Residual ISI Cancellation for OFDM with Applications to HDTV Broadcasting

Dukhyun Kim, *Student Member, IEEE,* and Gordon L. Stüber, *Senior Member, IEEE*

***Abstract*— An iterative technique is developed for orthogonal frequency division multiplexing (OFDM) systems to mitigate the residual intersymbol interference (ISI) that exceeds the length of the guard interval. The technique, called residual ISI cancellation (RISIC), uses a combination of tail cancellation and cyclic restoration and is shown to offer large performance improvements. The effects of imperfect channel estimation are also considered. The RISIC algorithm is applied to a typical terrestrial high-definition television (HDTV) broadcasting system that uses a concatenated coding scheme for error control. Results show that the RISIC algorithm can effectively mitigate residual ISI on static or slowly fading ISI channels.**

***Index Terms*— HDTV broadcasting, interchannel interference (ICI), intersymbol interference (ISI), orthogonal frequency division multiplexing (OFDM).**

I. Introduction

ORTHOGONAL frequency division multiplexing (OFDM) is an attractive technique for terrestrial broadcast channels because the intersymbol interference (ISI) introduced by the channel can be mitigated by using a guard interval, rather than the complex equalization schemes that are required for single carrier modulation [1], [2]. High-definition television (HDTV) broadcasting is among the many applications where OFDM is considered, due to the very high data rate (20 Mb/s) of HDTV programs and the limited bandwidth (6 MHz in North America and Japan) [3]. Moreover, successful HDTV broadcasting requires extremely low error rates, since the errors tend to propagate over many packets, due to source coding with a very high compression ratio. Typically, a packet error rate (PER) of less than 10^{-5} is necessary for satisfactory HDTV reception.

Along with ISI, the interchannel interference (ICI) that arises from the loss of subchannel orthogonality in an OFDM block is known to limit the performance of OFDM systems. The error floors created by ISI and ICI are severe challenges for the success of OFDM-based HDTV broadcasting. Trellis-coding and antenna diversity schemes can significantly lower the error floors due to ICI. However, the very long delay spreads or deployment of single-frequency simulcast networks (SFN's) poses the possibility that duration of the ISI exceeds the length of the guard interval [4]. Such ISI is called *residual* ISI and can be devastating, even in small amounts. Increasing the length of the guard interval to reduce the residual ISI has its limitations because it introduces a bandwidth penalty. An attempt to equalize residual ISI has been suggested in [5], but the technique has high complexity and provides only a limited improvement.

This paper introduces a technique to mitigate residual ISI, called residual ISI cancellation (RISIC). The RISIC technique is based on a very efficient method for echo cancellation [6]. The RISIC technique can be thought of as an iterative version of the echo cancellation method in [6]. The RISIC technique will be shown to be highly effective in combating residual ISI with reasonable complexity.

The remainder of this paper is organized as follows. Section II describes the OFDM system and channel models. In Section III, the effect of residual ISI is evaluated for different channels, both analytically and by computer simulation. The RISIC technique is introduced in Section IV, and its performance without coding is evaluated in Section V. In Section VI, the RISIC technique is applied to a terrestrial HDTV broadcasting scenario where a concatenated coding scheme is employed. Finally, Section VII provides some concluding remarks.

II. System Model

We assume a data rate of 20 Mb/s where the data are independent bits. A 16-quadrature amplitude modulation (QAM) constellation is used, which requires about 5 MHz of bandwidth. As in [8], an inverse fast Fourier transform/fast Fourier transform (IFFT/FFT) pair is used as a modulator/demodulator. For the ith block, the N-point IFFT output sequence is

$$x_{i,k} = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} X_{i,n} \exp\left\{ j\frac{2\pi nk}{N} \right\}, \qquad 0 \le k \le N-1 \tag{1}$$

where $\{X_{i,n}\}_{n=0}^{N-1}$ is the transmitted symbol sequence and N is the block size. If the guard interval is a cyclic extension of the IFFT output sequence [2], then the IFFT output sequence with the addition of the guard interval is

$$x_k^g = x_{(k+N-G)_N}, \qquad 0 \le k \le N+G-1 \tag{2}$$

where G is the length of the guard interval, and $(k)_N$ is the residue of k modulo N. The samples $\{x_k^g\}_{k=0}^{N+G-1}$ are passed through a digital-to-analog (D/A) converter with rate $1/T_s$ where T_s is the sample period of the OFDM signal with the guard interval.

We will consider both static and fading ISI channels. For both channels, a maximum delay spread of MT_s is

Manuscript received September 1, 1997; revised February 1, 1998.

The authors are with the Department of Electrical and Computer Engineering, Georgia Institute of Technology, Atlanta, GA 30332 USA.

Publisher Item Identifier S 0733-8716(98)07904-9.

TABLE I
FOUR-TAP STATIC ISI CHANNEL MODELS

tap #	delay (μs)	frac. power (Ch.1)	frac. power (Ch.2)
0	0.0	0.15	0.39
1	0.2	0.65	0.16
2	0.4	0.15	0.26
3	0.6	0.05	0.19

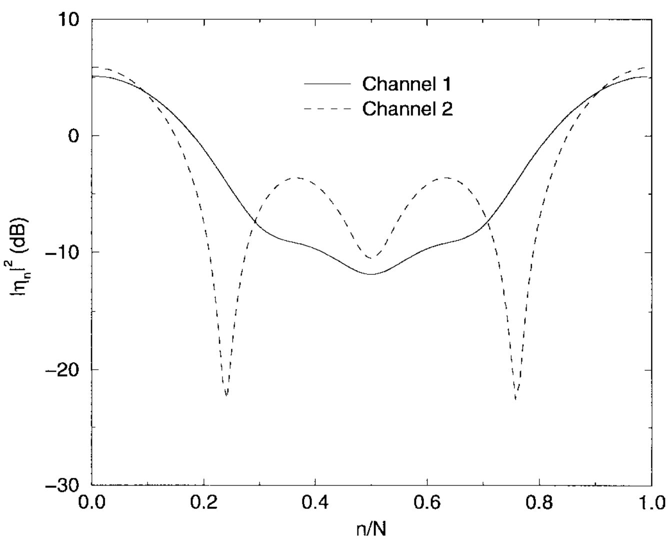


Fig. 1. Amplitude spectrum of static ISI channel with OFDM for $N = 128$.

assumed. Throughout the paper, perfect carrier and timing synchronization are assumed. In the absence of noise, if $G \geq M$, the received samples for the ith block after removal of the guard interval are

$$r_{i,k} = \sum_{m=0}^{M} h_{m,k} x_{i,(k-m)_N}, \qquad 0 \leq k \leq N-1 \tag{3}$$

where $h_{m,k}$ is the channel impulse response at lag m and instant k.

Two static ISI channels are considered, as shown in Table I. Channel 1 has subchannels with a moderate null, whereas Channel 2 has several subchannels with a severe null (Fig. 1). For fading channels, we choose the six-tap typical urban (TU) and hilly terrain (HT) power delay profiles from the COST-207 studies [9]. The tap gains are zero-mean complex Gaussian random processes that are generated using Jakes' method [10] corresponding to a two-dimensional (2-D) isotropic scattering environment.

