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Non-Biased Linear Approximation Method of CIR for ICI Mitigation in High Mobility OFDM System

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Abstract **Original Research Article**

In orthogonal frequency division multiplexing (OFDM) system, the property of the channel impulse response (CIR) is generally assumed to be time-invariant in one OFDM symbol. However, the assumption is proved wrong in mobile environments, thus leading to inter-carrier interference (ICI) by Doppler spread. A typical solution for this problem is a piece-wise linear approximation to estimate channel time-variations in pilot-aided OFDM mobile systems. Unfortunately, the solution causes biased linear approximation error, which yields the poor performance in the receiver.

In this paper, we propose novel non-biased linear approximation of CIR in OFDM mobile systems to mitigate ICI. Theoretical analysis and simulation show that the proposal satisfies the required correlation characteristics without a series of assumptions and outperforms the previous methods in terms of channel estimation.

Keywords: orthogonal frequency division multiplexing, channel estimation, inter-carrier interference mitigation, channel impulse response, non-biased linear approximation.

1. Introduction

Orthogonal frequency division multiplexing (OFDM) has been widely applied in broadband communications such as digital video broadcasting (DVB) [1], Long Term Evolution (LTE) [2], wireless local area networks [3], and underwater acoustic systems [4].

OFDM is a multicarrier technique for achieving high-spectral efficiency and tolerance for multi-path fading, however, it suffers from a loss of subcarrier orthogonality over fading channels with very high-mobility.

Conventionally, in OFDM systems, channel impulse response (CIR) is assumed to remain static in one OFDM symbol. However, this assumption is not working well in high-mobility environments such as broadcasting channels, cellular uplink/downlink channels on high-speed railway and underwater acoustic channels, because OFDM symbol duration exceeds channel coherent time. Consequently, time-varying channels with both delay and Doppler spread would induce inter-carrier interference (ICI), which degrades system performance due to lost subcarrier orthogonality [4, 5]. Therefore, it is very important to estimate CIR in OFDM receivers.

Many methods for the channel estimation have been proposed. They can be mainly divided into two parts: non-pilot-aided and pilot-aided. The estimation performance of pilot-aided methods has outperformed better than others. In pilot-aided OFDM systems, the pilot symbols are uniformly scattered both in time and frequency domain to estimate time-variant and frequency-selective channels characteristics [6, 12]. In addition, the pilot-aided methods are, in principle, classified into channel estimations using either one OFDM symbol [7] or multiple OFDM symbols [8-10]. The former uses a basis expansion model and a pilot block structure. The latter, like piece-wise linear approximation [8] and Least Square (LS) approximation [9, 10], Minimum Mean Square Error (MMSE) equalization [13] employs estimated time-averaged CIR for several OFDM symbols in order to increase the accuracy of channel estimation.

Among those, we focus on the linear model for channel variation within one or more OFDM symbols.

In [8], Y. Mostofi and D.C. Cox proposed two methods using a piece-wise linear channel model to approximate channel time-variation. While the first method derived channel time-variation information from the cyclic prefix, the second analyzed these variations by using the next symbol. They proved that channel power difference can be negligible when the normalized Doppler frequency is 15% and that the piece-wise linear model shows a good estimation for the normalized Doppler frequency of up to 20% [8]. Although those methods can greatly mitigate ICI effect, those for channel time-variation may cause biased linear approximation errors [11, 12].

In this paper, we use non-biased linear approximation of CIR in order to get a better performance than piece-wise linear approximation and demonstrate that it provides the required correlation characteristics of CIR without some assumptions unlike in [8].

The main contribution of this work is summarized as follows:

•We estimate ICI using non-biased linear approximation which can further reduce approximation error compared to the linear approximation.

•Unlike the linear approximation, we perform linearization without any assumptions.

• Simulation results show that the proposed method satisfies the needed correlation characteristics even without some assumptions.

The rest of this paper is organized as follows. In Section 2, the previous methods are reviewed. In Section 3, a novel non-biased piece-wise linear approximate model for channel time-variation is proposed and the ICI estimation performance of the proposed model is analyzed. The simulation result is shown in Section 4, followed by conclusion in Section 5.

