OFDM for High Speed Wireless Networks

Richard van Nee
Lucent Technologies Bell Labs Research

NOTICE

ABSTRACT

A comparison is made between the implementation complexity of an OFDM modem and a comparable single carrier system for high speed (>20 Mbps) wireless networks. It is demonstrated that OFDM has a significant lower complexity than a single carrier system with an equalizer. Further, OFDM can achieve a significantly better spectral efficiency than a constant envelope modulation technique like GMSK. It is shown that OFDM needs only a small training overhead. Regarding the power amplifier, the required backoff values for OFDM are about the same as needed for other linear modulation schemes such as QPSK. Finally, it is concluded that for the proposed OFDM parameters, phase noise and frequency offsets give a negligible signal-to-noise ratio degradation of less than 0.1 dB.

1 INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) has several attractive properties which make it a suitable modulation choice for high speed wireless networks. One of these properties is that for a certain delay spread, the complexity of an OFDM modem versus bit rate does not grow as fast as the complexity of a single carrier system with an equalizer. The reason is that when the bit rate is doubled, an equalizer has to be made twice as long at twice the speed, so its complexity grows quadratically with the bit rate. For OFDM, the FFT size has to be doubled, which gives a complexity increase of $2(1+\log_2 N)/\log_2 N$, where N is the original FFT size[1]. Hence, the FFT complexity grows only slightly faster than linear, which makes it easier to implement OFDM in modems which have to handle more than 50 ns of delay spread at data rates exceeding 20 Mbps.

Another attractive property of OFDM is its easy scaleability to different environments, bandwidths or bit rates, by simply changing parameters like the number of subcarriers, guard time, coding rate

[1] *Using radix-2 algorithm, an N-point FFT requires* $(N/2)\times\log_2(N)$ *complex multiplications*

or constellation size. This gives the possibility to use one modulation technique for various applications.

Commonly quoted disadvantages or concerns about OFDM are the peak-to-average power ratio, sensitivity to phase noise and the amount of training overhead when using a large number of subcarriers. This document shows that neither of these concerns provides a real disadvantage for OFDM.

2 OFDM TRANSCEIVER

Figure 1 shows a block diagram of an OFDM transceiver. In the transmitter path, binary input data is first encoded using a convolutional encoder. The coded output data is interleaved to get the benefit of time and frequency diversity. After interleaving, the binary data is mapped on Quadrature Amplitude Modulation (QAM) symbols. These symbols are then converted from serial to parallel, with a block length equal to the number of subcarriers. For each block of data, the Inverse Fast Fourier Transform (IFFT) is calculated, with a size that is larger than the number of subcarriers to make an output spectrum with low enough out-of-band radiation. The IFFT output is converted from parallel to serial, after which the final OFDM symbol is formed by adding a cyclic extension and a windowing function. The cyclic extension should be at least two times the expected delay spread in order to reduce intersymbol interference to an acceptable level.

