A Blind Closed Loop Subband Channel Equalizer

Nordberg, Jörgen[1]
Dam, Hai Huyen[2]
Nordholm, Sven[2]

26th February 2004

[1]Blekinge Institute of Technology
372 25 Ronneby, Sweden.
Jorgen.Nordberg@bth.se

[2]Western Australia Telecommunication Research Institute
35 Stirling Hwy, Crawley WA 8009, Australia.

Abstract

The wireless communication channel are characterized by its multipath richness, specially in the unlicensed 60 GHz band where many of the modern and future wireless communication system will work. To reverse the effects of even a moderate delay spread, in the linear equalization case, will require an estimation of high order filters. Thus, there will be a demand for low complexity high quality receivers that should be able to work during sever multipath conditions. In this paper, a new blind closed loop subband equalizer is proposed to reduce the computational complexity of the equalizer and also to improve the channel tracking capability. Simulation results show a significantly improvement, 12 times, of the convergence rate without any significant performance loss.

Keywords: *Channel Equalization, Wireless Communication, Multi-rate Signal Processing.*

1 Introduction

It is predicted that Wireless Local Area Networks (WLAN's) will be a key element in the next generation wireless communication systems, i.e. 4G systems. They will used as an interface to Internet, i.e. connect a wide range of terminals to Internet Access (IA) points. The target date rates in 4G are 100 Mbit/s for the pedestrian traffic case and 1 Gbit/s for the stationary traffic case. Furthermore, will probably the unlicensed frequency band around 60 GHz be utilized in order to get access to the necessary system bandwidth to

1

support the 4G target data rates. WLAN supports high speed data traffic over rather short distances. Since, the carrier frequency lies in the GHz band, the transmitted wave will not easily penetrate obstacles and thus the receiver will rarely have any Line-Of-Sight (LOS) signal. Consequently, the received signal will mainly consist of reflected signals. This fact in combination with the high data rate, results in severe frequency selective fading. To recover the transmitted signal from the effects of fading, an equalization scheme must be employed at the receiver. The equalization is usually made adaptive to be able to track channel variations. Equalizers are usually using a pre-known data sequence (pilot/training sequence) in each data block to estimate the equalizer filter weights, but there is also a variety of blind equalizers. Blind equalizers use properties in the data format to recover the transmitted symbol. Traditionally implemented blind channel equalizers are characterized by slow convergence. However, recently subband techniques have been considered as means to reduce both the computational complexity and improve the convergence speed of the channel equalizer [1].

In this paper, a closed loop subband Constant Modulus Algorithm (CMA) equalizer will be presented and compared with its fullband counterpart. The CMA algorithm is one of the most well known blind channel equalization algorithms [2, 3]. The subband equalization scheme consists of two parallel main parts: one fullband equalization filter part that continuously equalize the received signal and one subband part where the fullband equalization filter weights are estimated [4] using the CMA algorithm. The estimated equalization coefficients in each subband are transformed into a fullband equivalent filter[4, 5, 6]. The outline of this paper is as follows; Section 2 introduced the used system model , Section 3 presents the proposed blind subband equalizer, Section 4 illustrates the good properties of the proposed scheme through Monto-Carlo simulations and Section 5 presents some final conclusions and future work.

2 System Model

Consider the communication system given in block diagram form in Figure 1. The transmitted bit $b(n)$, where n denotes the discrete time index, is independently generated with equal probabilities and modulated using Quadrature Phase Shift Keying (QPSK). In QPSK are the following four different signal alternatives used

$$s_i(t) = \begin{cases} \sqrt{\frac{2E}{T}} \cos\left[2\pi f_c(2i-1)\frac{\pi}{4}\right] & 0 \le t \le T \\ \\ 0 \end{cases}, \qquad (1)$$

2

where $i = 1, 2, 3, 4$, E is the transmitted signal energy per symbol, T is the symbol duration and f_c is the carrier frequency. The transmitted bits are mapped to the four signal alternatives by Gray coding. The modulated signal $s(t)$ is then transmitted through a channel modelled as a time varying Finite Impulse Response (FIR) filter with impulse response $\mathbf{c}(t) = [c_0(t), \cdots, c_{K-1}(t)]^T$, where K and $[\cdot]^T$ denote the length of the channel and the vector transpose, respectively.
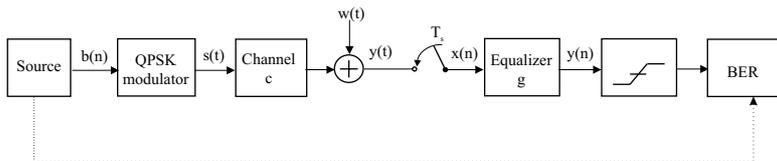

Figure 1: Communication system model.

