ICI Mitigation for Mobile OFDM Receivers

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Abstract

Orthogonal Frequency Division Multiplexing (OFDM) is robust against Inter-Symbol-Interference (ISI) due to the increase of the symbol duration. However, for mobile applications, channel variations during one OFDM symbol introduce Inter-Carrier-Interference (ICI), which degrades the performance. This gets more severe as mobile speed, carrier frequency or OFDM symbol duration increases. We introduce two new methods to mitigate ICI in an OFDM system with coherent channel estimation. Both methods use a piece-wise linear model to approximate channel variations. The first method extracts channel variations information from the cyclic prefix. The second method estimates these variations utilizing the next symbol. Since mathematical analysis of the whole system becomes intractable in a long delay spread environment, we provide a mathematical analysis that investigates the effect of linearization for the case of a time-variant one path channel. Then our simulation results show how these methods would improve the performance in a highly time-variant environment with a long delay spread.

INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) handles channel frequency selectivity, resulted from the delay spread, by expanding the symbol duration [1]. By adding a guard interval[1] to the beginning of each OFDM symbol, the effect of delay spread (provided that there is perfect synchronization) would appear as a multiplication in the frequency domain, for a time-invariant channel. This allows for higher data rates and has resulted in the recognition of OFDM as a standard for Digital Audio and Video Broadcasting (DAB and DVB). However, a transmission in a mobile communication environment is impaired by both delay and Doppler spread. Due to the increase of the symbol duration, time variations degrade the performance of an OFDM system more severely [2]-[4]. It will introduce Inter-Carrier-Interference (ICI) which results in the loss of orthogonality among carriers. Therefore, to improve the performance of an OFDM system in a time-variant environment, there is a need for ICI mitigation. In this paper, we propose two new methods for ICI mitigation in a pilot-aided OFDM system. For coherent channel estimation pilot tones should be transmitted. If maximum channel delay spans ν sampling periods, Negi [5] has shown that only $L = \nu + 1$ pilots are needed for perfect channel estimation in the absence of AWGN and mobility. However, in the presence of Doppler spread, these pilots are not enough to estimate the channel. Then we show how to estimate channel variations using either the cyclic prefix or the next OFDM symbol. Finally, our mathematical analysis and simulation results show the performance of the proposed methods, indicating how they can reduce the required amount of transmitted power.

SYSTEM MODEL

Fig. 1 shows the discrete baseband equivalent system model. We assume perfect synchronization in this paper. The available bandwidth is divided into N sub-channels and the length of the guard interval[2] is G. X_i represents the transmitted data in the ith sub-channel and is related to x_i as $X_i = \sum_{k=0}^{N-1} x_k e^{-\frac{j2\pi ki}{N}}$. $\vec{x}_p$ is the added cyclic prefix vector with length G and is related to x as follows:

$$\vec{x}_p(i) = x_{N-G+i} \quad 0 \le i \le G-1 \qquad (1)$$

Let T be the time duration of one OFDM symbol after adding the guard interval. Then, $h_k^{(i)}$ represents the kth channel path at time $t = i \times T_s$ where $T_s = \frac{T}{N+G}$. In our notation $h_k^{(i)}$ for $-G \le i \le -1$ and $0 \le i \le N-1$ represents kth channel path in the guard and data interval respectively[3]. Then the data part of the channel output can be expressed as follows:

$$y_i = \sum_{k=0}^{N-1} h_k^{(i)} x_{((i-k))_N} + w_i \quad 0 \le i \le N-1 \qquad (2)$$

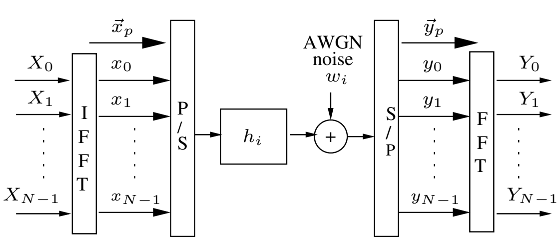

Fig. 1 Discrete baseband equivalent model

Where $x_{((i-k))_N}$ represents cyclic shift in the base of N and w_i represents a sample of additive white Gaussian noise.

