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SYSTEM LEVEL DESIGN OF BASEBAND OFDM FOR WIRELESS LAN

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ABSTRACT

Multicarrier or Orthogonal Frequency Division Multiplexing (OFDM) has gain recently an increased interest with the development of faster signal processing components and technologies. OFDM has been shown to be a very efficient scheme for mitigating the adverse effects of inter-symbol interference, squeezing multiple modulated carriers tightly together but keeping the modulated signals orthogonal so they do not interfere with each other's. This paper describes the system level model of the OFDM baseband and uses it to evaluate the BER performance.

1. INTRODUCTION

The new demand for wireless broadband access networks is imposed by the emergence of multimedia applications and high speed wireless access. Orthogonal Frequency Division Multiplexing has been proven suited for this kind of applications, being a method of data modulation which mitigates the effects of inter-symbol interference (ISI) caused by multipath propagation encountered in high-speed wireless communications. OFDM is used in wireless environment such as mobile communication systems, but also in other applications such as European digital audio broadcast (DAB). OFDM based on discrete multitone (DMT) systems are also employed for broadband digital communication on the existing copper networks, being exploited for high-bit-rate and asymmetric digital subscribers lines (HDSL and ADSL) [1], [2].

In wireless applications, OFDM can be found in several existing wireless local area network (WLAN) standards, like IEEE 802.11 in the United States, ARIB MMAC in Japan, or HIPERLAN/2, defined by the European Telecommunication Standards Institute Project on Broadband Radio Access Networks (ETSI BRAN). This system provides high-speed communications up to 54Mbit/s between portable computing devices and access points, attached to an Ethernet, ATM or UMTS backbone network [2], [7].

The main idea behind OFDM is to split data stream to be transmitted into N parallel streams of reduced data rate and to transmit each of them on a separate subcarrier. These carriers are made orthogonal by appropriately choosing the frequency spacing between them. Therefore, spectral overlapping among subcarriers is allowed, since the orthogonality will ensure that the receiver can separate the OFDM subcarriers, and a better spectral efficiency can be achieved than by using simple frequency division multiplex.

OFDM combats frequency selective fading and randomizes the burst errors caused by the wideband radio frequency channel. To prevent inter-symbol interference, a cyclic prefix is appended to each OFDM symbol at the transmitter and then removed at the receiver, before the detection process. However, there are proposed some techniques and algorithms, as in [11], of non-conventional OFDM systems without cyclic prefix. The cyclic prefix, which is a repetition of the last chips of the OFDM symbol, can be exploited also for both timing and frequency synchronization [13]. The synchronization technique based on the cyclic prefix extension is particularly suited to tracking or blind synchronization in a circuit-switched connection, when no special training signals are available. For packet transmission may be used several special OFDM training symbols for which data content is known at the receiver.

Channel estimation at the receiver is achieved using the training symbols provided in the preamble. A priori knowledge of the transmitted preamble facilitates the generation of a vector defining the channel state information. This is used to remove the channel distortion by multiplying the received OFDM vector by the pre-computed channel estimation. Another method to compensate for the fluctuation of amplitude and phase due to fading is the insertion of pilot chips at fixed time intervals at the transmission, used at receiver to estimate the channel characteristics [5].

The key advantages of the OFDM transmission scheme, according to [16], are:

- OFDM is an efficient way to deal with multipath; for a given delay spread, the implementation complexity is significantly lower than that of a single carrier system with an equalizer.
- In relatively slow time-varying channels, it is possible to significantly enhance the capacity by adapting the data rate per subcarrier according to the signal-to-noise ratio of the particular subcarrier.
- OFDM is robust against narrowband interference, because such interference affects only a small percentage of the subcarriers.

Nevertheless, OFDM has also some negatives associated compared with single-carrier modulation:

- OFDM is more sensitive to frequency offset and phase noise.
- The orthogonal nature leads to signals with a large dynamic range, imposing a relatively large peak-to-average power ratio, which tends to reduce the power efficiency of the RF amplifier.

In the following we give a mathematical description of the OFDM signal and present typical OFDM system. Finally, we show the results obtained by simulating an OFDM system in a slightly modified HIPERLAN/2 context.

