Space-Time Turbo Equalization Scheme Employing Trellis Coded Modulation over Frequency Selective Fading Channels

Shinsuke IBI, Seiichi SAMPEI
Graduate School of Engineering, Osaka University
2-1, Yamada-oka, Suita, Osaka 565-0871, JAPAN
e-mail: ibi@wireless.comm.eng.osaka-u.ac.jp

Norihiko MORINAGA
Dept. of Information Technology,
Hiroshima International University
5-1-1, hiro-koshingai, Kure, Hiroshima 737-0112, JAPAN

***Abstract*— To support bandwidth-efficient broadband systems using a single carrier transmission scheme over frequency selective fading channels, this paper proposes a new space-time turbo equalization scheme that employs trellis coded modulation (TCM). This paper aims to further improve in the spectral efficiency of the conventional soft interference canceller followed by MMSE filter (SC/MMSE) which is kind of turbo equalization technique. To this end, we introduce multilayer set-partitioning and multistage SC/MMSE. Computer simulation confirms that the proposed system is effective in improving spectral efficiency without extra bandwidth in the high symbol rate transmission. In addition, this paper also discusses the performance sensitivity of the proposed scheme to channel estimation error.**

***Keywords*— turbo equalization, trellis coded modulation, iterative channel estimation, soft canceller, MMSE filter**

I. Introduction

In recent years, broadbandization for down-link transmission has been a central issue of the development of wireless communication systems. With its progress, orthogonal frequency division multiplexing (OFDM) systems have been under intense development. However, that for up-link transmission will also be necessary with the progress of future multimedia and ubiquitous communication systems. In this regard, power-efficient systems are desirable due to the constraint on terminals. In addition, terminals with low hardware complexity are preferable. Consequently, these aspects bring us to a concept based on single carrier transmission.

To support high bit rate transmission services, the single carrier transmission needs to overcome severe frequency selective fading that produces inter-symbol interference (ISI). As a means of mitigating ISI, turbo equalization techniques with iterative receiver manner are available. Turbo equalization was first proposed in Ref. [1]. In addition, a computationally efficient algorithm for turbo equalization called soft interference cancellation followed by minimum mean square error filtering (SC/MMSE) was proposed in Ref. [2]. However, both systems are based on BPSK transmission.

In general, the systems can not satisfy demands for broadband services without a significant increase in communication spectral efficiency because the available radio spectral is limited. As the key technique, trellis coded modulation (TCM) is well known that set-partitioning applied higher order constellations are capable of increasing in the spectral efficiency without an excessive enlargement of the trellis diagram for channel encoding [3], [4].

To further improve spectral efficiency in the computationally efficient SC/MMSE, we will propose the following two techniques.

- Multilayer set-partitioning for SC/MMSE
- Multistage SC/MMSE

Applying these techniques in the Ref. [2]'s SC/MMSE algorithm allows the data rate to be enhanced because of increasing in the spectral efficiency by TCM. We call the proposed scheme as trellis coded SC/MMSE (TC-SC/MMSE).

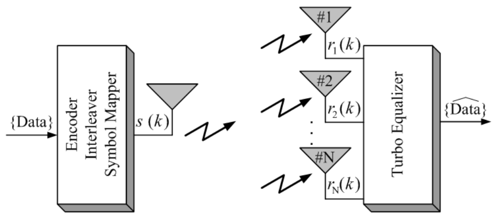

Fig. 1. The channel model

II. The Channel Model

Fig. 1 shows a communication-channel model; the receiver is equipped with N antennas. At the transmitter, the source bit stream is encoded and interleaved with a random interleaver. Then, the coded words are mapped onto symbols. Let us assume that each symbol is transmitted from an antenna where the transmitted baseband signal at discrete k-th timing is $s(k)$ (sampling period is T_s), and a channel impulse response between transmit antenna and n-th receive antennas is $h_n(t = lT_s) = h_n(l)$. In this case, discrete time measurement at the n-th antenna yields the sampled value series $r_n(k)$ of the antenna output as

$$r_n(k) = \sum_{l=0}^{L-1} h_n(l)s(k-l) + \nu_n(k) \qquad (n = 1, 2, \cdots N) \quad (1)$$

where L denotes channel memory length, and $\nu_n(k)$ is the noise component at the n-th receive antenna, which is subject to zero-mean complex white Gaussian noise with its variance of σ^2. In addition, Eq. (1) can be rewritten in spatial sampled vector form as

$$\boldsymbol{r}(k) = \sum_{l=0}^{L-1} \boldsymbol{h}(l)s(k-l) + \boldsymbol{\nu}(k) \quad (2)$$

where

$$\begin{aligned} \boldsymbol{r}(k) &= [r_1(k), r_2(k), \cdots, r_N(k)]^T, && (3)\\ \boldsymbol{h}(l) &= [h_1(k), h_2(k), \cdots, h_N(k)]^T, && (4)\\ \boldsymbol{\nu}(k) &= [\nu_1(k), \nu_2(k), \cdots, \nu_N(k)]^T. && (5) \end{aligned}$$