*Authors would like to thank Sirius Satellite Radio for its financial support

[1]This will also prevent inter-symbol-interference

[2]We assume that the length of the channel is always less than or equal to G in this paper

[3]We assume a constant channel over the interval $i \times T_s \le t < (i+1) \times T_s$

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Then Y, the FFT of sequence y, will be as follows:

$$Y_i = H_{i,0}X_i + \underbrace{\sum_{d=1}^{N-1} H_{i,d}X_{((i-d))_N}}_{ICI} + W_i \quad 0 \le i \le N-1 \tag{3}$$

Where W denotes the FFT of w. Furthermore, the second term on the right hand side of Eq. 3 represents the ICI introduced by mobility. It can be easily shown that $H_{i,d}$ is as follows for $0 \le i, d \le N-1$:

$$H_{i,d} = \frac{1}{N}\sum_{m=0}^{N-1}\sum_{u=0}^{N-1} h_m^{(u)} e^{-\frac{j2\pi}{N}(d\times(u-m)+m\times i)} \tag{4}$$

Let $h_m^{ave} = \frac{1}{N}\sum_{u=0}^{N-1} h_m^{(u)}$ represent the time average of the mth channel path over the data part of the symbol[4]. Then $H_{i,0} = \sum_{m=0}^{N-1} h_m^{ave} e^{-\frac{j2\pi mi}{N}}$ is the FFT of h_m^{ave}. Define f_d as the maximum Doppler shift. Then the normalized Doppler shift will be $f_{d,norm.} = f_d N T_s$. As $f_{d,norm.}$ increases, the second term on the right hand side of Eq. 3 can not be neglected.

PILOT EXTRACTION

Let $\nu \le G$ be the maximum predicted normalized channel delay [5]. Then we insert $L = \nu + 1$ equally spaced pilots, P_{l_i}, at sub-channels $l_i = \frac{i\times N}{L}$ for $0 \le i \le L-1$. An estimate of $H_{i,0}$ can then be acquired at pilot tones as follows:

$$\hat{H}_{l_i,0} = \frac{Y_{l_i}}{P_{l_i}} = H_{l_i,0} + \frac{(I_{l_i} + W_{l_i})}{P_{l_i}} \quad 0 \le i \le L-1 \tag{5}$$

Where I_{l_i} denotes ICI at $l_i th$ sub-carrier. Through an IFFT of length L, the estimate of h_k^{ave} would be,

$$\hat{h}_k^{ave} = \frac{1}{L}\sum_{i=0}^{L-1} \hat{H}_{l_i,0} e^{\frac{j2\pi ik}{L}} \quad 0 \le k \le L-1 \tag{6}$$

It can be shown that, for the kth channel path, $|h_k^{ave} - h_k^{(l)}|^2$ is minimized for[6] $l = \frac{N}{2} - 1$. Therefore, we use h_k^{ave} to represent an estimate of the channel at time instant $(\frac{N}{2} - 1) \times T_s$. We will have,

$$\hat{h}_k^{(\frac{N}{2}-1)} = \hat{h}_k^{ave} \tag{7}$$

In the absence of mobility, L pilots would have been enough to estimate the channel. However, in the presence of Doppler, more information is needed to track channel variations.

PIECE-WISE LINEAR APPROXIMATION

We approximate variations of each channel path with a piece-wise linear model that has a constant slope over a time duration of length T. In this section, we will derive the frequency domain relationship, equivalent to Eq. 3, when linear model is applied. Let α_k denote the slope of the kth channel path in the current symbol[7]. Consider linearization around $h_k^{(\frac{N}{2}-1)}$. Then $h_k^{(i)}$ can be approximated as follows:

$$h_k^{(i)} \approx h_k^{(\frac{N}{2}-1)} + (i+1-\frac{N}{2}) \times \alpha_k \times T_s \quad 0 \le i \le N-1 \tag{8}$$

Inserting Eq. 8 in Eq. 2, we will have,

$$\vec{y} \approx (\mathrm{h}_{mid} + \mathrm{M} \times \mathrm{A}) \times \vec{x} + \vec{w} \tag{9}$$