The received continuous time signal $x(t)$ is given by

$$x(t) = \sum_{n=0}^{K} c_k(t)s(t - \tau_k) + w(t), \qquad (2)$$

where τ_k denotes the delay associated with the kth path and $w(t)$ is additive white Gaussian noise. In this study a HIgh PErformance Radio Local Area Network II (HIPERLAN II) [7] inspired channel model is considered since HIPERLAN II is working in the GHz range. This channel model is a quasi-static sparse frequency selective channel with additive white Gaussian noise $w(t)$. Thus, (2) will reduce to

$$x(t) = \sum_{k=0}^{K} c_k s(t - \tau_k) + w(t). \qquad (3)$$

The channel has 100 taps of which 40 have a significant value (non-zero value). The significant taps are randomly generated and placed in the impulse response.

At the receiver $x(t)$ is sampled at symbol rate T_s and the discrete-time received signal is obtained

$$x(n) = x(nT_s), \quad 0 \le n \le \infty. \qquad (4)$$

The estimated fullband equalizer impulse response is denoted by $\mathbf{g} = [g(0), \cdots, g(L-1)]^T$, and the filter length L is chosen so that the signal $b(n)$ can be recovered. Then, the output signal from the equalizer is given by

$$y(n) = \sum_{l=0}^{L-1} g(l)^* x(n - l). \qquad (5)$$

where $(\cdot)^*$ denotes the complex conjugate.

3

3 Closed Loop Subband CMA Equalizer

The blind algorithm used in this study is the low computationally complex CMA algorithm [2, 3]. In order for the CMA to work, the signal must fulfill the Constant Modulus (CM) criterion. Therefore is the QPSK modulation technique used in this study. However, the CM criteria cannot be guaranteed in the different subbands, so the CMA cost function is applied to the fullband signal, and a CM fullband error is created. Subsequently, the error signal is feed back through the filter bank to the subband filter adaptation algorithms, i.e. closed loop implementation, see Figure 2.
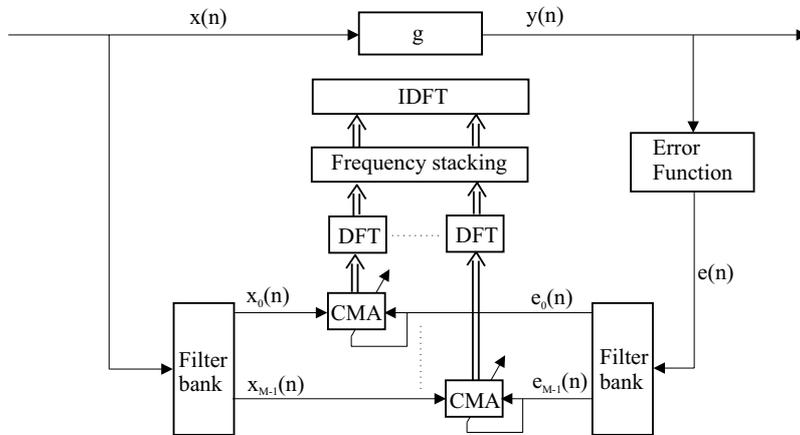

Figure 2: A closed–loop subband CMA equalizer.

3.1 Subband Decomposition

The received signal $x(n)$ is divided into M frequency bands by using a Uniformly Modulated Filter bank (UMF). The UMF is formed by several modulated versions of a prototype filter [4, 8]. Polyphase decomposition is used to efficiently implement the UMF. Due to polyphase decomposition the decimation can be moved before the subband filtering. By dividing the received signal into several frequency bands it is possible to decimate the signal with a decimation factor D. To reduce the impact of aliasing the filter bank is over sampled with a factor of two, i.e. the subband signals are decimated by a factor $D = \frac{M}{2}$. After including $D = \frac{M}{2}$ and polyphase implementation, the mth subband signal in the frequency domain, $X_m(z)$, can be expressed as:

$$X_m(z) = X(z)H_m^{\frac{M}{2}}(z),\tag{6}$$

where

$$H_m^{\frac{M}{2}}(z) = \sum_{d=0}^{\frac{M}{2}-1} (W^m z)^{-d} \sum_{n=-\infty}^{\infty} h(n\frac{M}{2} + d)(W^m z)^{-nM/2}$$

and

$$W^{-mnM/2} = (e^{j\pi m})^n = \begin{cases} (-1)^n & \text{m odd} \\ 1 & \text{m even} \end{cases}.$$

To further reduce the effects of aliasing, a filter bank design method is used that minimizes the in-band aliasing effects in the subband signals [8].

