A New Channel Estimator Design for Next Generation High-Speed Mobile Data Communications based on OFDM

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Abstract: In this paper, we present a new channel estimator design of a receiver for the next generation high-speed mobile data communications based on OFDM. First an overview of the OFDM system suitable for mobile communications is described. The effects of non-ideal transmission conditions of the OFDM system including channel estimation errors, symbol timing offset, carrier and sampling clock offset, phase noise and time selective fading are analyzed. We then propose a new channel estimator design of receiver in which all of these issues relevant to the mobile transmission environment are addressed. Novel techniques for symbol timing and frequency synchronization are proposed. The architecture of the new subband channel estimator and resulting implementation complexity are then analyzed. The overall performance of the proposed receiver is simulated and evaluated in various channel conditions.

Keywords: Channel estimator, OFDM receiver, Subband adaptive filtering.

1 Introduction

Orthogonal Frequency Division Multiplexing (OFDM) is well known as a highly spectral efficient transmission scheme capable of dealing with severe channel impairments encountered in mobile environment. OFDM has been chosen as the air interface standards for digital audio broadcasting (DAB) [2] and digital terrestrial television broadcasting (DVB-T) [3] in Europe which support digital broadcasting at megabits data rates. It is also the physical layer modulation scheme chosen by the IEEE 802.11a standard to allow wireless LANs operating at bit rates up to 54 Mb/s at 5 GHz. OFDM also found applications in broadband fixed wireless access. OFDM can largely eliminate the effects of inter-symbol interference for high-speed transmission in highly dispersive channels with a relatively low implementation cost by separating a single high speed bit stream into a multiplicity of much lower speed bit streams each modulating a different sub-carrier. Recently there is a growing interest in applying OFDM to provide high-speed data access in mobile cellular systems [4]. However OFDM is known to be vulnerable to synchronization errors due to the narrow spacing between sub-carriers. Additionally in order to achieve high spectrum efficiency coherent demodulation with multi-level modulations such as QAM is employed which mandates accurate channel estimation and tracking in the receiver.

There exist many publications dealing with the synchronization and channel estimation issues of OFDM receiver. However, a majority of the work addressed the fore mentioned problems for different applications; with even fewer works lend them readily to integrated circuit implementations suitable for mobile handheld devices.

2 OFDM System Model

The objective of the OFDM receiver is to synchronize and demodulate the signal properly and then hand the soft and hard decisions to the outer receiver for further processing such as diversity combining, deinterleaving and channel decoding.

2.1 OFDM Signal Model

An OFDM system forms its symbol by taking $K$ complex QAM symbols $X_{i,k}$ each modulating a
sub-carrier with frequency \( f_k = k/T_u \) where \( T_u \) is the sub-carrier symbol period. The modulation is accomplished by means of a \( N \) -point \((N \geq K)\) IFFT with a sampling period of \( T = T_u/N \). A pulse-shaping filter \( g(t) \) is used to further limit the transmission spectrum. To avoid intersymbol interference (ISI) caused by the channel multipaths, a cyclic prefix of length \( T_g = N_gT \) is pre-appended to the OFDM symbol thus the transmitted complex baseband signal is described by

\[
x(t) = \sum_{l=-\infty}^{\infty} \sum_{k=-\infty}^{\infty} X_{l,k} e^{j2\pi f_k (t - T_0 - lT_g)} g(t - lT_0)
\]

where \( T_0 = T_g + T_u \) is the resultant OFDM symbol period corresponding to \( N_0 = N_g + N \) samples.

The signal is transmitted over a frequency selective multipath fading channel and sampled at the receiver to yield

\[
y(nT) = \sum_{m=0}^{M-1} h_m(nT)x(nT - \tau_m) + n(nT)
\]

where \( h_m(t) \) is the equivalent low pass impulse response for the \( m \)-th multipath component. It is assumed that the channel impulse response is quasi static during one symbol period.