III. PERFORMANCE OF OFDM ON ISI CHANNELS

A. Static ISI Channel

If $M > G$, then the received samples in (3) can be modified as the sum of two components, *viz.*

$$r_{i,k} = r_{i|i-1,k} + r_{i|i,k} \tag{4}$$

where $r_{i|i-1,k}$ is the received sample component with contributions only from block $i-1$ and $r_{i|i,k}$ is the received sample component with contributions only from block i. Then

$$r_{i|i-1,k} = \sum_{m=G+1}^{M} h_m x_{i-1,(k-m+G)_N} (1 - u(k-m+G)) \tag{5}$$

$$r_{i|i,k} = \sum_{m=0}^{M} h_m x_{i,(k-m)_N} u(k-m+G) \tag{6}$$

where $h_{m,k}$ is replaced by h_m, since the channel is static, and $u(n)$ is the unit step function. The received sample sequence $\{r_{i,k}\}_{k=0}^{N-1}$ is demodulated by taking the N-point FFT

$$\mathrm{FFT}\{\boldsymbol{r}_i\} = \mathrm{FFT}\{\boldsymbol{r}_{i|i-1}\} + \mathrm{FFT}\{\boldsymbol{r}_{i|i}\} \tag{7}$$

where $\boldsymbol{r}_i$ represents a vector with elements from $\{r_{i,k}\}_{k=0}^{N-1}$. We can express the nth sample of $\mathrm{FFT}\{\boldsymbol{r}_{i|i-1}\}$, $\mathrm{FFT}\{\boldsymbol{r}_{i|i-1}\}(n)$ as

$$\begin{aligned}\frac{1}{N} \sum_{m=G+1}^{M} h_m \sum_{l=0}^{N-1} X_{i-1,l} \exp\left\{-j\frac{2\pi lm}{N}\right\} \\ \cdot \sum_{k=0}^{N-1} u(m-k-G-1) \exp\left\{j\frac{2\pi(l-n)k}{N}\right\} \\ \cdot \exp\left\{j\frac{2\pi lG}{N}\right\}\end{aligned} \tag{8}$$

and $\mathrm{FFT}\{\boldsymbol{r}_{i|i}\}(n)$ by

$$\begin{aligned}X_{i,n}\left\{\sum_{m=0}^{G} h_m \exp\left\{-j\frac{2\pi nm}{N}\right\} + \sum_{m=G+1}^{M} h_m \right. \\ \left. \cdot \exp\left\{-j\frac{2\pi nm}{N}\right\}\left(1 + \frac{G}{N} - \frac{m}{N}\right)\right\} \\ - \frac{1}{N} \sum_{m=G+1}^{M} h_m \sum_{\substack{l=0 \\ l\neq n}}^{N-1} X_{i,l} \exp\left\{-j\frac{2\pi lm}{N}\right\} \\ \cdot \sum_{k=0}^{N-1} u(m-k-G-1) \exp\left\{j\frac{2\pi(l-n)k}{N}\right\}.\end{aligned} \tag{9}$$

For symbol n of block i, (8) is the ISI contribution from block $i-1$, the top half of (9) is the useful signal term, and the bottom half of (9) is the ICI term. We can express (7) as

$$Z_n = \mathrm{FFT}\{\boldsymbol{r}_i\}(n) \tag{10}$$

$$= \eta_n X_n + I_n \tag{11}$$

where

$$\begin{aligned}\eta_n = \sum_{m=0}^{G} h_m \exp\left\{-j\frac{2\pi nm}{N}\right\} + \sum_{m=G+1}^{M} h_m \\ \cdot \exp\left\{-j\frac{2\pi nm}{N}\right\}\left(1 + \frac{G}{N} - \frac{m}{N}\right)\end{aligned} \tag{12}$$

$$I_n = I_{i,n} + I_{i-1,n} \tag{13}$$

and where $I_{i,n}$ is the ICI term and $I_{i-1,n}$ is the ISI term.

Next we find the signal-to-interference ratio (SIR) for symbol n defined by

$$\mathrm{SIR}(n) = P_u(n)/P_I(n) \tag{14}$$

where the useful signal power is

$$P_u(n) = \tfrac{1}{2}\, E|\eta_n X_n|^2 \tag{15}$$

where $E(\cdot)$ implies expectation. Since the input symbols $\{X_{i,n}\}_{n=0}^{N-1}$ are assumed to be independent, X_n and I_n are also independent. Furthermore, $I_{i,n}$ and $I_{i-1,n}$ are zero mean and independent, too. Then, the interference power is

$$\begin{aligned} P_I(n) &= \tfrac{1}{2}\, E|I_n|^2 \\ &= \tfrac{1}{2}\, E|I_{i-1,n}|^2 + \tfrac{1}{2}\, E|I_{i,n}|^2 \\ &= P_{\mathrm{ISI}}(n) + P_{\mathrm{ICI}}(n). \end{aligned} \tag{16}$$

The signal power in (15) can be expressed as

$$\begin{aligned} P_u(n) = {} & P_s \left| \sum_{m=0}^{G} h_m \exp\left\{-j\frac{2\pi nm}{N}\right\} + \sum_{m=G+1}^{M} h_m \right. \\ & \left. \cdot \exp\left\{-j\frac{2\pi nm}{N}\right\}\left(1 + \frac{G}{N} - \frac{m}{N}\right)\right|^2 \\ \approx {} & P_s \left| \sum_{m=0}^{M} h_m \exp\left\{-j\frac{2\pi nm}{N}\right\}\right|^2, \\ & \text{if } M - G \ll N \end{aligned} \tag{17}$$

where $P_s = \frac{1}{2}\, E|X_n|^2$. Expressions for (16) can be described as [7]

$$\begin{aligned} P_{\mathrm{ISI}}(n) = {} & \frac{P_s}{N}\left\{ \sum_{m=G+1}^{M} \sum_{m'=m}^{M} 2(m-G) \right. \\ & \cdot \mathrm{Re}\left[h_m h_m^* \exp\left\{-j\frac{2\pi n(m-m')}{N}\right\}\right] \\ & \left. - \sum_{m=G+1}^{M} |h_m|^2 (m-G)\right\} \end{aligned} \tag{18}$$

$$\begin{aligned} P_{\mathrm{ICI}}(n) = {} & P_{\mathrm{ISI}}(n) - \frac{P_s}{N^2} \\ & \cdot \left| \sum_{m=G+1}^{M} h_m \exp\left\{-j\frac{2\pi nm}{N}\right\}(m-G)\right|^2. \end{aligned} \tag{19}$$

Note that the second term in (19) is relatively small when $M - G \ll N$, in which case $P_I(n) \approx 2P_{\mathrm{ISI}}(n)$. The symbol error rate (SER) for 16-QAM is [11]

$$\mathrm{SER} = 3Q\left(\sqrt{\tfrac{1}{5}\gamma_s}\right)\left[1 - \tfrac{3}{4}\, Q\left(\sqrt{\tfrac{1}{5}\,\gamma_s}\right)\right] \tag{20}$$

where γ_s is the received SNR per symbol and $Q(\cdot)$ is the Q function. With OFDM, the SER on static ISI channel is

$$\mathrm{SER} = \frac{1}{N}\sum_{n=0}^{N-1} \mathrm{SER}\,(n) \tag{21}$$

where $\mathrm{SER}\,(n)$ is obtained from (20) with γ_s, replaced by $\mathrm{SIR}\,(n)$ from (14).