In general, the energy of the transmitted signal is dispersed not only in space domain, but also in time domain. Therefore, for gathering the dispersed energies as much as possible, Eq. (2) is rewritten in space-time domain as

$$\underline{\boldsymbol{r}}(k) = \underline{\boldsymbol{H}}\,\underline{\boldsymbol{s}}(k) + \underline{\boldsymbol{\nu}}(k) \quad (6)$$

where

$$\underline{\boldsymbol{r}}(k) = [\boldsymbol{r}(k+L-1), \boldsymbol{r}(k+L-2), \boldsymbol{r}(k)]^T, \quad (7)$$

$$\underline{\boldsymbol{H}} = \begin{bmatrix} \boldsymbol{h}(0) & \cdots & \boldsymbol{h}(L-1) & & \boldsymbol{0} \\ & \ddots & & \ddots & \\ \boldsymbol{0} & & \boldsymbol{h}(0) & \cdots & \boldsymbol{h}(L-1) \end{bmatrix}, \quad (8)$$

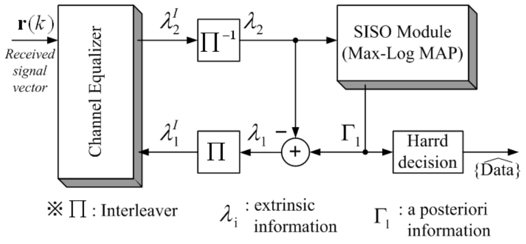

Fig. 2. The configuration of turbo equalizer

$$\begin{aligned}\underline{\boldsymbol{s}}(k) &= [s(k+L-1), \cdots, s(k+1), s(k) \\ &\qquad , s(k-1), \cdots, s(k-L+1)]^T, \quad (9)\\ \underline{\boldsymbol{\nu}}(k) &= [\boldsymbol{\nu}(k+L-1), \boldsymbol{\nu}(k+L-2), \cdots, \boldsymbol{\nu}(k)]^T. \quad (10)\end{aligned}$$

Note that under bar in the matrix or vector means space-time signal, and we will call $\underline{\boldsymbol{H}}$ as the space-time channel matrix.

III. Trellis Coded SC/MMSE (TC-SC/MMSE)

A. Motivation

Turbo equalization was first proposed by Douillard *et al.* for a serially concatenated convolutional-coded BPSK system in 1995 [1]. The equalization, which is based on trellis-based equalizing, is available to mitigate the effect of ISI when it has perfect channel impulse response information. The configuration of receiver is illustrated in Fig. 2. Soft-in/soft-out (SISO) algorithms such as the maximum a posteriori (MAP) algorithm [5], is employed at the channel equalizer block which outputs extrinsic information as shown the figure. The equalization and decoding co-operate via iterative process in order to overcome the channel's frequency selectivity.

However, this method requires extremely huge computational complexity due to the enlargement of trellis diagram for the frequency selectivity when the channel has large channel memory. To solve the issue, Reynolds and Wang have proposed a computationally efficient iterative equalization algorithm in 2001 [2]. Ref. [2]'s turbo equalizer (SC/MMSE) consists of a soft interference canceller and a linear adaptive filter based on the MMSE criterion. SC/MMSE is applied to the channel equalizer block in Fig. 2. Most important to note in the configuration is that the channel equalizer requires no information on trellis diagram.

Now, we briefly review the SC/MMSE algorithm. Note that the available modulation scheme is only BPSK. At first, replicas of interference components $\tilde{s}(j)$ based on log-likelihood ratio (LLR) λ_1^I which is fed back from channel decoder are produced by the soft interference canceller as

$$\tilde{s}(j) = \tanh\left(\frac{\lambda_1^I[s(j)]}{2}\right), \quad (k-L+1 \le j \le k+L-1) \quad (11)$$

where

$$\lambda_1^I[s(k)] = \ln\left(\frac{\Pr[s(k)=+1]}{\Pr[s(k)=-1]}\right). \quad (12)$$

After that, they are subtracted from received signals given in Eq. (6) as

$$\begin{aligned}\underline{\hat{\boldsymbol{r}}}(k) &= \underline{\boldsymbol{r}}(k) - \underline{\boldsymbol{H}}\underline{\tilde{\boldsymbol{s}}}(k) \\ &= \underline{\boldsymbol{H}}[\underline{\boldsymbol{s}}(k) - \underline{\tilde{\boldsymbol{s}}}(k)] + \underline{\boldsymbol{\nu}}(k) \quad (13)\end{aligned}$$

where

$$\begin{aligned}\underline{\tilde{\boldsymbol{s}}}(k) &= [\tilde{s}(k+L-1), \cdots, \tilde{s}(k+1), 0, \\ &\qquad \tilde{s}(k-1), \cdots, \tilde{s}(k-L+1)]^T. \quad (14)\end{aligned}$$

Note that L-th component in vector $\underline{\tilde{\boldsymbol{s}}}(k)$ is zero since it is not an interference component. Consequently, this process is capable of suppressing only interference components included in $\underline{\boldsymbol{r}}(k)$.

