expressed in terms of the equalizer input r_k by replacing z_k in Equation (12) with

$$z_k = r_k^* e(k) = \mathbf{r}_k^* \mathbf{e} \tag{14}$$

where '*' denotes the convolution operator. In this way, we obtain from (13)

$$\max|\mathbf{e}^* \mathbf{C}_4^{yr} \mathbf{e}| = \begin{cases} \text{subject to} \\ \mathbf{e}^* \mathbf{R}_{rr} \mathbf{e} = \sigma_u^2 \end{cases} \tag{15}$$

where the Hermitian $(l+1) \times (l+1)$ cross-cumulant matrix

$$\begin{aligned} \mathbf{C}_4^{yr} \cong\ & E\{|y_k|^2 \mathbf{r}_k \mathbf{r}_k^*\} \\ & - E\{|y_k|^2\} E\{\mathbf{r}_k \mathbf{r}_k^*\} \\ & - E\{y_k \mathbf{r}_k\} E\{y_k^* \mathbf{r}_k^*\} \\ & - E\{y_k^* \mathbf{r}_k\} E\{y_k \mathbf{r}_k^*\} \end{aligned} \tag{16}$$

contains the fourth order cross-cumulant

$$[\mathbf{C}_4^{yr}]_{i_1,i_2} = c_4^{yr}(-i_1, 0, -i_2). \tag{17}$$

in row i_1 and column i_2 with $i_1, i_2 \in (0, 1, \ldots, \ell)$. The quality function (15) is quadratic in the equalizer coefficients. Its optimization leads to a closed-form expression in guise of the generalized eigenvector problem [15]

$$\mathbf{C}_4^{yr}\mathbf{e}_{\mathrm{EVA}} = \lambda \mathbf{R}_{rr}\mathbf{e}_{\mathrm{EVA}} \tag{18}$$

which we term 'EVA equation'. The coefficient vector

$$\mathbf{e}_{\mathrm{EVA}} \cong [e_{\mathrm{EVA}}(0), \ldots, e_{\mathrm{EVA}}(\ell)]^{\mathrm{T}} \tag{19}$$

obtained by choosing the eigenvector of $\mathbf{R}_{rr}^{-1}\mathbf{C}_4^{yr}$ associated with the maximum magnitude eigenvalue λ is called the 'EVA-(ℓ) solution' to the problem of blind equalization.

4. MULTILEVEL TURBO DECODER

The problem of decoding turbo codes involves the joint estimation of two Markov processes, one for each constituent code. While in theory, it is possible to model a turbo code as a single Markov process, such a representation is extremely complex and does not lend itself to computationally tractable decoding algorithms. In Turbo decoding, there are two Markov processes which are defined by the same set of data, hence, the estimation can be refined by sharing the information between the two decoders in an iterative fashion. More specifically, the output of one decoder can be used as *a priori* information by the other decoder. If the outputs of the individual decoders are in the form of hard-bit decisions, then there is little advantage to sharing information. However, if soft-bit decisions are produced by the individual decoders, considerable performance gains can be achieved by executing multiple iterations of decoding. The received signal can be shown as,

$$r_k = \sum u_k h_{k-l}^{\mathrm{c}} + n_k \tag{20}$$

where r_k is noisy received signal, u_k is transmitted M-PSK signal, h_k^{c} is the slice of the time-varying composite channel's impulse response and n_k is Gaussian noise at time k.

The maximum *a posteriori* (MAP) algorithm can calculate the *a posteriori* probability of each bit with a perfect performance. Let $\bar{\gamma}^{\mathrm{st}}(s_k \to s_{k+1})$ denote the natural logarithm of branch metric $\gamma^{\mathrm{st}}(s_k \to s_{k+1})$, where s_k is state at time k and st is the decoding stage, then

$$\bar{\gamma}^{\mathrm{st}}(s_k \to s_{k+1}) = \ln \gamma^{\mathrm{st}}(s_k \to s_{k+1})$$

$$\ldots = \ln P[d_k] + \ln P[r_k|x_k] \tag{21}$$

and

$$\ln P[d_k] = z_k d_k - \ln(1 + \mathrm{e}^{z_k}) \tag{22}$$

z_k is the *a priori* information which is obtained from the output of the other decoder.