Fig. 2 shows the SER floors due to ISI without noise, for different block sizes, when a guard interval is not used $(G = 0)$ and the channel impulse response is assumed to be known. As expected, the error floor due to the ISI decreases with increasing block size. However, when there exist deep null subchannels, as in Channel 2, the improvement from increasing block size is quite small. Therefore, increasing block size is not always a very efficient countermeasure against ISI.

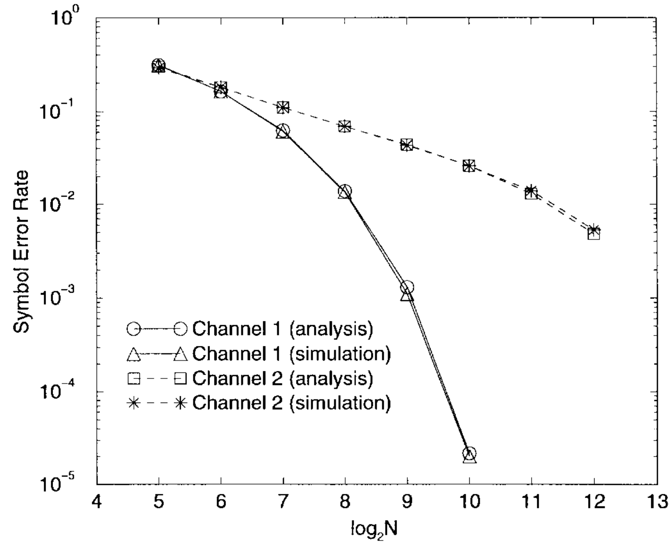


Fig. 2. Performance of OFDM signaling on static ISI channels with different block sizes $G = 0$.

B. Fading ISI Channel

On fading ISI channels, channel time variations during a block cause ICI. For large block sizes, the central limit theorem can be invoked and the ICI can be treated like additive white Gaussian noise (AWGN) [8]. If we assume 2-D isotropic scattering and Rayleigh fading, then for large N the SIR is [7]

$$\mathrm{SIR} = \frac{P_u}{P_s - P_u} \tag{22}$$

where P_u is

$$\begin{aligned} P_u = \frac{P_s}{N^2}\Bigg\{ & \sum_{m=0}^{G} E|h_m|^2 \left(N + 2\sum_{i=1}^{N-1}(N-i)\, \cdot\, J_0(2\pi f_D T_s i)\right) \\ & + \sum_{m=G+1}^{M} E|h_m|^2 \Bigg(N - m + G \\ & + 2\sum_{i=1}^{N-m-1+G}(N-m-i+G) J_0(2\pi f_D T_s i)\Bigg)\Bigg\} \end{aligned} \tag{23}$$

and where f_D is the maximum Doppler frequency and $J_0(\cdot)$ is the zero-order Bessel function of the first kind.

The SER for a Rayleigh fading channel is obtained by averaging (20) over the probability density function

$$p(\gamma_s) = \frac{1}{\overline{\gamma}_s}\exp\left\{-\frac{\gamma_s}{\overline{\gamma}_s}\right\}, \qquad \gamma_s > 0 \tag{24}$$

where $\overline{\gamma}_s$ is replaced by the SIR. The SER obtained by using the SIR in (22) is actually an upper bound when the interference caused by ISI is a dominant factor. This is due to the correlation between the useful signal and interference term. An intuitive explanation is as follows. If the channel varies slowly, then we can assume the channel impulse response is constant over a duration of a block. Hence, from (14), the conditional SIR for the subchannel n, given the channel impulse response $\boldsymbol{h}, \mathrm{SIR}\,(n)_{\boldsymbol{h}}$, is

$$\mathrm{SIR}\,(n)_{\boldsymbol{h}} = \frac{E[|\eta_n X_n|^2|\boldsymbol{h}]}{E[|I_n|^2|\boldsymbol{h}]}. \tag{25}$$

For $M - G \ll N$, $\mathrm{SIR}\,(n)_{\boldsymbol{h}}$ is well approximated by

$$\begin{aligned}\mathrm{SIR}\,(n)_{\boldsymbol{h}} &= \frac{P_s}{2}\left\{\frac{\sum_{m=0}^{G}\sum_{m'=0}^{G} h_m h_{m'}^* \exp\left\{-j\frac{2\pi n}{N}(m-m')\right\}}{P_{\mathrm{ISI}}(n)_{\boldsymbol{h}}}\right. \\ &\quad + \frac{\sum_{m=0}^{G}\sum_{m'=G+1}^{M} h_m h_{m'}^* \exp\left\{-j\frac{2\pi n}{N}(m-m')\right\}}{P_{\mathrm{ISI}}(n)_{\boldsymbol{h}}} \\ &\quad \left. + \frac{\sum_{m=G+1}^{M}\sum_{m'=G+1}^{M} h_m h_{m'}^* \exp\left(-j\frac{2\pi n}{N}(m-m')\right)}{P_{\mathrm{ISI}}(n)_{\boldsymbol{h}}}\right\}\end{aligned} \tag{26}$$

where $P_{\mathrm{ISI}}(n)_{\boldsymbol{h}}$ is from (18) but expressed differently as

$$\begin{aligned}P_{\mathrm{ISI}}(n)_{\boldsymbol{h}} = \frac{P_s}{N}\sum_{m=G+1}^{M}\sum_{m'=G+1}^{M} h_m h_{m'}^* \min\,[m-G, m'-G] \\ \cdot \exp\left\{-j\frac{2\pi n(m-m')}{N}\right\}.\end{aligned} \tag{27}$$

In (26), the first and second fractions represent the portion of the SIR for which the Rayleigh assumption is valid in computing the SER because the useful signal term and the interference term are uncorrelated. The last fraction, however, shows that the useful signal term and the interference term are correlated. Hence, when the useful signal term is faded so is the interference. Consequently, the Rayleigh assumption leads to pessimistic performance estimates, i.e., the use of (22) to compute the SER gives an upper bound. On the other hand, if the portion of the energy contained within the guard interval is relatively large, then the last fraction becomes insignificant relative to the first two fractions in (26) and, hence, the SER found by using the SIR from (22) is accurate.

Fig. 3 shows the performance on a fading ISI channel that is based on the COST-207 six-tap TU channel model [9]. Similar to the case of static ISI channels, the performance improves as the block size is increased if the normalized Doppler frequency, $f_D N T_s$, is small. Unlike the static ISI channels, however, the block size should not be made too large, because of the ICI caused by channel time variations over block (Fig. 3). Therefore, the block size must be small enough to keep the ICI small, while it must be large enough to keep the ISI small for channels with a long impulse response. Also, as explained above, the SER from the analysis when $GT_s = 2\,\mu\mathrm{s}$ agrees very well with simulation result, but gives an upper bound when $G = 0$.

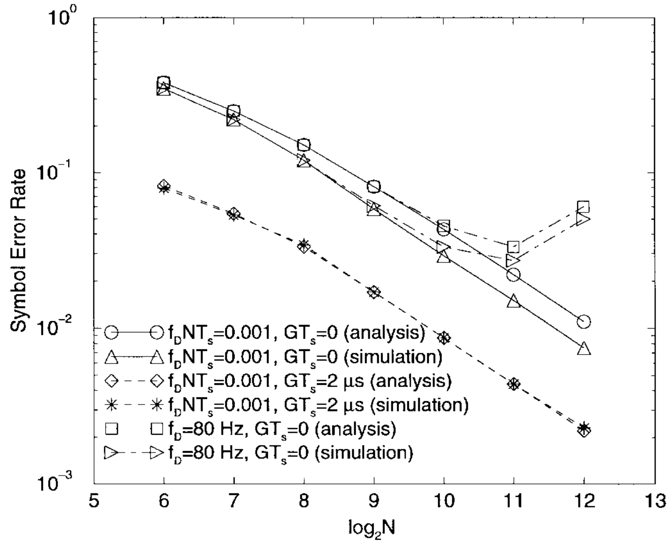


Fig. 3. Performance of OFDM signaling on fading ISI channels with different block sizes.

