LOW-COMPLEXITY CHANNEL ESTIMATION FOR TRANSMITTER DIVERSITY OFDM SYSTEMS

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ABSTRACT

Orthogonal Frequency Division Multiplexing (OFDM) has recently been proposed for achieving high-bit rate communication required in wireless systems. Transmitter diversity can be used with OFDM system for performance improvement in a fading environment. For these systems the receiver requires channel state information for decoding. In this paper, we propose a novel low complexity, frequency domain channel estimation algorithm for transmitter diversity OFDM systems. The performance of the proposed channel estimation is demonstrated by computer simulations.

1. INTRODUCTION

OFDM (Orthogonal Frequency Division Multiplexing) is a multi-carrier block modulation scheme which is highly efficient since it allows for spectral overlap. OFDM transforms a frequency selective fading channel into multiple narrow flat fading parallel sub-channels. This increases the symbol duration and mitigates inter-symbol interference (ISI) caused due to multipath [1]. OFDM has been incorporated in high-bit rate wireless LANs like IEEE 802.11a and HiperLAN-2. It is also being strongly considered for the emerging IEEE 802.16a and IEEE-ISTO BWIF , broadband fixed wireless access standards.

Space diversity using multiple-element antenna arrays can be used combat the detrimental effects of fading in wireless environments. Generally it is economical to employ an antenna array at the base station than at the subscriber station. Traditionally, spatial diversity has been implemented at the receiver using multiple receive antennas (Receiver Diversity) which helps in upstream transmission to the base station. Likewise, it is economical to achieve diversity in the downstream transmission from the base station using multiple transmit antenna (Transmitter Diversity) which is being investigated recently.

For transmitter diversity OFDM systems channel parameters have to be estimated at the receiver to facilitate decoding. In [2], a time domain based Minimum Mean Square Error (MMSE) channel estimation for OFDM Transmitter Diversity system was proposed. The main drawback of the MMSE channel estimation approach is its high computational complexity. In this paper, we investigate a low complexity channel estimation which takes advantage of the inherent frequency domain simplicity of an OFDM system.

2. OFDM SYSTEM

A block diagram of a conventional OFDM system is as shown in Fig. 1. In the transmitter, the serial-to-parallel converter collects blocks of K serial data symbols to be modulated by the Inverse Discrete Fourier Transform (IDFT). The serial-to-parallel converter

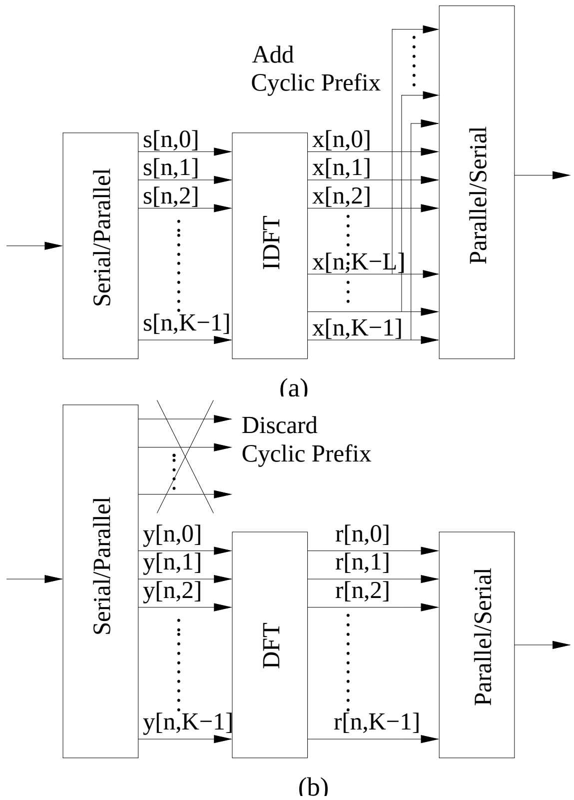

Figure 1: (a) Generic OFDM Transmitter (b) Generic OFDM Receiver

can also be viewed as a time-to-frequency mapper. K is usually a power of 2 to facilitate the use of the Fast Fourier Transforms (FFT). Denoting $s[n,k]$ as the signal modulating the k^{th} tone during the n^{th} block, the IDFT generates the required time domain OFDM symbol as,