Where $\vec{y}$, $\vec{x}$ and $\vec{w}$ are $N \times 1$ vectors containing samples of y_i, x_i and w_i for $0 \le i \le N-1$. Furthermore, $\mathrm{h}_{mid}(k,m) = h_{((k-m))_N}^{(\frac{N}{2}-1)}$, $\mathrm{A}(k,m) = \alpha_{((k-m))_N}$ and M is a diagonal matrix with diagonal elements of $\mathrm{M}(k,k) = \beta_{k-1} = T_s \times (k - \frac{N}{2})$ for $1 \le k, m \le N$. Let $\vec{y_1} = \mathrm{A}\vec{x}$. Then we will have $\vec{Y_1} = \mathrm{H}_{slope}\vec{X}$ Where Y_1 represents the FFT of y_1 and H_{slope} is defined as:

$$\mathrm{H}_{slope} = diag\{FFT([\ \alpha_0 \quad \alpha_1 \quad \cdots \quad \alpha_{N-1}\ ])\} \tag{10}$$

Where $FFT(\vec{z})$ represents the FFT of the vector $\vec{z}$. Let $\vec{y_2} = \mathrm{M}\vec{y_1}$. Taking an FFT of it, we will have $Y_2(k) = \frac{B_k \otimes_N Y_1(k)}{N}$ Where B is the FFT of β and $\otimes_N$ represents circular convolution in base N. It can be easily shown that B_k is as follows [6] :

$$B_k = T_s \times N \times \begin{cases} -\frac{1}{1-e^{-\frac{j2\pi k}{N}}} & k \ne 0 \\ 0.5 & k = 0 \end{cases} \tag{11}$$

Let $\vec{W}$ be the vector containing the FFT of noise samples, W_i, and $\mathrm{C}_{n,m} = \frac{\mathrm{B}_{n-m}}{N}$. Therefore, we will have,

$$\vec{Y} \approx \mathrm{H}_{mid}\vec{X} + \mathrm{C} \times \mathrm{H}_{slope}\vec{X} + \vec{W} \tag{12}$$

Where,

$$\mathrm{H}_{mid} = diag\{FFT([\ h_0^{(\frac{N}{2}-1)} \quad h_1^{(\frac{N}{2}-1)} \quad \cdots \quad h_{N-1}^{(\frac{N}{2}-1)}\ ])\} \tag{13}$$

To solve Eq. 12 for $\vec{X}$, both H_{mid} and H_{slope} should be estimated. An estimate of H_{mid} is readily available from Eq. 5-7. In the following sub-sections, we show how to estimate H_{slope} using the redundancy of the cyclic prefix or the information of the next symbol.

Method I: ICI mitigation exploiting the cyclic prefix

The output prefix vector, $\vec{y}_p$ of Fig. 1, can be written as follows:

$$\vec{y}_p = \mathrm{Q} \times \vec{p} + \vec{w}_p \tag{14}$$

[4]Note that h_m^{ave} solely refers to a time averaging over the data part of the symbol and is different from the channel ensemble average, unless $N \times T_s$ is at least 20-30 times the channel coherence time, which is considerably higher than any existing application

[5]Normalized channel delay refers to the channel delay divided by the sampling period

[6]Without loss of generality, we assume an even N in this paper

[7]Since the maximum length of the channel is G (maximum number of paths is $G+1$), both α_k and $h_k^{(i)}$ are zero for $G+1 \le k \le N-1$

Where $\vec{p} = \begin{bmatrix} \vec{x}_p^{pre} \\ \vec{x}_p \end{bmatrix}$, $\vec{w}_p$ contains AWGN samples and

$$\mathrm{Q}(i,j) = h^{(-G+i-1)}_{((i-j+G))_{2\times G}} \quad 1 \le i \le G \quad 1 \le j \le 2 \times G \tag{15}$$

$\vec{x}_p$ is a $G \times 1$ vector that includes samples of the transmitted cyclic prefix, $x_p(i)$, as defined in Eq. 1. $\vec{x}_p^{pre}$ is defined in a similar manner for the transmitted cyclic prefix of the previous OFDM symbol. Define $\vec{\zeta}$ as $\vec{\zeta} = \begin{bmatrix} \alpha_0 & \alpha_1 & \cdots & \alpha_G \end{bmatrix}^t$. Inserting $h_k^{(i)}$ from Eq. 8 in Eq. 14, we can show that Eq. 14 can be written as follows:

$$\vec{y}_p - \mathrm{R} \times \vec{p} \approx \mathrm{D} \times \mathrm{X}_{p_{mat}} \times \vec{\zeta} + \vec{w}_p \tag{16}$$

Where $\mathrm{R}(i,j) = h^{(\frac{N}{2}-1)}_{((i-j+G))_{2\times G}}$, D is a diagonal matrix with $\mathrm{D}(i,i) = T_s \times (-\frac{N}{2} + i - G)$ for $1 \le i \le G$ and $1 \le j \le 2 \times G$ and $\mathrm{X}_{p_{mat}} = \begin{bmatrix} Rev(\vec{p}^t\,[1:G+1]) \\ Rev(\vec{p}^t\,[2:G+2]) \\ \vdots \\ Rev(\vec{p}^t\,[G:2\times G]) \end{bmatrix}$.