For every decoding stage of ML-TC, noisy BPSK inputs are evaluated. These noisy input sequences are obtained from the 1 and 0 probabilities of the received signal, which are given below.

$$P_{k,0}^{\mathrm{st}} = \sum_{j=0}^{M/(2\mathrm{st})-1} \frac{1}{(r_k - u_{0,j}^{\mathrm{st}})^2} \tag{23a}$$

$$P_{k,1}^{\mathrm{st}} = \sum_{j=0}^{M/(2\mathrm{st})-1} \frac{1}{(r_k - u_{1,j}^{\mathrm{st}})^2} \tag{23b}$$

$P_{k,0}^{\mathrm{st}}$ and $P_{k,1}^{\mathrm{st}}$ indicate 0 and 1 probabilities of the received signal at time k and stage st. st is decoding stage and $\mathrm{st} \in \{1, 2, \ldots, \log_2 M\}$. In ML-TC scheme, each digit of binary correspondence of MPSK signals, matches to one stage from most significant to least significant while stage level st increases. Signal set is partitioned into the subsets due to the each binary digit matching stage depending on whether it is 0 or 1 as revealed in Figure 2 .The subsets $u_{0,j}^{\mathrm{st}}$ and $u_{1,j}^{\mathrm{st}}$ are chosen from the signal set at stage st according to the partitioning rule. In partitioning rule, the subscripts $\{0, 1\}$ of $u_{0,j}^{\mathrm{st}}$ and $u_{1,j}^{\mathrm{st}}$ indicate the subset, which is related to the binary digit 0 or 1 at stage st. When decoding stage increases, MPSK signals are chosen among the adequate signal set regarding to the previous decoding stage as in Figure 2.

After computing the 1 and 0 probabilities as in Equations (23a) and (23b), received signal is mapped to one-dimensional BPSK signal as

$$\xi_k^{\mathrm{st},q} = 1 - \frac{2P_{k,0}^{\mathrm{st}}}{P_{k,0}^{\mathrm{st}} + P_{k,1}^{\mathrm{st}}} \tag{24}$$

These probability computations and mapping are executed in every stage of decoding process according to the signal set, which is shown in Figure 2. Thus, Equation (24) becomes

$$\bar{\gamma}^{\mathrm{st}}(s_k \to s_{k+1}) = \ln P[d_k] - \frac{1}{2}\ln(\pi N_0/E_s) - \frac{E_s}{N_0}\sum_{q=0}^{n-1}[\xi_k^{\mathrm{st},q} - (2x^q - 1)]^2 \tag{25}$$

Now let $\bar{\alpha}^{\mathrm{st}}(s_k)$ be the natural logarithm of $\alpha^{\mathrm{st}}(s_k)$,

$$\begin{aligned}\bar{\alpha}^{\mathrm{st}}(s_k) &= \ln \alpha^{\mathrm{st}}(s_k)\\ &= \ln\left\{\sum_{s_{k-1}\in A} \exp[\bar{\alpha}^{\mathrm{st}}(s_{k-1}) + \bar{\gamma}^{\mathrm{st}}(s_{k-1} \to s_k)]\right\}\end{aligned} \tag{26}$$

where A is the set of states s_{k-1} that are connected to the state s_k. Now let $\bar{\beta}^{\mathrm{st}}(s_k)$ be the natural logarithm of $\beta^{\mathrm{st}}(s_k)$,

$$\begin{aligned}\bar{\beta}^{\mathrm{st}}(s_k) &= \ln \beta^{\mathrm{st}}(s_k)\\ \ldots &= \ln\left\{\sum_{s_{k+1}\in B} \exp[\bar{\beta}^{\mathrm{st}}(s_{k+1}) + \bar{\gamma}^{\mathrm{st}}(s_k \to s_{k+1})]\right\}\end{aligned} \tag{27}$$

where B is the set of states s_{k+1} that are connected to state s_k , and we can calculate the log likelihood ratio (LLR) by using

$$\Lambda_k^{\mathrm{st}} = \ln\frac{\sum_{S_1} \exp[\bar{\alpha}^{\mathrm{st}}(s_k) + \bar{\gamma}^{\mathrm{st}}(s_k \to s_{k+1}) + \bar{\beta}^{\mathrm{st}}(s_{k+1})]}{\sum_{S_0} \exp[\bar{\alpha}^{\mathrm{st}}(s_k) + \bar{\gamma}^{\mathrm{st}}(s_k \to s_{k+1}) + \bar{\beta}^{\mathrm{st}}(s_{k+1})]} \tag{28}$$

where $S_1 = \{s_k \to s_{k+1} : d_k = 1\}$ is the set of all state transitions associated with a message bit of 1, and $S_0 = \{s_k \to s_{k+1} : d_k = 0\}$ is the set of all state transitions associated with a message bit of 0. At the last iteration we make the hard decision by using the second decoder output $\Lambda(1)$,

$$\hat{d}_k = \begin{cases} 1 & \text{if } \Lambda(2) \geqslant 0 \\ 0 & \text{if } \Lambda(2) < 0 \end{cases} \tag{29}$$

