Space–Time Turbo Equalization in Frequency-Selective MIMO Channels

Tetsushi Abe and Tadashi Matsumoto, *Senior Member, IEEE*

***Abstract*—A computationally efficient space–time turbo equalization algorithm is derived for frequency-selective multiple-input–multiple-output (MIMO) channels. The algorithm is an extension of the iterative equalization algorithm by Reynolds and Wang for frequency-selective fading channels and of iterative multiuser detection for code-division multiple-access (CDMA) systems by Wang and Poor. The proposed algorithm is implemented as a MIMO detector consisting of a soft-input–soft-output (SISO) linear MMSE detector followed by SISO channel decoders for the multiple users. The detector first forms a soft replica of each composite interfering signal using the log likelihood ratio (LLR), fed back from the SISO channel decoders, of the transmitted coded symbols and subtracts it from the received signal vector. Linear adaptive filtering then takes place to suppress the interference residuals: filter taps are adjusted based on the minimum mean square error (MMSE) criterion. The LLR is then calculated for adaptive filter output. This process is repeated in an iterative fashion to enhance signal-detection performance. This paper also discusses the performance sensitivity of the proposed algorithm to channel-estimation error. A channel-estimation scheme is introduced that works with the iterative MIMO equalization process to reduce estimation errors.**

***Index Terms*—Iterative channel estimation; multiple-input–multiple-output (MIMO) system, turbo equalization.**

I. Introduction

POST-IMT-2000 broad-band mobile-communication systems will be required to transmit signals at 10–100 Mb/s and to accommodate as many users as possible within the available frequency bandwidth. To realize these goals, spectrally efficient broad-band signaling schemes have to be created. Multiple-input–multiple-output (MIMO) communication systems allow multiple users to send information sequences using the same time and frequency slots. References [3] and [4] show that MIMO channels offer enormous channel-capacity gains over independent narrow-band channels having the same total bandwidth. Hence, the MIMO concept is seen as one of the key technologies for future wireless communication systems.

Since intersymbol interference (ISI) and multiple-access interference (MAI) occur in the multiuser broad-band communication scenario, MIMO receivers must suppress the ISI and MAI effects to adequately detect the multiple users' signals. The iterative equalization technique [5] and [6], which is an extension of the turbo-coding concept, is attracting much attention for realizing ISI equalization. However, when it is used in MIMO channels, the computational complexity required to derive the *a posteriori* log likelihood ratio (LLR) is excessive. This is because the number of states in the trellis diagram for the frequency-selective MIMO channels increases exponentially against the product of the number of users and their channel memory lengths.

Reynolds and Wang recently proposed a computationally efficient iterative equalization algorithm for severe ISI channels [1], which was derived from an iterative multiuser detector for code-division multiple-access (CDMA) systems [2]. Its conceptual basis is to replicate the ISI components by using the LLR of the interferers' coded bits fed back from their channel decoder and to subtract the soft replica from the received composite-signal vector. Adaptive linear filtering is used to remove the interference residuals: taps of the linear filter are determined adaptively in order to minimize the mean square error (MSE) between the filter output and the signal point corresponding to the desired user's coded symbol. The LLR of the filter output is then calculated. After deinterleaving, the LLR values of the filter output are brought to a channel decoder as extrinsic information. Soft-input–soft-output (SISO) decoding is performed by the channel decoder. The process discussed above is repeated in an iterative manner. The key point of this scheme is that it offers much lower computational complexity than turbo equalizers using a trellis diagram of the channel.

This paper aims to create a new ISI and MAI cancellation technique for frequency-selective MIMO channels based on Reynolds and Wang's method. Since the received composite signal suffers both ISI and MAI in frequency-selective MIMO channels, the detector first forms the soft replica of the ISI and MAI components and subtracts it from the received composite signal vector. Residual ISI and MAI components are then removed by a linear adaptive filter. SISO channel decoding takes place for each user independently. It is shown that the proposed detector can substantially increase BER performance.

Despite the computational simplicity of Reynolds and Wang's algorithm, it assumes that the detector knows all of the parameters related to the channels. In practice, however, the parameters have to be estimated and the estimation error can degrade the detector's performance. This paper assumes non-blind parameter estimation: each user's information sequence is headed by a unique word sequence whose waveform and timing are known to the detector. This paper also proposes a new iterative channel-estimation technique that enhances the performance of the MIMO detector by reducing the severity of channel-estimation errors.

Manuscript received February 25, 2001; revised July 25, 2001 and April 22, 2002.

T. Abe is with Wireless Laboratories, NTT DoCoMo, Kanagawa 239-8536, Japan.

T. Matsumoto is with the Center for Wireless Communications, University of Oulu, FIN-90014, Finland.

Digital Object Identifier 10.1109/TVT.2003.813227

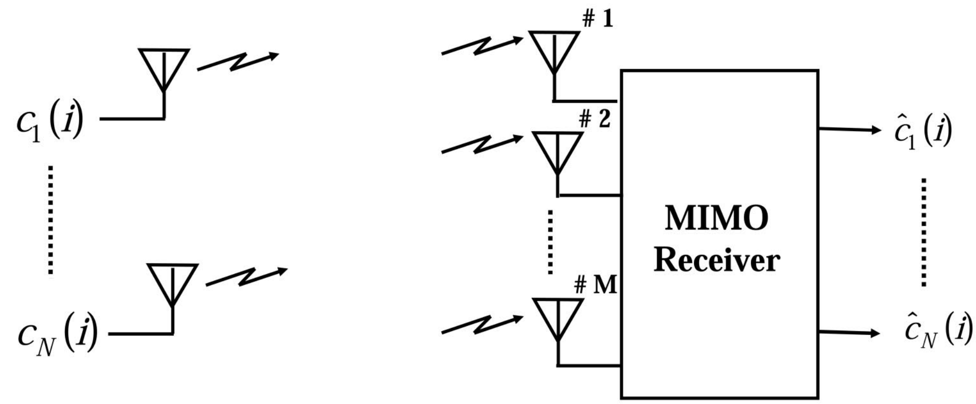


Fig. 1. MIMO channel model.