3.2   Closed Loop Subband CMA Equalizer

In Figure 3, the basic architecture of an fullband adaptive CMA equalizer is outlined. The output signal of the CMA equalizer $y(n)$ is given by

$$y(n) = \mathbf{g}^H(n)\mathbf{x}(n), \tag{7}$$

where $(\cdot)^H$ is the complex conjugate transpose. The input vector to the equalizer is denoted as

$$\mathbf{x}(n) = [x(n), x(n-1), \cdots, x(n-L+1)]^T.$$

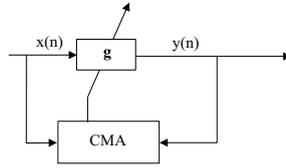The objective of CMA is to restore the output signal $y(n)$ to a constant

Figure 3: An adaptive CMA equalizer.

signal constellation. This is accomplished by adjusting the weight vector $\mathbf{g}(n)$ in order to minimize the cost function $J_{p,q}$, which provides a measure of the amplitude fluctuations. The cost function $J_{p,q}$ is given by

$$J_{p,q} = E\left[\left(|y(n)|^p - 1\right)^q\right] \tag{8}$$

The cost function $J_{p,q}$ can be minimized by using a gradient search algorithm. Typically, to ensure convergence and reasonably small tap-gain fluctuation, (p,q) are set to (2,2) By replacing the ensemble average operation with an instantaneous value in the cost function $J_{2,2}$, the update equation for $\mathbf{g}(n)$ is obtained as

$$\begin{aligned} \mathbf{g}(n+1) &= \mathbf{g}(n) - \mu\varepsilon(n)^*\mathbf{x}(n) \tag{9} \\ \varepsilon(n) &= y(n)\left(|y(n)|^2 - 1\right) \tag{10} \end{aligned}$$

5

where μ is the step-size constant. It is clear from the definition of $J_{p,q}$ that the algorithm is insensitive to phase fluctuations. Thus, the carrier phase must be recovered after the adaptive processing by some kind of phase locked loop (PLL).

The main drawback of the CMA is the relatively slow convergence rate. In order to improve the speed of convergence of the CMA the following normalization factor has been used [9]

$$\eta(n) = \mathbf{x}^H(n)\mathbf{x}(n). \tag{11}$$

The Normalized CMA (NCMA) weight update equation is now given by

$$\mathbf{g}(n+1) = \mathbf{g}(n) - \frac{\mu\varepsilon(n)^*\mathbf{x}(n)}{\eta(n)}. \tag{12}$$

In the subband case, see Figure 2, the error signal is created from the fullband signal and feeds back into the subbands, thus giving a local subband weight update given as

$$\mathbf{g}_m(n+1) = \mathbf{g}_m(n) - \frac{\mu\varepsilon_m(n)^*\mathbf{x}_m(n)}{\eta_m(n)}, \tag{13}$$

where $m, 0 \leq m \leq M - 1$ is the subband index, $\varepsilon_m(n)$ denotes the mth subband component of $\varepsilon(n)$, $\eta_m(n) = \mathbf{x}_m^H(n)\mathbf{x}_m(n)$ denotes the normalization factor in the mth subband and

$$\mathbf{x}_m(n) = [x_m(n), x_m(n-1), \cdots, x_m(n-L_s+1)]^T$$

is the mth subband component of $\mathbf{x}(n)$, where L_s is the length of the subband equalizer. Since, it is desirable to have the main energy of the equalization filter close to the center of the filter, the mid tap of both the subband- and fullband- equalization filters are initialized to 1.

The impulse response for the fullband equalizer is the Inverse Discrete Fourier Transform (IDFT) of the fullband frequency response, which is obtained by stacking the Discrete Fourier Transforms (DFT's) of the subband responses [4, 5, 6] together. In [5] the number of points for the DFT is doubled compared to original method outlined in [6]. Due to the usage of longer DFT lengths, finite block effects are avoided in [5].

