Robust OFDM in Fast Fading Channels

Sungsoo Kim and Gregory J. Pottie
Department of Electrical Engineering
University of California, Los Angeles
Los Angeles, CA 90095
Email: sungskim@ee.ucla.edu, pottie@ee.ucla.edu

***Abstract*— Techniques for reducing interchannel interference (ICI) of orthogonal frequency division multiplexing (OFDM) systems under a fast fading channel are presented. Combined frequency domain equalization and bit-interleaved coded modulation (BICM) are investigated. Using the fact that ICI energy is concentrated in adjacent subchannels, the complexity of the frequency domain equalization can be significantly reduced without much performance degradation. New bit metrics for the BICM are derived to improve the performance when the frequency domain equalizer and BICM are used together. Computer simulation demonstrates the robustness of the suggested techniques even when the normalized Doppler frequency is higher than 0.1.**

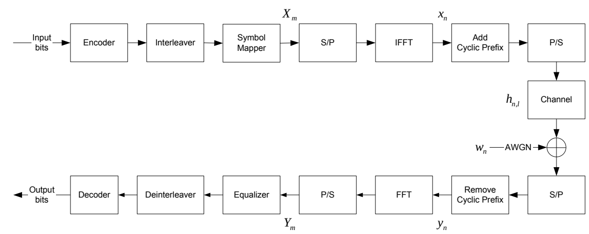

Fig. 1. A baseband equivalent model for an OFDM system.

I. Introduction

Orthogonal frequency division multiplexing (OFDM) is known as an effective modulation technique in highly frequency selective channel conditions. The entire transmission channel of the OFDM systems is divided into many narrow subchannels, and each subchannel is modulated by orthogonal subcarriers [1]. In time-selective channel environments, on the other hand, time variations of the channel within an OFDM symbol interval lead to a loss of subchannel orthogonality, resulting in interchannel interference (ICI) and an irreducible error floor in conventional OFDM receivers [2]. The performance degradation due to ICI becomes significant as normalized Doppler frequency increases, which is the ratio of maximum Doppler frequency shift to the subchannel bandwidth [3].

In [4], a simplified frequency domain equalization is suggested to reduce ICI. The complexity of the equalizer can be reduced significantly by using the fact that the energy of the ICI is concentrated in adjacent subchannels – in other words, only a few adjacent subchannels are major interferers to a desired subchannel. Bit-interleaved coded modulation (BICM) is first introduced in [6], and further analyzed in [7]. The performance of the coded modulation over a fast fading channel can be improved by bit-wise interleaving at the encoder output, and by using an appropriate soft-decision metric as an input to a Viterbi decoder. Even though the metric is suboptimal, the BICM outperforms the conventional trellis-coded modulation (TCM) over a fast fading channel , mainly due to the additional diversity induced by the bit-wise interleaver.

In this paper both frequency domain equalization and BICM techniques are combined for robust and reliable reception in OFDM systems in a fast fading channel environment. A new design method for simplified frequency domain equalization using the minimum mean-squared error (MMSE) criterion is obtained. Also, new bit metrics are derived to improve the performance of the BICM in the presence of the equalizer.

This paper is organized as follows. In Section II, the overall system model is described and properties of ICI are investigated. A simplified MMSE equalizer and bit metrics for BICM are designed in Section III. The performance of the suggested techniques is demonstrated by computer simulation in Section IV, and finally conclusions are made in Section V.

II. System Model and Properties of ICI

In this section the mathematical model of an OFDM system for a time-varying channel is described. Properties of ICI are shown, and these properties will be used for designing a hardware-efficient MMSE equalizer in the next section.

A. System Model under Time-Varying Channel

Fig. 1 shows an overall baseband equivalent model for an OFDM system. The output of IFFT x_n at time n is given by

$$x_n = \frac{1}{\sqrt{N}} \sum_{m=0}^{N-1} X_m e^{j2\pi nm/N}, \tag{1}$$

where N, and X_m are the number of subchannels of the OFDM system, and the outputs of encoder at subchannel m, respectively. By assuming that the channel consists of L multipath components, the output of the channel can be given by [8]

$$y_n = \sum_{l=0}^{L-1} h_{n,l} x_{n-l} + w_n, \tag{2}$$

where $h_{n,l}$ and w_n represent the channel impulse response (CIR) of l^{th} path and additive white Gaussian noise (AWGN) at time n, respectively. From (1), y_n can be written as

$$y_n = \frac{1}{\sqrt{N}} \sum_{m=0}^{N-1} X_m H_n^{(m)} e^{j2\pi nm/N}, \tag{3}$$

where $H_n^{(m)} \equiv \sum_{l=0}^{L-1} h_{n,l} e^{-j2\pi lm/N}$ is the Fourier transform of the channel impulse response for subchannel m at time n. After removing the cyclic prefix, the output of the fast Fourier transform (FFT) is

$$Y_m = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} y_n e^{-j2\pi nm/N}. \tag{4}$$