$$x[n,l] = \frac{1}{\sqrt{K}} \sum_{k=0}^{K-1} s[n,k] \exp\left(\frac{j2\pi lk}{K}\right),$$

$$l = 0,1,..,K-1 \qquad (1)$$

The OFDM symbol duration is KT_s where T_s is the incoming data symbol duration. A cyclic prefix (CP) of length L which is the repetition of the last L samples of the IDFT is prepended for each block. This causes the overall symbol duration of the OFDM block to be PT_s, where, $P = K + L$. The CP acts like a guard interval between successive OFDM symbols and prevents intersymbol interference (ISI) if the channel impulse response length is less than or equal to the length of the CP [1]. The received time domain signal $y[n,l]$ is a linear convolution of $x[n,l]$ and the channel impulse response $h[n,l]$. The CP tranforms this linear convolution into cyclic convolution. Thus,

$$y[n,l] = x[n,l](*)h[n,l] + w_t[n,l] \qquad (2)$$

where, $(*)$ denotes cyclic convolution and $w_t[n,k]$ is Additive White Gaussian Noise.

In the receiver, the CP is first discarded and then a K-point DFT is applied. The DFT demodulates the time domain OFDM signal generating $r[n,k]$ the received signal in the k^{th} tone and n^{th} block as follows,

$$r[n,k] = \frac{1}{\sqrt{K}} \sum_{l=0}^{K-1} y[n,l] \exp\left(-\frac{j2\pi lk}{K}\right),$$

$$k = 0,1,..,K-1 \qquad (3)$$

The parallel-to-serial converter transforms blocks into serial output symbol stream. It also can be viewed as a frequency-to-time mapper. From (1), (2) and (3), we get,

$$r[n,k] = H[n,k]s[n,k] + w[n,k],$$

$$k = 0,1,..,K-1 \qquad (4)$$

where,

$$H[n,k] = \frac{1}{\sqrt{K}} \sum_{l=0}^{K-1} h[n,l] \exp\left(-\frac{j2\pi lk}{K}\right),$$

$$w[n,k] = \frac{1}{\sqrt{K}} \sum_{l=0}^{K-1} w_t[n,l] \exp\left(-\frac{j2\pi lk}{K}\right),$$

$$k = 0,1,..,K-1$$

This implies the frequency selective channel is tranformed into K parallel flat sub-channels with gains given by the DFT of $h[n,k]$. This makes frequency domain equalization simple at the receiver.

3. OFDM TRANSMITTER DIVERSITY FOR WIRELESS SYSTEMS

3.1. OFDM Transmitter Diversity System

A block diagram of a OFDM tranmsitter diversity system is as shown in Fig. 2. Extending

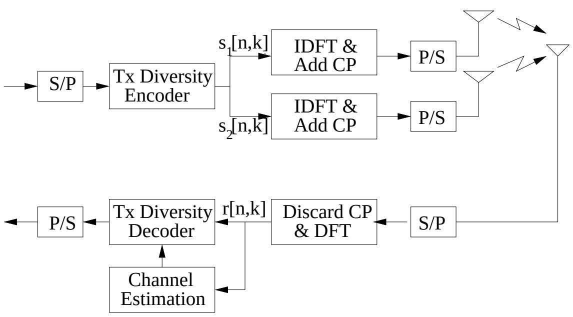

Figure 2: Transmitter Diversity OFDM System

(4) to two antenna case, the received signal at each tone is a superposition of the two transmitted signal attenuated by the frequency response of the corresponding channel at that tone, i.e.,

$$\begin{aligned} r[n,k] &= H_1[n,k]s_1[n,k] + \\ &\quad H_2[n,k]s_2[n,k] + w[n,k], \\ &\qquad k = 0,1,..,K-1 \end{aligned} \tag{5}$$

where, $H_j[n,k]$ is the channel frequency response at the k^{th} tone and n^{th} block corresponding to the j^{th} transmit antenna. A joint Maximum Likelihood (joint ML) decoding is used to separate the signals at each tone by finding the least metric from the various possible metrics given by

$$|r[n,k] - (\tilde{H}_1[n,k]s_1[n,k] + \tilde{H}_2[n,k]s_2[n,k])|^2$$

where, $\tilde{H}_1[n,k]$ and $\tilde{H}_2[n,k]$ are frequency domain channel estimates. Hence, channel state information is required for decoding in transmitter diversity OFDM systems.