2. System model

In pilot-aided OFDM systems, the available bandwidth is divided into N sub-channels and the guard interval with G sampling periods is added. The duration of one OFDM symbol is $T_B = (N+G)T_S$ and the sampling period is $T_S = \frac{T_B}{N+G}$.

Let $h_l^{(i)}$ $(-G \le i \le -1,\ 0 \le i \le N-1)$ represent CIR, then the received signal y can be expressed as follows:

$$y_i = \sum_{l=0}^{L-1} h_l^{(i)} x(i-l) + w_i, \quad 0 \le i \le N-1 \quad (1)$$

where w_i are additive white Gaussian noises and $L \le G$ is the length of CIR. The Fast Fourier Transform (FFT) of sequence y_i brings the following equation.

$$Y_i = H_{i,0} X_i + \sum_{d=1}^{N_c-1} H_{i,d} X(i-d) + W_i, \quad 0 \le i \le N-1 \quad (2)$$

In Eq.(2), W_i denotes the FFT of w_i and $\sum_{d=1}^{N_c-1} H_{i,d} X(i-d)$ represents ICI.

Define $F_l(k)$ as the FFT of the l^{th} channel tap with respect to time-variation.

$$F_l(k) = \sum_{u=0}^{N-1} h_l^{(u)} e^{-\frac{j2\pi \cdot u \cdot k}{N}}, \quad (3)$$
$$0 < l < L-1 \le G-1,\ 0 \le k \le N-1$$

In Eq.(3), $F_l(k)$ is different from the channel frequency response (CFR) defined by FFT in Eq.(5) of [8]. They assumed that $h(t,\tau)$ is constant with respect to time t, and obtained CFR by using FFT of $h(t,\tau)$ with respect to τ. But in the paper, $F_l(k)$ is obtained by FFT of $h(t,\tau)$ with respect to *t*, in the case that $h(t,\tau)$ is time-variant.

Then, $H_{i,d}$ can be represented as follows.

$$H_{i,d} = \frac{1}{N} \sum_{l=0}^{L-1} F_l(d) e^{-\frac{j2\pi \cdot l(i-d)}{N}}, \quad 0 \le i,d \le N-1 \quad (4)$$

$$H_{i,0} = \sum_{l=0}^{L-1} h_l^{ave} e^{-\frac{j2\pi \cdot l \cdot i}{N}} \quad (5)$$

In above equation, $h_l^{ave} = \frac{1}{N}\sum_{u=0}^{N-1} h_l^{(u)}$ is the average of l^{th} channel tap over the time duration of $0 \le t < N \cdot T_S$, and $H_{i,0}$ represents the FFT of this average.

When the maximum Doppler shift f_d increases, ICI term on the right hand side of Eq.(2) cannot be neglected.

If $L(L \le G)$ denotes the maximum length of CIR and *L* pilots P_{l_i} are equally spaced per sub-channel, then $l_i = (i \cdot N)/L$. Hence, we can get the estimation of $H_{i,0}$ by using these pilot tones:

$$\hat{H}_{l_i,0} = \frac{Y_{l_i}}{P_{l_i}} = H_{l_i,0} + \frac{I_{l_i} + W_{l_i}}{P_{l_i}}, \quad 0 \le i \le L-1 \quad (6)$$

where I_{l_i} is the second term in Eq.(2) and the estimation of h_l^{ave} can be written as follows.

$$\hat{h}_l^{ave} = \frac{1}{L} \sum_{i=0}^{L-1} \hat{H}_{l_i,0} e^{\frac{j2\pi i l}{L}}, \quad 0 \le l \le L-1 \quad (7)$$

3. Non-biased linear model

Fig. 1 shows the non-biased linear model of time-variation channel compared to the biased piece-wise linear model in [8].

As shown in Fig. 1, piece-wise linear model in [8] adopted the linear approximation in tangential direction at the middle of the OFDM symbol, thus resulting in errors when compared to actual CIR. That is, piece-wise linear model can be called biased linear approximation model.