By inserting (3) into (4), Y_m can be rewritten as [4]

$$Y_m = \alpha_m X_m + \beta_m + W_m, \tag{5}$$

where

$$\begin{aligned} W_m &= \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} w_n e^{-j2\pi nm/N} \\ \alpha_m &= \frac{1}{N} \sum_{n=0}^{N-1} H_n^{(m)} \\ \beta_m &= \frac{1}{N} \sum_{k=0, k\neq m}^{N-1} X_k \sum_{n=0}^{N-1} H_n^{(k)} e^{-j2\pi(m-k)n/N}. \end{aligned} \tag{6}$$

Here, W_m, α_m, and β_m represent the Fourier transform of w_n, the multiplicative distortion of a subchannel m, and the ICI caused by a time-varying channel, respectively. Note that α_m is the average frequency response of the CIR over one OFDM symbol period. In other words, in a time-invariant channel $H_n^{(k)}$ is not a function of n, and α_m simply becomes the frequency response of CIR as usual.

We can express (5) in a vector-matrix form as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{w} \tag{7}$$

where $\mathbf{y} = [Y_0, \ldots, Y_{N-1}]^T$, $\mathbf{x} = [X_0, \ldots, X_{N-1}]^T$, $\mathbf{w} = [W_0, \ldots, W_{N-1}]^T$, and

$$\mathbf{H} = \begin{bmatrix} H_{0,0} & H_{0,1} & \cdots & H_{0,N-1} \\ H_{1,0} & H_{1,1} & \cdots & H_{1,N-1} \\ \vdots & \vdots & \ddots & \vdots \\ H_{N-1,0} & H_{N-1,1} & \cdots & H_{N-1,N-1} \end{bmatrix}. \tag{8}$$

Here, $H_{m,k}$ in (8) is defined as

$$H_{m,k} \equiv \frac{1}{N} \sum_{n=0}^{N-1} H_n^{(k)} e^{-j2\pi(m-k)n/N}. \tag{9}$$

In an OFDM system over a time-varying channel, the ICI can be characterized by *normalized Doppler frequency* $f_d T$ where f_d is the maximum Doppler frequency and T is the time duration of one OFDM symbol. Hence we can think of the normalized Doppler frequency as a maximum cycle change of the time-varying channel per symbol duration in a statistical sense.

The β_m's in (5), or off-diagonal elements of $\mathbf{H}$ in (8) represent the ICI caused by the time-varying nature of the channel. In a time-invariant channel, one can easily see that β_m is zero, or $\mathbf{H}$ becomes a diagonal matrix, due to the orthogonality of the multicarrier basis waveforms. In a slowly time-varying channel (i.e. the normalized Doppler frequency $f_d T$ is small), we can assume $|\beta_m|^2 \approx 0$. On the other hand, when the normalized Doppler frequency is high, the ICI power cannot be ignored and produces an irreducible error floor in conventional OFDM receivers.

B. Property of ICI

In [8], an explicit mathematical expression for ICI power is derived. By assuming that the multipath intensity profile has an exponential distribution, and the inverse Fourier transform of the Doppler spectrum is the zeroth-order Bessel function of the first kind, the autocorrelation function of the channel is

$$E\left\{h_{n_1,l_1} h_{n_2,l_2}^*\right\} = c \cdot J_0\left(\frac{2\pi f_d T\,(n_1 - n_2)}{N}\right) \cdot e^{-l_1/L} \delta\,(l_1 - l_2) \tag{10}$$

where c, a normalization constant, is chosen to satisfy $c \sum_l e^{-l/L} = 1$, $J_0(\cdot)$ denotes the zeroth-order Bessel function of the first kind. Assuming the data on each subchannel is uncorrelated, and $E\{|X_m|^2\} = 1$, the normalized ICI power of subchannel m caused by subchannel k is [8]