Residual components, which are still remain due to canceling imperfections, are further suppressed by an adaptive filter based on MMSE after the soft cancellation process. When the tap weight vector for the MMSE filter at k-th timing is expressed as $\boldsymbol{w}(k)$, and its optimum value is given by Weiner-Hopf's equation as follows:

$$\boldsymbol{w}(k) = \arg\min_{\boldsymbol{w}(k)} E\left\{||s(k) - \boldsymbol{w}^H(k)\underline{\hat{\boldsymbol{r}}}(k)||^2\right\} \quad (15)$$

where $\cdot^H$ and $||\cdot||$ denote complex conjugate and Euclidian norm, respectively. From Eq. (13), we can obtain

$$\begin{aligned}E\left\{\underline{\hat{\boldsymbol{r}}}(k)\underline{\hat{\boldsymbol{r}}}^H(k)\right\} &= \underline{\boldsymbol{H}}\boldsymbol{\Lambda}(k)\underline{\boldsymbol{H}}^H + \sigma^2\boldsymbol{I}, \quad (16)\\ E\left\{\underline{\hat{\boldsymbol{r}}}(k)s^*(k)\right\} &= \underline{\boldsymbol{H}}^H\boldsymbol{e}_L \quad (17)\end{aligned}$$

where

$$\begin{aligned}\underline{\boldsymbol{\Lambda}}(k) &= \mathrm{Cov}\left\{\underline{\boldsymbol{s}}(k) - \underline{\tilde{\boldsymbol{s}}}(k)\right\} \\ &= \mathrm{diag}\left[1-||\tilde{s}(k+L-1)||^2, \cdots, 1-||\tilde{s}(k+1)||^2, 1,\right. \\ &\qquad \left. 1-||\tilde{s}(k-1)||^2, \cdots, 1-||\tilde{s}(k-L+1)||^2\right]. \quad (18)\end{aligned}$$

$E\{\cdot\}$ denotes the emsamble average, $\mathrm{Cov}\{\cdot\}$ is covariance, $\boldsymbol{I}$ is $(2L-1)\times(2L-1)$ identity matrix, and $\boldsymbol{e}_L$ is $(2L-1)\times 1$ vector with its L-th element of one and the others are zeros. It is important to note that the covariance matrix can be calculated only in the case that $||s(k)||^2 = 1$ such as BPSK and QPSK with this manner. As the result, optimum $\boldsymbol{w}(k)$ and MMSE filter output $z(k)$ can be obtained as

$$\begin{aligned}\boldsymbol{w}(k) &= \left[\underline{\boldsymbol{H}}\boldsymbol{\Lambda}(k)\underline{\boldsymbol{H}}^H + \sigma^2\boldsymbol{I}\right]^{-1}\underline{\boldsymbol{H}}^H\boldsymbol{e}_L, \quad (19)\\ z(k) &= \boldsymbol{w}^H(k)\underline{\hat{\boldsymbol{r}}}(k) \quad (20)\end{aligned}$$

Lastly, LLR which is delivered to the channel decoder can be obtained by the calculation of LLR λ_2^I based on $z(k)$ as

$$\lambda_2^I[s(k)] = \frac{4\Re\{z(k)\}}{1-\mu(k)} \quad (21)$$

where

$$\begin{aligned}\mu(k) &= E\left\{z(k)s^*(k)\right\} \\ &= \boldsymbol{e}_L^T\underline{\boldsymbol{H}}^H\left[\underline{\boldsymbol{H}}\boldsymbol{\Lambda}(k)\underline{\boldsymbol{H}}^H + \sigma^2\boldsymbol{I}\right]^{-1}\underline{\boldsymbol{H}}^H\boldsymbol{e}_L \\ &= \boldsymbol{e}_L^T\underline{\boldsymbol{H}}^H\boldsymbol{w}(k) \quad (22)\end{aligned}$$

and $\mu(k)$ denotes amplitude of the MMSE filter outputs.

In the case of BPSK and QPSK, sampled values of the transmitted symbol energies are constant. On the other hand, they are not constant in the case of 16QAM or 64QAM. In other words, they depend on coded words which will be mapped onto symbols. It means that we cannot directly apply the aforementioned algorithm as it is for 16QAM or 64QAM because we cannot accurately calculate the covariance matrix $\underline{\boldsymbol{\Lambda}}(k)$.

Therefore, this paper aims to apply higher order constellation (TCM) into SC/MMSE algorithm by obtaining the covariance matrix with the other manner. It is achieved by multilayer set-partitioning for SC/MMSE and multistage SC/MMSE which are proposed in this paper.

