THROUGHPUT OF DS-CDMA HARQ WITH OVERLAP FREQUENCY-DOMAIN EQUALIZATION

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ABSTRACT

High speed packet access is the core technology of the next generation mobile communication systems. Turbo coded hybrid ARQ (HARQ) is known as one of the promising error control techniques. For direct sequence code division multiple access (DS-CDMA), frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion can replace the well known rake combining with much improved transmission performance. However, the insertion of guard interval (GI), which is required for conventional FDE, reduces the throughput. Recently, overlap FDE that requires no GI insertion was proposed for DS-CDMA. In this paper, we evaluate the throughput of DS-CDMA HARQ with overlap FDE.

1. INTRODUCTION

In today's wireless mobile communication systems, high-speed and high-quality data services are demanded [1]. To realize the high speed packet access, turbo coded hybrid ARQ (HARQ) is known as one of the promising error control techniques [2]. In the 3rd generation mobile communication systems, direct sequence code division multiple access (DS-CDMA) with rake combining is adopted to provide higher data rate services of around a few Mbps [3]. However, since the wireless channel is composed of many propagation paths having different time delays, severe frequency-selective fading channel is produced [4]; this degrades the throughput performance of DS-CDMA HARQ with rake combining. It was shown that frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion can significantly improve the transmission performance of DS-CDMA [5, 6]. To avoid the inter-block interference (IBI), MMSE-FDE needs the insertion of guard interval (GI). However, the GI insertion reduces the throughput. Recently, overlap FDE that needs no GI insertion was proposed [7-9]. Overlap FDE can be applied to the present DS-CDMA mobile radio systems to improve its transmission performance. In this paper, we evaluate the throughput performance of DS-CDMA HARQ with overlap FDE and compare that with rake combining.

2. HARQ WITH CC

In this paper, we use the turbo encoded HARQ with Chase combining (CC) [10, 11]. Turbo coding with original coding rate R=1/3 is considered. A conceptual diagram for HARQ with CC is shown in Fig. 1. At the transmitter, an information sequence is encoded and the systematic bit sequence and two parity bit sequences are generated. After puncturing the parity bit sequences and bit-interleaving, the transmit bit sequence is obtained. At the receiver, after deinteleaving and depuncturing, turbo decoding and error detection is performed on the received packet. If no error is detected in the received packet, the receiver sends the ACK signal to the transmitter, and a new packet is transmitted. If some errors are detected in the received packet, the receiver sends the NACK signal and the same packet is retransmitted. When the same packet is received at the receiver, packet combining is performed. In CC, the time diversity gain can be obtained when the retransmitted packets are combined. In this paper, ideal error detection and an error-free feedback channel are assumed.

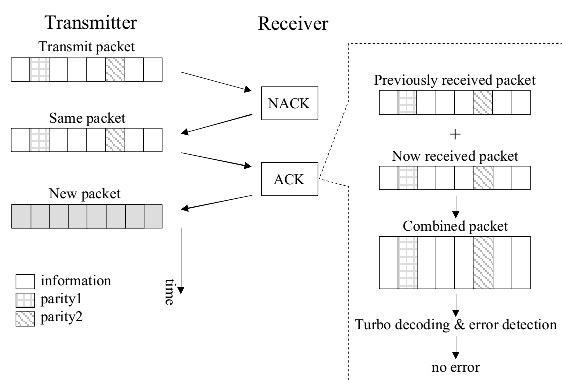

Fig. 1 Conceptual diagram for HARQ with CC.

3. DS-CDMA TRANSMISSION SYSTEM WITH FDE

Figure 2 shows the overall transmission system model of multicode DS-CDMA HARQ with overlap FDE [8]. In this paper, chip-spaced discrete-time representation of signals is used. The bit sequence to be transmitted is transformed into the data-modulated symbol sequence and then serial/parallel (S/P) converted to U parallel symbol sequences, $\{d_u(i);\ i=\ldots,-1,0,1,\ldots\}$, $u=0\sim U-1$.