5. PERFORMANCE OF ML-TC SIGNALS OVER WSSUS MULTIPATH CHANNEL

In this section, we evaluate the performance of multilevel-turbo coded signals for WSSUS multipath channels. In our study, 4PSK 2L-TC scheme with 1/3 rate RSC encoder and over all 2/3 coding rate, the generator matrix $g = [111 : 101]$ and random interleavers are used. The frame size of information symbols (bits) is chosen as $N = 256$. Information bits are composed of data bits and training bits for non-blind (LMS and RLS) channel equalization. For blind equalization, almost the entire information bits (256 bits) are used without training bits.

As for the echo delay time τ, we use standard COST-207 TU, BU and HT profiles. The samples of magnitude impulse responses $|h^c(\tau, t)|$, are obtained from Equation (5) of COST207 (TU, BU and HT). Both time axes are normalized to the GSM symbol (bit) period $T \approx 3.7\,\mu s$. The velocity of the mobile unit v is 100 km/h. Assuming a carrier frequency of 950 MHz, this leads to a maximum Doppler shift of $f_{d,\max} = 88$ Hz. Equation (5) is evaluated over a t range covering one minimum Doppler period $T_{d,\min} = 1/f_{d,\max} = 3080T = 11.4$ ms. For simulation, seventeen the channel coefficients of COST207 (TU, BU and HT) are used as shown in Table I for one path delay ($\tau/T = 0.25$).

Table I. Channel coefficients of $h^c(\tau = 0.25, t)$ of TU, BU and HT types of COST207.

	COST207		
Path number	TU	BU	HT
1	−0.1651 + 0.5534i	−0.0484−0.2943i	−1.0607 + 0.2118i
2	−0.0506 + 0.7480i	−0.0000−0.4282i	−1.3775−0.0015i
3	0.0849 + 0.9113i	0.0447−0.5339i	−1.5213−0.1868i
4	0.2169 + 1.0292i	0.0815−0.6063i	−1.4957−0.3254i
5	0.3215 + 1.0907i	0.1064−0.6414i	−1.3252−0.4056i
6	0.3778 + 1.0881i	0.1162−0.6368i	−1.0507−0.4232i
7	0.3702 + 1.0178i	0.1090−0.5919i	−0.7242−0.3814i
8	0.2910 + 0.8806i	0.0839−0.5077i	0.4014−0.2902i
9	0.1403 + 0.6815i	0.0418−0.3872i	−0.1342−0.1641i
10	−0.0733 + 0.4294i	−0.0152−0.2351i	0.0356−0.0203i
11	−0.3338 + 0.1373i	−0.0835−0.0577i	0.0820 + 0.1237i
12	−0.6196−0.1790i	−0.1582 + 0.1375i	−0.0021 + 0.2523i
13	−0.9057−0.5011i	−0.2335 + 0.3418i	−0.2047 + 0.3541i
14	−1.1663−0.8090i	−0.3030 + 0.5460i	−0.4962 + 0.4224i
15	−1.3771−1.0820i	−0.3599 + 0.7407i	−0.8341 + 0.4562i
16	−1.5177−1.3001i	−0.3978 + 0.9165i	−1.1686 + 0.4600i
17	−1.5736−1.4452i	−0.4108 + 1.0649i	−1.4497 + 0.4420i

In a mobile environment, transmission channel is time-variant so that the equalizer needs to be adjusted repeatedly. Time-variance of the channel is relatively slow in many applications, so that it can be assumed time-invariant over a certain period of time (piecewise or quasi time-invariant). WSSUS channels have fading statistics that remain constant over short periods of time [16]. It is assumed that the channel coefficients are quasi (stationary) during transmitting of one frame size ($N = 256$ bits).