$$\gamma_{m,k} \equiv \frac{N + 2\sum_{n=1}^{N-1} (N-n)\, J_0\left(\frac{2\pi f_d T n}{N}\right) \cos\left(\frac{2\pi n(m-k)}{N}\right)}{N + 2\sum_{n=1}^{N-1} (N-n)\, J_0\left(\frac{2\pi f_d T n}{N}\right)}, \tag{11}$$

and the total ICI power for subchannel m is then given by

$$\gamma_m = \sum_{k=0, k\neq m}^{N-1} \gamma_{m,k}. \tag{12}$$

Fig. 2 shows the distribution of normalized ICI power among subchannels for different $f_d T$ values. Note that the overall normalized ICI power level increases as the normalized Doppler frequency increases. Also note that the ICI power tends to concentrate in the neighborhood of the desired subchannel which is set to be zero in the Fig. 2. In other words, $\gamma_{m,k_1} > \gamma_{m,k_2}$ if $|m - k_1| < |m - k_2|$ for any $0 \le m, k_1, k_2 \le N-1$ and $k_1, k_2 \neq m$. Because the ICI power decreases significantly as $|m - k|$ increases, it is inefficient to use the entire set of subchannels to equalize a particular desired subchannel. This idea is the key for designing a hardware-efficient equalizer in the next section.

III. Reduced-tap MMSE Equalizer and BICM

The conventional detection of OFDM signals using a single tap equalizer exhibits relatively good performance at low values of $f_d T$. However, in an environment where the normalized Doppler frequency is high, orthogonality between subchannels breaks down, and there is an irreducible error floor due to the interference induced between subchannels. In this section we design simplified MMSE equalizers in the frequency domain

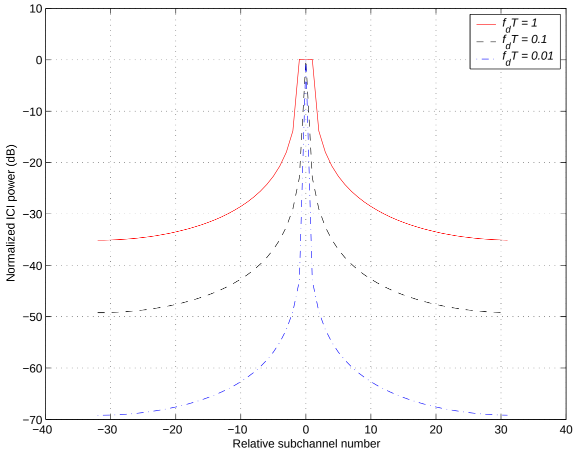

Fig. 2. Distribution of normalized ICI power, N=64

and appropriate bit metrics for BICM. First, a traditional MMSE equalizer design approach is described, and a new design method using a reduced-tap MMSE equalizer in the frequency domain is presented. Then we propose improved bit metrics for BICM in the presence of the equalizer.

A. Reduced-tap MMSE Equalizer

Our OFDM system model is

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{w} \tag{13}$$

as given in (7). In this problem, we want to find the N-by-N equalizer matrix $\mathbf{G}$ which minimizes the cost function $E\{|\mathbf{x}-\hat{\mathbf{x}}|^{\mathbf{2}}\}$, where $\hat{\mathbf{x}} = \mathbf{G}\mathbf{y}$ is the equalizer output vector. This is the classical MMSE design problem, and the solution is given as

$$\mathbf{G} = \mathbf{R}_{\mathbf{xy}}\mathbf{R}_{\mathbf{y}}{}^{-1} \tag{14}$$

where $\mathbf{R}$ denotes the covariance matrix, which is defined as $\mathbf{R}_{\mathbf{xy}} = E\left\{\mathbf{x}\mathbf{y}^{\mathrm{H}}\right\}$ and $\mathbf{R}_{\mathbf{y}} = E\left\{\mathbf{y}\mathbf{y}^{\mathrm{H}}\right\}$. Here the superscript H denotes complex conjugate transpose. The resulting MMSE is then

$$\mathrm{MMSE} = \mathrm{Tr}\left(\mathbf{R}_{\mathbf{x}} - \mathbf{R}_{\mathbf{xy}}\mathbf{R}_{\mathbf{y}}^{-1}\mathbf{R}_{\mathbf{yx}}\right) \tag{15}$$