2. SYSTEM MODEL

The following description of the OFDM signal is based on [4] and [6]. In order to give a mathematical description of an OFDM system we assume a system with N subcarriers, a bandwidth of B Hz and an OFDM symbol length of T_S seconds, of which T_{CP} is the length of the cyclic prefix. The spacing between subcarriers is given by (1), as in Figure 1.

$$T = \frac{1}{\Delta f} = \frac{N}{B} = T_S - T_{CP} \tag{1}$$

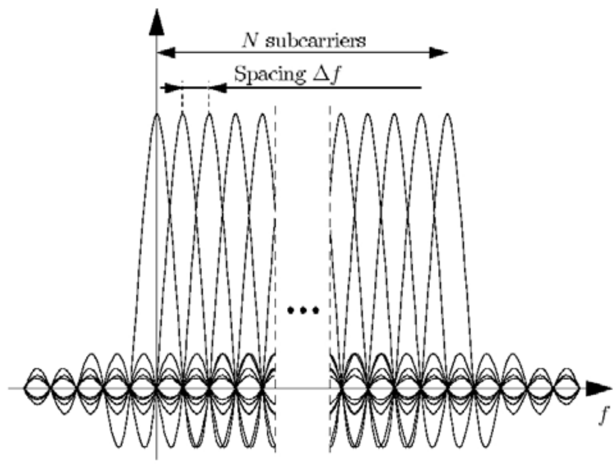

Figure 1. Individual subchannels for an OFDM system.

Figure 2 illustrates the baseband OFDM model mathematically described bellow. Every n^{th} OFDM symbol of the transmission stream can be written as a set of modulated carriers transmitted in parallel. Relations (2) express the waveforms used in modulation.

$$\phi_k(t) = \begin{cases} \dfrac{1}{\sqrt{T_S - T_{CP}}} \cdot e^{j2\pi f_k (t - T_{CP})} & , t \in [0, T_S) \\ 0 & , otherwise \end{cases}, \quad where$$

$$f_k = f_C + \left(k - \frac{N-1}{2}\right) \cdot \frac{1}{T}, \quad k = 0, ..., N-1, \quad for\ passband \quad or \tag{2}$$

$$f_k = \frac{k}{T}, \quad k = 0, ..., N-1, \quad for\ baseband\ echivalent$$

Note that nonzero term of $\phi_k(t)$ has the period $[T_{CP}, T_S)$ and

$$\phi_k(t) = \phi_k(t + \frac{N}{B}), \quad for\ t \in [0, T_{CP}) \tag{3}$$

If $d_{n,0}, ..., d_{n,N-1}$ denotes the complex symbols, obtained by QAM mapping of the input data stream, the n^{th} OFDM symbol $s_n(t)$ is expressed by (4) and the infinite sequence of OFDM symbols transmitted is obtained by juxtaposition of the individual ones.

$$s(t) = \sum_{n=-\infty}^{\infty} s_n(t) = \sum_{n=-\infty}^{\infty} \sum_{k=0}^{N-1} d_{k,n} \phi_k(t - nT_S) \tag{4}$$

Assuming the impulse response $ch(\tau;t)$ of the physical channel (possibly time variant) is restricted to the length of cyclic prefix $\tau \in [0, T_{CP})$, the received signal becomes (5), where $n(t)$ is complex, additive and white Gaussian (AWGN) channel noise.