B. Definition of Log-Likelihood Vector

LLR is actually useful to express soft decision value obtained by Turbo principle. However, it can manage only binary sequence. In the higher order constellations, it is obvious that the LLR is unavailable. In order to express the soft decision value even in the higher order constellations, we define a posteriori log-likelihood vector (LLV) given by

$$\begin{aligned}\boldsymbol{\Gamma}_2^I[c(k)] &= [\ln\{\Pr[c(k)=0|\boldsymbol{r}(k)]\}, \ln\{\Pr[c(k)=1|\boldsymbol{r}(k)]\} \\ &\qquad , \cdots, \ln\{\Pr[c(k)=M-1|\boldsymbol{r}(k)]\}] \\ &= \boldsymbol{\lambda}_2^I[c(k)] + \boldsymbol{\lambda}_1^I[c(k)] \quad (23)\end{aligned}$$

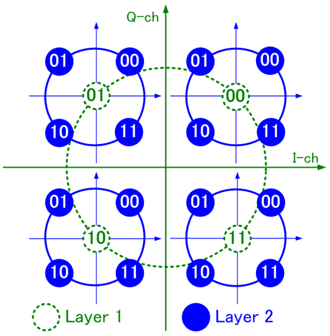

Fig. 3. Multilayerd constellation for SC/MMSE (16QAM)

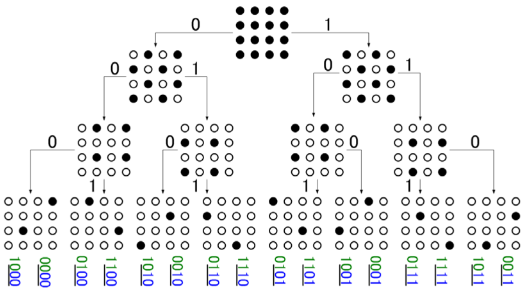

Fig. 4. Multilayerd set-partitioning for SC/MMSE (16QAM)

where $c(k)$ is coded word at the k-th timing and M is number of constellation points. On top of that, the first term in Eq. (23), λ_2^I, represents the extrinsic LLV delivered to the SISO decoder and the second term, λ_1^I, denotes a priori LLV provided by SISO decoder. The superscript $\cdot^I$ indicates interleaved domain. On the other hand, using a priori LLV about all coded word in a burst, $\boldsymbol{\lambda}_1[c(j)]_{j=0}^{B-1}$, SISO decoder provides a posteriori LLV as

$$\begin{aligned} \boldsymbol{\Gamma}_1[c(k)] &= \big[\ln\{\Pr[c(k)=0|\boldsymbol{\lambda}_1[c(j)]_{j=0}^{B-1}]\}, \\ &\quad ,\cdots,\ln\{\Pr[c(k)=M-1|\boldsymbol{\lambda}_1[c(j)]_{j=0}^{B-1}]\}\big] \\ &= \boldsymbol{\lambda}_1[c(k)]+\boldsymbol{\lambda}_2[c(k)] \end{aligned} \quad (24)$$

where $\boldsymbol{\lambda}_1$ represents the extrinsic LLV fed back to the soft canceller, $\boldsymbol{\lambda}_2$ denotes de-interleaved a priori LLV of $\boldsymbol{\lambda}_2^I$, and B is the burst length. These parameters correspond to Fig. 2. Similar definition is used in turbo trellis coded modulation [6]. However, the deference is that LLV is not based on source words but on coded words generated in the channel encoder.

C. Configuration of Transmitter (Multilayer Set-Partitioning)

In our proposed system, the source bit stream is encoded by TCM encoder (Ungerboec's code [3], [4]) at the transmitter, by which coded word stream is produced. The coded words are interleaved with a word-specific random interleaver. Then, they are mapped onto symbols based on constellation points shown in Fig. 3 where it demonstrates the case of 16QAM. In this paper, we will explain a proposed scheme in the case of 16QAM. Observe that the constellation is given by summation of QPSK symbols with natural mapping at layer 1 and 2. It means that the transmitted symbol is given by

$$\begin{aligned} s[c(k)] &= [2,1]\cdot[S^Q[c_1(k)],S^Q[c_2(k)]]^T \\ &= s_1(k)+s_2(k) \end{aligned} \quad (25)$$

where QPSK constellation is given by

$$\begin{aligned} \boldsymbol{S}^Q &= [S^Q[0],S^Q[1],S^Q[2],S^Q[3]] \\ &= [1+j,-1+j,-1-j,1-j]. \end{aligned} \quad (26)$$

On top of that, $c_1(k)$ and $c_2(k)$ denote coded words for layer 1 and 2, respectively. It means, assuming the coded word $c(k)$ in binary is expressed as $[b_3(k),b_2(k),b_1(k),b_0(k)]$, $c_1(k)$ and $c_2(k)$ correspond to the lower two bits $[b_1(k),b_0(k)]$ and upper two bits $[b_3(k),b_2(k)]$ in decimal, respectively. On the other hand, it is important to note that the proposed constellation arrows set-partitioning shown as Fig. 4. It clearly shows that the Euclidian distance among constellation points is increased at every partitioning step. We call the constellation constituting set-partitioning with multilayer structure as multilayer set-partitioning. Consequently, applying SC/MMSE scheme into every layer, the calculation of covariance matrix can be independent from 16QAM constellation points which have deferent energies because the energies at each layer, $||s_1(k)||^2$ and $||s_2(k)||^2$, are constant. As the result, ISI can be effectively suppressed by the multistage process even for higher order constellation cases.