This paper is organized as follows: Section II describes the MIMO communication-channel model used in this paper. A mathematical space–time representation of the channel is given. Section III proposes a new MIMO turbo-equalization algorithm based on [1] and [2] and Section IV proposes a new iterative channel-estimation technique. Section V shows the results of the computer simulations conducted to verify the effectiveness of the proposed algorithm. Finally, in Section VI we make a conclusion.

II. MIMO Channel Model

Fig. 1 shows a MIMO communication-channel model; there are N users and the receiver is equipped with M antennas. All N users transmit information symbols at the same time and frequency slots without spreading their signals in the frequency domain.

This paper assumes a coded MIMO system as shown in Fig. 2. The information symbols $c_n(i)$s are first encoded by each user's channel encoder where i and n denotes the symbol and user indices, respectively. The coded symbols are then interleaved and modulated according to the modulation format used. The modulated symbols $b_n(k)s$ are then transmitted over frequency-selective channels. k denotes the symbol index of modulated symbols. At the receiver, discrete time measurement at the mth antenna yields the sampled value series $r_m(k)$ of the antenna output as

$$r_m(k) = \sum_{l=0}^{L-1} \sum_{n=1}^{N} h_{mn}(l) b_n(k-l) + v_m(k) \tag{1}$$

where L is the channel-memory length.[1] Without loss of generality, the channel-memory length is assumed to be identical for all N users. $h_{mn}(l)$ is a discrete time representation of the channel between the nth user and the mth receiver antenna, $b_n(k-l)$ is the nth user's transmitted symbol, and $v_m(k)$ is additive white Gaussian noise (AWGN).

Stacking those measurements into a vector form, which is equivalent to sampling in the space domain, results in

$$\boldsymbol{r}(k) \equiv [r_1(k), r_2(k) \dots r_M(k)]^T \tag{2}$$

$$= \sum_{l=0}^{L-1} \boldsymbol{H}(l) \boldsymbol{b}(k-l) + \boldsymbol{v}(k) \tag{3}$$

where

$$\boldsymbol{H}(l) = \begin{bmatrix} h_{11}(l) & \dots & h_{1N}(l) \\ \vdots & \ddots & \vdots \\ h_{M1}(l) & \dots & h_{MN}(l) \end{bmatrix} \tag{4}$$

$$\boldsymbol{b}(k-l) = [b_1(k-l), b_2(k-l) \dots b_N(k-l)]^T \tag{5}$$

and

$$\boldsymbol{v}(k) = [v_1(k), v_2(k) \dots v_M(k)]^T. \tag{6}$$

Finally, temporal sampling to capture the multipath signals for diversity combining yields the following space–time representation of the received signal $\boldsymbol{y}(k)$

$$\boldsymbol{y}(k) \equiv \left[\boldsymbol{r}^T(k+L-1), \boldsymbol{r}^T(k+L-2) \dots \boldsymbol{r}^T(k)\right]^T \tag{7}$$

$$= \boldsymbol{H} \cdot \boldsymbol{u}(k) + \boldsymbol{n}(k) \tag{8}$$

where

$$\boldsymbol{H} = \begin{bmatrix} \boldsymbol{H}(0) & \dots & \boldsymbol{H}(L-1) & & \boldsymbol{O} \\ & \ddots & & \ddots & \\ \boldsymbol{O} & & \boldsymbol{H}(0) & \dots & \boldsymbol{H}(L-1) \end{bmatrix} \tag{9}$$

is the channel matrix with $\boldsymbol{u}(k)$ and $\boldsymbol{n}(k)$ being

$$\boldsymbol{u}(k) = \left[\boldsymbol{b}^T(k+L-1) \dots \boldsymbol{b}^T(k) \dots \boldsymbol{b}^T(k-L+1)\right]^T \tag{10}$$

and

$$\boldsymbol{n}(k) = \left[\boldsymbol{v}^T(k+L-1) \dots \boldsymbol{v}^T(k)\right]^T. \tag{11}$$

III. Space–Time Turbo Equalization

A. System Model

Fig. 2 shows a block diagram of the space–time turbo equalizer for frequency-selective MIMO channels. The equalizer is comprised of a SISO MMSE detector and SISO channel decoders. The basic concept of this configuration follows [1] and [2]; our aim is to extend their concept to cover frequency-selective MIMO channels.

[1]Equation (1) is valid regardless the synchronization in symbol timing among the users. In fact, impulse responses of the channel and filters in the transmitter and the receiver are folded into $h_{mn}(l)$ (see [7]).