IV. RISIC

If the channel changes little during the block duration and the guard interval is sufficiently large, i.e., $G \geq M$, then the channel output is

$$\tilde{r}_{i,k} = \sum_{m=0}^{M} h_m x_{i,(k-m)_N}, \qquad 0 \leq k \leq N-1 \tag{28}$$

where $\tilde{r}_{i,k}$ represents the desired channel output that is free of ISI. To achieve the desired channel output $\tilde{r}_{i,k}$ in the presence of residual ISI two steps must be taken. The first is to remove the residual ISI from the received signal, and the second is to use reconstruction to restore cyclicity and avoid ICI. These two procedures are called tail cancellation and cyclic reconstruction, respectively [6]. The procedure can be described by

$$\begin{aligned}\tilde{r}_{i,k} = r_{i,k} - r_{i|i-1,k} + \sum_{m=G+1}^{M} h_m x_{i,(k-m)_N} \\ \cdot (1 - u(k-m+G)).\end{aligned} \tag{29}$$

The residual ISI is removed from the received signal by subtracting the second term in (29). Cyclicity is restored by the last term in (29).

The feasibility of implementing the tail cancellation and cyclic reconstruction procedures depends on the availability of the transmitted signal $\{x_k\}$ at the receiver. Echo cancellers have exact knowledge of the transmitted symbols and, therefore, the above procedures have been successfully implemented [6]. The vast majority of communication applications that require mitigation of ISI, however, do not enjoy this luxury. We now describe the method for reducing the effect of ISI, using the aforementioned procedures, when the transmitted symbols are not available to the receiver *a priori*.

A. RISIC Algorithm

We assume that the channel impulse response is constant over a block period, i.e., $h_{m,k} = h_m, 0 \le k \le N-1$. The RISIC algorithm proceeds as follows.

1) An estimate of the channel impulse response, $\hat{h}_m$, is obtained from a training sequence and updated in a decision-directed mode. Channel estimation will be treated in more detail in Section V-C.
2) Decisions on the transmitted symbols $\{\hat{X}_{i-1,n}\}_{n=0}^{N-1}$ from block $i-1$ are obtained for use in tail cancellation. Since the decisions are affected by residual ISI, some may be erroneous. These symbols are converted back to the time domain using IFFT giving $\{\hat{x}_{i-1,k}\}_{k=0}^{N-1}$.
3) For the block of index i we perform tail cancellation by calculating the residual ISI and subtract it from $r_{i,k}$, i.e.,

$$\tilde{r}_{i,k}^{(0)} = r_{i,k} - \sum_{m=G+1}^{\hat{M}} \hat{h}_m \hat{x}_{i-1,(k-m+G)_N} \cdot (1-u(k-m+G)), \qquad 0 \le k \le N-1 \quad (30)$$

where $\hat{M}$ is the estimate of the maximum channel impulse response length.
4) The $\{\tilde{r}_{i,k}^{(0)}\}_{k=0}^{N-1}$ obtained in Step 3 are converted to the frequency domain using FFT and decisions are made. Afterwards, the decisions are converted back to the time domain to give $\{\hat{x}_{i,k}^{(0)}\}_{k=0}^{N-1}$.
5) Next, we perform cyclic reconstruction by forming

$$\tilde{r}_{i,k}^{(I)} = \tilde{r}_{i,k}^{(0)} + \sum_{m=G+1}^{\hat{M}} \hat{h}_m \hat{x}_{i,(k-m)_N}^{(I-1)} \cdot (1-u(k-m+G)), \qquad 0 \le k \le N-1 \quad (31)$$

where I represents an iteration number with an initial value of $I=1$.
6) The $\{\tilde{r}_{i,k}^{(I)}\}_{k=0}^{N-1}$ are converted to the frequency domain and decisions are made yielding $\{\hat{X}_{i,n}^{(I)}\}_{n=0}^{N-1}$. This completes the Ith iteration in the RISIC algorithm.
7) To continue iterations, convert the $\{\hat{X}_{i,n}^{(I)}\}_{n=0}^{N-1}$ to $\{\hat{x}_{i,k}^{(I)}\}_{k=0}^{N-1}$ and repeat Steps 5–7 with $I \leftarrow I+1$.
8) The end of the RISIC algorithm for block i.

V. Performance of the RISIC Algorithm

We now evaluate the performance of the RISIC algorithm on both static and fading ISI channels. The stability of the RISIC algorithm and the effect of using imperfect estimates of the channel impulse response is also investigated. First, it is assumed that $\hat{M} = M$, but $\hat{M} \ne M$ will be considered later.

A. Static ISI Channel

Suppose that the receiver has perfect channel information. The two static ISI channels in Section II are considered.

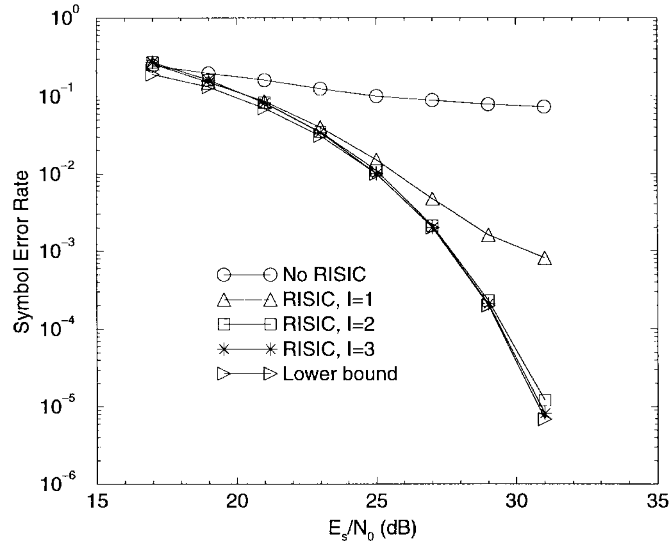


Fig. 4. Performance of the RISIC technique on Channel 1: $G = 0$, $N = 128$.

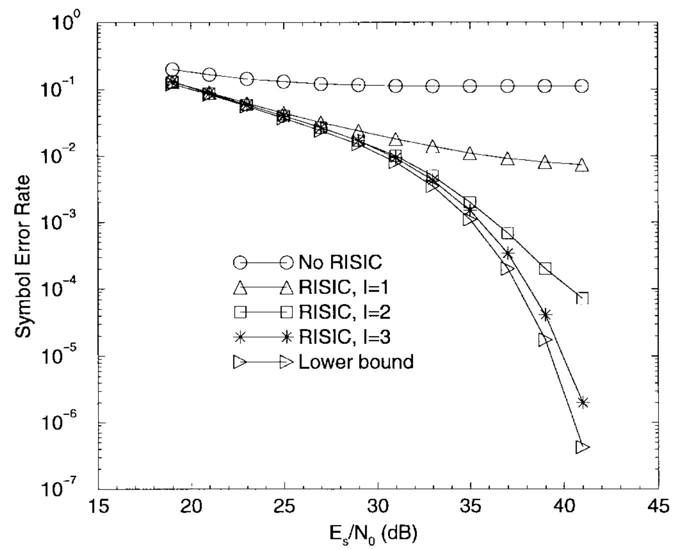


Fig. 5. Performance of the RISIC technique on Channel 2: $G = 0$, $N = 128$.