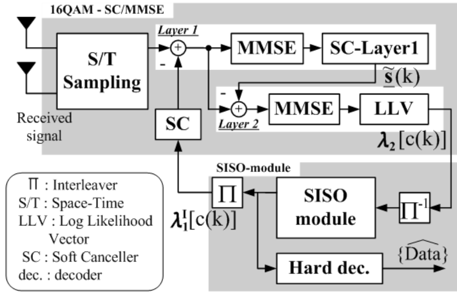

Fig. 5. Configuration of multistage SC/MMSE

D. Configuration of Receiver (Multistage SC/MMSE)

Fig. 3 shows the configuration of the SC/MMSE receiver with multistage soft canceling process. It takes advantage of multistage set-partitioning. We call it as multistage SC/MMSE. In the case that available modulation scheme is 16QAM, the received signal after space-time sampling in Eq. (6) is rewritten as

$$\underline{\boldsymbol{r}}(k)=\underline{\boldsymbol{H}}\,[\underline{\boldsymbol{s}}_1(k)+\underline{\boldsymbol{s}}_2(k)]+\underline{\boldsymbol{\nu}}(k) \quad (27)$$

where

$$\begin{aligned} \underline{\boldsymbol{s}}_1(k) &= [s_1(k+L-1),\cdots,s_1(k-L+1)]^T, \quad (28) \\ \underline{\boldsymbol{s}}_2(k) &= [s_2(k+L-1),\cdots,s_2(k-L+1)]^T. \quad (29) \end{aligned}$$

Since soft cancellation is also applied in each layer, the provided LLV must be translated into LLV for each layer. Converting it into 4×4 matrix as

$$\boldsymbol{\lambda}_1'[c(k)]=\begin{bmatrix} \boldsymbol{\lambda}_1[c(k)=0] & \cdots & \boldsymbol{\lambda}_1[c(k)=12] \\ (00|00) & & (11|00) \\ \vdots & \ddots & \vdots \\ \boldsymbol{\lambda}_1[c(k)=3] & \cdots & \boldsymbol{\lambda}_1[c(k)=15] \\ (00|11) & & (11|11) \end{bmatrix}, \quad (30)$$

the translation can be easily. In this matrix, the first column corresponds to the log-likelihood probability of $c_1(k)=0$. With the same manner, the second, third, and fourth columns correspond to probabilities of $c_1(k)=1$, 2, and 3, respectively. Therefore, each likelihood probability in layer 1 is given by

$$\begin{aligned} \Pr[c_1(k)=m_1] &= \sum_{i=0}^{3}\exp\left(\lambda_1[c(k)=4m_1+i]\right), \\ &\quad (0\le m_1\le 3). \end{aligned} \quad (31)$$

On the other hand, since each likelihood probability in layer 2 corresponds to each row in the matrix, the probability in layer 2 is given by

$$\begin{aligned} \Pr[c_1(k)=m_2] &= \sum_{i=0}^{3}\exp\left(\lambda_1[c(k)=4i+m_2]\right), \\ &\quad (0\le m_2\le 3). \end{aligned} \quad (32)$$

From Eq. (31) and (32), soft decision values of layer 1 and 2 can be obtained as

$$\tilde{s}_1(j) = \sum_{m_1=0}^{3} 2S^Q[m_1]\Pr[c_1(k) = m_1], \quad (33)$$

$$\tilde{s}_2(j) = \sum_{m_2=0}^{3} S^Q[m_2]\Pr[c_2(k) = m_2]. \quad (34)$$

At the stage of layer 1 process, soft replicas of ISI components are produced and it is subtracted from Eq. (27) in order to mitigate ISI components included in layer 1.

$$\begin{aligned} \underline{\hat{\boldsymbol{r}}}(k) &= \underline{\boldsymbol{r}}(k) - \underline{\boldsymbol{H}}\left[\underline{\boldsymbol{s}}_1(k) + \underline{\boldsymbol{s}}_2(k)\right] \quad (35) \\ &= \underline{\boldsymbol{H}}\left[(\underline{\boldsymbol{s}}_1(k) - \underline{\tilde{\boldsymbol{s}}}_1(k)) + (\underline{\boldsymbol{s}}_2(k) - \underline{\tilde{\boldsymbol{s}}}_2(k))\right] \end{aligned}$$

where

$$\begin{aligned} \underline{\tilde{\boldsymbol{s}}}_1(k) &= [\tilde{s}_1(k+L-1), \cdots, \tilde{s}_1(k+1), 0, \\ &\quad \tilde{s}_1(k-1), \cdots, \tilde{s}_1(k-L+1)]^T, \quad (36) \\ \underline{\tilde{\boldsymbol{s}}}_2(k) &= [\tilde{s}_2(k+L-1), \cdots, \tilde{s}_2(k+1), \tilde{s}_2(k), \\ &\quad \tilde{s}_2(k-1), \cdots, \tilde{s}_2(k-L+1)]^T. \quad (37) \end{aligned}$$