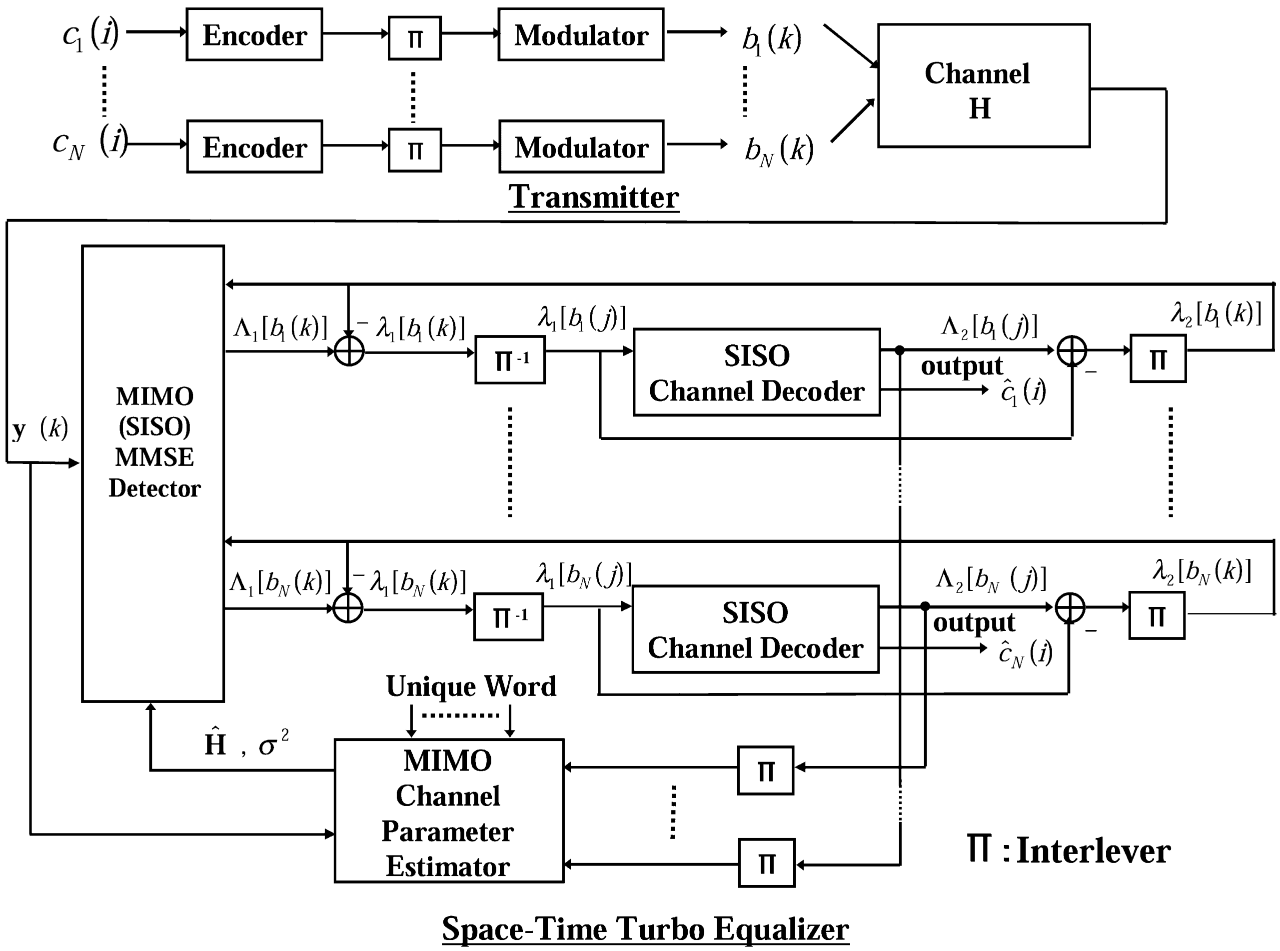


Fig. 2. Block diagram for MIMO equalizer.

Binary phase-shift keying (BPSK) is assumed to be a modulation scheme that was used. The detector produces the LLR for each coded bit as

$$\Lambda_1\left[b_n(k)\right] = \log\frac{Pr\left[b_n(k)=+1|\boldsymbol{y}(k)\right]}{Pr\left[b_n(k)=-1|\boldsymbol{y}(k)\right]} \tag{12}$$
$$\equiv \lambda_1\left[b_n(k)\right] + \lambda_2^p\left[b_n(k)\right] \tag{13}$$

where $\lambda_1[b_n(k)]$ is the extrinsic information fed to the nth user's channel decoder following the MMSE detector and $\lambda_2^p[b_n(k)]$ is the a priori information provided by the nth user's channel decoder.

The channel decoders derive the LLR for each coded bit as

$$\Lambda_2\left[b_n(j)\right] = \log\frac{Pr\left[b_n(j)=+1|\lambda_1\left[b_n(j)\right], j=0,\ldots,B-1\right]}{Pr\left[b_n(j)=-1|\lambda_1\left[b_n(j)\right], j=0,\ldots,B-1\right]} \tag{14}$$
$$\equiv \lambda_2\left[b_n(j)\right] + \lambda_1^p\left[b_n(j)\right] \tag{15}$$

where, with j as the symbol index after de-interleaving, $\lambda_2[b_n(j)]$ is the extrinsic information fed back to the MMSE detector and $\lambda_1^p[b_n(j)]$ is the a priori information provided by the channel detector. B is the burst length.