Comparable statements can be made for the performance of multilevel-turbo coding: For $\tau/T = 0.25$, bit error ratio (BER) versus signal-to-noise ratio (SNR) curves of 4PSK 2 level-turbo coding system (2L-TC) are obtained for COST207 (TU, BU and HT) by using various algorithms (RLS, LMS and EVA) as in Figures 5–7. We realize that without equalization, it is impossible to estimate information bits. Thus equalization block is necessary for all schemes in WSSUS channels. BER performance improvement of our proposed schemes with equalizers increases exponentially after BER of 10^{-1} in all figures. The performance of 2L-TC_RLS (2L-TC with RLS equalizer) is better then the performance of 2L-TC_LMS (2L-TC with LMS equalizer) and 2L-TC_EVA (2L-TC with EVA equalizer). In Figure 5, the performance of 2L-TC_LMS is better than 2L-TC_EVA except 8 dB. After 7.8 dB, the curves of 2L-TC_EVA cross over the curves of 2L-TC_LMS and approaches to 2L-TC_RLS. In Figures 6 and 7, the performance of 2L-TC_EVA is better than 2L-TC_LMS at 9 and 12 dB, respectively. In general, equalization of 2L-TC with RLS, LMS and EVA exhibit good performance as reaching 10^{-5} BER values at 8 dB for COST207 TU, 10^{-5} BER at 9 dB for COST207 BU and 10^{-4} BER at 12 dB for COST207 HT.

In order to evaluate performance of 4PSK 2L-TC, we compare our results with 8PSK-TTCM, since both of their overall coding rate is 2/3. From Figures 5–7, bit error rates of 8PSK-TTCM (no equalization) and 4PSK 2L-TC (no equalization) are quite bad at all SNR values. Thus equalization block is necessary for WSSUS environments. All types of 2L-TC have better error performance than TTCM after 2 dB for TU and BU and after 4 dB for HT types of COST 207. As an example, in TU type of COST 207 channels, 2L-TC_RLS (3. iteration) reaches bit error rate of 10^{-3} at 4.5 dB, while TTCM_RLS (3. iteration) reaches the same BER at 10 dB, thus coding gain is about 5.5 dB (Figure 5). In BU type of COST 207 channels, 2L-TC_RLS (3. iteration) reaches bit error rate of 10^{-3}at 6 dB, while TTCM_RLS (3. iteration) reaches the same BER at 10.5 dB, thus coding gain is about 4.5 dB (Figure 6). In HT type of COST 207 channels, 2L-TC_RLS (3. iteration) reaches bit error rate of 10^{-3} at 9.7 dB, while TTCM_RLS (3. iteration) reaches the same BER at 13 dB, thus coding gain is about 3.3 dB (Figure 7).

The performance of EVA is superior to LMS at high SNR values, because EVA has a specific feature which updates a closed-form solution iteratively similar to RLS. As it is mentioned in Reference [16], after some iterations, EVA converges to the optimum linear equalizer solution and exhibits excellent convergence. For constant modulus signals, this blind approach converges as fast as the non-blind RLS algorithm. When convergence performance of blind (EVA) is compared to non-blind (RLS); EVA uses the fourth order cumulants so closely approaches the convergence performance of RLS for which the frame size > 140 [17]. We have chosen frame size $N = 256$ in the simulations. EVA is robust with respect to errors in the correlation and cumulants estimates, because they cancel each other, to some extent in the EVA equation (24). It has been demonstrated that some performance gain can be obtained by using a lattice prewhitening filter followed by a cascade of first order all-pass systems [18]. However, it has found no way approximate RLS' performance as closely as EVA does. As such equalizer structures suffer from several principal disadvantages (such as the additional time the