$$r(t) = (ch * s)(t) = \int_0^{T_{CP}} ch(\tau;t) s(t - \tau) + n(t) \tag{5}$$

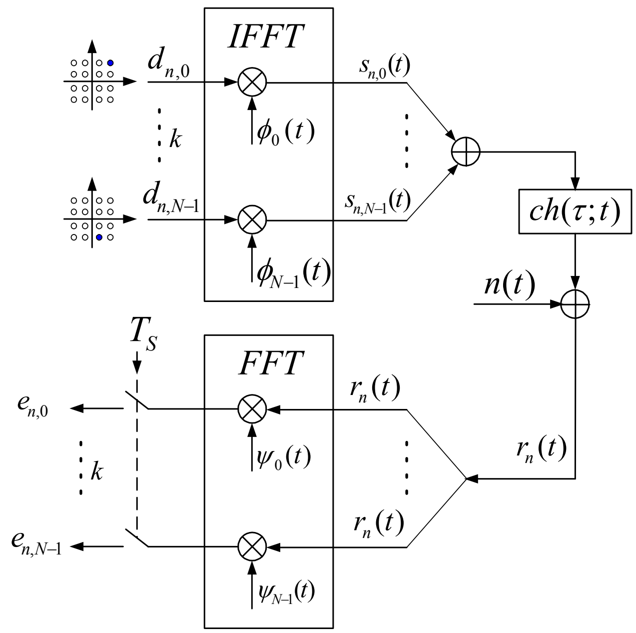

Figure 2. Baseband OFDM system model.

The filter from the receiver is matched to the last part $[T_{CP}, T_S)$ of the transmitter waveform (6), the cyclic prefix being this way effectively removed in the receiver. Since the cyclic prefix contains the ISI, the sample output from the receiver filter bank contains no ISI. Also, we can ignore the time index n when calculating the sampled output at the k^{th} matched filter (7).

$$\psi_k(t) = \begin{cases} \phi_k^*(T_S - t) & , t \in [0, T_S - T_{CP}) \\ 0 & , otherwise \end{cases} \tag{6}$$

$$e_k = (r * \psi_k)(t)\big|_{t=T_S} = \int_{-\infty}^{\infty} r(t) \cdot \psi_k(T_S - t) \cdot dt \tag{7}$$

Considering the channel to be fixed over the OFDM symbol interval, denoting it by $ch(\tau)$ and taking into account the orthogonality condition expressed by (8), we obtain after some mathematical operations the output data, given by (9).

$$\int_{T_{CP}}^{T} \phi_l(t) \cdot \phi_k^*(t) = \delta(k - l) \tag{8}$$

$$\begin{aligned} e_k &= h_k \cdot d_k + n_k, \quad where \\ h_k &= \int_0^{T_{CP}} ch(\tau) \cdot e^{-j2\pi k \tau \frac{B}{N}} \cdot d\tau \ and \\ n_k &= \int_{T_{CP}}^{T_S} n(T_S - t) \cdot \phi_k^*(t) \cdot dt \end{aligned} \tag{9}$$

By sampling the low-pass equivalent signal of (2) and (4) at a rate N times higher than the subcarrier symbol rate *1/T*, we can obtain the discrete model of the baseband OFDM system, where the modulation/demodulation with waves ϕ/ψ can be replaced with iDFT/DFT (or practically with IFFT/FFT) and the channel model with discrete-time convolution.

In figure 3, a block diagram of an OFDM system is presented, based on [9], [16], and the practical way of generating the OFDM signal is described bellow. The binary input data are encoded using a FEC block. This is realized by using a single convolutional code or a concatenated code. Usually it implies an outer block code with hard decision, which may be a shortened Reed-Solomon [17], followed by a convolutional code with a rate of ½ or higher by puncturing. The encoders are accompanied by bit interleaving blocks and some times by a scrambler. At the reception, the inverse operations are made, using as pair for the convolutional encoder, a Viterbi decoder with soft decision.

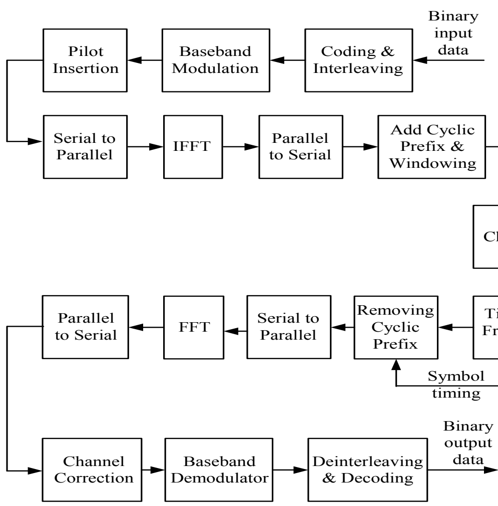

Figure 3. Block diagram of an OFDM system.