Estimates of each user's information symbols can be obtained as

$$\hat{c}_n(i) = sign\left(\Lambda_2\left[c_n(i)\right]\right) \quad (n=1,\ldots,N) \tag{16}$$

after sufficient times of iterations, where

$$\Lambda_2\left[c_n(i)\right] = \log\frac{Pr\left[c_n(i)=+1|\lambda_1\left[b_n(j)\right], j=0,\ldots,B-1\right]}{Pr\left[c_n(i)=-1|\lambda_1\left[b_n(j)\right], j=0,\ldots,B-1\right]}. \tag{17}$$

B. SISO MMSE Detector

Utilizing the extrinsic information provided by the nth user's channel decoder, the MMSE detector first forms a soft symbol estimate of the nth user's kth symbol as

$$\tilde{b}_n(k) = tanh\left[\frac{\lambda_2\left[b_n(k)\right]}{2}\right], \quad (n=1,\ldots,N) \tag{18}$$

which is used to form the soft replica $\boldsymbol{H}\cdot\tilde{\boldsymbol{u}}(k)$ of the MAI and ISI components. The soft replica is subtracted from the received signal vector $\boldsymbol{y}(k)$ to produce the nth user's signal estimate vector as

$$\tilde{\boldsymbol{y}}_n(k) = \boldsymbol{y}(k) - \boldsymbol{H}\cdot\tilde{\boldsymbol{u}}_n(k) \tag{19}$$

where

$$\tilde{\boldsymbol{u}}_n(k) = \left[\tilde{\boldsymbol{b}}^T(k+L-1)\ldots\tilde{\boldsymbol{b}}^T(k)\ldots\tilde{\boldsymbol{b}}^T(k-L+1)\right]^T \quad (20)$$

with

$$\tilde{\boldsymbol{b}}(k+l) = \left[\tilde{b}_1(k+l)\ldots\tilde{b}_N(k+l)\right]^T. \quad (21)$$

If $l = 0$

$$\tilde{\boldsymbol{b}}(k) = \left[\tilde{b}_1(k)\ldots 0\ldots\tilde{b}_N(k)\right]^T \quad (22)$$

with the nth element being zero. Equation (19) yields soft interference cancellation. Note that [1]'s equalizer aims to soft-cancel only ISI components, whereas the MIMO detector aims to soft-cancel both ISI and MAI components.

The objective of the rest of the algorithm is to suppress the ISI and MAI residuals left after soft interference cancellation. An adaptive linear filter is used for this purpose: the $M \times L$ vector $\boldsymbol{w}_n(k)$ of the filter taps is determined so that the MSE between the filter output and the signal point corresponding to the detected desired user's symbol is minimized as

$$\boldsymbol{w}_n(k) = \arg\min_{\boldsymbol{w}_n(k)} \left\|\boldsymbol{w}_n^H(k)\tilde{\boldsymbol{y}}_n(k) - b_n(k)\right\|^2. \quad (23)$$

Since the derivation of the optimum vector $\boldsymbol{w}_n(k)$ follows [1] and [2], only the results are shown below as

$$\boldsymbol{w}_n(k) = \left[\boldsymbol{H}\boldsymbol{\Lambda}_n(k)\boldsymbol{H}^H + \sigma^2\boldsymbol{I}\right]^{-1}\boldsymbol{h}_n \quad (24)$$

where

$$\boldsymbol{h}_n \equiv [h_{1n}(L-1)\ldots h_{Mn}(L-1)\ldots h_{1n}(0)\ldots h_{Mn}(0)]^T \quad (25)$$

$$\boldsymbol{\Lambda}_n(k) = diag\,[\boldsymbol{D}(k+L-1)\ldots\boldsymbol{D}(k)\ldots\boldsymbol{D}(k-L+1)], \quad (26)$$

with

$$\boldsymbol{D}(k+l) = diag\left[1-\tilde{b}_1^2(k+l)\ldots 1-\tilde{b}_n^2(k+l)\ldots 1-\tilde{b}_N^2(k+l)\right]. \quad (27)$$

For $l = 0$

$$\boldsymbol{D}(k) = diag\left[1-\tilde{b}_1^2(k)\ldots 1\ldots 1-\tilde{b}_N^2(k)\right] \quad (28)$$

with the (n, n) element being one. By approximating the error at the MMSE filter output by a Gaussian process [1], the extrinsic information to be delivered to the channel decoder can be derived as

$$\lambda_1[b_n(k)] = \log\frac{Pr[\boldsymbol{y}(k)|b_n(k)=+1]}{Pr[\boldsymbol{y}(k)|b_n(k)=-1]} \quad (29)$$

$$= \frac{4Re[z_n(k)]}{1-\mu_n(k)} \quad (30)$$

where $z_n(k)$s are the filter outputs

$$z_n(k) = \boldsymbol{w}_n^H(k)\tilde{\boldsymbol{y}}_n(k) \quad (31)$$

and

$$\mu_n(k) = \boldsymbol{h}_n^H\left[\boldsymbol{H}\boldsymbol{\Lambda}_n(k)\boldsymbol{H}^H + \sigma^2\boldsymbol{I}\right]^{-1}\boldsymbol{h}_n. \quad (32)$$

IV. Iterative Channel Estimation

The sensitivity of turbo-equalization performance to channel-estimation error was reported in [6]. It is obvious that increasing the accuracy of channel estimation enhances the performance. In deriving the algorithm described in Section III, all elements of channel matrix H are assumed to be known. This section proposes a new channel-estimation scheme that effectively utilizes the iterative mechanism of the MIMO turbo equalizer.