Figs. 4 and 5 illustrate the performance of the RISIC for Channels 1 and 2, respectively, where E_s represents the symbol energy and N_o represents the power spectral density of AWGN.

RISIC offers a huge improvement in the SER, even after the first iteration and especially at high SNR. Several orders of magnitude improvement in the SER are achievable for both channels after two or three iterations. The lower bounds shown in these figures are obtained by computing the SER using (21), without ISI and with noise only. On Channel 1, two iterations are required to effectively achieve the lower bound, while three iterations are required on Channel 2.

A periodic pilot sequence may be needed, even if perfect channel information is available, in order to prevent instability when feeding back a highly erroneous signal in the tail cancellation procedure. Table II shows the performance of RISIC on Channel 1 with $N = 64$ for $E_s/N_o = 35$ dB

TABLE II
EFFECT OF USING PILOT SEQUENCE WITH RISIC TECHNIQUE

Overhead	No RISIC	SER ($I = 1$)	SER ($I = 2$)	SER ($I = 3$)
0 %	0.17	0.84	0.84	0.85
2 %	0.17	7.5×10^{-3}	2.9×10^{-4}	5.8×10^{-5}
5 %	0.17	7.5×10^{-3}	2.6×10^{-4}	3.1×10^{-5}

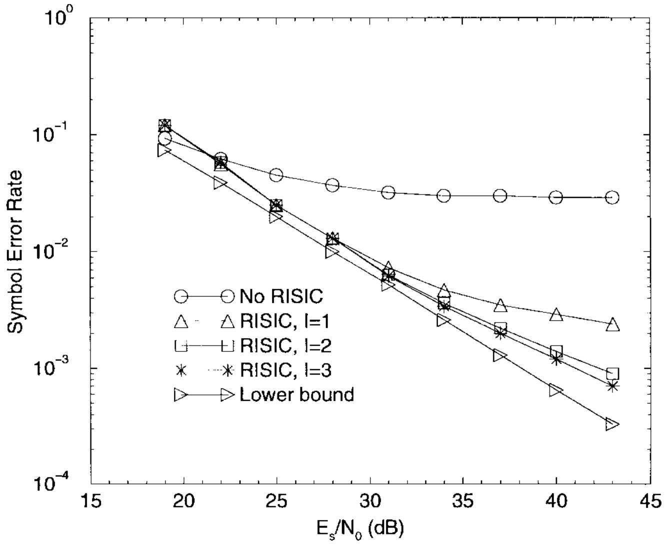


Fig. 6. Performance of the RISIC technique on a fading ISI channel: $G = 0$, $N = 1024$, $f_D N T_s = 0.001$.

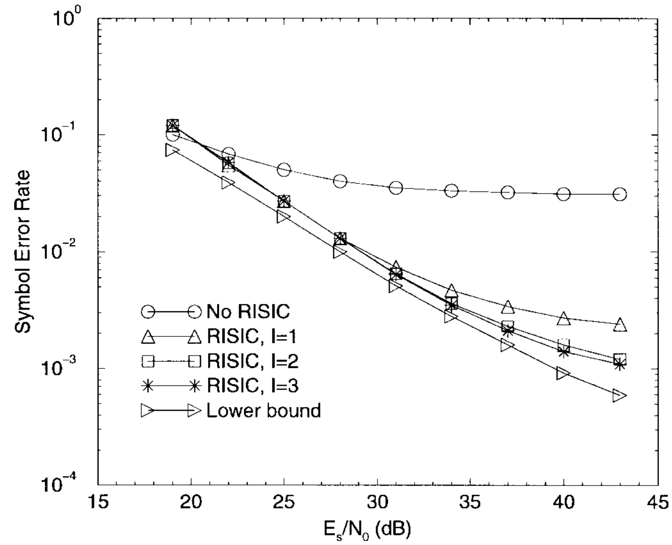


Fig. 7. Performance of the RISIC technique on a fading ISI channel: $G = 0$, $N = 1024$, $f_D N T_s = 0.005$.

and various training sequence overheads. Even with small overhead (2%) the SER improves dramatically, whereas the algorithm diverges when a pilot sequence is not used. Furthermore, only a small degradation was observed with 2% overhead as compared to 5% overhead. If the block size is increased to $N = 128$, the RISIC algorithm converged for Channel 1, although a pilot sequence was not used. Since many communications systems do utilize periodic training sequences for the purpose of synchronization or channel estimation, additional overhead to maintain stability when applying RISIC may be unnecessary.

B. Fading ISI Channel

For fading ISI channels, the channel impulse response varies with time. If the channel changes little over a block duration, however, and $M - G \ll N$, then

$$h_m \approx \frac{1}{\sqrt{N}} \mathrm{IFFT}\{\boldsymbol{\eta}\}(m). \tag{32}$$

When the channel variations are rapid, Fig. 3 shows that the performance is dominated by ICI rather than residual ISI. Assume η_n is known at the receiver and the channel impulse response is found by (32). Figs. 6 and 7 show the performance of RISIC on a TU channel with $f_D N T_s = 0.001$ and $f_D N T_s = 0.005$, respectively, and 5% training overhead.

C. Channel Estimation

All the remarkable improvements that RISIC has shown so far are achieved under the premise of perfect channel information. For a strictly static ISI channel, the channel estimation process needs to be carried out only once. In fact, almost perfect channel estimates can be obtained using a training sequence [6]. Time-varying channels present the real challenge for channel estimation.

For fading ISI channels, we propose a channel estimation technique that can provide accurate estimates, even in the presence of residual ISI. A special OFDM training block is used where a chirp sequence of $N/2$ symbols are used [6]

$$C_n = \exp\left(j \frac{2\pi}{N} n^2\right), \qquad 0 \le n \le \frac{N}{2} - 1. \tag{33}$$

The chirp sequence provides a desirable peak-to-average power ratio of one and weighs each subchannel equally in estimating the channel, since the FFT of a chirp sequence is also chirp sequence. Unlike [6], however, we modify the training sequence to increase resilience to residual ISI, by inserting zeros in every odd subchannel. The training block symbols are

$$D_n = \begin{cases} \sqrt{2} C_{n/2}, & n = 0, 2, \cdots, N-2 \\ 0, & n = 1, 3, \cdots, N-1 \end{cases} \tag{34}$$

where $\sqrt{2}$ is a normalization factor. The N-point IFFT of D_n is[1]

$$d_k = c_{(k)_{N/2}}, \qquad 0 \le k \le N-1. \tag{35}$$

From (35), the first half of the time domain training sequence $\{d_k\}$ is identical to the second half. This is a valuable property for long channel impulse responses because the first half of $\{d_k\}$ can be used just like a guard interval, while $\{d_k\}$ still possesses a peak-to-average power ratio of one. The following

[1] The c_n can also be obtained by the $N/2$-point IFFT of $\{C_n\}$.

is the channel estimation procedure using the proposed training block.