The baseband modulation performed can be from BPSK to 256 QAM. Conforming [16], in fading channels is preferable to use high-order constellations in combination with low-rate coding schemes. The usage of Trellis structures as convolutional encoders designed for specific constellations requires Gray-encoded QAM. In order to estimate the fading channel's effects on the received signal, especially suited for continuous transmissions is the insertion at fixed time intervals of pilot subcarriers with polarity generated by a pseudoaleatory sequence. The FFT block's hardware is very similar to IFFT and it may be used the same block for both if the transmitter and the receiver does not have to work simultaneously. According to [16], an IFFT can be efficiently built using a *radix-4 butterfly* algorithm. The cyclic prefix is important for synchronization, especially in continuous transmissions, and for ISI caused by multipath propagation avoidance. The out-of-band spectrum decreases rather slowly, so that to accelerate it, windowing is applied to the OFDM signal. It can be used conventional windows, such as raised cosine, Hamming, Blackman, Kaiser, or special designed windows, such as Lawrey window [10].

3. SIMULATION RESULTS

The main parameters of the HYPERLAN/2 standard are expressed in table 1 [8]. We chose to simulate a system at *8Mbps* rate, with *64QAM* modulation. The guard interval (CP) has to be *3-5* times greater than the delay spread, in order to mitigate the ISI. Assuming a tolerable delay spread of *200ns* for both indoor and microcellular outdoor [16], a guard time of $T_{CP} = 800ns$ meets the requirement. Also, the SNR loss introduced by the CP and given by (10) is $\Delta_{SNR} = 0.9691dB$, which is an acceptable value, being smaller than $1dB$.

$$\Delta_{SNR} = -10 \cdot \log_{10}\left(1 - \frac{T_{CP}}{T_S}\right) \tag{10}$$

Data rate	*6, 9, 12, 18, 24, 36, 48, 54 Mbps*	OFDM symbol duration	$4\mu s$
Modulation	*BPSK, QPSK, 16QAM, 64QAM*	Guard interval	*800ns*
Coding rate	*1/2, 2/3, 3/4*	Subcarrier spacing	*312.5 kHz*
Number of subcarriers	*52*	-3dB Bandwidth	*16.56 MHz*
Number of pilots	*4*	Channel spacing	*20 MHz*

Table 1. Main parameters of the HYPERLAN2 standard.

Because the number of subcarriers is *52*, the FFT length will be *64*, matched by the subcarrier spacing (*20MHz/64 = 312.5kHz = 1/3.2µs*). To obtain a *48Mbps* bit rate with a symbol period of $T_S = 4\mu s$, the number of bits per OFDM symbol should be *192*. Taking into account that the modulation level is *6* ($2^6 = 64$), we used a coding rate of *2/3*, matched by the number of data subcarriers *52-4=48* (*2/3 = 192/288 = 32/48*). We can use an outer shortened Reed-Solomon(36,32) encoder and an inner Trellis encoder with constrained length *7* and generator polynomials $133_{(6)}$ and $171_{(6)}$, with a ½ rate and punctured to ¾ .

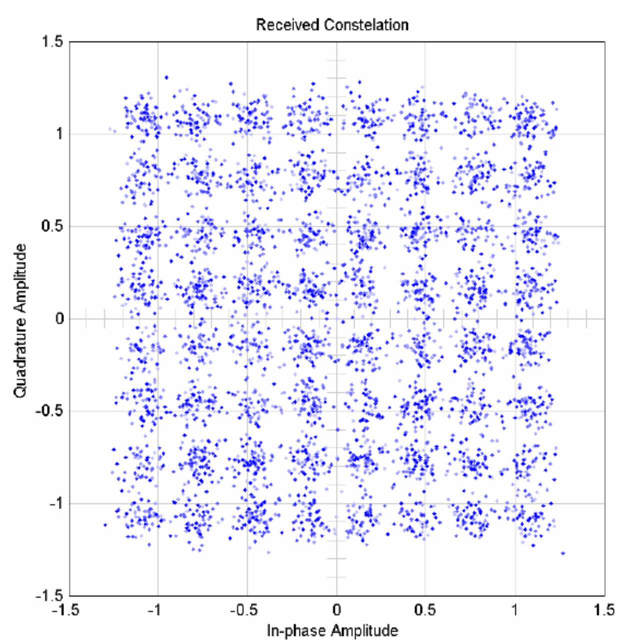

Figure 4. Received signal constellation for *64QAM* and $E_b/N_0 = 19.5dB$.