The objective of the channel estimation is to find the optimal estimates of the channel-impulse responses of multiple users given (1)'s received signal-sample sequence $r_m(k)$ of the mth $(m = 1, \ldots, M)$ receiver antenna. The estimation optimality in this case is in the least-square (LS) sense expressed as

$$\hat{\boldsymbol{\alpha}}_m = arg\min_{\boldsymbol{\alpha}_m}\left\|\boldsymbol{\alpha}_m^H\boldsymbol{\beta}(k) - r_m(k)\right\|^2 \quad (33)$$

where

$$\boldsymbol{\alpha}_m \equiv [h_{m1}(0),\ldots,h_{m1}(L-1),\ldots,h_{mN}(0),\ldots,h_{mN}(L-1)]^T \quad (34)$$

and

$$\boldsymbol{\beta}(k) \equiv \left[\boldsymbol{\beta}_1^T(k), \boldsymbol{\beta}_2^T(k), \ldots, \boldsymbol{\beta}_N^T(k)\right]^T \quad (35)$$

with

$$\boldsymbol{\beta}_n(k) \equiv [b_n(k), b_n(k-1), \ldots, b_n(k-L+1)]^T \quad (36)$$

being nth $(n = 1, \ldots, N)$ user's reference signal. The recursive least-square (RLS) algorithm can be used to properly solve (33)'s optimization problem. It is assumed that each user's information sequence is headed by a unique word sequence, whose waveforms and timings are known to the receiver. Prior to the first iteration for the MIMO channel equalization, the channel-impulse responses are estimated adaptively by using the known unique word waveforms as the signal references. All the users are assumed to be frame synchronized.[2] Hence, an estimate of H is obtained at the end of the unique word period. The detector then runs the first iteration for the MIMO equalization described in Section III. The first iteration produces initial soft estimates of the N users' transmitted coded bits as

$$\tilde{b}_n(k) = \tanh\left[\frac{\Lambda_2[b_n(k)]}{2}\right]. \quad (37)$$

Obviously, the larger the $|\tilde{b}_n(k)|$ values, the more reliable they are, which suggests that the hard decision results of $\tilde{b}_n(k)$ having relatively large $|\tilde{b}_n(k)|$ values can be used as additional signal references for the channel estimation. Thresholding may properly identify the reliable soft estimates.

Prior to the second iteration, the RLS parameter-estimation algorithm is run again. The unique word waveform as well as the information symbols can be used if, as illustrated in Fig. 3, the period during which all users' information symbols are identified as being reliable is larger than the ISI length. The estimates of the channel-impulse responses are then updated. The detector runs the second iteration for the MIMO equalization using the updated channel estimates. This process is repeated.

[2]This assumption does not impose a requirement on the synchronization of symbol timing. Unique word sequences on the earliest paths may be received at different times, which may spread over the unique word periods. The receiver has to know their unique word-reception timings.

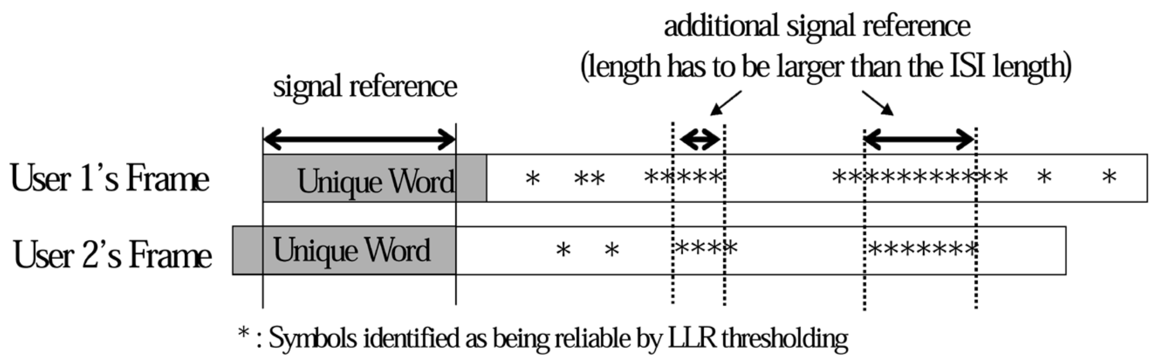


Fig. 3. An example of reliable symbol determination for the proposed iterative channel estimation for $N = 2, L = 3$.

Due to the turbo principle, the $|\tilde{b}_n(k)|$ values increase with the iteration number, thereby yielding more additional reference signals. This results in better estimates of the channel-impulse responses.

V. Simulation Results

This section presents the results of computer simulations conducted to evaluate the performance of the proposed MIMO turbo equalizer. All simulations assumed that channel-frequency selectivity was the result of L-path propagation with each path experiencing frequency-flat Rayleigh fading. A rate 1/2 nonsystematic convolutional code with the constraint length of 3 and generators $[G_1, G_2] = [5, 7]_{oct}$ was used. There are $L \cdot M \cdot N$ path components and all are assumed to have the same average power. One burst has 900 coded symbols headed by a unique word sequence for channel estimation. It is assumed that all N users are symbol and frame synchronized for simplicity. A random interleaver with size 900 was used, which is the same size as the frame length. The forgetting factor of the RLS algorithm was set at 0.99. The max-log maximum *a posteriori* probability (MAP) algorithm was used in the SISO channel decoders.