$Rev(\vec{J}[i:j])$ represents the vector produced by reversing the order of elements i through j of vector $\vec{J}$. Eq. 12 and 16 provide enough information to solve for $\vec{X}$. It is possible to combine both equations to form a new one that is only a function of $\vec{X}$. However, the complexity of solving such an equation would be high. Therefore, we use a simpler iterative approach to solve for $\vec{X}$. We start with an initial value for $\vec{X}$ and $\vec{\zeta}$. Through each iteration we improve the estimate of $\vec{X}$ through Eq. 12 followed by improving the estimate of $\vec{\zeta}$ through Eq. 16. This procedure is summarized in Table I:

Step	Command
1	Set the initial estimate of H_{slope} to zero
2	Use Eq. 5-7 & 13 to estimate H_{mid} from pilots
3	Solve Eq. 12 for $\vec{X}$
4	Solve Eq. 16 for $\vec{\zeta}$
5	Use Eq. 10 to get an estimate of H_{slope}
6	If not converged or timed out, go to step 3

Table I. Procedure for Method I

Method II: ICI mitigation using the next OFDM symbol

For those applications that can tolerate a delay of one OFDM symbol[8], the slopes of the channel variations can be obtained through using the next OFDM symbol. This is shown in Fig. 2. In this case, pilots of the current symbol will provide an estimate of the channel at the mid-point of the current symbol, $\hat{h}_k^{(\frac{N}{2}-1)}$. Upon processing of the next symbol, an estimate of the channel at the mid-point of the next symbol, $\hat{h}_k^{(\frac{N}{2}-1),next}$, becomes available. An estimate of the slopes in region 2 (see Fig. 2) can then be obtained as follows:

$$\hat{\alpha}_k^{r_2} = \frac{\hat{h}_k^{(\frac{N}{2}-1),next} - \hat{h}_k^{(\frac{N}{2}-1)}}{T} \quad 0 \le k \le N-1 \tag{17}$$

Where $\alpha_k^{r_2}$ represents the slope of the kth channel path in region 2. Call the estimate of the slope in region 1 $\alpha_k^{r_1}$. This estimate is acquired in the same manner during the processing of the previous OFDM symbol and is stored in the system. For Method II, it is also possible to estimate the channel variations in one OFDM symbol with only one set of slopes, i.e. either $\alpha_k^{r_1}$ or $\alpha_k^{r_2}$. However, this would result in a performance degradation, since it assumes a constant slope over a longer time duration of $1.5 \times T$. By dividing each OFDM symbol into two regions, constant slope is only assumed over a time length of T.[9] Utilizing two slopes would then introduce a minor change in Eq. 12. It can be shown that in this case we will have,

$$\vec{Y}_{method_{II}} \approx \mathrm{H}_{mid}\vec{X} + (\mathrm{C}^{r_1} \times \mathrm{H}^{r_1}_{slope} + \mathrm{C}^{r_2} \times \mathrm{H}^{r_2}_{slope}) \times \vec{X} + \vec{W} \tag{18}$$

Where $\mathrm{H}^{r_i}_{slope}$ is the diagonal matrix defined in Eq. 10 for slopes of region r_i.[10]

An estimate of X can then be obtained from Eq. 18 after the reception of the next symbol. Equivalent of Eq. 9 can also be written for this case.

For both methods, it is also possible to first obtain an estimate of x from the time domain relationship, i.e. Eq. 9 followed by an FFT to get the estimate of X. Depending on the length of the channel delay and the amount of Doppler spread, solving in one of the domains will be less computationally complex.