3.2. Wireless Channel Model

The complex baseband channel representation of the wireless channel impulse response can be described as in [2] by,

$$h(t,\tau) = \sum_k \nu_k(t)\delta(\tau - \tau_k)$$

where, τ_k is the delay of the k^{th} path and $\nu_k(t)$ is the corresponding complex amplitude. Discretising the above model, i.e., $h[n,l] = h(nT_f, lT_s)$ and applying DFT yields

$$H[n,k] = \frac{1}{\sqrt{K}} \sum_{l=0}^{K_0-1} h[n,l] \exp\left(-\frac{j2\pi kl}{K}\right)$$

where, $H[n,k] = H(nT_f, k\Delta f)$, K is the number of tones of an OFDM block, T_f and Δf are the block duration and tone spacing of the OFDM system, respectively, and T_s is the sample interval of the system that relates to Δf by $Ts = \frac{1}{K\Delta f}$. In the above expression, K_0 is the channel delay spread in samples or channel impulse response length which is usually much less than K ($K_0 << K$). For simulations, first order Rayleigh fading channel model with fade variable ρ is generated by summing up real and imaginary components, obtained individually by passed white Gaussian random variable through first order AR process $\sqrt{(1-\rho^2)}/(1-\rho z^{-1})$ for each $l = 0,1,..,K_0-1$.

4. CHANNEL ESTIMATION

4.1. Training

In [2], a time domain based MMSE approach was proposed. The main disadvantage is its high computational complexity. In the proposed low complexity approach, pilot symbols are not transmitted on all tones through each antenna as in [2]. Instead, in a two transmit antenna system, pilot symbols are transmitted through one antenna for half of the tones and through the other antenna for remaining half of the tones. Thus at any tone there is contribution from only one antenna avoiding any form of interference from the other antenna. The authors recently realized that a similar approach has been proposed independently in [3]. However, the issue of channel tracking in the presence of fading and residual frequency offset was not addressed in [3].

For example, in the even tones pilot symbols $t_1[n, k]$ are transmitted from first antenna, which results in having only one term in (5) in these tones during training, i.e.,

$$r[n, k] = H_1[n, k]t_1[n, k] + w[n, k]$$
$$k = 0, 2, .., K - 2. \qquad (6)$$

Hence, in the proposed frequency domain approach, channel estimates for the even tones corresponding to the first transmit antenna are easily obtained by simply dividing the received signal by the known pilot signal. Channel estimates for odd tones can be obtained by interpolation in the frequency domain since the channel is time-limited; i.e., it is non-zero only for a small portion in an OFDM block ($K_0 << K$). However, the disadvantage of the interpolation in the frequency domain is the order of interpolating filter should be high to obtain accurate estimates. Instead, upsampling by a factor 2 in the frequency domain would result in two replicas of the required impulse response in the time domain, which can then be windowed to get the required copy. This requires an additional IDFT and DFT, but is much better in performance.

4.2. Tracking

In the presence of fading the channel changes substantially between the training symbols. Also uncompensated frequency offset leads to phase errors in the channel estimates which has to be corrected. Hence, we require channel tracking schemes to improve channel estimation between training symbols. We consider the use of Common Phase Error (CPE) tones to identify the phase changes of the channel estimates in the frequency domain. A frame comprises of N OFDM symbols with the first used for training and the rest used for data transmission. During data transmission, CPE tones are inserted in a few known frequency bins to facilitate tracking. Also these are made non-overlapping between antennas, i.e., different CPE tones are assigned for different antennas. This prevents interference and hence, frequency domain channel estimates can be easily obtained by simply dividing the received signal by the known CPE tone signal. This estimate is then used to find out the phase error between the previous estimates and the current channel parameters. The phase errors obtained from all the CPE tones are then averaged and used for correcting the channel estimates for rest of the tones. Also we replace the estimates in the CPE tones with the new estimates and repeat the frequency domain estimation to track in fading environments. We propose to use progressively increasing number of these tones in a frame with few CPE tones near the training symbol and more CPE tones further away from the training symbol. This scheme helps to track well in a fading environment since channel estimates are more degraded away from the training symbol.