Figs. 4 and 5 show for $L = 5$ and $M = 2$ bit-error rate (BER) performance curves for $N = 2$ and $N = 3$, respectively. Each curve was obtained by averaging over all users' BERs. Channel matrix H was assumed to be known in this case. Very slow fading was assumed for each user and, hence, channel matrix H was fixed during each burst. E_b is defined as the average per-information bit energy of each user's signal received by one antenna element. The figure also shows, as a performance reference, the performance curve of an order-10 $(L \times M)$ maximum ratio combining (MRC) diversity followed by Viterbi decoding for mono antenna receiver added with 3 dB antenna gain $(= 10log_{10}2)$. This curve corresponds to the case that the derived MIMO equalizer can fully exploit the $L \times M$-order diversity gain due to frequency-selective fading, while completely eliminating the MAI and ISI components from the received signal $\boldsymbol{y}(k)$. Results show that the derived MIMO equalizer offers substantial iteration gains and achieves BER performances almost equivalent to the MRC BER curve.

Fig. 6 shows the BER performances of the MIMO equalizer for $L = 5$, $M = 4$, and $N = 2$. It is found from the figure

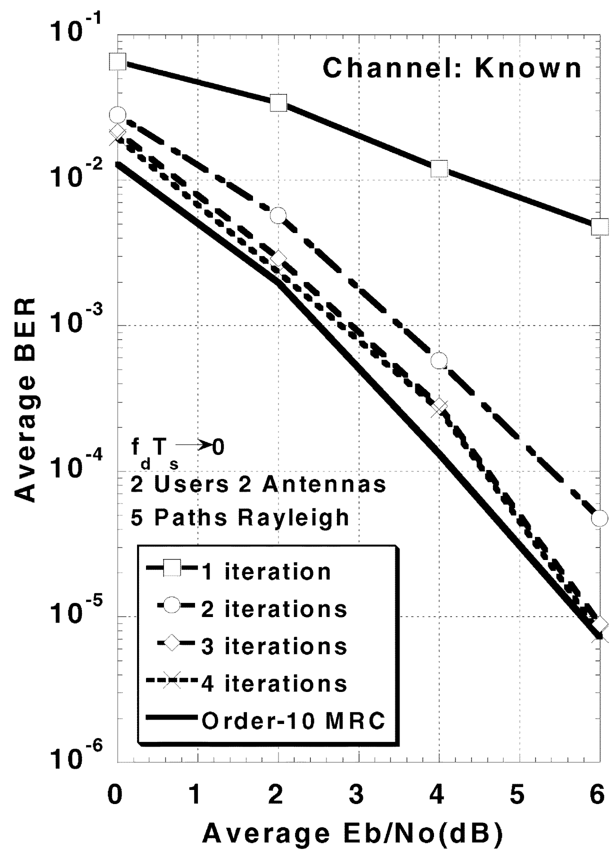


Fig. 4. BER performance: $N = 2, L = 5$, and $M = 2$, channel known.

that increasing the number of antenna elements improves BER performance.

Fig. 7 shows the BER performance of the proposed MIMO equalizer with the proposed iterative channel-estimation scheme versus threshold value $(= Th)$ for $L = 5$, $M = 2$, $N = 2$, and $E_b/N_o = 4$ dB. The normalized maximum Doppler frequency $f_d T_s$, normalized by the symbol duration T_s, is 1/20 000. Each unique word sequence is 25 symbols long. The RLS algorithm was used to estimate the channel matrix H. The loss of signal-to-noise ratio (SNR) due to unique word sequence is not taken into account.[3] Note that with $Th = 1.0$, the channel estimator uses just the unique word sequences and with

[3]The E_b/N_o values in Figs. 7 and 8 do not reflect the bandwidth expansion factor due to the unique word. This is because a major purpose of Figs. 7 and 8 is to compare performances with and without channel estimation.

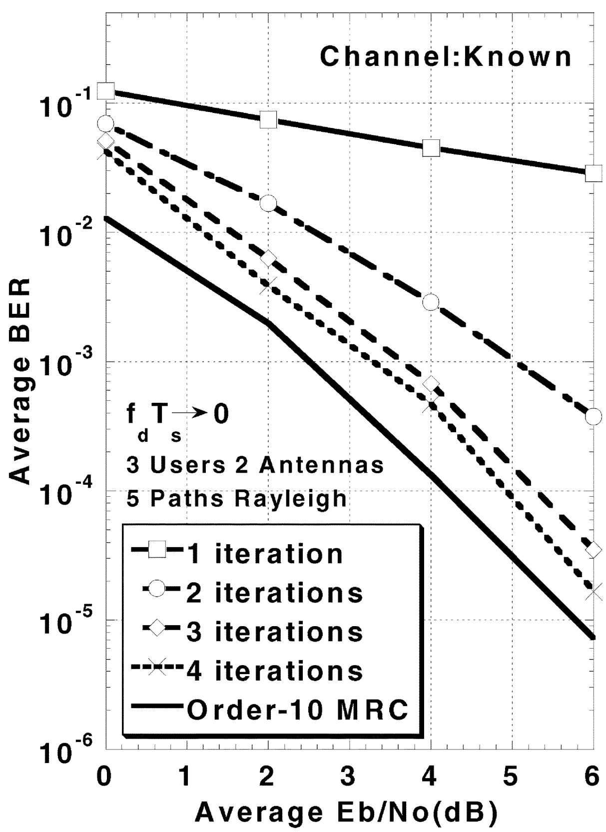


Fig. 5. Performance: $N = 3$, $L = 5$, and $M = 2$, channel known.