4   Simulation Results

The HIPERLAN II inspired channel model is used to evaluate and compare the performance of a closed loop subband NCMA equalizer with a fullband

NCMA equalizer. In the filter bank design the length of the prototype filter is chosen to be four times the total number of subbands. To evaluate the convergence properties of the subband equalizer, the average fullband amplitude fluctuation $\varepsilon_A(m)$ from the constellation magnitude R_s is being used and is defined as

$$\varepsilon_A(m) = 10 \log_{10} \left(\frac{1}{N_B} \sum_{n=0}^{N_B-1} ||u(n)| - R_s|^2 \right), \tag{14}$$

where N_B is the block length over which the amplitude fluctuations are averaged. The block length is set to 1000 symbols and the signal to noise ratio (SNR) is 30 dB. In Figure 4, the fullband equalizer (M = 1) and subband equalizers are compared for a varying number of subbands. Both equalizers are using $L = 256$ taps, and it is clear from the results shown in Figure 4 that the major gain in using a subband update approach lies in the improved convergence rate. The 64 subband equalizer is roughly 12 times faster than the fullband equalizer.
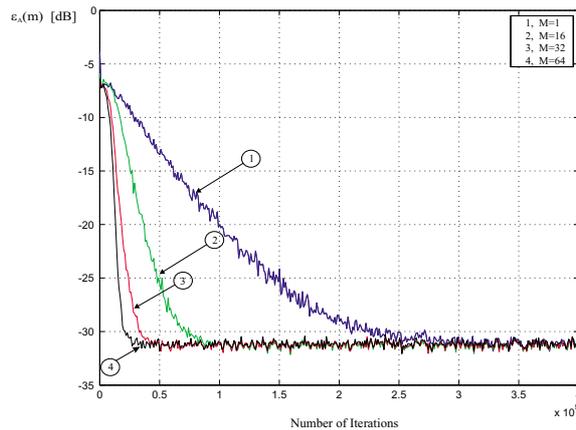

Figure 4: The performance of the subband equalizer for different number of subbands, with an $L = 256$ taps equalizer and SNR = 30 dB.

5 Conclusion and Future Work

In this paper, a new closed loop subband NCMA equalizer has been presented. The subband equalizer was evaluated using Monto-Carlo simulation and HIPERLAN II inspired channel model. The simulation showed that for a $L = 256$ tap equalizer using $M = 64$ subbands, a 12 times faster convergence was achieved compared to its fullband counterpart without any

significant performance loss. Future works includes studies of the tracking performance of the blind subband channel equalizer, i.e. using fading channel coefficients.

References

[1] R. W. Stewart, S. Weiss, D. Garcia-Alis, and G. C. Freeland, *Subband Adaptive Equalization of Time-Varying Channels*, Conference Record of the Asilomar Conference on Signals, Systems and Computers,Monterey, CA, USA, vol. 1, pp. 534-538, 1999.

[2] D. N. Godard, *Self-Recovering Equalization and Carrier Tracking in Two Dimensional Data Communication Systems*, IEEE Transaction on Communication, vol. 28, no. 11, pp.1867-1875, November 1980.

[3] J. R. Treichler and B. G. Agee, *A New Approach to Multipath Correction of Constant Modulus Signals*, IEEE Transaction Acoustics Speech and Signal Processing, vol. 31, no. 2, pp. 459-471, April 1985.

[4] H. H. Dam, J. Nordberg and S. Nordholm, *Non-Causal Delayless Subband Adaptive Equalizer*,Proceedings of IEEE International Conference on Acoustics, Speech and Signal Processing (ICASSP) 2002, Orlando, USA, vol. 3, pp. 2649–2652, May 2002.

[5] J. Huo, S. Nordholm and Z. Zang, *New Weight Transform Schemes for Delayless Subband Adaptive Filtering*, Proceedings of IEEE Global Telecommunications Conference, Texas, USA, vol. 1, pp. 197–201, November 2001.

[6] D. R. Morgan, and J. C. Thi, *A Delay-less Sub-band Adaptive Filter Architecture*, IEEE Transaction on Signal Processing, vol. 43, no. 8, pp. 1819-1830, August 1995.

[7] J. Medbo and P. Schramm, *Channel Models for HIPERLAN/2 in Different Indoor Scenarios*, ETSI/BRAN document for discussion, no. 3ERI085B, 1998.

[8] J. M. De Haan, N. Grbic, I. Claesson, and S. Nordholm, *Design of Oversampled Uniform DFT Filter Banks with Delay Specifications using Quadratic Optimization*, Proceedings of IEEE International Conference on Acoustics, Speech and Signal Processing, Salt Lake City, USA, ICASSP-01, vol. 6, pp. 3633–3636, May 2001.

[9] D. L. Jones, *A normalized constant-modulus algorithm*, Proceedings of 29th Asilomar Conference on Signals, Systems and Computers, Pacific Grove, CA, USA, vol. 2, pp. 694-697, October 1999.