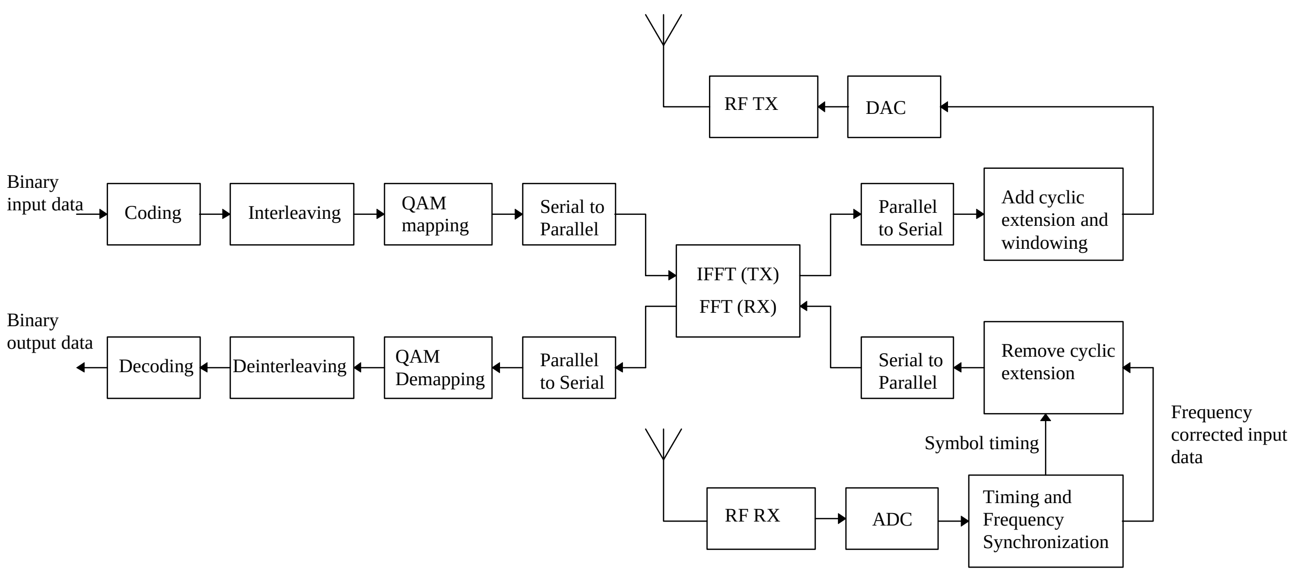

Figure 1: Block diagram of an OFDM transceiver.

Figure 2 shows the OFDM symbol structure. Here, T is the FFT duration and T_G is the guard time. Each OFDM symbol is windowed by a raised cosine window to reduce the out-of-band radiation. The purpose of the guard time and the cyclic prefix is to prevent both inter symbol interference (ISI) and inter carrier interference (ICI). To illustrate this, 3 subcarriers are depicted in more detail in figure 3. An OFDM receiver uses only a part of this signal to calculate the FFT. In the FFT interval, every subcarrier has exactly an integer number of cycles, which ensures orthogonality. In a multipath fading channel, the input signal will be a sum of delayed and scaled replicas of those depicted in figure 3. However, for each multipath component, there will still be an integer number of cycles within the FFT interval, as long as the multipath delay does not exceed the guard time. Hence, there is no interference between symbols or between subcarriers. Thanks to the guard time and cyclic prefix, the wideband multipath fading is experienced in OFDM as a set of narrowband fading subcarriers without ISI or ICI. The effect of narrowband fading is that the received subcarriers have different amplitudes, and some may be almost lost in deep fades. In order to become insensitive to such deep fades, forward error correcting coding is used. By proper coding and interleaving across the subcarriers, the OFDM link performance is depending on the average received power rather than the worst case lowest power in deep fades.

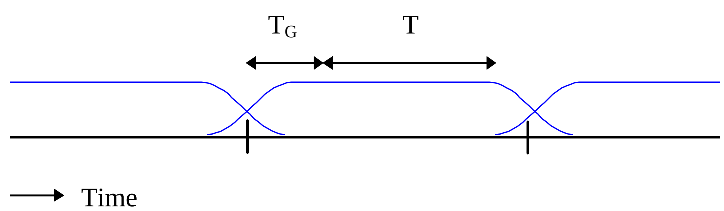

Figure 2: OFDM symbol structure showing guard time T_G and FFT interval T.

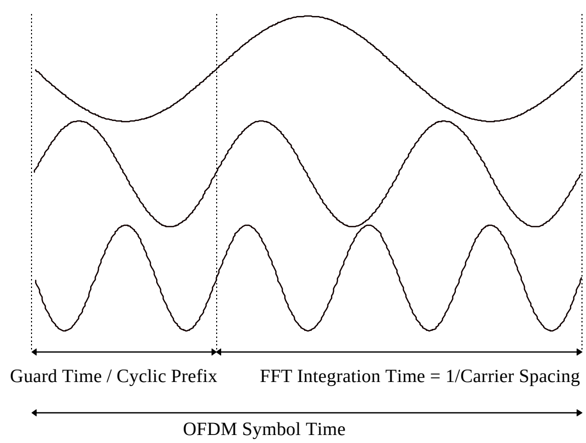

Figure 3 : OFDM symbol structure showing 3 specific subcarriers.