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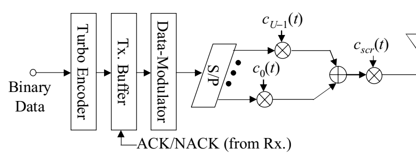
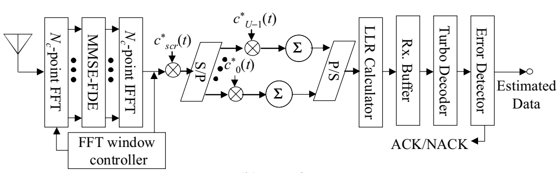
Then, each symbol sequence is spread by using the orthogonal spreading code $\{c_u(t);\ t=0\sim SF-1\}$ with spreading factor SF. After multiplexing U chip sequences, the multicode chip sequence is multiplied by a common scramble code $\{c_{scr}(t);\ t=\ldots,-1,0,1,\ldots\}$ and transmitted.

The transmitted multicode DS-CDMA signal is received at the receiver via a frequency-selective fading channel. At the receiver, the received chip sequence is divided into a sequence of M-chip blocks. For FDE of each M-chip block, the fast Fourier transform (FFT) interval is extended to an N_c-chip block containing the M-chip block of interest in its center as shown in Fig. 3. After decomposing the N_c-chip block into N_c frequency components, MMSE-FDE [5] is performed. After equalization, N_c-point inverse-FFT (IFFT) is applied to obtain a time-domain N_c-chip sequence. Since the impulse response of MMSE-FDE filter is present only at a vicinity of $t=0$ as in Fig. 4, the IBI after FDE exists only near the both ends of N_c-chip block. Therefore, the IBI can be reduced by picking up the M-chip sequence. However, overlap FDE increases the number of FFT/IFFT operation by N_c/M times. Therefore, M should be maximized to avoid an excessive increase in the computational complexity while sufficiently suppressing the IBI.



(a)Transmitter



(b)Receiver

Fig. 2 DS-CDMA HARQ with overlap FDE.

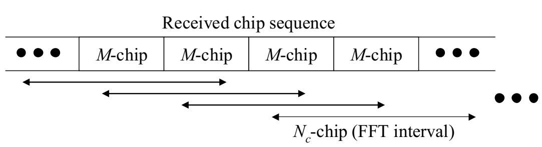


Fig. 3 Overlap FDE.

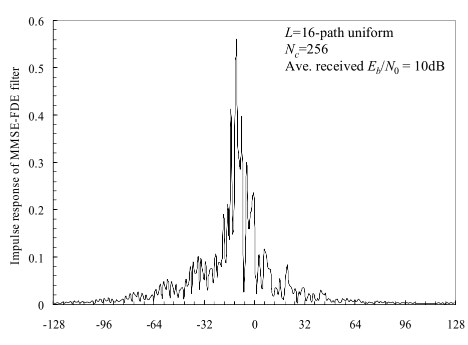


Fig. 4 Impulse response of MMSE-FDE filter.

1.1. Received Signal

We assume a chip-spaced L-path frequency-selective block fading channel having the impulse response as

$$h(\tau)=\sum_{l=0}^{L-1} h_l\delta(\tau-\tau_l)\ , \qquad (1)$$

where h_l and τ_l denote the complex valued path gain and delay time of the lth path, respectively, and $\sum_{l=0}^{L-1} E[|h_l|^2]=1$. The trth received chip sequence is given by

$$r^{(tr)}(t)=\sum_{l=0}^{L-1} h_l^{(tr)} s((t-\tau_l) \bmod N_c)+\nu^{(tr)}(t)+\eta^{(tr)}(t) \qquad ,(2)$$

where $s(t)$ is the transmitted multicode DS-CDMA chip sequence, $h_l^{(tr)}$ and τ_l are respectively the path gain and the delay time of the lth path, $\nu^{(tr)}(t)$ is the IBI and $\eta^{(tr)}(t)$ is the noise due to the additive white Gaussian noise (AWGN) having the one-sided power spectrum density N_0. $s(t)$ can be expressed as

$$s(t)=\sqrt{\frac{2E_c}{T_c}}\sum_{u=0}^{U-1} d_u(\lfloor t/SF\rfloor)c_u(t \bmod SF)c_{scr}(t)\ , \qquad (3)$$