where $\mathbf{R}_{\mathbf{x}} = E\left\{\mathbf{x}\mathbf{x}^{\mathrm{H}}\right\}$ and $\mathbf{R}_{\mathbf{yx}} = E\left\{\mathbf{y}\mathbf{x}^{\mathrm{H}}\right\}$, and $\mathrm{Tr}(\cdot)$ denotes trace function. Assuming $\mathbf{H}$ is known, $\mathbf{x}$ is a zero-mean i.i.d. random vector with variance $\sigma_x{}^2$, and $\mathbf{w}$ is an AWGN vector with variance $\sigma_w{}^2$, then $\mathbf{R}_{\mathbf{xy}}$ is

$$\mathbf{R}_{\mathbf{xy}} = \sigma_x^2\mathbf{H}^{\mathrm{H}}, \tag{16}$$

and $\mathbf{R}_{\mathbf{y}}$ is

$$\mathbf{R}_{\mathbf{y}} = \sigma_x^2\mathbf{H}\mathbf{H}^{\mathrm{H}} + \sigma_w^2\mathbf{I}_N, \tag{17}$$

where $\mathbf{I}_N$ is the N-by-N identity matrix. Then (14) can be rewritten as

$$\mathbf{G} = \mathbf{H}^{\mathrm{H}}\left(\mathbf{H}\mathbf{H}^{\mathrm{H}} + \frac{\sigma_w{}^2}{\sigma_x{}^2}\mathbf{I}_N\right)^{-1}. \tag{18}$$

Likewise, (15) becomes

$$\mathrm{MMSE} = \sigma_x^2\mathrm{Tr}\left(\mathbf{I} - \mathbf{G}\mathbf{H}\right) \tag{19}$$

As can be seen from (18), the MMSE equalizer is too complex to be implemented, especially when N is a large number. First, an N-by-N matrix inversion is required to obtain the equalizer coefficient matrix $\mathbf{G}$, and N^2 complex multipliers are needed to equalize N symbols.

Using the fact that the ICI power is localized to the neighborhood of a desired subchannel, only a few neighborhood subchannels can be used for equalization without much performance penalty. Derivation of the q-tap MMSE equalizer is similar to the MMSE case. This time, however, we find the solution for each desired subchannel individually. The problem is to find the equalizer coefficient vector $\mathbf{g}_m = \left[g_{m,0}, \ldots, g_{m,q-1}\right]$ which minimizes the mean-squared error $E\left\{\left|X_m - \hat{X}_m\right|^2\right\}$ where $\hat{X}_m = \mathbf{g}_m\mathbf{y}_m$ and $\mathbf{y}_m = \left[Y_{(m-(q-1)/2)_N}, \ldots, Y_{(m+(q-1)/2)_N}\right]^{\mathrm{T}}$. Here $(\cdot)_N$ denotes modular function with modulus N. $\mathbf{y}_m$ is then

$$\mathbf{y}_m = \mathbf{H}_m\mathbf{x} + \mathbf{w}_m \tag{20}$$

where $\mathbf{w}_m = \left[W_{(m-(q-1)/2)_N}, \ldots, W_{(m+(q-1)/2)_N}\right]^{\mathrm{T}}$ and

$$\mathbf{H}_m = \begin{bmatrix} H_{(m-(q-1)/2)_N,0} & \cdots & H_{(m-(q-1)/2)_N,N-1} \\ \vdots & \vdots & \vdots \\ H_{m,0} & \cdots & H_{m,N-1} \\ \vdots & \vdots & \vdots \\ H_{(m+(q-1)/2)_N,0} & \cdots & H_{(m+(q-1)/2)_N,N-1} \end{bmatrix}. \tag{21}$$

From (14), the MMSE solution is

$$\mathbf{g}_m = \mathbf{R}_{X_m\mathbf{y}_m}\mathbf{R}_{\mathbf{y}_m}^{-1}, \tag{22}$$

and, by the same assumption as in the previous section, we have

$$\begin{aligned} \mathbf{R}_{X_m\mathbf{y}_m} &= E\left\{X_m\mathbf{y}_m^{\mathrm{H}}\right\} \\ &= \sigma_x^2\mathbf{h}_m^{\mathrm{H}} \end{aligned} \tag{23}$$

where $\mathbf{h}_m$ is the m^{th} column of the matrix $\mathbf{H}_m$, i.e.,

$$\mathbf{h}_m = \left[H_{(m-(q-1)/2)_N,m}, \ldots, H_{(m+(q-1)/2)_N,m}\right]^{\mathrm{T}}. \tag{24}$$