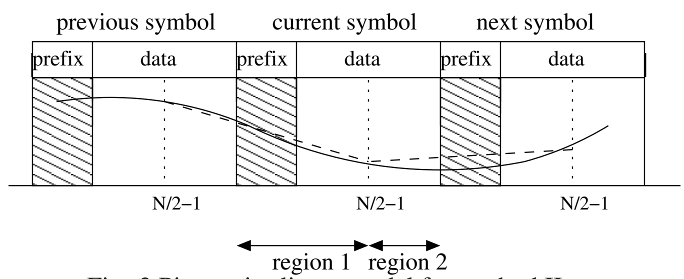


Fig. 2 Piece-wise linear model for method II

MATHEMATICAL ANALYSIS OF THE EFFECT OF LINEARIZATION

[8]DAB is an example of such applications

[9]Note that in Method I, constant slope is assumed over one OFDM symbol which has a length of T. In Method II, constant slope is assumed in between mid-points of two consecutive symbols, which has a length of T as well

[10]It can be shown that $\mathrm{C}^{r_1}(n,m) = T_s \times \begin{cases} \frac{-.5}{1-e^{\frac{-j2\pi(n-m)}{N}}} + \frac{1-(-1)^{n-m}}{N\times(1-e^{\frac{-j2\pi(n-m)}{N}})^2} & n \neq m \\ \frac{1}{4} - \frac{N}{8} & n = m \end{cases}$ and $\mathrm{C}^{r_2}(n,m) = T_s \times \begin{cases} \frac{-.5}{1-e^{\frac{-j2\pi(n-m)}{N}}} - \frac{1-(-1)^{n-m}}{N\times(1-e^{\frac{-j2\pi(n-m)}{N}})^2} & n \neq m \\ \frac{1}{4} + \frac{N}{8} & n = m \end{cases}$

In this section we provide a mathematical analysis of the effect of linearization in mitigating ICI. We assume a one path time-variant channel to make the analysis tractable and leave the case of wide-band channels to our simulations in the next section. It is of interest to calculate SIR, Signal to Interference Ratio, when ICI mitigation is applied and compare it with SIR of the no mitigation case.

In the absence of AWGN noise and for a one path time-variant channel, $h^{(i)}$, Eq. 2 can be written as:

$$y_i = h^{(i)} \times x_i \qquad (19)$$

The estimate of the time domain input will then be,

$$\hat{x}_i = \frac{y_i}{\hat{h}^{(i)}} = x_i + \frac{e_i}{\hat{h}^{(i)}} \times x_i \qquad (20)$$

Where $e_i = h_i - \hat{h}_i$. There is an implicit assumption in Eq. 20 that $\hat{h}^{(i)} \neq 0$. Since $\hat{h}^{(i)}$,as estimated through either Method I or II, has a complex Gaussian distribution, we need to exclude the possibility of a zero $\hat{h}^{(i)}$, to make the analysis meaningful. Therefore, we introduce a slight modification in the pdf of $|\hat{h}^{(i)}|$. Let $prob(|\hat{h}^{(i)}|^2 \leq Threshold) = \epsilon$. Then for a near to zero ϵ, we take the pdf of $|\hat{h}^{(i)}|$ to be zero for $|\hat{h}^{(i)}|^2 \leq Threshold$.

Through an FFT of Eq. 20, we will have,

$$\hat{X}_i = X_i + A_i \qquad (21)$$

Where $a_i = \frac{e_i}{\hat{h}^{(i)}} \times x_i$ and A_i is the FFT of it. A_i is approximately the residual interference after linearization[11].

For one realization of $\hat{h}^{(i)}$ and e_i, the instantaneous Signal to Interference Ratio (SIR) will be as follows:

$$SIR_{inst.} = \frac{\sigma_X^2}{\sigma_A^2} \qquad (22)$$

Where σ_X^2 is the power of X and σ_A^2 can be calculated as follows:

$$\sigma_A^2 = E(A_i A_i^*) = \sum_{m=0}^{N-1}\sum_{n=0}^{N-1} \overline{a_n a_m^*} e^{-\frac{j2\pi i(n-m)}{N}} \qquad (23)$$

Where $\overline{z}$ represents the average of z. Since $\overline{x_n x_m^*} = \frac{\sigma_X^2}{N}\delta_{n,m}$ then,

$$\sigma_A^2 = \frac{\sigma_X^2}{N} \times \sum_{m=0}^{N-1} |\frac{e_m}{\hat{h}^{(m)}}|^2 \qquad (24)$$