where E_c and T_c respectively denote the energy per chip and the chip duration and $\lfloor x\rfloor$ represents the largest integer smaller than or equal to x. $\nu^{(tr)}(t)$ can be expressed as [8,9]

$$\nu^{(tr)}(t)=\sum_{l=0}^{L-1} h_l^{(tr)}\begin{Bmatrix} s(t-\tau_l) \\ -s((t-\tau_l) \bmod N_c)\end{Bmatrix}\{u(t)-u(t-\tau_l)\} \qquad ,\ (4)$$

where $u(t)$ is the step function given by

$$u(t)=\begin{cases} 0, & t<0 \\ 1, & t\geq 0 \end{cases}. \qquad (5)$$

At the receiver, N_c-point FFT is applied to the received chip sequence to decompose into N_c frequency components $\{R^{(tr)}(k);\ k=0\sim N_c-1\}$. $R^{(tr)}(k)$ is given by

$$\begin{aligned} R^{(tr)}(k)&=\sum_{t=0}^{N_c-1} r^{(tr)}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ &=H^{(tr)}(k)S(k)+N^{(tr)}(k)+\Pi^{(tr)}(k) \end{aligned} \qquad , \qquad (6)$$

where $S(k)$ is the kth frequency-component of $s(t)$, and $H^{(tr)}(k)$, $N^{(tr)}(k)$, and $\Pi^{(tr)}(k)$ are respectively the channel gain, the IBI component, and the noise component at the kth frequency and are given as

$$\begin{cases} S(k)=\sum_{t=0}^{N_c-1} s(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ H^{(tr)}(k)=\sum_{l=0}^{L-1} h_l^{(tr)}\exp\left(-j2\pi k\frac{\tau_l}{N_c}\right) \\ N^{(tr)}(k)=\sum_{t=0}^{N_c-1} \nu^{(tr)}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ \Pi^{(tr)}(k)=\sum_{t=0}^{N_c-1} \eta^{(tr)}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \end{cases}. \tag{7}$$

1.2. Packet combining using overlap FDE

In CC, the time diversity gain can be obtained. Assuming that the same packet has been received Q times (the number of retransmissions are Q–1), the packet combining based on CC is carried out on each frequency as

$$\begin{aligned}\hat{R}(k)&=\sum_{tr=0}^{Q-1} w^{(tr)}(k)R^{(tr)}(k) \\ &=\hat{H}(k)S(k)+\hat{N}(k)+\hat{\Pi}(k)\end{aligned}, \tag{8}$$

where

$$\begin{cases} \hat{H}(k)=\sum_{tr=0}^{Q-1} w^{(tr)}(k)H^{(tr)}(k) \\ \hat{N}(k)=\sum_{tr=0}^{Q-1} w^{(tr)}(k)N^{(tr)}(k) \\ \hat{\Pi}(k)=\sum_{tr=0}^{Q-1} w^{(tr)}(k)\Pi^{(tr)}(k) \end{cases}. \tag{9}$$

$\hat{H}(k)$ is called the equivalent channel gain after FDE and CC. $w^{(tr)}(k)$ is the MMSE weight, which is given by [Appendix A]

$$w^{(tr)}(k)=\frac{\left\{H^{(tr)}(k)\right\}^*}{\sum_{q=0}^{Q-1}\left|H^{(q)}(k)\right|^2+\frac{2}{N_c}\sum_{l=0}^{L-1}\left|h_l^{(tr)}\right|^2\tau_l+\left(U\frac{E_c}{N_0}\right)^{-1}} ,\tag{10}$$

After equalization and packet combining, N_c-point IFFT is applied to $\{\hat{R}(k); k=0\sim N_c-1\}$ to obtain a time-domain chip sequence $\{\hat{r}(t); t=0\sim N_c-1\}$. $\hat{r}(t)$ is given as

$$\begin{aligned}\hat{r}(t)&=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{R}(k)\exp\left(j2\pi k\frac{t}{N_c}\right) \\ &=\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}(k)\right)s(t)+\mu(t)+\hat{\nu}(t)+\hat{\eta}(t)\end{aligned}, \tag{11}$$

where

$$\begin{cases} \mu(t)=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}(k)\left[\sum_{\substack{\tau=0\\ \neq t}}^{N_c-1} s(\tau)\exp\left(j2\pi k\frac{t-\tau}{N_c}\right)\right] \\ \hat{\nu}(t)=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{N}(k)\exp\left(j2\pi k\frac{t}{N_c}\right) \\ \hat{\eta}(t)=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{\Pi}(k)\exp\left(j2\pi k\frac{t}{N_c}\right) \end{cases}. \tag{12}$$