3 PROCESSING LOAD OF OFDM VERSUS SINGLE CARRIER

Figure 4 shows the block diagram of a decision feedback equalizer with symbol spaced taps. From references [1, 2], it can be learned that at least 8 feedforward and 8 feedback taps are required to handle a delay spread of 100 ns at a data rate of 24 Mbps. To get the same delay spread performance as the 33 Mbps OFDM proposal which can handle delay spreads up to 200 ns using a 64-point FFT [3], the equalizer would have to run at 33 MHz instead of 24 MHz, and the number of taps has to be increased to about 22 to get an effective doubling of the equalizer span (8 taps at 24 MHz gives a span of 333 ns, 22 taps at 33 MHz give 666 ns). Fortunately, for GMSK, only the real outputs of the complex multiplications are used, so each multiplier has to perform 2 real multiplications per sample. Hence, the number of real multiplications per second becomes $2{\cdot}22{\cdot}33{\cdot}10^6 = 1452{\cdot}10^6$. For the OFDM system, a 64-point FFT has to be processed every 2.6 µs. With a radix-4 algorithm, this requires 64 complex multiplications[2], which gives a processing load of about $100{\cdot}10^6$ real multiplications per second. So, in terms of multiplications per second, the OFDM system has a processing advantage of more than a factor of 14!

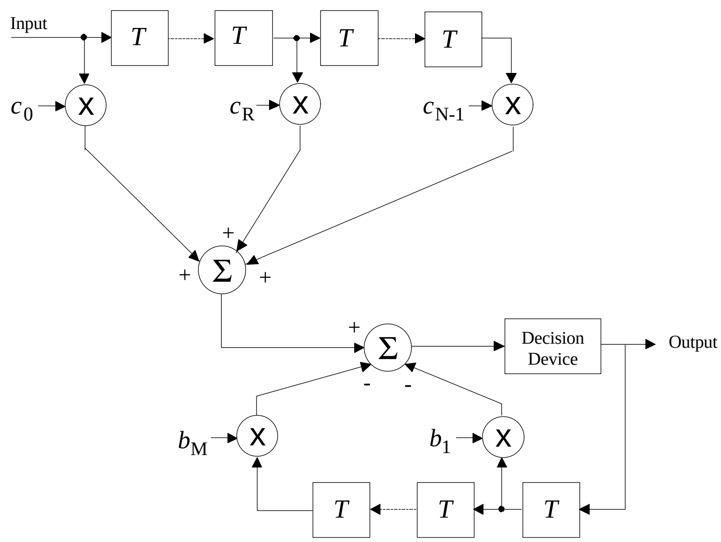

Figure 4: Decision-feedback equalizer

Even if the comparison is made with a 24 Mbps single carrier system with reduced delay spread tolerance, there is a distinct advantage for OFDM. In this case, a single carrier system needs an 8-taps equalizer to cope with delay spreads up to 100 ns, which gives a processing load of $2{\cdot}8{\cdot}24{\cdot}10^6 = 384{\cdot}10^6$ real multiplications per second. The above mentioned OFDM system with a 64-point FFT can be scaled down from 33 Mbps to 24 Mbps by simply lowering the clock rate (and bandwidth) by a factor of 24/33. This increases the symbol duration to 3.58 µs - and the delay spread tolerance to 275 ns - so the processing load becomes $72{\cdot}10^6$ real multiplications per second. Thus, the single carrier system still needs 5 times more processing, even though the OFDM system in this example is able to deal with almost 3 times larger delay spread values!