To find the average SIR, an averaging over the joint distribution of $\hat{h}^{(i)}$ and e_i should be performed:

$$\overline{SIR} = \frac{N}{\sum_{m=0}^{N-1} \overline{(\frac{|e_m|^2}{|\hat{h}^{(m)}|^2})}} \qquad (25)$$

Then $\overline{(\frac{|e_m|^2}{|\hat{h}^{(m)}|^2})}$ should be evaluated. Both e and $\hat{h}$ are sum of complex Gaussian variables. Therefore, it is reasonable to assume that they are jointly Gaussian[12]. Then $\overline{|\frac{e_m}{\hat{h}^{(m)}}|^2}$ can be calculated as follows [7]:

$$\overline{(|\frac{e_m}{\hat{h}^{(m)}}|^2)} = \frac{\rho_m^2 \sigma_{e_m}^2}{\sigma_{\hat{h}_m}^2} + \sigma_{e_m^2}(1-\rho_m^2)\overline{(\frac{1}{|\hat{h}^{(m)}|^2})} \qquad (26)$$

Also[13] $\overline{(\frac{1}{|\hat{h}^{(m)}|^2})} = \frac{Ei(-ln(1-\epsilon))}{\sigma_{\hat{h}_m}^2}$. Furthermore, $\sigma_{\hat{h}^{(m)}}^2 = \overline{|\hat{h}^{(m)}|^2}$, $\sigma_{e_m}^2 = \overline{|e_m|^2}$ and $\rho_m = \frac{\overline{e_m \hat{h}^{(m)*}}}{\sigma_{e_m}\sigma_{\hat{h}^{(m)}}}$.

Using the formula of $\hat{h}^{(m)}$ from the previous sections, these averages can be obtained. Inserting Eq. 26 in Eq. 25, the average SIR will be,

$$\overline{SIR} = \frac{N}{\sum_{m=0}^{N-1} \frac{\sigma_{e_m}^2 \times \rho_m^2 + Ei(-ln(1-\epsilon)) \times (1-\rho_m^2) \times \sigma_{e_m}^2}{\sigma_{\hat{h}_m}^2}} \qquad (27)$$

Fig. 3 shows $\overline{SIR}$, obtained from the analysis, as a function of $f_{d,norm}$ at $\epsilon = 10^{-6}$ for Method II. Also, the result of simulating the same case can be seen on the graph for comparison. A good match is observed between the analysis and simulation. For comparison, $\overline{SIR}$ for the case of no mitigation is plotted as well. This shows how mitigation through linearization increases $\overline{SIR}$.

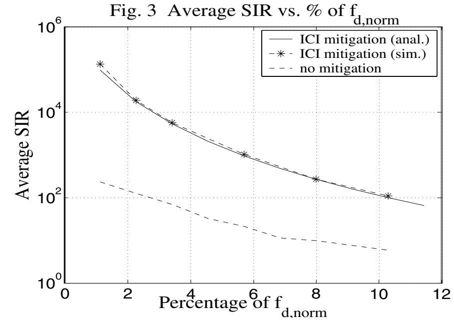


Fig. 3 Average SIR vs. % of $f_{d,norm}$

SIMULATION RESULTS

We simulate an OFDM system in a time-variant environment with high delay spread. System parameters are summarized in Table[14] II.

Parameter	Value
Input Modulation	8PSK
bit rate	7.3Mbps
# of sub-carriers (N)	892
# of pilots (L)	223
length of the guard interval (T_g)	44.4μs
length of one OFDM symbol (T)	273.5μs
$T_s = \frac{T}{N+L}$	.26μs
$G = \frac{T_g}{T_s}$	173

Table II. Parameters of the simulated system

[11] A_i is not purely interference and contains a term that depends on X_i. However, the power of that term is considerably small. Therefore, to reduce the complexity of the analysis we take A_i as the interference term which makes the analysis a tight approximation

[12] They are correlated

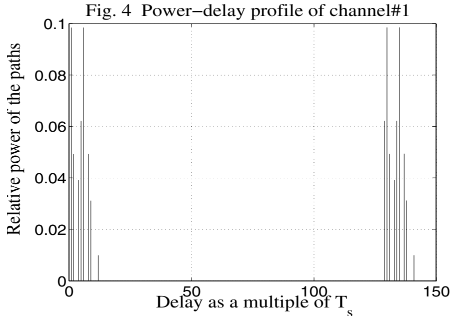
[13] Ei stands for the exponential integral