The cyclic prefix is chosen to be longer than the maximum channel multipath delay so that the current symbol is not contaminated by previous symbols. Assuming perfect receiver synchronization, stripping away the cyclic prefix and applying an \( N \)-point FFT yields the demodulated sub-carrier Symbols

\[
Y_{l,k} = X_{l,k} H_{l,k} + \tilde{n}_{l,k}
\]

where \( H_{l,k} = \sum_{m=0}^{M-1} h_{1,m} e^{-j2\pi f_k \tau_m} \), is the channel frequency response at sub-carrier frequency \( f_k \) during the \( l \)-th symbol. Assuming

\[
E\{X_{l,k}\} = E\{H_{l,k}\} = 1
\]

and the sampled AWGN \( \tilde{n}_{l,k} \), has variance of \( \sigma_n^2 \), the signal to noise ratio (SNR) per sub-carrier symbol is then

\[
SNR = 1/\sigma_n^2.
\]

2.2 Wideband OFDM Testbed

To demonstrate the feasibility of a flexible, low-cost and robust OFDM-based system that could be employed in future mobile systems, a wideband OFDM testbed is currently being developed. Figure 1 illustrates the detailed frame structure used in the testbed. Five OFDM symbols transmitted in 1 ms comprise one signaling unit or burst. A long burst guard interval follows the last symbol which eliminates any ISI between adjacent bursts.

Dedicated pilot symbols are embedded in the 2nd and 4th OFDM symbols as shown in Figure 2 where \( D_i \) and \( D_f \) indicate the distance between pilots in the time and frequency direction respectively. The pilot symbols are BPSK modulated and derived from a maximum length binary sequence (MLBS) with the polynomial \( G(D) = D_{10} + D_1 + 1 \). TABLE 1 summarizes the basic transmission parameters of the OFDM testbed system in our study. The 1024 IFFT samples represent the transform of maximum 640 QPSK modulated sub-carriers. The sub-carrier spacing is 6.35kHz and the total bandwidth is thus 4 MHz. The OFDM symbol length is 188.3\( \mu \)s after appending a cyclic extension of 30.8\( \mu \)s. The maximum raw channel data rate is 2.56 MBaud. With QPSK modulation and rate 1/2 channel coding plus framing and control overhead, this can easily attain a peak end user capacity of 2 Mb/s.
### TABLE 1: OFDM System Parameters

<table>
<thead>
<tr>
<th>$N$</th>
<th>$N_g$</th>
<th>$1/T$ (MHz)</th>
<th>$T_u$ (µs)</th>
<th>$1/T$ (kHz)</th>
<th>$K$</th>
<th>Modulation</th>
</tr>
</thead>
<tbody>
<tr>
<td>1024</td>
<td>200</td>
<td>6.5</td>
<td>157.5</td>
<td>6.35</td>
<td>640</td>
<td>QPSK</td>
</tr>
</tbody>
</table>

### 3 Effects of Non-Ideal Transmission

A variety of impairments exist in an OFDM transmission system that could cause performance degradation. Among them, symbol timing offset, carrier frequency and sampling clock frequency offset, phase noise and time-variant channels are the ones that need special attention, as the receiver performance heavily depends on them.

#### 3.1 Symbol Timing Offset

The timing synchronization unit of the receiver performs the task of identifying the correct symbol start and removal of the cyclic prefix/suffix. As illustrated in Figure 3, as long as the FFT window start position falls within the allowed region no ISI occurs. Under this condition, the presence of a positive symbol timing offset $\Delta t = \Delta nT$ causes the FFT output to become

$$Y_{l,k} = X_{l,k}H_{l,k} e^{-j\pi \frac{k\Delta n}{N}} + \tilde{n}_{l,k}$$

(4)

It is evident that the presence of timing offset causes a phase rotation proportional to the sub-carrier index $k$ and the timing offset $\Delta n$. However, if the timing offset causes the FFT window start position to fall outside the allowed region, both ISI and inter-carrier interference (ICI) occurs due to the disturbance of adjacent symbols. The post-FFT signal is then given by [5]

$$Y_{l,k} = e^{-j\pi \frac{k\Delta n}{N}} \alpha(\Delta t) X_{l,k}H_{l,k} + \tilde{n}_{l,k} + n_{sf,l,k}$$

(5)

where the ISI and ICI are modeled as additional noise $n_{sf,l,k}$ while the attenuation is $\alpha(\Delta t)$ close to 1 for large $N$.

#### 3.2 Carrier and Sampling Clock Frequency Offset

The effect of carrier frequency offset in OFDM is analyzed in detail in [5] [6]. Assume in addition to a frequency offset of $\Delta f = \Delta f_u$, $y(t)$ is sampled at an interval of $\tilde{T} = (1 + \delta)T$ where $\delta$ is the sampling clock frequency offset. The FFT output becomes (6)

$$Y_{l,k} = X_{l,k}H_{l,k} \frac{\sin \pi \varepsilon}{N \sin(\pi \varepsilon / N)} e^{-j\pi \frac{N-1}{N} + \tilde{n}_{l,k}} + n_{sf,l,k}$$

where $\varepsilon = (1+\delta) + k\delta$ (and the ICI noise term is given by (7)

$$n_{sf,l,k} = \sum_{m=-K/2+1}^{K/2-1} X_{l,m}H_{l,m} \frac{\sin \pi (m-k+\varepsilon)}{N \sin(\pi (m-k+\varepsilon) / N)} e^{-j\pi (m-k+\varepsilon)/N}$$