If we take into account processing differences in other parts of the modem, probably the only significant difference is made by the channel estimation - estimating and correcting for the amplitudes and phases of the various OFDM subcarriers. In a single carrier system, coherent detection is implicitly done by the equalizer. In OFDM, all subcarriers must be explicitly corrected using the reference phases obtained during the training phase. This requires an extra complex

---

[2] *With the radix-4 algorithm, an N-point FFT can be calculated with $\frac{1}{2}N(\frac{1}{2}\log_2(N)-1)$ multiplications for values of N equal to a power of 4 [4].*

multiplication for each subcarrier containing data. In the system with a 64-point FFT, 43 subcarriers carry data, so an extra 43 multiplications per 2.6 μs have to be performed for coherent detection. This increases the previously mentioned processing loads by a factor of 1.67, which means that OFDM has a processing advantage of 8 or 3 compared to a single carrier system at 33 Mbps or 24 Mbps, respectively.

Notice that in the above comparison, we did not include equalizer training, which will even further increase the gap between OFDM and single carrier systems. With the simplest possible training using the least mean squares (LMS) technique, the amount of multiplications per second is doubled during the training phase. For other techniques, like recursive least squares (RLS) or direct tap calculation techniques, the processing load is roughly an order of magnitude higher.

To illustrate the feasibility of the proposed OFDM system, it is worthwhile to mention he European Digital Video Broadcasting standard, which uses OFDM to deliver data rates of more than 30 Mbps using a 8192 point FFT [5]. Commercial chips are now available which include an entire OFDM demodulator for DVB [6]. Since the 8192 point FFT has to be processed within 1 ms, the processing load for the FFT is about $100 \cdot 10^6$ real multiplications per second, which is exactly the same as for the proposed OFDM system with a 64-point FFT in 2.6 μs. In fact, the DVB chip is more complicated because it needs to do much more buffering.

4 DISADVANTAGES OF OFDM?

In all discussions about OFDM versus other modulation schemes, the following three points are often quoted as being disadvantages for OFDM:

- Phase noise and frequency offset
- Peak-to-average power ratio
- Training overhead when using a large (>32) number of subcarriers

The next sections take a detailed look at these effects to see how much of a disadvantage they really are.

4.1 Phase Noise and Frequency Offset

In an OFDM link, the subcarriers are perfectly orthogonal only if transmitter and receiver use exactly the same frequencies. Any frequency offset or phase noise will immediately result in inter carrier interference (ICI). This is the reason that phase noise and frequency offset are often mentioned as disadvantages of OFDM. However, as this section will show, for the OFDM parameters needed, the effects of phase noise and frequency offset are minimal.

In [7], the degradation in signal-to-noise ratio caused by phase noise is given as:

$$D_{phase} \cong \frac{11}{6 \ln 10}\left(4\pi N \frac{\beta}{B}\right)\frac{E_s}{N_o} \qquad (1)$$

Here, b is the -3 dB one-sided bandwidth of the power density spectrum of the carrier. Figure 5 shows the signal-to-noise ratio degradation in dB as a function of b. For a negligible degradation of less than 0.1 dB, the -3 dB phase noise bandwidth has to be about 0.1 percent of the subcarrier spacing. For the proposed OFDM system with a 64-point FFT, the subcarrier spacing is about 455 kHz [3], so the tolerable -3 dB phase noise bandwidth becomes 455 Hz. Practical papers relating to phase noise often quote -100 dBc/Hz spectral density figures rather than the -3 dB linewidth. For instance, in [8], a phase noise spectrum at 5 GHz is shown with a -100 dBc/Hz value at an offset of less than 100 kHz and -40 dBc/Hz at an offset of about 10 Hz. To relate this to the -3 dB linewidth, figure 6 shows a modeled phase noise spectrum with a -3 dB linewidth of 1 Hz. The -100 dBc/Hz value of this curve corresponds to the measured value of [8], while the -40 dBc/Hz value is actually slightly larger than that reported in [8]. Hence, 1 Hz is a conservative figure for the -3 dB linewidth. Since the tolerable -3 dB bandwidth is 455 times larger, it can be concluded that phase noise does not cause significant degradation.