In Eq. (11), the first, second, third and fourth terms represent the desired signal, the residual inter-chip interference (ICI), the IBI and the noise, respectively. To suppress the IBI, we pick up only the M-chip sequence $\{\tilde{r}(t)\,;\ t=N_c/2-M/2\sim N_c/2+M/2-1\}$ from the equalized N_c-chip sequence.

The above operation is repeated for obtaining a sequence of equalized M-chip blocks. Then, despreading is applied to obtain a soft decision symbol sequence as

$$\hat{d}_u(i)=\frac{1}{SF}\sum_{t=iSF}^{(i+1)SF-1}\tilde{r}(t)c_{scr}^*(t)c_u^*(t \bmod SF). \tag{13}$$

1.3. Log-Likelihood Ratio (LLR)

A sequence of log-likelihood ratio (LLR) is generated using the soft decision symbol sequence for turbo decoding. In this paper, we assume M is set small enough to avoid the IBI, so the IBI is not included in LLR calculation. ICI is assumed to be approximated as a complex Gaussian noise. Then, the sum of ICI and noise is treated as a new complex Gaussian noise process with variance $2\sigma^2$. The LLR is given by

$$L(b_m)=\frac{1}{2\sigma^2}\left(\left|\hat{d}_u(i)-\sqrt{\frac{2E_c}{T_c}}\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}(k)\right)d_{b_m=0}^{\min}\right|^2-\left|\hat{d}_u(i)-\sqrt{\frac{2E_c}{T_c}}\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}(k)\right)d_{b_m=1}^{\min}\right|^2\right) ,\tag{14}$$

where $d_{b_m=0,1}^{\min}$ represents a candidate symbol within a set of $\{d_{b_m}\}$ which maximizes the LLR, and $2\sigma^2$ are given as [12]

$$2\sigma^2=\frac{2}{SF}\frac{N_0}{T_c}\left[\frac{1}{N_c}\sum_{k=0}^{N_c-1}\sum_{tr=0}^{Q-1}\left|w^{(tr)}(k)\right|^2+\left(U\frac{E_c}{N_0}\right) \times\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}\left|\hat{H}(k)\right|^2-\left|\frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{H}(k)\right|^2\right)\right] .\tag{15}$$

2. SIMULATION RESULTS

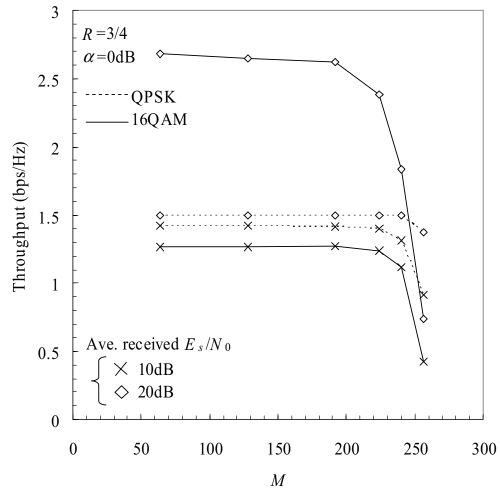
The simulation conditions are summarized in Table 2. We assume QPSK or 16QAM data-modulation. For turbo coding, rate R=3/4 turbo encoder having two (13, 15) recursive systematic component encoders followed by puncturer is used. The puncturing matrix is given by

$$\begin{bmatrix} 1 & 1 & 1 & 1 & 1 & 1 \\ 1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 0 & 0 \end{bmatrix}. \qquad (16)$$