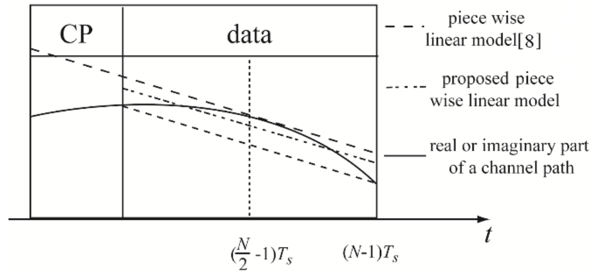

Fig.1. Non-biased piece-wise linear model of time-variation channel

However, the proposed linear approximation is modeled as a straight line parallel to the dotted line connecting two points which represent the path gain at the beginning and end of the data symbol, satisfying the following equations.

$$\int_0^{(N-1)T_s} \mathrm{Re}\{h_l(\tau)\}d\tau = \int_0^{(N-1)T_s} \mathrm{Re}\{\hat{h}_l(\tau)\}d\tau \quad (8)$$

$$\int_0^{(N-1)T_s} \mathrm{Im}\{h_l(\tau)\}d\tau = \int_0^{(N-1)T_s} \mathrm{Im}\{\hat{h}_l(\tau)\}d\tau \quad (9)$$

The geometric meaning of the above equations is that the area of the original real channel curve $h_l(\tau)$ in Fig.1 is equal to the area of the estimated channel curve $\hat{h}_l(\tau)$ in the proposed linear approximation. Thus, our proposal can be regarded as the non-biased estimation without linear approximation error during the piece-wise linear modeling process.

As the real or imaginary part of all CIR over the time duration $0 \le t < N \cdot T_S$ is unknown, the above equations can be rewritten as follows.

$$\sum_{i=0}^{N-1} \mathrm{Re}\{h_l(i)\} = \sum_{i=0}^{N-1} \mathrm{Re}\{\hat{h}_l(i)\},$$

$$\sum_{i=0}^{N-1} \mathrm{Im}\{h_l(i)\} = \sum_{i=0}^{N-1} \mathrm{Im}\{\hat{h}_l(i)\} \quad (10)$$

3.1 Correlation characteristics of CIR

In general, the channel is characterized using correlation and Doppler delay. Similar to [8], it is necessary to analyze how correlation characteristics of the proposed approximation changes with respect to Doppler delay.

As shown in Fig. 2, let both the channel impulse response *h* and its non-biased piece-wise linear approximation h_{nb} be wide-sense stationary (WSS) processes.

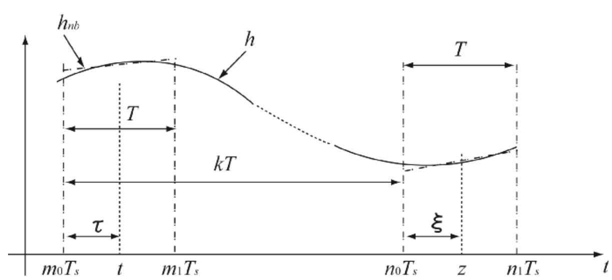

Fig.2. Non-biased piece-wise linear approximation of a random process *h*

In order to analyze the performance of the proposed approximation, we compare auto-correlation function of h_{nb} with that of *h*.

$$R_{nb}(t = mT_s, z = nT_s) = E\left[h_{nb}^{(m)} \cdot h_{nb}^{(n)^*}\right] \quad (11)$$

, where $h_{nb}^{(m)}$ represents h_{nb} at time instant mT_s.