5. SIMULATION AND RESULTS

An OFDM Transmitter Diversity System was simulated for a two transmit antenna and one receive antenna in a first order Rayleigh fading wireless channel model. QPSK signals are assumed as input to the system shown in Fig. 2. The parameters of the OFDM system are identical to those in [2]. A wireless channel of bandwidth 800 kHz is divided into $K = 128$ tones or subchannels yielding a tone spacing $\Delta f = 6.25$ kHz. To make the tones orthogonal, the OFDM symbol duration is $\frac{1}{\Delta f} = 160\mu s$. An additional $40\mu s$ is used as guard interval i.e. CP of $L = 32$. A channel delay spread of $20\mu s$ is assumed yielding $K_0 = 17$. A frame is composed of $N = 5$

OFDM symbols with the first being the training symbol. In the proposed scheme, for each antenna, the first symbol following the training symbol has 2 CPE tones, the second symbol has 4 tones, the third symbol has 8 tones and the fourth symbol has 16 tones equally spaced. For comparison, an equal CPE tone scheme with 8 tones in all symbols was also simulated [1].

The channel estimation was done using the proposed frequency domain approach and compared with the time domain MMSE approach in [2]. The raw BER obtained using a joint ML decoding was found to be identical in both the approaches as shown in Fig. 3. However, the floating point operations required in the time domain approach during each training symbol is 3 times that of the frequency domain approach for $K_0 = 17$. The BER for the proposed tracking scheme is 2.0 dB better than tracking with equal CPE tones for $\rho = 0.995$ at $BER = 8\text{x}10^{-2}$ or for $\rho = 0.999$ at $BER = 3\text{x}10^{-2}$ as shown in Fig. 4. Note that if error control coding is employed across the OFDM tones, all the curves in Fig. 4 will shift significantly to lower SNR values. However, we expect the relative difference between the various curves in Fig. 4 to remain unaltered.

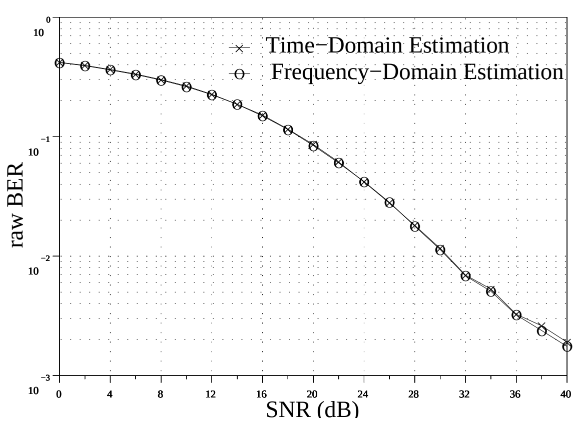

Figure 3: Performance with fade variable $\rho = 1$ i.e. no fading

[1]For detailed Time-Frequency Grid Diagram refer www.tenet.res.in website

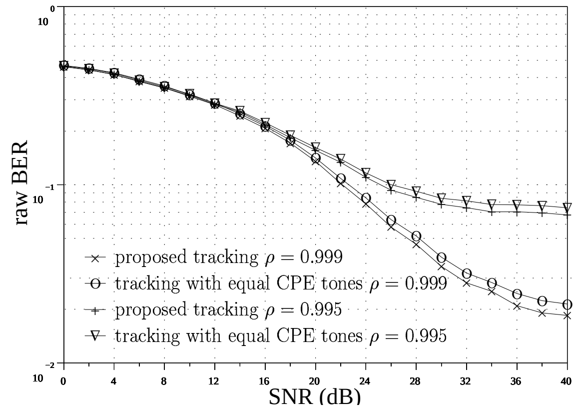

Figure 4: Performance with fade variable $\rho = 0.995$ and $\rho = 0.999$

6. CONCLUSIONS

In this paper we have presented a novel low complexity, frequency domain channel estimation algorithm for OFDM transmitter diversity systems. The performance of the proposed scheme is comparable to existing techniques while its computational complexity is significantly lower. Hence, it is a suitable alternative for channel estimation in transmitter diversity OFDM systems.

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