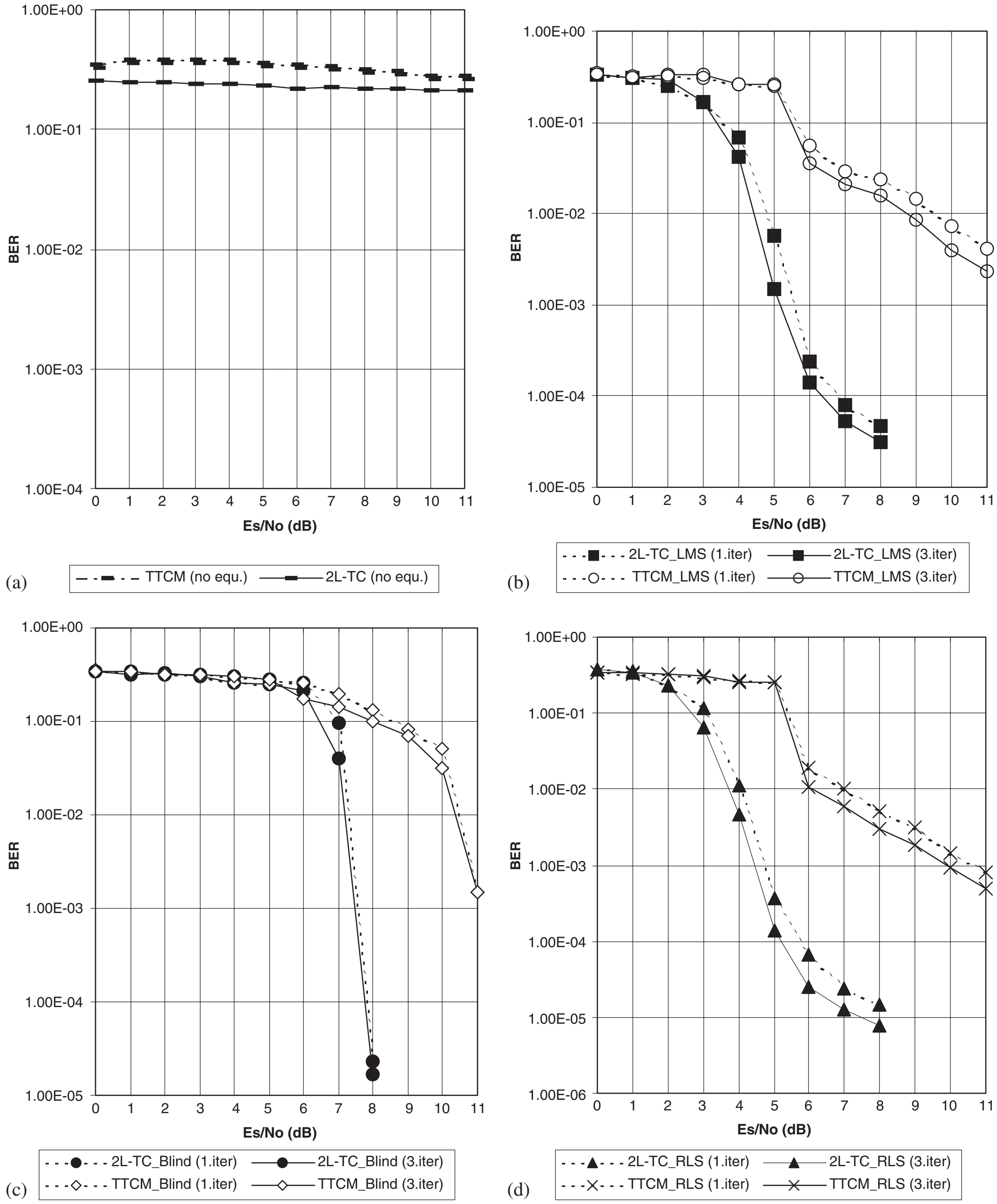

Figure 5. Performance of 2L-TC and TTCM for TU type of COST207: (a) no equalization; (b) with LMS; (c) with Blind; and (d) with RLS.