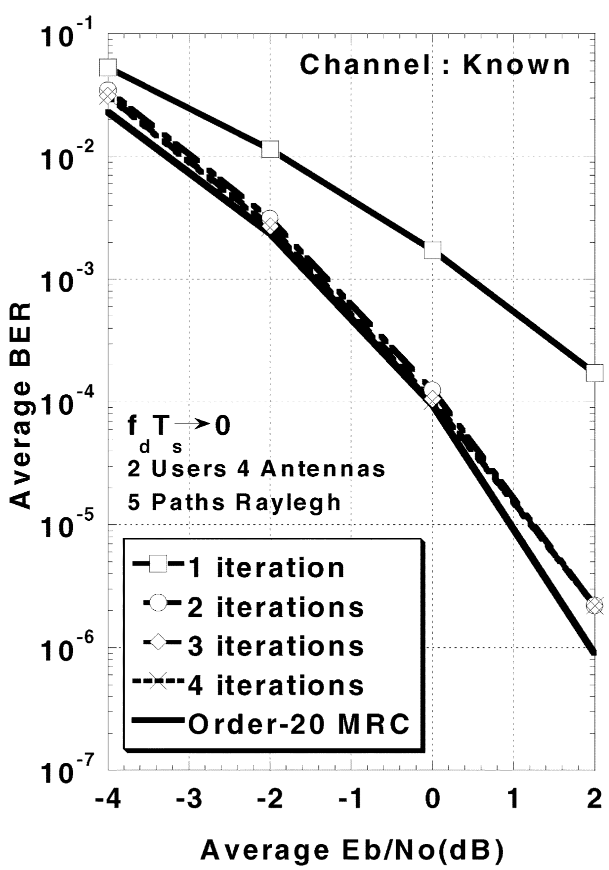


Fig. 6. BER performance: $N = 3$, $L = 5$, and $M = 4$, channel known.

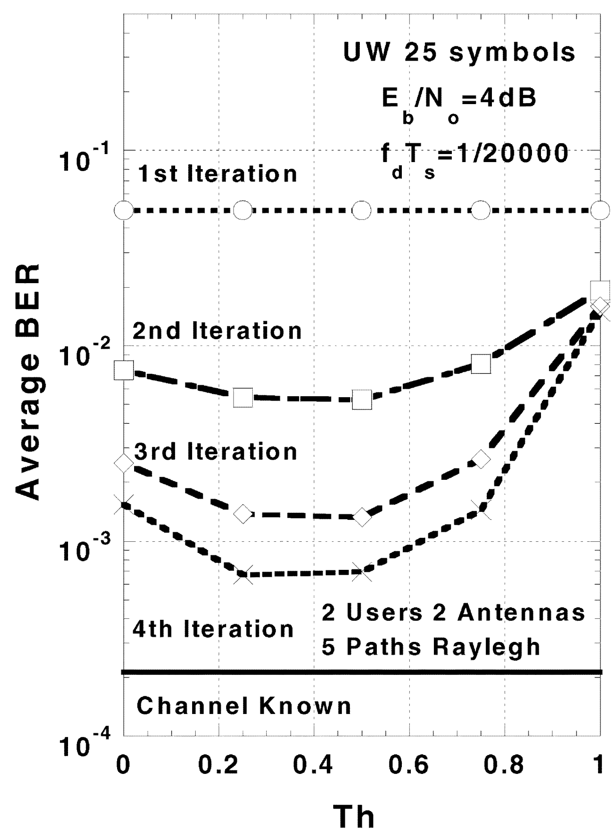


Fig. 7. BER performance versus threshold value of the proposed equalizer with iterative channel estimation:$N = 2$, $L = 5$, and $M = 2$.

$Th = 0.0$, it uses entire portion of the information sequences. With a smaller Th value, more symbols in the information sequences can be used for channel estimation, but they are less reliable. Hence, it is obvious that there is an optimal threshold value that achieves the best performance. Obviously, the BER with iterative estimation is lower-bounded by a certain value, which corresponds to the case in which perfect knowledge about the channel is available. It is found from Fig. 7 that the optimal threshold is approximately $Th = 0.25$.

Fig. 8 shows BER performances with and without iterative channel estimation and with perfect knowledge about the channel. Without iterative channel estimation, channel was estimated only once by using the unique word sequence before the first iteration and the equalizer used the same channel estimates later on (channel estimates were not updated). With iterative channel estimation, channel was re-estimated at the beginning of every iteration using the technique described in Section IV. The optimal threshold value $Th = 0.25$ was used for the iterative estimation. Four iterations were performed. It is found that the iterative channel-estimation scheme can significantly improve system performance.

VI. Conclusion

We extended Reynolds and Wang's iterative ISI equalizer to create a space–time turbo equalizer for frequency-selective MIMO channels. Its performance was evaluated by computer simulations. Results show that the MIMO equalizer offers