From Fig. 2, the relationship between *h* and its approximation h_{nb} can be written as follows.

$$h_{nb}^{(m)} = h(m_0T_s) + \tau \times \frac{h_{nb}^{(m_1)} - h_{nb}^{(m_0)}}{T} + \Delta(m_0, m_1), \quad m_0 \le m \le m_1$$
$$h_{nb}^{(n)} = h(n_0T_s) + \tau \times \frac{h_{nb}^{(n_1)} - h_{nb}^{(n_0)}}{T} + \Delta(n_0, n_1), \quad n_0 \le n \le n_1 \quad (12)$$

The key difference between Eq.(12) here and Eq.(38) in [8] is that the Eq.(38) in [8] was formulated under the assumption of

$h_{lin}^{(m_0)} = h(m_0 T_s)$, $h_{lin}^{(n_0)} = h(n_0 T_s)$ (In reality, the equality can never be established, as can be seen in Fig. 1) using $h_{lin}^{(m_0)}$, $h_{lin}^{(n_0)}$ instead of $h(m_0 T_s)$, $h(n_0 T_s)$ respectively, whereas in Eq.(12) here we have formulated $h_{nb}^{(m)}$, $h_{nb}^{(n)}$ as

$h_{nb}^{(m)} = h(m_0 T_s) + \Delta(m_0, m_1), h_{nb}^{(n)} = h(n_0 T_s) + \Delta(n_0, n_1)$.

Note that the piece-wise linear approximation (Eq.(38) in [8]) can be applied to linear approximation during several OFDM symbols, but the proposed approximation of Eq. (12) can be applied to linear approximation during only one OFDM symbol. When non-biased linear approximation of Eq.(12) is applied to adjacent OFDM symbols, the approximation accuracy can be increased, but the negative effect is inevitable due to discontinuity of h_{nb} at the first and end of data symbol.

In Eq.(12), considering $h_{nb}^{(m_1)} - h_{nb}^{(m_0)} = h(m_1 T_s) - h(m_0 T_s)$, $h_{nb}^{(n_1)} - h_{nb}^{(n_0)} = h(n_1 T_s) - h(n_0 T_s)$,

the autocorrelation function of h_{nb}, $R_{nb}(t,z)$, can be represented as follows.

$$
\begin{aligned}
R_{nb}(t = mT_s, z = nT_s) &= E\left[h_{nb}^{(m)} \cdot h_{nb}^{(n)^*}\right] = \\
&E\left[\left((1-\frac{\tau}{T})h(m_0) + \frac{\tau}{T}h(m_1) + \Delta(m_0,m_1)\right)\times\right. \\
&\left.\times\left((1-\frac{\xi}{T})h(n_0) + \frac{\xi}{T}h(n_1) + \Delta(n_0,n_1)\right)\right] = \\
&= (1-\frac{\tau}{T})(1-\frac{\xi}{T})E\left[h^{(m_0)} \cdot h^{(n_0)^*}\right] + \left(\frac{\xi}{T} - \frac{\xi\tau}{T^2}\right)\times \\
&\times E\left[h^{(m_0)} \cdot h^{(n_1)^*}\right] + \left(\frac{\tau}{T} - \frac{\xi\tau}{T^2}\right)E\left[h^{(m_1)} \cdot h^{(n_0)^*}\right] + \\
&+ \frac{\xi\tau}{T^2}E\left[h^{(m_1)} \cdot h^{(n_1)^*}\right] + \\
&+ E[\Delta(m_0,m_1)]E\left[(1-\frac{\xi}{T})h(n_0)^* + \frac{\xi}{T}h(n_1)\right] + \\
&+ E[\Delta(n_0,n_1)^*]E\left[(1-\frac{\xi}{T})h(m_0) + \frac{\tau}{T}h(m_1)\right] + \\
&+ E[\Delta(m_0,m_1)\cdot\Delta(n_0,n_1)^*]
\end{aligned}
\quad (13)
$$

Let $R(t = mT_s, z = nT_s) = R((m-n)T_s) = E[h^{(m)} h^{(n)^*}]$ be auto-correlation of random process h, Eq.(13) is simplified as follows ($k = m_0 - n_0 = m_1 - n_1$).

$$
\begin{aligned}
R_{nb}(t,z) = R_{nb}(\tau,\xi,k) &= (\frac{\tau}{T} - \frac{\tau\xi}{T^2})R((k-1)T) + \\
&+ (\frac{\xi}{T} - \frac{\tau\xi}{T^2})R((k+1)T) + \\
&+ (1 - \frac{\tau}{T} - \frac{\xi}{T} + 2\frac{\tau\xi}{T^2})R(kT)
\end{aligned}
\quad (14)
$$