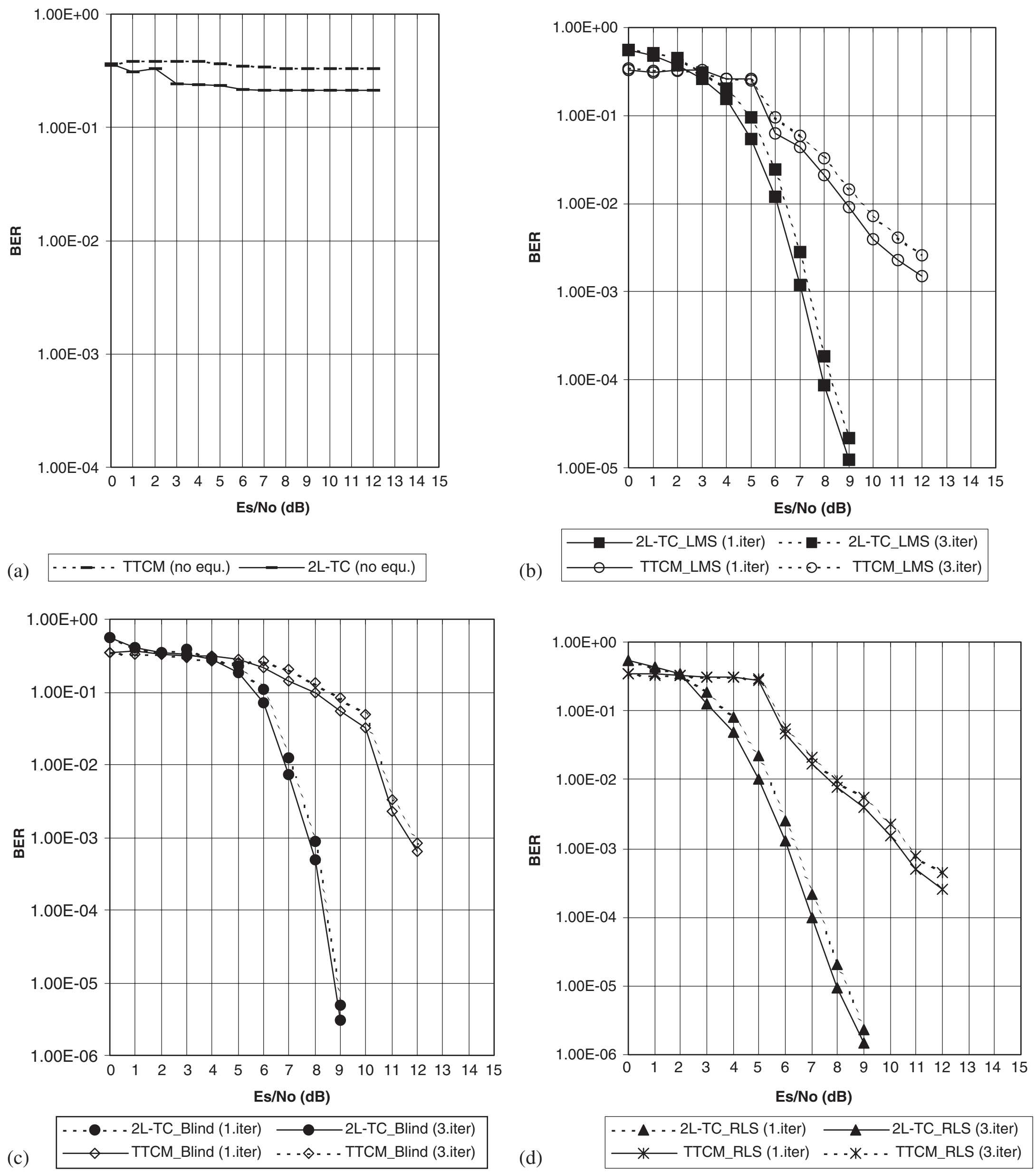

Figure 6. Performance of 2L-TC and TTCM for BU type of COST207: (a) no equalization; (b) with LMS; (c) with Blind; and (d) with RLS.

prewhitening filter requires for convergence and the suboptimum solution in presence of additive Gaussian noise). It has been proven that, RLS will always outperform LMS [19]. As a result, the performance of EVA is close to RLS but better than LMS as obtained in our study.