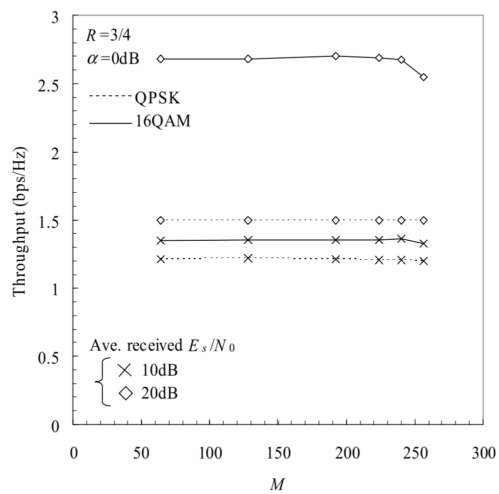
Table 2 Simulation conditions

Turbo coding	R=3/4 (13,15) RSC encoder Log-MAP decoding with 8 iterations	
Channel interleaver	Block interleaver & S-random interleaver	
Data modulation	QPSK, 16QAM	
DS-CDMA	No. of FFT points	N_c=256
	Spreading sequences	Walsh and PN sequence
	Spreading Factor	SF=16
ARQ	Chase combining	
Channel model	Frequency-selective block Rayleigh fading τ_l=l, f_DT=0.001	
	Power delay profile	L=16-path exponential power delay profile
	Decay factor	α=0, 6dB
Channel Estimation	Ideal	

Figure 5 plots the throughput of overlap FDE as a function of M. The average received energy per symbol-to-noise power spectrum density ratio E_s/N_0 is set as 10dB and 20dB. Since the throughput degrades due to IBI, M should be small enough to sufficiently suppress the IBI for the case of the strongest frequency-selectivity (α=0dB). On the other hand, when the channel frequency-selectivity is weak (α=6dB), M can be set to as large as M=200 (or even M=256) and the computational complexity increase can be minimized. It is seen from Fig. 5 that the choice of M=128 gives a good throughput regardless of channel frequency-selectivity and therefore, M=128 is used in the following simulation. Figure 6 shows the throughput comparison of overlap FDE (M=128), conventional FDE using N_g=32-chip GI, and the rake combining. Overlap FDE can always provide better throughput performance than conventional FDE irrespective of the channel selectivity (α=0, 6dB) and data modulation scheme (QPSK, 16QAM). Also seen is that the throughput performance with rake combining is significantly lower than with overlap FDE and conventional FDE due to strong inter-path interference (IPI).

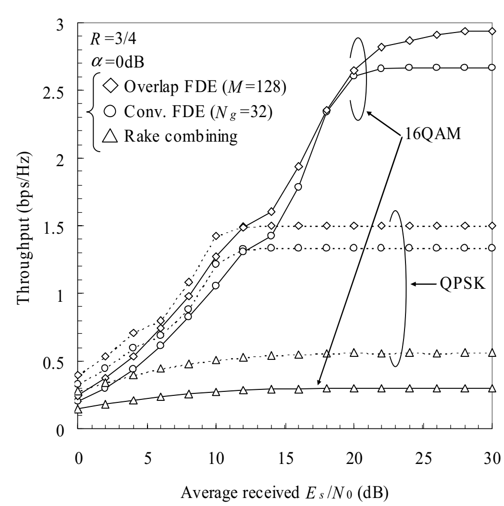


(a)α=0dB

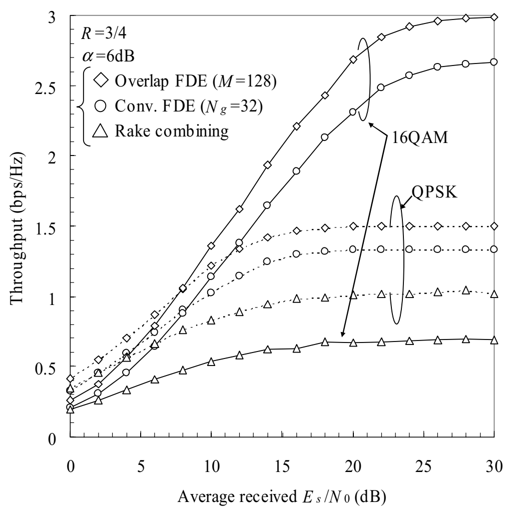


(b)α=6dB

Fig. 5 Impact of M.