Since h is a wide-sense stationary (WSS) process, we can get Eq. (15).

$$R(t,z) = R(t-z) = R(\tau - \xi + kT) \quad (15)$$

Similar to the method proposed in [8], we also define P as the average power of the difference of $R(t,z)$ and $R_{nb}(t,z)$ over D OFDM symbols.

$$P = \frac{1}{D}\sum_{k=0}^{D-1}\int_0^T\int_0^T (R_{nb}(\tau,\xi,k) - R(\tau - \xi + kT))^2 d\tau d\xi \quad (16)$$

Then the normalized P_{norm} can be defined as follows.

$$P_{norm} = \frac{P}{\frac{1}{D}\sum_{k=0}^{D-1}\int_0^T\int_0^T R^2(\tau - \xi + kT)d\tau d\xi} \quad (17)$$

Because $R(t)$ can be approximated to $J_0(2\pi f_d t)$, P_{norm} becomes the function of $f_d T$, as shown in Fig. 3.

Fig. 3 represents the characteristic of P_{norm} when P_{norm} increase.

The graph suggests that for $f_d T$ of up to 20%, P_{norm} is negligible. For instance, for $f_d T$ =15%, P_{norm} is 1%.

Although Fig. 3 is obtained without any approximation and assumption, the result is the same as that in [8] which is obtained by some assumptions and approximations.

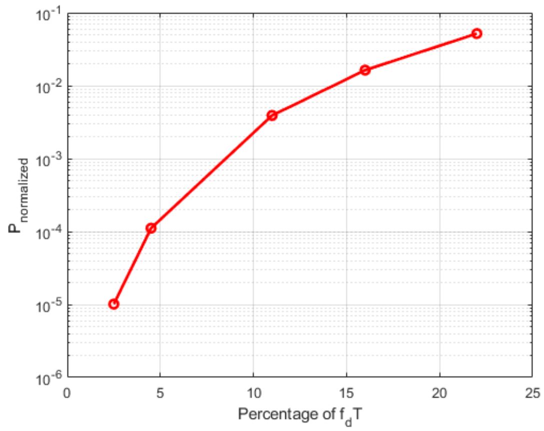

Fig.3. P_{norm} versus % of $f_d T$

3.2 Performance of Non-biased linearly ICI estimation

Like in [8], we can define ICI power of signal model P_{ICI} as follows.

$$P_{ICI} = E\left[\left|\sum_{d=0}^{N-1} H_{i,d} X(i-d)\right|^2\right] =$$

$$= \sum_{d=0}^{N-1} E[|H_{i,d}|^2] E[|X(i-d)|^2] = E[|X_i|^2] \sum_{d=0}^{N-1} E[|H_{i,d}|^2] \quad (18)$$

From Eq.(4), we can obtain the following equation.

$$E[|H_{i,d}|^2] = \frac{1}{N^2} \sum_{l=0}^{L-1} E[|F_l(d)|^2] \quad (19)$$

As $F_l(d) = \sum_{u=0}^{N-1} h_l^{(u)} e^{-\frac{j2\pi \cdot u \cdot d}{N}}$, $(0 < l < L-1 \le G - -1, 0 \le k \le N-1)$ represents the time-variation of $h(t,\tau)$, it is the same as $F_l(d)$ of $h_l^{(u)}$.