$\underline{\hat{\boldsymbol{r}}}(k)$ is then delivered to the MMSE filter for layer 1. The optimum weight vector for layer 1 is given as

$$\begin{aligned} \boldsymbol{w}_1(k) &= \left[\underline{\boldsymbol{H}}\left(\underline{\boldsymbol{\Lambda}}_1(k) + \underline{\boldsymbol{\Lambda}}_2(k)\right)\underline{\boldsymbol{H}}^H + \sigma^2 \boldsymbol{I}\right]^{-1} \\ &\quad \cdot \underline{\boldsymbol{H}}\boldsymbol{e}_L||s_1||^2 \quad (38) \end{aligned}$$

where

$$\begin{aligned} \underline{\boldsymbol{\Lambda}}_1(k) &= \mathrm{diag}\left[||s_1||^2 - ||\tilde{s}_1(k+L-1)||^2, \cdots, \right. \\ &\quad ||s_1||^2, \cdots, \\ &\quad \left. ||s_1||^2 - ||\tilde{s}_1(k-L+1)||^2\right], \quad (39) \\ \underline{\boldsymbol{\Lambda}}_2(k) &= \mathrm{diag}\left[||s_2||^2 - ||\tilde{s}_2(k+L-1)||^2, \cdots, \right. \\ &\quad ||s_2||^2 - ||\tilde{s}_2(k)||^2, \cdots, \\ &\quad \left. ||s_2||^2 - ||\tilde{s}_2(k-L+1)||^2\right]. \quad (40) \end{aligned}$$

Note that covariance matrices can be calculated since

$$||s_1||^2 = ||s_1(k)||^2 \quad : \quad constant, \quad (41)$$

$$||s_2||^2 = ||s_2(k)||^2 \quad : \quad constant. \quad (42)$$

As the result, the MMSE filter output is given by

$$z_1(k) = \boldsymbol{w}_1^H(k)\underline{\hat{\boldsymbol{r}}}(k). \quad (43)$$

The amplitude of the MMSE filter output $\mu_1(k)$ and its variance $\kappa_1^2(k)$ are given by

$$\begin{aligned} \mu_1(k) &= E\left\{z_1(k)s_1^*(k)\right\}/||s_1||^2 \\ &= \boldsymbol{e}_L^T\underline{\boldsymbol{H}}\boldsymbol{w}_1^H(k)/||s_1||^2, \quad (44) \\ \kappa_1^2(k) &= \mathrm{var}\{z_1(k)\} = ||s_1||^2[\mu(k) - \mu^2(k)]. \quad (45) \end{aligned}$$

From Eq. (44) and (45), extrinsic information to be fed to the soft canceller for layer 2 can be obtained as

$$\Pr[c_1(k) = m_1] = \frac{1}{2\pi\kappa_1^2(k)} \exp\left(-\frac{||z_1(k) - \mu_1(k)\beta(m_1)||^2}{2\kappa_1^2(k)}\right) \quad (46)$$

where $\beta(m_1)$ denotes baseband signal in the case of $c_1(k) = m_1$. Substituting Eq. (46) into Eq. (31), soft decision values of layer 1 are updated before the stage of layer 2 process. In the same manner, SC/MMSE is applied to layer 2. However, calculation of likelihood probabilities for the all coded words is required because following SISO decoder requires soft input for the coded words.

Tab. 1. Computer simulation parameters

Modulation (Coding Rate)	QPSK (1/2), 16QAM (3/4) 64QAM (5/6)
Symbols in a burst	1024 syms
Interleaver	Random
Number of antennas	Tx : 1 , Rx : 2
Symbol rate	10 Msyms/sec
Tx and Rx filters	Root Nyquist filter (roll-off factor = 0.5)
Channel model	5-path Rayleigh fading ($f_D \cdot T_s$ = 1/12000)
TCM	8 states Ungerboeck's code
Decoder	Max-Log-MAP

Consequently, $\boldsymbol{w(k)}$, $\mu(k)$, and $\kappa(k)$ about coded words are given by

$$\begin{aligned} \boldsymbol{w}(k) &= \left[\underline{\boldsymbol{H}}\left(\underline{\boldsymbol{\Lambda}}_1(k) + \underline{\boldsymbol{\Lambda}}_2(k)\right)\underline{\boldsymbol{H}}^H + \sigma^2\boldsymbol{I}\right]^{-1}\underline{\boldsymbol{H}}\boldsymbol{e}_L, \quad (47) \\ \mu(k) &= E\left\{z(k)s^*(k)\right\} = \boldsymbol{e}_L^T\underline{\boldsymbol{H}}\boldsymbol{w}^H(k), \quad (48) \\ \kappa^2(k) &= \mathrm{var}\{z(k)\} = \mu(k) - \mu^2(k). \quad (49) \end{aligned}$$