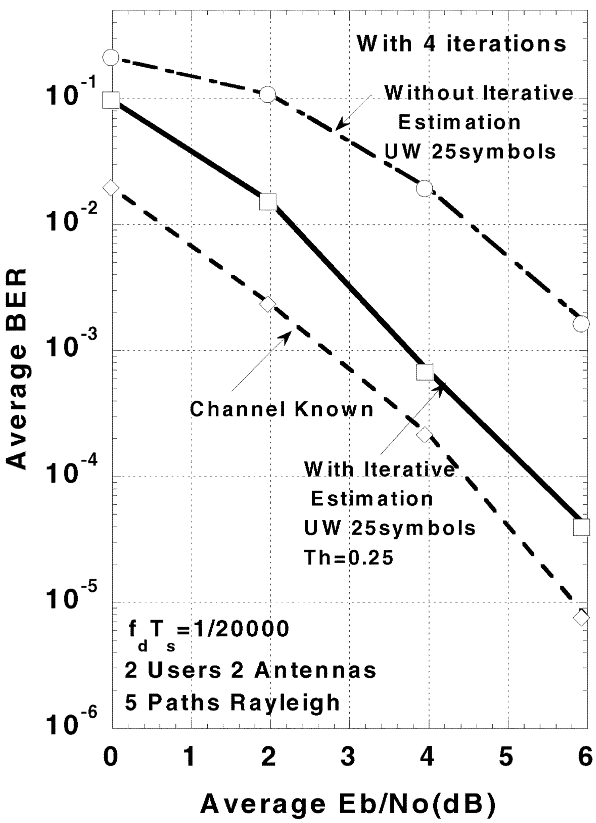
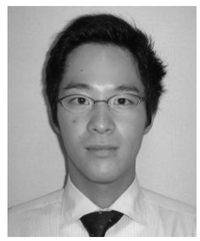


Fig. 8. BER performance with perfect, iterative, and without iterative channel estimation: $N = 2$, $L = 5$, and $M = 2$.

substantially enhanced BER performance even in the presence of both multiple users and propagation paths. Since the complexity of the MMSE detector is not prohibitive and the SISO channel decoders' computational complexity is comparable to that of practical Viterbi channel decoders, the derived MIMO equalizer is, as a whole, computationally efficient. We also proposed a new channel-estimation scheme that effectively works within the iterative-equalization process. Computer simulations showed that the proposed channel-estimation scheme significantly improves MIMO turbo equalizer performance.

REFERENCES

[1] D. Reynolds and X. Wang, "Low complexity turbo-equalization for diversity channels," *Signal Processing*, vol. 81, no. 5, pp. 989–995, 2001.
[2] X. Wang and H. V. Poor, "Iterative (turbo) soft interference cancellation and decoding for coded CDMA," *IEEE Trans. Commun.*, vol. 47, pp. 1046–1061, July 1999.
[3] G. J. Foschini and M. J. Gans, "On the limits of wireless communications in a fading environment," *Wireless Commun.*, no. 6, pp. 315–335, 1998.
[4] G. J. Foschini, "Layered space-time architecture for wireless communication in a fading environment when using multi-element antennas," *Bell Syst. Tech. J.*, vol. 1, no. 2, pp. 41–59, 1996.
[5] C. Douillard *et al.*, "Iterative correction of intersymbol interference: turbo equalization," *European Trans. Telecommun.*, vol. 6, no. 5, pp. 507–511, 1995.
[6] G. Bauch and V. Franz, "A comparison of Soft-In/Soft-Out algorithms for turbo-detection," in *Proc. Int. Conf. Telecommun.*, Porto Carras, Greece, June 1998, pp. 259–263.
[7] V. Poor and G. Wornell, *Wireless Communications*. Englewood Cliffs, NJ: Prentice-Hall, pp. 132–138.



Tetsushi Abe was born in Ehime, Japan, on October 6, 1975. He received the B.S. and M.S. degrees in electrical and electronic engineering from Tokyo Institute of Technology, Tokyo, Japan, in 1998 and 2000, respectively.

From 1998 to 1999, he studied electrical and computer engineering at University of Wisconsin, Madison, WI, under the scholarship exchange-student program offered by the Japanese Ministry of Education. Currently, he is with the Wireless Laboratory at NTT DoCoMo, Inc. His research interests include optimization theory and space–time signal processing for wireless communications.

Tadashi Matsumoto (S'84–SM'95) received the B.S., M.S., and Ph.D. degrees in electrical engineering from Keio University, Yokohama-shi, Japan, in 1978, 1980, and 1991, respectively.

He joined Nippon Telegraph and Telephone Corporation (NTT), Tokyo, Japan, in April 1980. From April 1980 to May 1987, he researched signal transmission technologies, such as modulation/demodulation schemes, as well as radio link design for mobile communications systems. He participated in the R&D project of NTT's high-capacity mobile communications system. From May 1987 to February 1991, he researched error-control strategies for digital mobile radio channels. He was involved in the development of a Japanese TDMA digital cellular mobile communications system (PDC). In July 1992, he transferred to NTT DoCoMo, Inc. From February 1991 to April 1994, he researched code division multiple access (CDMA) mobile communication systems. From 1992 to 1994, he served as a part-time Lecturer at Keio University. In April 1994, he transferred to NTT America, where he served as a Senior Technical Advisor of a joint project between NTT and NEXTEL Communications, Lafayette, CA. In March 1996, he returned to NTT DoCoMo and was a head of the radio-signal processing laboratory at NTT DoCoMo until August of 2001, where he researched adaptive signal processing, MIMO turbo signal detection, interference cancellation, and space–time coding techniques for broadband mobile communications. In May 2002, he moved to the Oulu University, Finland, where he is a Professor at the Center for the Wireless Communications.

Dr. Matsumoto was recognized for his hard work for the organization of the 2000 Spring IEEE Vehicular Technology Conference in Tokyo, where he received an outstanding service award from the IEEE Vehicular Technology Society. Presently, he is serving as a Board-of-Governor of the IEEE Vehicular Technology Society for a term from January 2002 to December 2004.