1) After removal of the guard interval, the received samples are rearranged as
$$\overline{r}_{ts,k} = r_{ts,k+(N/2)}, \qquad 0 \le k \le \frac{N}{2} - 1 \tag{36}$$
where the subscript ts indicates that the received samples are for a training block.

2) $N/2$ channel estimates $\tilde{\eta}_n$ are calculated by
$$\tilde{\eta}_n = \frac{Z_{ts,n}}{C_n}, \qquad 0 \le n \le \frac{N}{2} - 1 \tag{37}$$
where $\{Z_{ts,n}\}_{n=0}^{(N/2)-1}$ is the $N/2$-point FFT of $\{\overline{r}_{ts,k}\}_{k=0}^{(N/2)-1}$.

3) The estimates $\tilde{\eta}_n$ are converted to $\tilde{h}_m$ by
$$\tilde{h}_m = \frac{1}{\sqrt{\frac{N}{2}}} \mathrm{IFFT}\{\tilde{\boldsymbol{\eta}}\}(m) \tag{38}$$
where an $N/2$-point IFFT is used. Then, $\{\tilde{h}_m\}_{m=0}^{(N/2)-1}$ is passed through a rectangular window to zero $\tilde{h}_m, m > \hat{M}$ and the result is $\{\hat{h}_m\}_{m=0}^{(N/2)-1}$, which is used for RISIC.

4) $\{\hat{h}_m\}_{m=0}^{(N/2)-1}$ is padded by zero sequence of length $N/2$ and then converted to $\{\hat{\eta}_n\}_{n=0}^{N-1}$ by
$$\hat{\eta}_n = \sqrt{N}\, \mathrm{FFT}\{\hat{\boldsymbol{h}}\}(n), \qquad 0 \le n \le N - 1 \tag{39}$$
where, this time, an N-point FFT is used. The channel estimates for all N subchannels are now available for data demodulation.

When M is not known, the window size must be large enough to include the entire channel impulse response yet as small as possible to minimize the effect of noise. The impact of $\hat{M} > M$ is considered in Section VI.

On static ISI channels, the channel estimation is performed only once, at the beginning of each simulation run, and the channel estimates are averaged over four blocks. On fading ISI channels, the channel estimation is performed by periodically sending one training block out of every 20 blocks transmitted, while updating the channel estimate in the remaining blocks in a decision-directed mode [12]. In the decision directed mode, the decisions made after the final iteration of the RISIC algorithm are used to update the channel estimates. Decision errors do not degrade the accuracy of the estimates excessively, because the rectangular window can smooth out the aberrations caused by the decision errors.

Fig. 8 shows the performance of the RISIC technique with channel estimation on static and very slowly fading ISI channels. For a static ISI channel, the SER obtained is virtually identical to that obtained with perfect channel estimation. For a slowly varying fading ISI channel ($f_D N T_s = 0.001$), the RISIC technique works well with the channel estimation, especially at high SNR. However, for faster fading ($f_D N T_s = 0.005$), the degradation is severe, as shown in Fig. 9. This suggests that the proposed RISIC technique with channel estimation is well suited for static and slowly time-varying ISI channels.

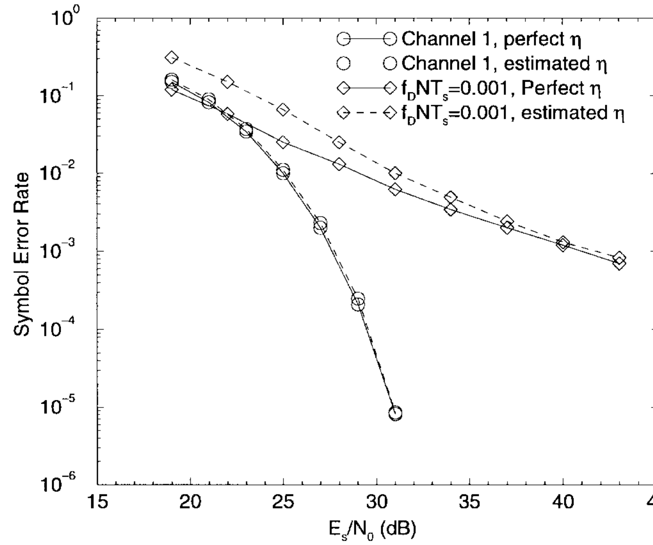


Fig. 8. Effect of imperfect channel estimation with the RISIC technique for static and slowly fading ISI channels with $N = 128$ and $N = 1024$, respectively: $G = 0, I = 3$.

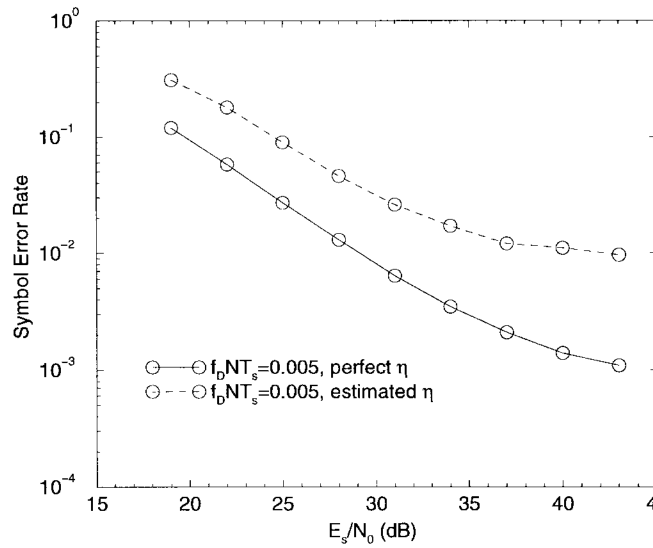


Fig. 9. Effect of imperfect channel estimation with the RISIC technique for a fading ISI channel: $N = 1024, G = 0, I = 3$.

VI. Application of RISIC to Terrestrial HDTV Broadcasting

Concatenated coding schemes have been suggested as a means to achieve the extremely low bit error rate (BER) that is required for HDTV programs [13], [14]. Since the bandwidth in North America and Japan is limited to 6 MHz, the inner code must be a bandwidth-efficient coding scheme, such as a trellis-code [15]. We first show how the RISIC technique can be used in conjunction with error correcting codes and then evaluate the link performance of a typical HDTV broadcasting system on a static ISI channel.

Trellis-codes can be generated by either feedforward or feedback encoders. Implementation of the RISIC technique requires block-by-block decisions. When trellis coding is used, this condition can be satisfied by terminating the codes at the end of each block. Therefore, we choose a feedforward encoder, since the trellis can be terminated easily with an all-zeroes tail sequence at the end of each block.

Choosing a trellis-code for a broadcast channel is a challenge because the channels are extremely diverse. It is impossible to select a single code that optimizes the performance on all static ISI channels. We can, however, choose a code that will perform well with expected channel characteristics. For television broadcasting, the channels can be categorized as a static (or quasi-static) ISI channel, where the channel is time invariant or varies very slowly in time. When OFDM is used on a static ISI channel, the frequency selectivity makes the channel behave like a flat fading channel, although there are no actual channel time variations. The reason is that the code symbols being transmitted on the different subchannels will experience different gains, i.e., some of the subchannels will appear to exhibit a fade while others do not. For interleaved flat fading channels, trellis-codes having a large minimum error event length perform better than those having large free Euclidean distances [16].