Note that these parameters are not for layers. Lastly, LLV, $\boldsymbol{\lambda}_2^I$, delivered to SISO decoder via de-interleaver can be obtained by Eq. (46) and Eqs. (47)-(49), where $\mu_1(k)$, $\kappa_1^2(k)$, and m_1 in Eq. (46) are replaced to $\mu(k)$, $\kappa^2(k)$, and m $0 \leq m \leq 15$ respectively.

After an extrinsic LLV is calculated by SISO decoder, it is delivered to the soft canceller via interleaver. This process consisted of the multistage SC/MMSE and SISO decoder is iterated a certain times. As the result, a hard decision of the LLV enables a high reliability detection of the source bit stream with high spectral efficiency. In this section, we have explained the proposed system for 16QAM. For QPSK and 64QAM, the system is available by following in a similar scheme.

E. Channel Estimation

Since SC/MMSE algorithm requires MMSE filtering based on channel impulse response, errors of channel estimation cause degradation of system performance. In other words, increasing the accuracy of channel estimation enhances the performance. In the channel estimation process, least-square (LS) approach is effective in the optimization problem expressed as

$$\hat{\boldsymbol{h}}_n^H = arg \min_{\boldsymbol{h}_n^H} ||\boldsymbol{h}_n\boldsymbol{s}_{ref}(k) - r_n(k)||^2 \quad (50)$$

where

$$\begin{aligned} \boldsymbol{h_n} &= [h_n(0), h_n(1), \cdots, h_n(1)], \quad (51) \\ \boldsymbol{s}_{ref}(k) &= [s_{ref}(k), s_{ref}(k-1), \cdots, s_{ref}(k-L+1)]^T \\ &\quad : \text{reference sequence}. \quad (52) \end{aligned}$$

In Eq. (50), a recursive least-square (RLS) scheme can be available to the optimization problem [7], [8]. The channel estimation is achieved by heading the reference sequence to information sequence. In this case, the system requires enough length of reference sequence since the accuracy of estimation depends on it. However, heading long reference sequence causes a degradation of throughput. It means that short one is preferable. As the key solution, the iterative channel estimation is available [9]. If bursts with high likelihood probability of coded words are included in LLV information after the first iteration process, additional reference signal at next iteration process can be obtained from them. In the proposed system, $L-1$ likelihood probabilities in a row whose absolute value exceeds a predetermined threshold value provide additional reference. As the result, channel estimation with high accuracy is achieved even in the case that short reference sequence is headed.

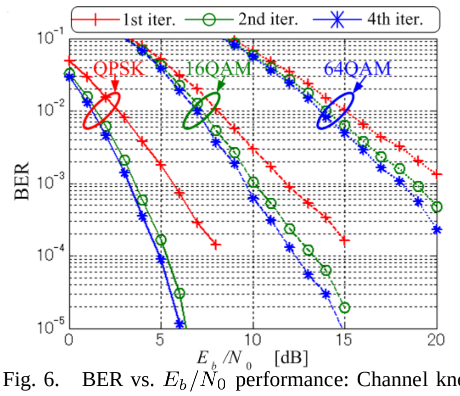

Fig. 6. BER vs. E_b/N_0 performance: Channel known

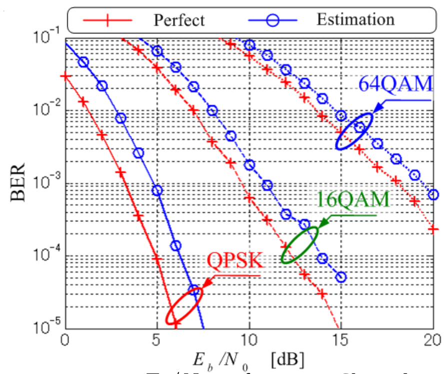

Fig. 7. BER vs. E_b/N_0 performance: Channel estimation

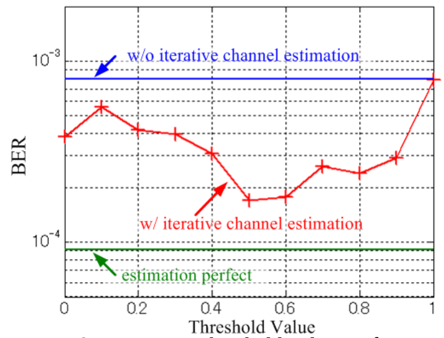

Fig. 8. BER vs. Threshold value performance : iterative channel estimation (QPSK)