Therefore, the $F_l(d)$ for $h_l^{(u)} - h_l^{ave}$ can be can be given as follows.

$$E[|F_l(d)|^2] = \sum_{u=0}^{N_c-1} \left|h_l^{(u)} - h_l^{ave}\right|^2 \quad (20)$$

Thus, ICI power of the received signal is rewritten by substituting Eq.(19) and Eq.(20) into Eq.(18):

$$P_{ICI} = P_X \sum_{d=1}^{N-1} E[|H_{i,d}|^2] = P_X \frac{1}{N^2} \sum_{d=1}^{N-1} \sum_{l=0}^{L-1} E[|F_l(d)|^2] =$$

$$= \frac{P_X}{N^2} \sum_{d=1}^{N-1} \sum_{l=0}^{L-1} \sum_{u=0}^{N-1} \left|h_l^{(u)} - h_l^{ave}\right|^2 = \frac{P_X}{N} \sum_{u=0}^{N-1} \sum_{l=0}^{L-1} \left|h_l^{(u)} - h_l^{ave}\right|^2 \quad (21)$$

where $P_X = E\left[\left|X_i^2\right|\right]$ is the average transmitter symbol power and $h_l^{ave} = \frac{1}{N} \sum_{u=0}^{N_c-1} h_l^{(u)}$ is the average of lth channel taps.

ICI power by change of l^{th} path, $P_{ICI,l}$ can be defined as follows.

$$P_{ICI,l} = \frac{P_X}{N_c} \sum_{u=0}^{N_c-1} \left|h_l^{(u)} - h_l^{ave}\right|^2 \quad (22)$$

$$P_{ICI} = \sum_{l=0}^{L-1} P_{ICI,l} \quad (23)$$

That is, ICI power by change of CIR time-variation is determined by change of time-variation of taps in every path. The problem is how much approximation errors can occur by this ICI estimation. Actually $P_{ICI,l}$ is determined by $\left|h_l^{(u)} - h_l^{ave}\right|^2$, it is needed to estimate $h_l^{(u)}$ and h_l^{ave} to exactly know $P_{ICI,l}$.

In [8], the researchers used $h_{lin,l}^{(u)}$ and $h_{lin,l}^{\frac{N_c-1}{2}}$ instead of linearly approximated $h_l^{(u)}$ and h_l^{ave} respectively. In contrast, we use $h_{nb,l}^{(u)}$ and $h_{nb,l}^{\frac{N_c-1}{2}}$ instead of $h_l^{(u)}$ and h_l^{ave} respectively, thus proving $h_{nb,l}^{\frac{N_c-1}{2}} = h_l^{ave}$.

Of course, $E_{pilot}[h_l] = h_l^{ave}$.

We analyze the additional ICI power due to CIR estimation in piece-wise linear approximation [8] and the proposed model.

$$\begin{aligned} P_{ICI,lin,l} &= \frac{P_X}{N_c}\sum_{u=0}^{N_c-1}\left(\left|h_{lin,l}^{(u)}-h_l^u\right|^2+\left|h_l^{ave}-h_{lin,l}^{\frac{N_c}{2}-1}\right|^2\right)= \\ &=\frac{P_X}{N_c}\sum_{u=0}^{N_c-1}\left(\left|h_{nb,l}^{(u)}-h_l^u+h_{lin,l}^{(u)}-h_{nb,l}\right|^2+\left|h_l^{ave}-h_{lin,l}^{\frac{N_c}{2}-1}\right|^2\right)\geq \\ &\geq\frac{P_X}{N_c}\sum_{u=0}^{N_c-1}\left(\left|h_{nb,l}^{(u)}-h_l^u+h_{lin,l}^{idea,u}-h_{nb,l}\right|^2+\left|h_l^{ave}-h_{lin,l}^{\frac{N_c}{2}-1}\right|^2\right)= \\ &=\frac{P_X}{N_c}\sum_{u=0}^{N_c-1}\left(\left|h_{nb,l}^{(u)}-h_l^u\right|^2+2\left|h_l^{ave}-h_{lin,l}^{\frac{N_c}{2}-1}\right|^2\right)= \\ &=P_{ICI,nb,l}+2P_X\left|h_l^{ave}-h_{lin,l}^{\frac{N_c}{2}-1}\right| \end{aligned} \tag{24}$$

4. Simulation results

We simulate an OFDM system in a time-variant environment with high delay spread. Table. 1 shows the system parameters and channel properties.