To transmit 4 bits/s/Hz over a fading channel we use independent inphase and quadrature (I-Q) coding with two rate-2/3 eight-pulse-amplitude modulation (PAM) trellis-codes that have shown good performance [8]. Since the encoders for these rate-2/3 eight-PAM codes are given in feedback realization in [8], we need to convert to a feedforward realization. This can be done by following the conversion rules in [16]. For a constraint length-five encoder, the generator matrix is

$$\mathbf{G} = \begin{bmatrix} D & 0 & D^2+D+1 \\ 1 & D^3+D^2+D & D^2+D \end{bmatrix}. \tag{40}$$

For the outer code, we use a shortened RS(204, 188) code with eight bits per symbol, that can correct eight symbol errors [17].

The RISIC technique with concatenated coding is essentially the same as described in Section IV-A. The main difference is that the decoders must be used in the decision-making process, and the feedback signal used in the cancellation procedure is obtained from the code symbols (with frequency-interleaving). Generally, the outer code will provide satisfactory performance if the inner code can provide an SER of 10^{-3} or lower. As we have seen earlier, the RISIC algorithm performs well, even when the SER is much higher than 10^{-2}. Therefore, the complexity involved in using the RISIC technique with concatenated coding can be reduced by using the decisions at the output of the inner trellis-code in the RISIC algorithm. The outer decoder uses the decisions at the output of the inner decoder after the last iteration of the RISIC algorithm is finished. The channel estimates can be updated using the final decisions from either the inner or outer decoder. Since the SER at the output of the outer decoder is extremely low in HDTV broadcasting, our simulations are only performed for the inner trellis-code, while the performance of the outer code is evaluated analytically [13], [17].

For the 20 Mb/s data rate assumed in this paper, a total bandwidth of about 5.73 MHz is required, including overhead for a training sequence (5%), coding, and trellis termination. This meets the 6-MHz bandwidth assigned for HDTV broadcasting in Japan and North America. Channel estimation is performed

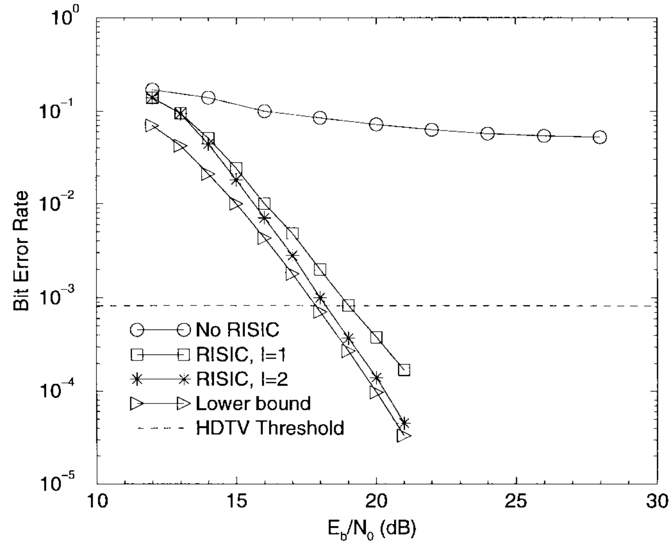


Fig. 10. BER of the inner trellis-code with the RISIC algorithm.

at the beginning of the simulation by averaging four training blocks. Periodic training blocks are also transmitted at a rate of one training block out of every 20 blocks sent.

The channel is based on the COST-207 six-tap typical HT model [9], with fixed tap coefficients. We choose a block size of $N = 1024$ with the guard interval of 5 μs. The guard interval is long enough to mitigate the ISI for the TU channel model, but much less than the maximum delay spread for the HT model which is 17.2 μs.

The HDTV viewing threshold is based on the PER. After the performance of the inner trellis-code has been found by simulation, the PER at the output of the RS decoder is approximated by [13]

$$\mathrm{PER} \approx \sum_{k=t+1}^{L} \binom{L}{k} \mathrm{SER}^k (1-\mathrm{SER})^{L-k} \tag{41}$$

where L is the RS codeword length in symbols, t is the error correction capability of the RS code in symbols, and SER is the probability of code symbol error. If sufficient interleaving is used between the inner and outer codes, then $\mathrm{SER} = 1-(1-\mathrm{BER})^n$, where BER is the bit error probability at the output of the inner decoder and n is the number of bits per RS symbol. The use of such interleaving is for analytical convenience and better performance can actually be achieved by not interleaving and exploiting the error trapping property of the RS code.

The threshold for satisfactory HDTV viewing is set at $\mathrm{PER} = 10^{-5}$ [18], requiring a BER at the output of the inner decoder of 8.2×10^{-4}. Figs. 10 and 11 show the performance of the inner and outer codes, respectively, with $\hat{M} = M$. Without RISIC, the result is catastrophic. With RISIC, however, HDTV quality can be achieved at $E_b/N_o \approx 19$ dB after one iteration and $E_b/N_o \approx 18$ dB after two iterations where E_b represents the energy per bit. The lower bound is achieved if the ISI is completely removed by using a long guard interval and the channel estimates are perfect. The

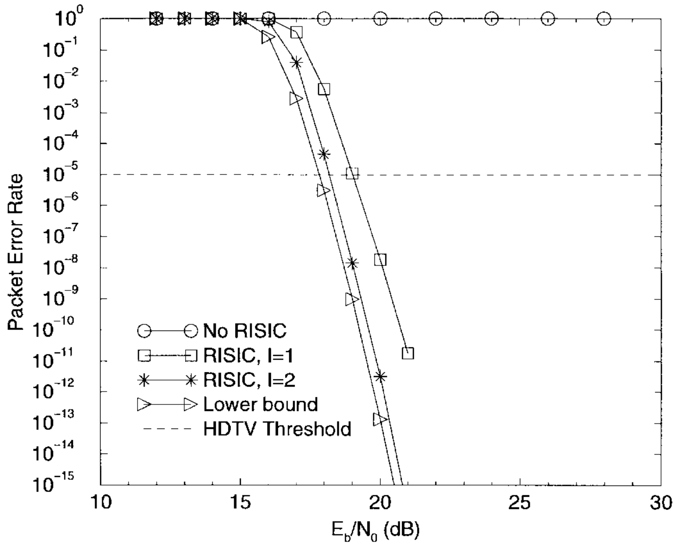


Fig. 11. PER's for concatenated coding with the RISIC algorithm.

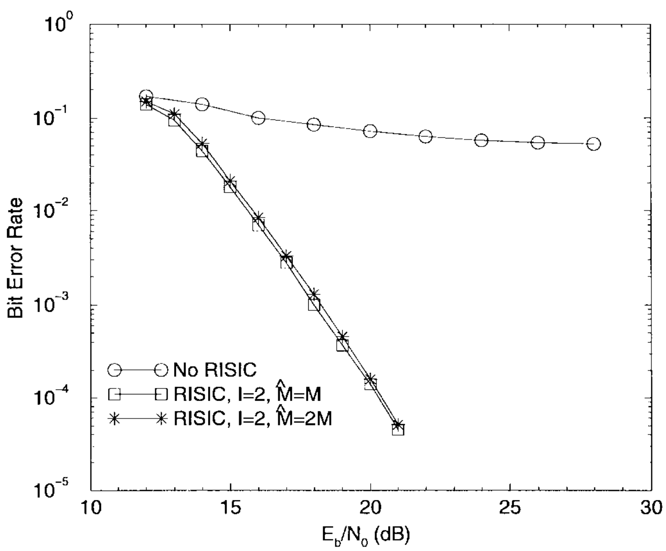


Fig. 12. Effect of using $\hat{M} > M$ with the RISIC algorithm.

lower bound is only about 0.3 dB better than the performance of the RISIC algorithm with two iterations (see Fig. 11).