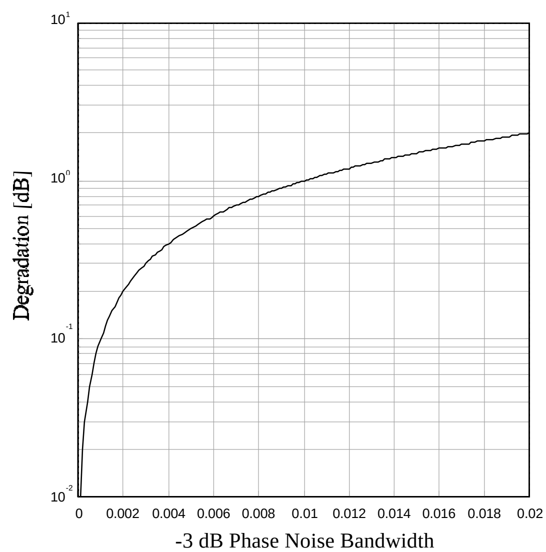

Figure 5: Signal-to-noise ratio degradation in dB versus the one sided 3-dB bandwidth of the phase noise spectrum.

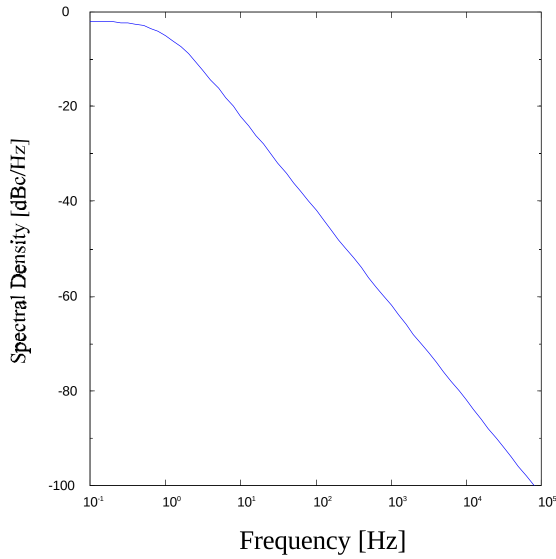

Figure 6: Phase noise spectral density with a -3 dB linewidth of 1 Hz and a -100 dBc/Hz density at 100 kHz offset.

In [7], the degradation in signal-to-noise ratio caused by a frequency offset is given as:

$$D_{freq} \cong \frac{10}{3\ln 10}\left(\pi \frac{N\Delta f}{B}\right)^2 \frac{E_s}{N_o} \qquad (2)$$

This degradation is depicted in figure 7 as a function of the frequency offset, normalized to the subcarrier spacing. It can be seen that for a negligible degradation of about 0.1 dB, the maximum tolerable frequency offset is about 1% of the subcarrier spacing. For the previously mentioned OFDM example, 1% equals 45 kHz, which translates into a relative accuracy of about 10 ppm at a carrier frequency of 5 GHz. The initial frequency error of a low-cost oscillator will normally not meet this requirement, which means that a frequency synchronization technique has to be applied prior to the FFT. One example of such a technique is to measure the phase difference between two OFDM training symbols. Figure 8 shows the simulated standard deviation of the frequency error for this technique, where the frequency error was measured by taking the phase difference between 2 training symbols with a spacing of 3 symbol intervals. It demonstrates that an accuracy of 1% of the subcarrier spacing can be achieved for an input signal-to-noise ratio of 4 dB.