#### 3.3 Effects of Phase Noise

OFDM has been shown to be sensitive to the phase noise in the oscillators. The effects of oscillator phase noise are two folded as shown in [7]. The first consequence due to phase noise is the loss of orthogonality between sub-carriers. The resultant ICI noise power on each sub-carrier is upperbounded by [7] (8)

$$\sigma^2_{pn} \leq \sum_{k=-K/2}^{K/2-1} \int_{-\infty}^{\infty} L_{qn}(f - [f_k - f_0]) \text{Sinc}^2(\pi f T_u) df$$

Thus the ICI noise power depends on the shape of the phase noise spectrum $L_{qn}(f)$ and increases with the number of sub-carriers $K$. The second
effect due to phase noise is the so-called common phase error (CPE) [7]. In addition to the ICI, a common phase drift $\Phi$ that is only a function of the OFDM symbol index $l$ also modulates each sub-carrier. Usually a near-in phase noise level less than $-60$ dB/Hz is required such that perturbation due to CPE is minimal.

3.4 Effects of Time Varying Channel

One major perturbation to the channel estimation is the loss of orthogonality among sub-carriers due to time-variant channels. The assumption that $h(t,\tau)$ is constant during one OFDM symbol is no longer true in fast fading environments.

Several publications have analyzed this effect [5][8]. The received sub-carrier symbols in this case should be revised as

$$Y_{i,k} = X_{i,k}H_{i,k} + \tilde{n}_{i,k} + n_{fad,i,k} \tag{9}$$

where the additional ICI noise $n_{fad,i,k}$ is caused by the time-variant channel and $H_{i,k}$ is the average channel gain factor.

The power of the ICI noise caused by a band-limited fading channel with the Jakes Doppler spectrum is lower bounded in [5] by

$$\sigma_{fad,i}^2 \approx \frac{\pi^2}{6} (f_D T_u)^2 \tag{10}$$

where $f_D$ is the maximum Doppler frequency. From the equation it is calculated that a maximum normalized Doppler frequency of 0.02 (which corresponds to $f_D = 130$ Hz) can be tolerated without significant performance loss at high SNR values.

4 Subband Channel Estimator Structure

In this section, we outline the OFDM mobile receiver channel estimation strategies based on the above analysis and present a new subband estimator structure. We assume all the channel multipaths are Rayleigh distributed according to the Jakes spectrum with maximum Doppler frequency $f_D$. The sampling theorem requires that the sampling rate in the time and frequency direction must satisfy

$$1/(D T_0) > 2(f_D + \Delta f) \tag{11}$$
$$T_u / D f > 2 \tau_{max} \tag{12}$$

where $\tau_{max}$ is the maximum channel delay spread and $D f$ is the frequency offset.

A 1-D FFT/IFFT based estimator is proposed in [11] while in [12] Li proposes a robust channel estimator using 2-D FFT and IFFT.

4.1 Structure of Subband Adaptive Filtering

An equivalent structure of the system identification model is given in Figure 4 where $H_0(z), H_1(z), F_0(z)$ and $F_1(z)$ are the analysis and synthesis filters of a perfect reconstruction filterbank. Here, the output signals from the filters $H_0(z)$ and $H_1(z)$ are divided into subbands, decimated, subtracted, and combined through an appropriate filter bank to form the error signal $e(n)$.

Where $\hat{S}(z)$ denote the channel estimation model that must be adapted. Let’s consider it can be decomposed into polyphase components as

$$\hat{S}(z) = \hat{S}_0(z^2) + z^{-1} \hat{S}_1(z^2) \tag{13}$$

Figure 4: Equivalent structure of channel estimator

Using this decomposition and Noble identities [14], the configuration of Figure 4 can be transformed into Figure 5.

When the filterbanks is lossless, the mean square error minimization of the individual residuals is equivalent to minimization of the overall residual after synthesis bank [15].