The effect of choosing $\hat{M} > M$ is shown in Fig. 12. With $\hat{M} = 2M$, the BER out of the inner decoder is almost the same as that with $\hat{M} = M$. Therefore, it is easy to choose $\hat{M}$ sufficiently large for the RISIC algorithm to work, without causing much performance degradation due to noise.

Finally, the computational burden of the RISIC algorithm is very high. The number of FFT operations (including IFFT) required for I iterations is $2I + 2$. On the other hand, there are very fast FFT processors commercially available in the market and even faster processors have been developed and demonstrated in the academy.[2] For example, the COBRA processor [19] spends only 9.5 μs for one 1024-FFT operation. With this speed, processing time required for three iterations of the RISIC algorithm is 76 μs. Note that one OFDM block period of our HDTV scheme is about 184 μs, which is much larger than the required processing time. The overall processing time will be larger since coding, decoding, and convolution procedures must also be considered. Nevertheless, implementation of the RISIC algorithm seems quite feasible with the rapid advancements in the related technologies.

[2] FFT Info Page, http://www-star.stanford.edu/~bbaas/fftinfo.html.

VII. Concluding Remarks

This paper has shown that residual ISI can severely degrade the performance of the OFDM systems. A very effective algorithm, RISIC, was introduced to cancel the residual ISI. The RISIC algorithm is capable of removing residual ISI almost completely on static or slowly fading ISI channels. The proposed technique can be applied to terrestrial HDTV broadcasting systems in order to successfully receive HDTV signals that would otherwise be impossible to receive.

The RISIC algorithm can provide future consumers of HDTV receivers with two options: a regular HDTV receiver where only a guard interval is used to mitigate the ISI, or a more costly receiver that uses the RISIC algorithm to suppress ISI that is longer than the guard interval.

References

[1] M. Alard and R. Lassalle, "Principles of modulation and channel coding for digital broadcasting to mobile receivers," *EBU Rev.*, vol. 224, pp. 168–190, Aug. 1987.

[2] A. Ruiz, J. M. Cioffi, and S. Kasturia, "Discrete multiple tone modulation with coset coding for the spectrally shaped channel," *IEEE Trans. Commun.*, vol. 40. pp. 1012–1029, June 1992.

[3] M. Sablatash, "Transmission of all-digital advanced television: State of the art and future directions," *IEEE Trans. Broadcast.*, vol. 40, pp. 102–121, Jan. 1994.

[4] G. Malmgren, "Single frequency broadcasting networks," Ph.D. dissertation, Royal Instit. Technol., Stockholm, Sweden, 1997.

[5] E. Viterbo and K. Fazel, "How to combat long echoes in OFDM transmission schemes: Subchannel equalization or more powerful channel coding," in *Proc. GLOBECOM '95*, Singapore, China, Nov. 1995, pp. 2069–2074.

[6] J. M. Cioffi and A. C. Bingham, "A data-driven multitone echo canceller," *IEEE Trans. Commun.*, vol. 42. pp. 2853–2869, Oct. 1994.

[7] D. Kim, "Orthogonal frequency division multiplexing for digital broadcasting," Ph.D. dissertation, Georgia Instit.Technol., Atlanta, Dec. 1998.

[8] M. Russell and G. L. Stüber, "Terrestrial digital video broadcasting for mobile reception using OFDM," *Kluwer Wireless Personal Commun.*, vol. 2, pp. 45–66, 1995.

[9] COST 207 Management Committee, "COST 207: Digital land mobile radio communications," Commission of the European Communities, Luxembourg, Belgium, Final Report, 1989.

[10] W. C. Jakes, Ed., *Microwave Mobile Communications.* New York: IEEE, 1994.

[11] J. G. Proakis, *Digital Communications*, 3rd ed. New York: McGraw-Hill, 1994.

[12] V. Mignone and A. Morello, "CD3-OFDM: A novel demodulation scheme for fixed and mobile receivers," *IEEE Trans. Commun.*, vol. 44. pp. 1144–1151, Sept. 1996.

[13] C. Heegard, S. A. Lery, and W. H. Paik, "Practical coding for QAM transmission of HDTV," *IEEE J. Select. Areas Commun.*, vol. 11, pp. 111–118, Jan. 1993.

[14] F. F. Tzeng, "Error protection for satellite broadcast of HDTV and conventional TV," in *Proc. GLOBECOM '93*, Houston, TX, Nov. 1993, pp. 1617–1621.

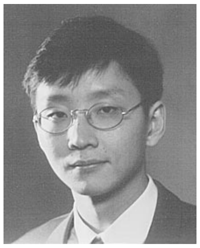
[15] G. Ungerboeck, "Channel coding with multilevel/phase signals," *IEEE Trans. Inform. Theory*, vol. 28, pp. 56–67, Jan. 1982.

[16] E. Biglieri, D. Divsalar, P. J. McLane, and M. K. Simon, *Introduction to Trellis-Coded Modulation with Applications.* New York: Macmillan, 1991.

[17] D. Kim and G. L. Stüber, "Performance of multiresolution OFDM in frequency-selective fading channels," in *Proc. GLOBECOM '97*, Phoenix, AZ, Nov. 1997, pp. 16–20.

[18] K. Ramchandran, A. Ortega, K. M. Uz, and M. Vetterli, "Multiresolution broadcast for digital HDTV using joint source/channel coding," *IEEE J. Select. Areas Commun.*, vol. 11, pp. 6–22, Jan. 1993.

[19] G. Sunada, J. Jin, M. Berzins, and T. Chen, "COBRA: An 1.2 million transistor expandable column FFT chip," in *Proc. IEEE Int. Conf. on Computer Design*, Cambridge, MA, Oct. 1994, pp. 546–550.



Dukhyun Kim (S'92) received the B.S. degree in electrical engineering from Rutgers University, Piscataway, NJ, in 1993 and the M.S. degree from the Georgia Institute of Technology, Atlanta, in 1995. He is currently working toward the Ph.D. degree in the School of Electrical and Computer Engineering at the Georgia Institute of Technology.

During the summer of 1992, he worked as an intern at the research and development center of Ericsson, Research Triangle Park, NC. Since 1993 he has been a research assistant at the Georgia Institute of Technology. His research interests include design and analysis of wireless and mobile communication systems, multicarrier modulation, and digital broadcasting.

Gordon L. Stüber (S'82–M'82–SM'96) received the B.A.Sc. and Ph.D. degrees in electrical engineering from the University of Waterloo, Waterloo, Ontario, Canada, in 1982 and 1986, respectively.

In 1986, he joined the School of Electrical and Computer Engineering, Georgia Institute of Technology, Atlanta, GA, where he is currently a Professor. His research interests are in wireless communications and communication signal processing. He is the author of *Principles of Mobile Communication* (Boston, MA: Kluwer, 1996).

Dr. Stüber served as an Editor for spread spectrum with the IEEE TRANSACTIONS ON COMMUNICATIONS from 1993 through 1997. He served as the Technical Program Chair for the 1996 IEEE Vehicular Technology Conference and the 1998 IEEE International Conference on Communications.