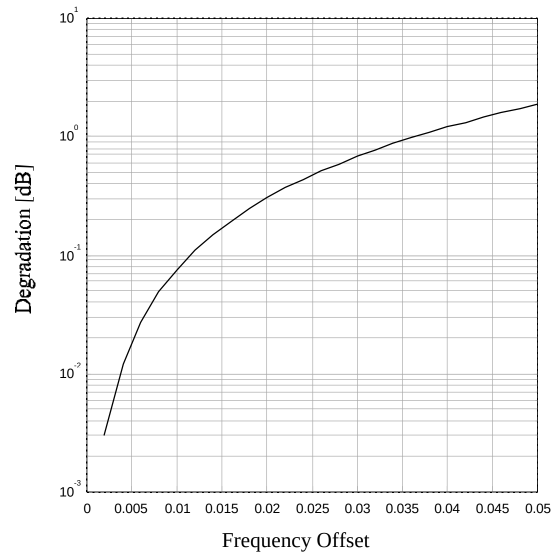

Figure 7: Signal-to-noise ratio degradation in dB caused by frequency offset, normalized to the subcarrier spacing.

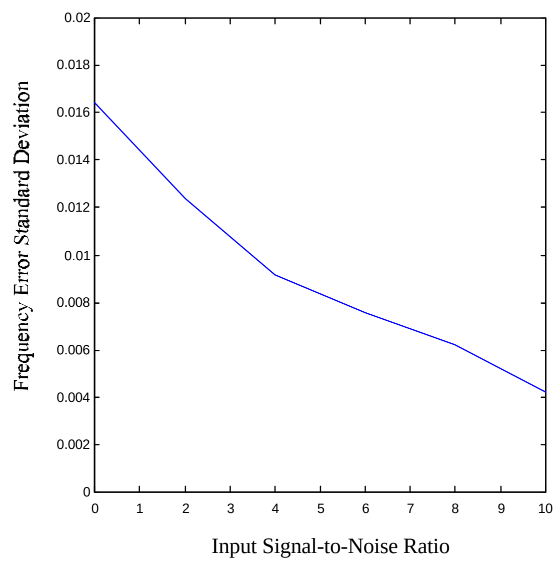

Figure 8: Simulated frequency estimation error (normalized to subcarrier spacing) versus input signal-to-noise ratio. Estimation is done over 4 OFDM symbols, by estimating phase difference of first and last symbol.

As a final comment on the issues of phase noise and frequency offset, we again refer to the Digital Video Broadcasting system [6]. This system uses OFDM with a subcarrier spacing of about 1 kHz. Clearly, since DVB does not appear to have a major difficulty with phase noise or frequency offsets, it should not be too difficult to make an OFDM system with subcarrier spacings that are more than 100 times larger!

4.2 Peak-to-Average Power Ratio

Many people view the peak-to-average power ratio as a major problem of OFDM, since it tends to decrease the efficiency of the RF power amplifier and also requires a large amount of quantization bits if any signal distortion is to be avoided. However, as will be shown in this section, it is possible to reduce the peak-to-average power ratio of OFDM to quite acceptable levels with little impact on the bit error ratio and without causing too much out-of-band radiation.

The most simplest way to reduce the PAP ratio is to clip the signal, but this significantly increases the out-of-band radiation. A different approach is to multiply large signal peaks with a certain window, like proposed in [9]. In [9], a Gaussian shaped window is used, but in fact any window can be used, provided it has good spectral properties. Since the OFDM signal is multiplied with several of these windows, the resulting spectrum is a convolution of the original OFDM spectrum with the spectrum of the applied window. So, ideally the window should be as narrowband as possible. On the other hand, the window should not be too long in the time domain, because that implies that many signal samples are affected, which increases the bit error ratio. Examples of suitable window functions are the cosine, Kaiser and Hamming window. An example of reducing the large peaks in OFDM with the use of windowing is given in figure 9.