Also we have

$$\begin{aligned} \mathbf{R}_{\mathbf{y}_m} &= E\left\{\mathbf{y}_m\mathbf{y}_m^{\mathrm{H}}\right\} \\ &= \sigma_x^2\mathbf{H}_m\mathbf{H}_m^{\mathrm{H}} + \sigma_w^2\mathbf{I}_q. \end{aligned} \tag{25}$$

After inserting (23) and (25) into (22), the q-tap equalizer vector $\mathbf{g}_m$ becomes

$$\mathbf{g}_m = \mathbf{h}_m^{\mathrm{H}}\left(\mathbf{H}_m\mathbf{H}_m^{\mathrm{H}} + \frac{\sigma_w^2}{\sigma_x^2}\mathbf{I}_q\right)^{-1}. \tag{26}$$

Similarly, we have

$$\mathrm{MMSE} = \sigma_x^2\sum_{m=0}^{N-1}\left(1 - \mathbf{g}_m\mathbf{h}_m\right). \tag{27}$$

By choosing an appropriate number q, we can reduce complexity of the equalizer significantly. For example, when $N = 64$,

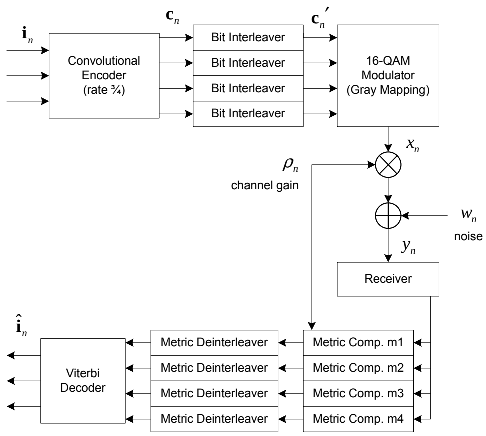

Fig. 3. Block diagram of a BICM system

the full-tap MMSE requires 64-by-64 matrix inversion with 4096 complex multipliers, while the 3-tap MMSE equalizer needs 64 3-by-3 matrix inversions with only 192 complex multipliers.

B. Bit-Interleaved Coded Modulation

OFDM is an attractive modulation technique for the systems employing convolutional channel coding because of its inherent orthogonality between subchannels. When an OFDM system does not suffer from ICI, each subchannel can be viewed as an independent flat fading channel. By independent we mean an impulse channel so that, when transmitting convolutional codes, the decoder's trellis has the same number of states as the transmitter's. Single carrier systems under multipath channels, on the other hand, suffer from intersymbol interference (ISI) so that much more complex algorithms are required for maximum likelihood decoding.

It is known that the BICM technique, based on a convolutional code followed by bit interleavers, yields higher coding gain over a Rayleigh fading channel than the original trellis-coded modulation (TCM) [5], [6]. The diversity of a coded system can be increased with this approach, and the diversity is proportional to free binary Hamming distance d_{free} of the code, and the error performance is governed by some function of d_{free}. First, the BICM decoding method is briefly described. And then new bit metrics are designed for proper interaction with the equalizer.

An example of the BICM system is shown in Fig. 3. In this example, a binary sequence $\mathbf{i}_n$ at time n is encoded into another binary sequence $\mathbf{c}_n$ using a rate $R = \frac{3}{4}$, convolutional code. The encoder outputs are fed into four independent ideal interleavers, resulting in a binary vector $\mathbf{c}_n{}' = \left[c_n^1{}', c_n^2{}', c_n^3{}', c_n^4{}'\right]$. A group of 4 bits at the output of the interleavers is mapped into the 16-QAM signal set x_n. The mapped signal points are digitally pulse shaped, and transmitted over the channel. At the receiver, a faded noisy version of the transmitted signal y_n can be written as