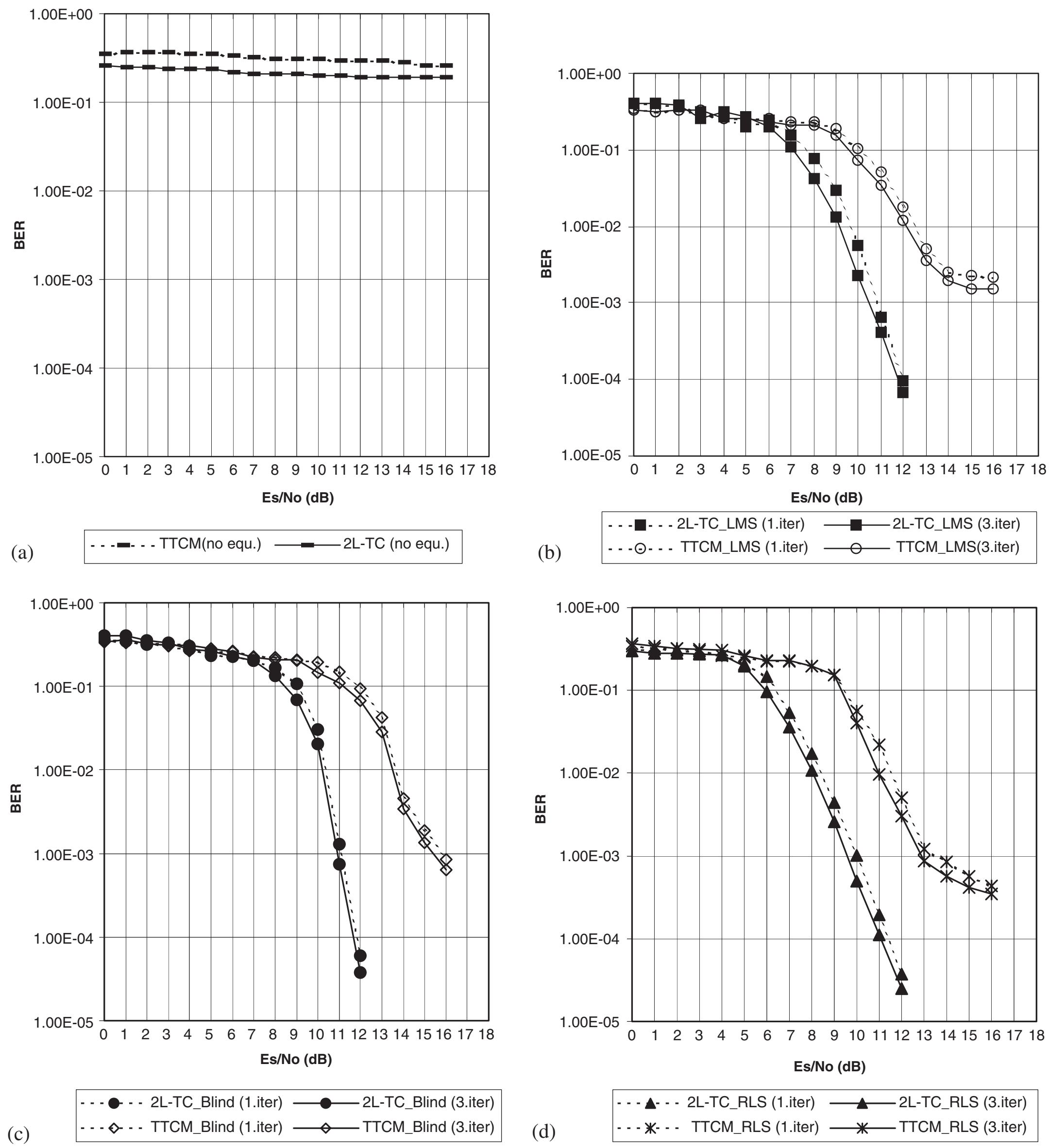

Figure 7. Performance of 2L-TC and TTCM for HT type of COST207: (a) no equalization; (b) with LMS; (c) with Blind; and (d) with RLS.

6. CONCLUSION

In this paper, multilevel-turbo codes (ML-TC) are introduced and the performance of 2 level-turbo codes (2L-TC) is studied over WSSUS channels using blind or non-blind equalization algorithms. They employ more than one turbo encoder/decoder blocks. Here, input bit streams are encoded and can be mapped to any modulation type. After 2L-TC modulated signals are passed through COST207 TU, BU and HT based on WSSUS channels. Before decoding process, 2L-TC modulated signals are equalized blind (EVA) and non-blind algorithms (RLS and LMS). For comparison, 2L-TC signals are also sent to the receiver without equalization. At the receiver side, soft decoding is achieved in each level by using the estimated input bit streams of previous levels. Simulation results are drawn for 2 level-turbo codes (2L-TC) and 8PSK-TTCM schemes with frame size 256. Bit error rate (BER) of non-equalized signals (8PSK-TTCM and 2L-TC) are fairly bad in comparison to equalized ones. For all three channels, the performance of 2L-TC_RLS is significantly better than 2L-TC_LMS and 2L-TC_EVA. The performance gain of 2L-TC_LMS over 2L-TC_EVA is better for lower values of SNR, but after certain values of SNR (8 dB for TU, 9 dB for BU and 12 dB for HT), 2L-TC_EVA outperforms 2L-TC_LMS. It is obvious that application of 2L-TC with equalization over WSSUS multipath exhibits a good performance at lower SNR values. Thus we conclude that ML-TC schemes can be considered as a compromising approach for WSSUS channels if they are accompanied by equalization blocks.

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