Table.1. system parameters and channel properties

Parameter	Value
Modulation	8PSK
Bit rate	7.3Mbps
The number of sub-carriers (N)	892
The number of pilots (L)	223
Length of the guard interval (T_g)	44.4μs
Length of one OFDM symbol (T)	273.5μs
$T_s = \frac{T}{N+L}$	0.26μs
$G = \frac{T_g}{T_s}$	173

We simulate the power-delay profile with two main taps separated each other by 20μs. For the simulated channels, the power of channel taps is normalized to make the total power to be 1.

Fig. 4 shows the average SIR as the function of $f_{d,norm}$, $f_{d,norm}$ is divided at sub-carrier intervals. This analytic result matches the corresponding simulation as can be seen from the graph. For comparison, the received SIR for the case of ICI mitigation in [8] is plotted as well. As shown in Fig. 4, ICI mitigation using the proposed method improves the received SIR.

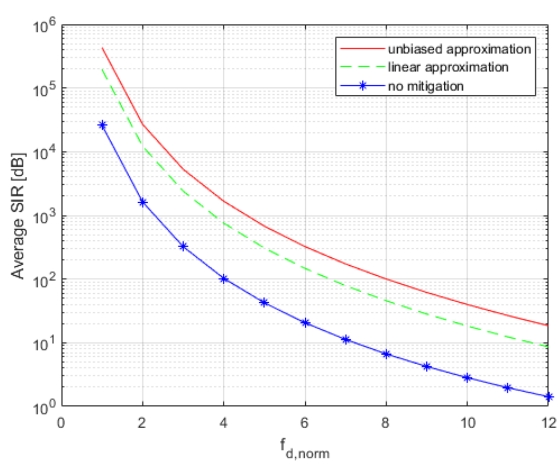

Fig.4. Average SIR versus % of $f_{d,norm}$ for a narrowband channel

Fig. 5 shows ICI power ratio with respect to CIR change. Here, ICI power is the additional ICI power caused by the linear approximation of CIR.

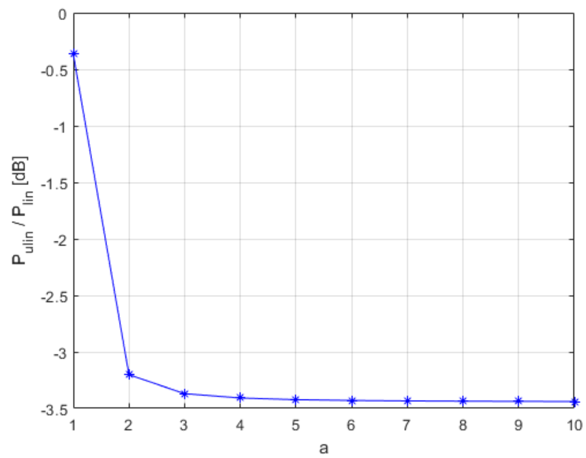

Fig.5. ICI power ratio with respect to CIR change

In Fig. 5, the horizontal axis, which is the inverse of curvature, represents the change of CIR, and the longitudinal axis represents the ratio of ICI power of the proposed method to that of the method in [8]. As shown in Figure.5, the performance of the proposed method is almost similar to that of method in [8], as the curvature gets lower, that is, CIR change is being linear. Of course, The ICI power using the proposed method is lower 0.2dB than the previous method [8], even in case that the average slope is $\tan^{-1}0.2$ during one symbol period.

Interval decreases with a decreasing SNR and optimal guard interval increases with an increasing τ_{RMS}.

5. Conclusion

This paper proposes the method how to approximate CIR non-biased linearly in OFDM and MC-CDMA system. The results show that the proposed method provides needed correlation characteristics without some assumptions, unlike in piece-wise linear approximation method and it has superior channel estimation performance to the previous method.

The extension of non-biased linear approximation of the CIR in one OFDM symbol interval to intervals between adjacent OFDM codes as in piece-wise linear approximation method may result in a discontinuity at the boundary point, thus leading to higher ICI noise. Therefore, the non-biased linear approximation method between adjacent OFDM codes will be our future research.

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