$$y_n = \rho_n x_n + w_n \tag{28}$$

where ρ_n is a random variable representing the random amplitude of the received signal, and w_n is a complex zero-mean Gaussian random variable with variance ${\sigma_w}^2$. The received signal is then passed through a demodulator, four metric computation units, and metric de-interleavers. Finally, the decision on the transmitted sequence is taken with the aid of the Viterbi decoder. At the receiver, the faded, noisy version of the transmitted symbol is passed through four metric computation units. An optimal decoder calls for a complicated metric which takes into account the apriori probabilities of transmission of all possible transmitted symbols associated with the output bit c_n^i. In selecting a decoding metric, a tradeoff exists between simplicity of implementation, robustness of the system, and error performance. In [6], the suboptimal metric

$$m_i\left(y_n^i, S_i^c; \rho_n^i\right) = \min_{x \in S_i^c} \left|y_n^i - \rho_n^i x\right|^2, \quad c = 0, 1 \tag{29}$$

is suggested.

We can combine the BICM with the q-tap MMSE equalizer designed previously to increase diversity as well as signal to interference noise ratio (SINR). The problem with this combination is to determine what metrics are appropriate in the presence of the equalizer. Without the equalizer, the metrics should be the same as (29), i.e.,

$$\text{Metric0} \equiv \min_{x \in S^c} \left|y_n - H_{n,n} x\right|^2, \quad c = 0, 1 \tag{30}$$

where $H_{n,n}$ in (9) is the channel gain at the subchannel n. Note that off-diagonal elements of $\mathbf{H}$ in (8) are all zeros since there is no ICI. However, in the presence of equalizers, Metric0 cannot be used directly since the input of the metric computation units is not the channel output y_n, but the equalizer output $\hat{x}_n$. Hence the simplest metrics in the presence of equalizers could be

$$\text{Metric1} \equiv \min_{x \in S^c} \left|\hat{x}_n - x\right|^2, \quad c = 0, 1. \tag{31}$$

Metric1, however, does not include the subchannel gain information which may improve the performance greatly. If we scale the metrics based on the power of each subchannel, then we have

$$\text{Metric2} \equiv \min_{x \in S^c} \left|\hat{x}_n - x\right|^2 |H_{n,n}|^2, \quad c = 0, 1. \tag{32}$$

We can improve further by examining the equalizer output,

$$\begin{aligned} \hat{x}_n &= \mathbf{g}_n \mathbf{y}_n \\ &= \mathbf{g}_n \left(\mathbf{H}_n \mathbf{x} + \mathbf{w}_n\right) \\ &= \mathbf{g}_n \mathbf{h}_n x_n + \mathbf{g}_n \left(\mathbf{w}_n + \mathbf{w}'_n\right) \end{aligned} \tag{33}$$

where $\mathbf{g}_n = [g_{n,0}, \ldots, g_{n,q-1}]$ is the equalizer coefficient vector, the channel matrix $\mathbf{H}_n$ is from (21), and $\mathbf{h}_n$ is n^{th} column of $\mathbf{H}_n$. Here $\mathbf{w}'_n$ denotes ICI associated with $\mathbf{y}_n$. Since

$$\begin{aligned} \left|\hat{x}_n - \mathbf{g}_n \mathbf{h}_n x_n\right|^2 &= \left|\mathbf{g}_n \left(\mathbf{w}_n + \mathbf{w}'_n\right)\right|^2 \\ &\simeq \left|\mathbf{g}_n\right|^2 \left|\mathbf{w}_n + \mathbf{w}'_n\right|^2, \end{aligned} \tag{34}$$

new metrics can be obtained as

$$\text{Metric3} \equiv \frac{1}{\left|\mathbf{g}_n\right|^2} \min_{x \in S^c} \left|\hat{x}_n - \mathbf{g}_n \mathbf{h}_n x\right|^2. \tag{35}$$