(a)α=0dB



(b)α=6dB

Fig. 6 Throughput as a function of average received E_s/N_0.

3. CONCLUSION

In this paper, the throughput performance of DS-CDMA HARQ with overlap FDE was evaluated by computer simulation. It was shown that although the GI insertion is not used, overlap FDE can sufficiently suppress the IBI, and achieve higher throughput than conventional FDE using the GI and rake combining. Since the GI insertion is not required, overlap FDE can be applied to the present wireless DS-CDMA mobile communication systems to significantly improve the throughput performance.

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5. APPENDIX A: MMSE-FDE WEIGHT

The equalization error of the kth frequency component $e^{(Q-1)}(k)$ is defined as the difference between $\hat{R}(k)$ and $S(k)$. From Eq. (8), $e^{(Q-1)}(k)$ is given by

$$e^{(Q-1)}(k)=\sum_{tr=0}^{Q-1} w^{(tr)}(k)R^{(tr)}(k)-S(k)\,. \qquad \text{(A1)}$$

Since $S(k)$, $N^{(tr)}(k)$ and $\Pi^{(tr)}(k)$ in Eq. (6) are independent, MSE can be given as

$$\begin{aligned}E\left[\left|e^{(Q-1)}(k)\right|^2\right]&=\left|\sum_{tr=0}^{Q-1} w^{(tr)}(k)H^{(tr)}(k)-1\right|^2 E\left[\left|S(k)\right|^2\right]\\&+\sum_{tr=0}^{Q-1}\left|w^{(tr)}(k)\right|^2 E\left[\left|N^{(tr)}(k)\right|^2\right]\\&+\sum_{tr=0}^{Q-1}\left|w^{(tr)}(k)\right|^2 E\left[\left|\Pi^{(tr)}(k)\right|^2\right]\end{aligned} \qquad \text{.(A2)}$$

The scramble code is white-noise like. We assume $E[c_{scr}(t)c^{*}_{scr}(t')]=\delta(t-t')$ and hence, the variance $2\sigma^2_{IBI}(i)$ of the IBI at the ith chip position in a N_c-point FFT block of the trth received packet is expressed as

$$2\sigma^2_{IBI}(i)=2U\frac{2E_c}{T_c}\sum_{l=0}^{L-1}\left|h_l^{(tr)}\right|^2\begin{Bmatrix}u(i)\\-u(i-\tau_l)\end{Bmatrix}. \qquad \text{(A3)}$$

To obtain the variance of IBI at the kth frequency, we apply Parseval's equality to Eq. (A3) as

$$\sum_{k=0}^{N_c-1}E[\left|N^{(tr)}(k)\right|^2]=N_c\sum_{i=0}^{N_c-1}2\sigma^2_{IBI}(i)\,. \qquad \text{(A4)}$$

Therefore, $E[|S(k)|^2]$, $E[|N^{(tr)}(k)|^2]$, and $E[|\Pi^{(tr)}(k)|^2]$ in Eq. (A2) are given by

$$\begin{cases}E\left[\left|S(k)\right|^2\right]=N_cU\dfrac{2E_c}{T_c}\\E\left[\left|N^{(tr)}(k)\right|^2\right]=2U\dfrac{2E_c}{T_c}\displaystyle\sum_{l=0}^{L-1}\left|h_l^{(tr)}\right|^2\tau_l\\E\left[\left|\Pi^{(tr)}(k)\right|^2\right]=N_c\dfrac{2N_0}{T_c}\end{cases}. \qquad \text{(A5)}$$

$w^{(tr)}(k)$ is the MMSE weight that satisfies $E[|e^{(Q-1)}(k)|^2]/w^{(tr)}(k)=0$. After some manipulation, $w^{(tr)}(k)$ is obtained as

$$w^{(tr)}(k)=\frac{\left\{H^{(tr)}(k)\right\}^*}{\displaystyle\sum_{q=0}^{Q-1}\left|H^{(q)}(k)\right|^2+\frac{2}{N_c}\sum_{l=0}^{L-1}\left|h_l^{(tr)}\right|^2\tau_l+\left(U\frac{E_c}{N_0}\right)^{-1}} \qquad \text{.(A6)}$$