The system was modeled with Mathworks Simulink using frame based signals, which speeds up the computation [12], [15]. Using Simulink blocks, data rates, power of the signal, signal waves, QAM symbol constellation, as in figure 4, signal spectrum, as in figure 5, and instantaneous BER can be explored. In order to obtain a *BER versus* E_b/N_0 for AWGN channel we used a matlab script that automatically modifies the E_b/N_0 parameter of the AWGN block and runs the simulation a specified simulation time. The result may be observed in figure 6.

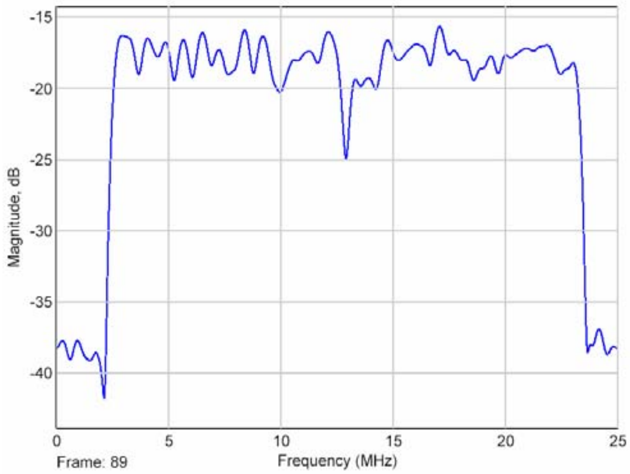

Figure 5. Transmitted signal spectrum.

4. CONCLUSIONS AND FUTURE WORK

According to [5] and [14], BER performance in conventional *64QAM* scheme and AWGN channel is given by (11).

$$BER_{Th} = \frac{7}{24} \cdot erfc\left(\sqrt{\frac{1}{7} \cdot \frac{E_b}{N_0}}\right) - \frac{49}{384} \cdot erfc^2\left(\sqrt{\frac{1}{7} \cdot \frac{E_b}{N_0}}\right) \qquad (11)$$

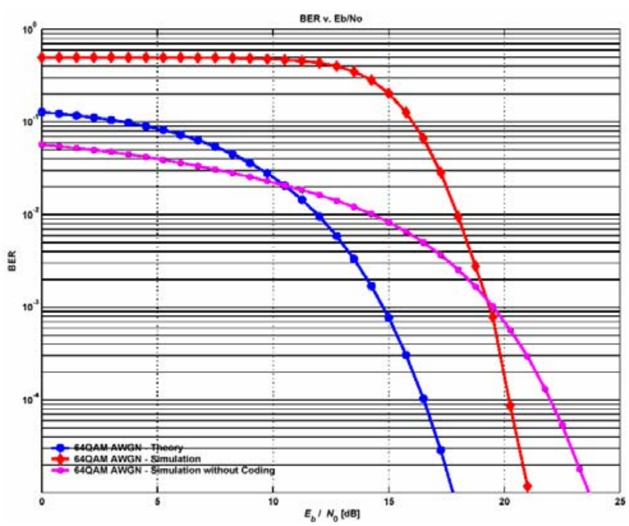

Figure 6. BER for *64QAM* and *AWGN*.

In figure 6, it can be observed the shape of BER for AWGN channel for the theoretical situation and the simulating one with and without coding. The difference between the simulated situation and the theoretical one against the bit energy-to-noise ratio (E_b/N_0) has to be observed. It must not be forgotten the *0.9691dB* SNR loss caused by cyclic extension as compared to the theoretical results.

As future work we intend to extend first the Simulink model in order to simulate the multipath fading effect; second, to introduce as channel model, the extended Suzuki model [3] and third, to parameterize the simulation model.

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