[14] Parameters are based on Sirius Satellite Radio second generation system specification proposal

We simulate two power-delay profiles. Power-delay profile of channel#1 is shown in Fig. 4 and has two main clusters with the total delay of 36.5μs to represent an SFN channel. Power-delay profile of channel#2 has two main paths that are separated by 20μs and represents a non-SFN channel. Each channel path is generated as a random process with Rayleigh distributed amplitude and uniformly distributed phase using Jakes model [8]. The auto-correlation of each path is zero-order Bessel function.

To reduce the effect of noise/interference in $\hat{h}_k^{ave}$ of Eq. 6, a noise reduction mechanism is deployed which would zero those paths with an amplitude below a pre-defined Threshold. This mechanism improves the performance in Doppler spread (or noisy) environments at the cost of a slight performance degradation in time-invariant low noise cases.

Fig. 4 Power–delay profile of channel#1



To see how ICI mitigation methods improve the error floor, Fig. 5 shows the average P_b in the absence of noise for both channels. In the "No ICI mitigation" case, pilots are used to estimate $\hat{h}_k^{ave}$ which would represent an estimate of the channel in one OFDM symbol with no compensation for time-variations. As can be seen, P_b increases considerably for "No ICI mitigation" case. Both methods reduce the error floor considerably. Method II shows a slight better performance than method I. This is due to the iterative way of solving for unknowns in method I. Also, channel#2 results in a lower error floor due to its lower amount of delay and number of paths, as expected. To see the effect of noise,

Fig. 5 Error floor vs. % of $f_{d,norm}$

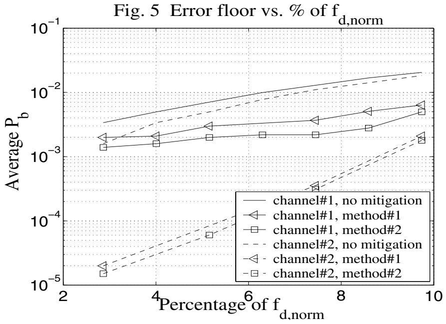


Fig. 6 shows the average probability of bit error as a function of received SNR for $f_{d,norm.} = 6.5\%$. Received SNR is defined as the ratio of the total received signal power (including the effect of channel paths) to the received noise power. Error rate for the case of no Doppler is also plotted for comparison[15]. To see how ICI mitigation methods reduce the required received SNR for achieving a specific bit error rate, Fig. 7 shows the required received SNR for reaching $P_b = .02$. The graph shows how performing ICI mitigation saves power. Required SNR for the case of no Doppler is 17.6dB for comparison. It can be seen that both methods reduce the amount of required power to a level close to that of the no Doppler case. Comparing to "No Mitigation" case, the amount of power saving increases considerably as $f_{d,norm.}$ increases.

Fig. 6 Average P_b vs. received SNR at $f_{d,norm}$=6.5%

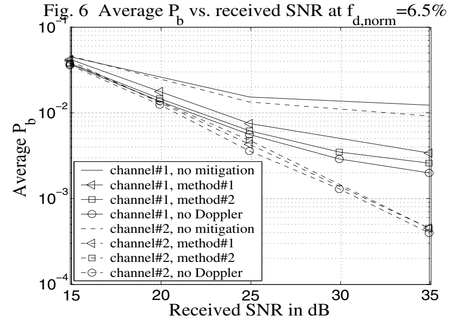
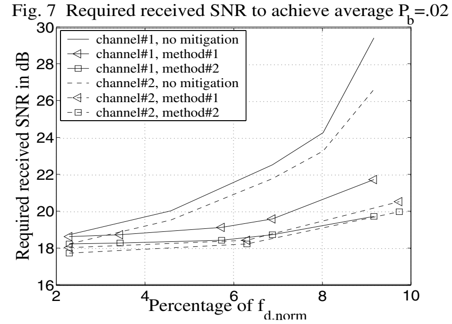


Fig. 7 Required received SNR to achieve average P_b=.02



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[8] W. Jakes, "Microwave mobile communications"

[15]Note that all the graphs present the error rate before decoding