Decision algorithm decides about received signal, this section consist of equalizer, echo canceller, matched filters and any others parts for decision. Several algorithms are proposed recently for this part.
4.2 Adaptive algorithm

The filters $\hat{S}_0(z)$ and $\hat{S}_1(z)$ are to be adapted. Note that $x_{00}(n), x_{01}(n), x_{10}(n)$ and $x_{11}(n)$ are the subband components of the input $y(n)$. We use $e_0(n)$ and $e_1(n)$ to adapt the coefficients of these filters. New cost function defined by [13] as

$$J(n) = E(\alpha_0 e_0^2(n) + \alpha_1 e_1^2(n))$$  \hspace{1cm} (14)

where $\alpha_0$ and $\alpha_1$ are proportional to the inverse of the powers of $b_0(n)$ and $b_1(n)$, $E(.)$ respectively, and denotes the expectation operator. This cost function gives higher weight to the error corresponding to the subband of lower signal power. As showed in [13], this cost function brings down the eigenvalue spread of the weighted sum of the correlation matrices of the input signals to the adaptive filter, thereby resulting in improved rate of convergence.

$$\hat{s}_{ik}(n+1) = \hat{s}_{ik}(n) + 2\mu_i \alpha_i e_i(n) x_{0i}(n-k)$$  \hspace{1cm} (15)

$$\hat{s}_{ik}(n+1) = \hat{s}_{ik}(n) + 2\mu_i \alpha_i e_i(n) x_{1i}(n-k)$$  \hspace{1cm} (16)

where $\alpha_0$ and $\alpha_1$ are adapted in each iteration.

$$p_0(n+1) = (1-\beta_0)p_0(n) + \beta_0 b_0^2(n)$$  \hspace{1cm} (17)

$$p_1(n+1) = (1-\beta_1)p_1(n) + \beta_1 b_1^2(n)$$  \hspace{1cm} (18)

$$\alpha_0(n+1) = \frac{1}{p_0(n+1)}$$  \hspace{1cm} (19)

$$\alpha_1(n+1) = \frac{1}{p_1(n+1)}$$  \hspace{1cm} (20)

where $\beta_0$ and $\beta_1$ are constants.

4.3 Extension to $M$ -band case

Here, we decompose the adaptive filter $\hat{S}(z)$ into $M$ polyphase components

$$\hat{S}(z) = \sum_{i=0}^{M-1} z^{-i} \hat{S}_i(z^M)$$  \hspace{1cm} (21)

In each subband there are $M$ filters, each of length $L/M$ , at the output of each analysis filter. The cost function in this case is the extension of (14)

$$J(n) = E[\alpha_0 e_0^2(n) + \alpha_1 e_1^2(n) + \ldots + \alpha_M e_M^2(n)]$$  \hspace{1cm} (22)

where $\alpha_0, \alpha_1, \ldots, \alpha_M$ constants are inversely proportional to the powers of $b_0, b_1, \ldots, b_M$, respectively.

This is, again, a weighted cost function, with more weight given to the error corresponding to the subband of lower signal power. Following the steps as in the two-band case, we can obtain the adaptation equations for the filter coefficients as

$$\hat{s}_{ik}(n+1) = \hat{s}_{ik}(n) + 2\mu_i \alpha_i e_i(n) x_{0i}(n-k)$$  \hspace{1cm} (23)

$$\hat{s}_{ik}(n+1) = \hat{s}_{ik}(n) + 2\mu_i \alpha_i e_i(n) x_{1i}(n-k)$$  \hspace{1cm} (24)

$$k = 0,1,\ldots,(M-1), i = 0,1,\ldots,(L/M-1)$$

Receiver hardware is optimum when $M$ is equal to number of carriers.

5 Performance Analysis of Proposed Subband Estimator

The complete OFDM estimator has been simulated with proposed subband channel estimator and correction algorithm. The performance achieved with full band and IFFT/FFT estimation and synchronization is also plotted as a comparison.

We used cosine modulated paraunitary filter banks. In particular, we used filters (analysis and synthesis) with lengths 20.

The normalized coefficient error vector norm and mean square error (in decibels) at time $n$, which is defined as $10\log_{10} \frac{\nu^T(n)\nu(n)}{s^T s}$ and

$10\log_{10} e^2(n)$ where $\nu(n)=[v_0(n), v_1(n)\ldots v_{M-1}(n)]$, $v_k(n)=s_k(n)-\hat{s}_k(n)$, and $s=[s_0, s_1\ldots s_{M-1}]$ is used to depict the convergence performance.
6 Conclusions

In this paper, a new channel estimator for OFDM receiver architecture suitable for high-speed mobile wireless communications has been proposed. The proposed architecture exhibits robust performance in various mobile channels with high Doppler frequency. Such a receiver will be incorporated on a testbed and used in field trial to demonstrate the feasibility of high-speed mobile data access based on OFDM for the next generation cellular system.

References