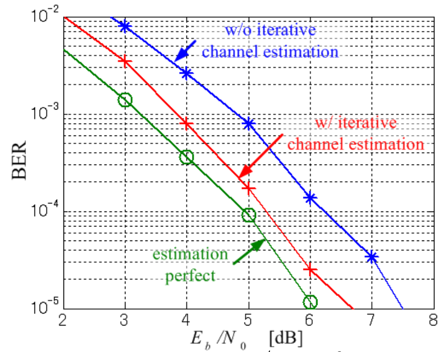

Fig. 9. BER vs. E_b/N_0 performance : iterative channel estimation (QPSK)

IV. Numerical Results

Performances of the proposed system with two received antennas in 5-path Rayleigh with uniform average tap profile are evaluated by computer simulation. We take the viewpoint that the channel time variation is negligible over a burst. It means that the channel taps changes burst-by-burst where the taps is assumed to be uncorrelated. Tab. 1 summarizes specifications of the simulated system.

Fig. 6 shows the BER vs. E_b/N_0 performance for QPSK, 16QAM, and 64QAM where space-time channel matrix $\underline{\boldsymbol{H}}$ is assumed to be known at the reciever. It clearly shows that excessive number of iterations cannot promise better performances. At the BER=10^{-3}, increasing number of iterations from one to four, required E_b/N_0 can be reduced by approximately 2 dB in any modulation schemes. On top of that, significant iterative gain can be obtained at the smaller BER. It means that the soft cancellation is effective. In addition, the figure also shows that the more number of iterations allows slightly better performances for the higher order constellations.

Fig. 7 shows BER vs. E_b/N_0 performances of channel known and channel estimation situations at the fourth iteration. In this case, the channel impulse response is estimated by only reference sequence where the number of unique word in it is 24 symbols and the forgetting factor of RLS is 0.99. The performance with channel estimation is actually worse than channel known case. Especially, the degradations are approximately 1.5 dB E_b/N_0 at the BER=10^{-3}, which is very large. Therefore, we evaluate the performance with iterative channel estimation since the smaller degradation is preferable. Fig. 8 shows the BER vs. threshold value performance for QPSK system. It is obvious that the optimal threshold value of iterative channel estimation is 0.5. Fig 9 shows BER performance of iterative channel estimation for QPSK when the thereshold value is set to 0.5. It obviously show the effectiveity of the itelative channel estimation. At the BER= 10^{-4}, the degradation is only 0.3 dB.

V. Conclusions

This paper has proposed the Trellis Coded SC/MMSE Turbo equalizer, in which multilayer setpartitioning and multistage SC/MMSE are employed. Computer simulation confirms that the proposed system is effective in improving spectral efficiency without extra bandwidth in the high symbol rate transmission. In addition, this paper also discusses the performance sensitivity of the proposed scheme to channel estimation error. As the result, we have confirmed the effectivity of the iteration channel estimation.

References

[1] C. Douillard, C. B. Michel Jezequel, C. Berrou, A. Picart, P. Didier, and A. Glavieux, "Iterative correction of intersymbol interference: turbo-equalization," *European Trans. Telecommun.*, vol. 6, no. 5, pp. 507-511, Sept. 1995.

[2] D. Reynolds and X. Wang, "Low-complexity turbo-equalization for diversity channels," *Signal Processing, Elsevier Science Publishers*, vol. 81, no.5, pp. 989-995, May 2000.

[3] G. Ungerboeck, " Trellis-coded modulation with redundant signal sets part 1: introduction," *IEEE Commun. Mag.*, vol. 25, no. 2, pp.5-11, Feb. 1987.

[4] G. Ungerboeck, " Trellis-coded modulation with redundant signal sets part 2: state of art," *IEEE Commun. Mag.*, vol. 25, no. 2, pp.12-21, Feb. 1987.

[5] G. Bauch, and V. Franz, "A comparison of soft-in/soft-out algorithms for 'turbo-detection'," *Proc. International Conference on Telecommunications (ICT98)*, Porto Carras, Greece, pp.259-263, June 1998.

[6] P. Robertson, and T. Wörz, "Bandwidth-efficient Turbo trellis-coded modulation using punctured component codes," *IEEE J. Select. Areas Commun.*, vol. 16, no. 2, pp.206-218, Feb. 1998.

[7] D. Godard, "Channel equalization using a Kalman filter for fast data transmission," IBM j. Res. Develop., vol. 18, No. 3, pp. 267-273, May 1974.

[8] J. G. Proakis, "*Digital Communications*," McGraw-Hill, 3rd Edition, 1995.

[9] T. Abe, and T. Matsumoto, "Space-time turbo equalization in frequency-selective MIMO channels," *IEEE Trans. Veh. Technol.*, vol. 52, no. 3, pp. 469-475, May 2003.

[10] M. Tüchler, R. Koetter, and A. C. Singer, " Turbo equalization: principles and new results," *IEEE Trans. Commun.*, vol. 50, no. 5, pp. 754-767, May 2002.