IV. Simulation Results

Fig. 4 illustrates the mean-squared error performance of the MMSE equalizers as a function of SNR when $f_dT = 0.1$. In this simulation, the number of subchannels N is 64, and 1-tap, 3-tap, 5-tap, and full 64-tap MMSE equalizers are under consideration. Note that the curve of the 64-tap MMSE equalizer is almost a straight line, meaning that the full-tap MMSE equalizer does not suffer from an irreducible error floor due to the ICI unlike other fewer-tap equalizers. Also note that the error floor decreases as the number of the equalizer taps increase. Fig. 5 compares BER performance for different metrics. In this simulation the number of subcarriers $N = 32$, the normalized Doppler frequency f_dT=0.4, 4-QAM modulation, and 64-state convolutional code with rate 1/2 are used. As expected, the performance order is Metric1 $<$ Metric2 $<$ Metric3. At BER=10^{-4}, Metric3 has 1.2 dB gain over Metric2, and 0.2 dB over Metric1 for the 3-tap MMSE equalizer. Fig. 6 shows the BER performance under various conditions. Only the systems having both equalizer and BICM do not suffer from serious ICI degradation. Note that both '1-tap with BICM' and 'BICM only' has an irreducible error floor due to the severe ICI impairments.

V. Conclusion

As demand for high speed communications under various mobile scenarios rises, the ICI problem of OFDM systems become an important issue. In this paper combined techniques for robust OFDM under fast fading channels are proposed – reduced-tap frequency domain equalization and bit-interleaved coded modulation. Complexity reduction compared to the MMSE equalizer is achieved with little performance penalty due to the energy concentration property of ICI. A new MMSE design method for reduced tap equalization, and bit metrics for BICM are developed. Computer simulation results show that the proposed technique is robust under very fast fading channel conditions.

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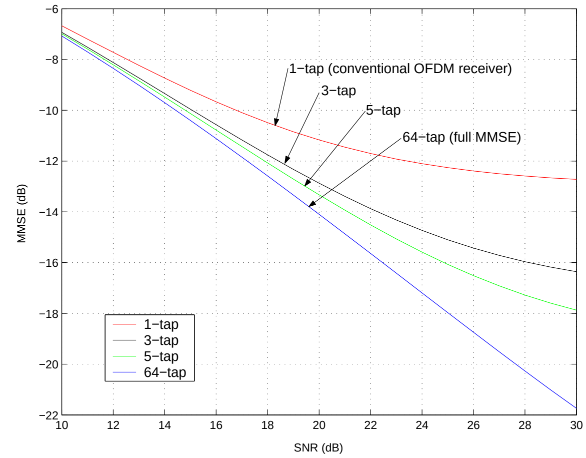

Fig. 4. Mean-Squared Error performance of proposed MMSE equalizer, $f_dT = 0.1$

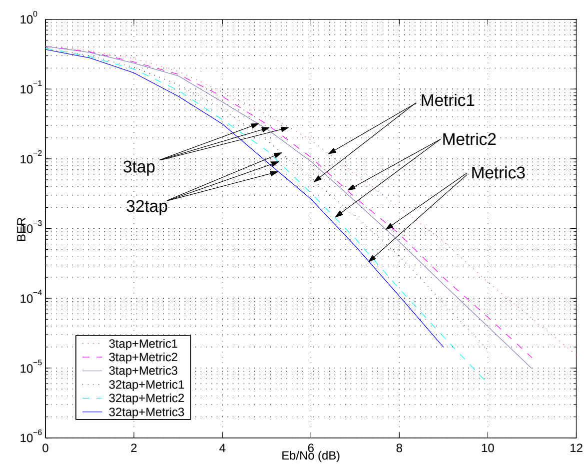

Fig. 5. BER performance for proposed metrics, $N = 32$, $f_dT = 0.4$, 4-QAM, 64 states, rate 1/2

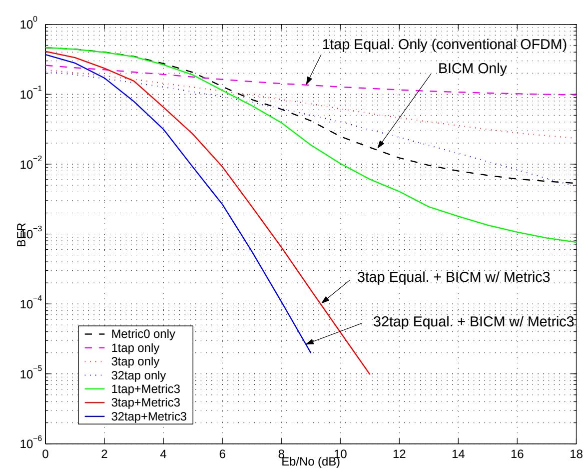

Fig. 6. BER performance comparisons under various conditions, $N = 32$, $f_dT = 0.4$, 4-QAM, 64 states, rate 1/2