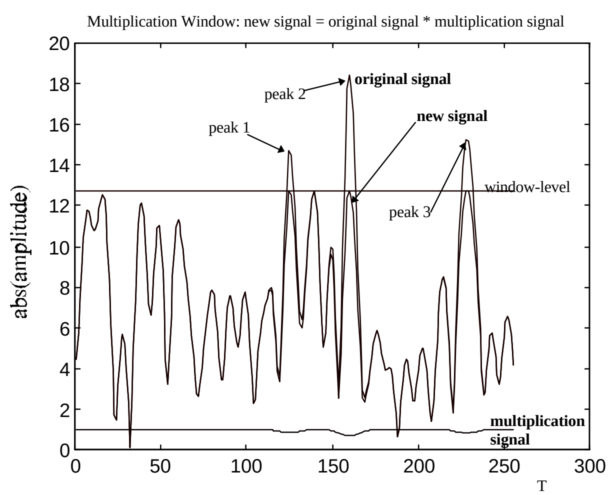

Figure 9:Windowing an OFDM time signal

From the previous analysis, it does not immediately follow what backoff is required for a practical power amplifier in order to get an acceptable level of out-of-band radiation. To simulate a power amplifier, the following model is used for the AM/AM conversion [10]:

$$g(A) = \frac{A}{\left(1 + A^{2p}\right)^{\frac{1}{2p}}} \qquad (3)$$

The AM/PM conversion of a solid state power amplifier is small enough to be neglected. Figure 10 gives some examples of the transfer function for various values of p. A good approximation of existing amplifiers is obtained by choosing p=3 [10]. For large values of p, the model converges to a clipping amplifier which is perfectly linear till it reaches its maximum output level.

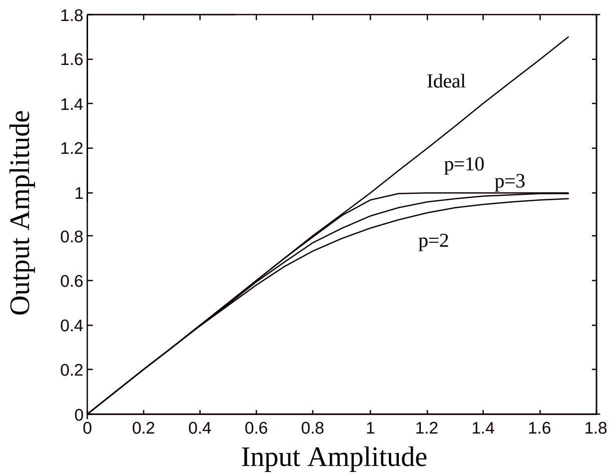

Figure 10: Rapp's model of AM/AM conversion.

Figure 11 shows OFDM spectra with and without peak windowing using Rapp's amplifier model with *p*=3. For a sideband suppression of at least 30 dB, peak windowing needs only 5 dB backoff relative to the maximum output power. Without peak windowing, about 7.4 dB backoff is required. These backoff values are about the same as those required for other linear modulation schemes such as QAM.

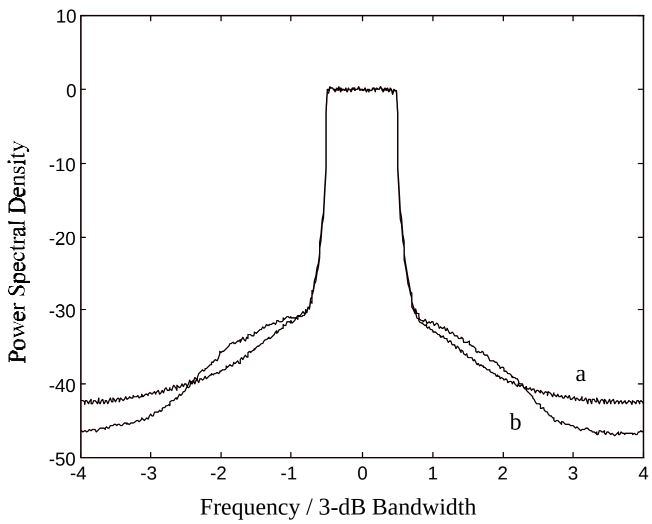

Figure 11: a) plain OFDM with 7.4 dB backoff,
b) peak windowing with 5 dB backoff
32 carriers, Rapp's model with p=3

Figure 12 shows similar plots as figure 11, but now for 1024 subcarriers. This demonstrates that the required backoff with or without peak windowing is independent from the number of subcarriers, as long as this number is large compared to one.

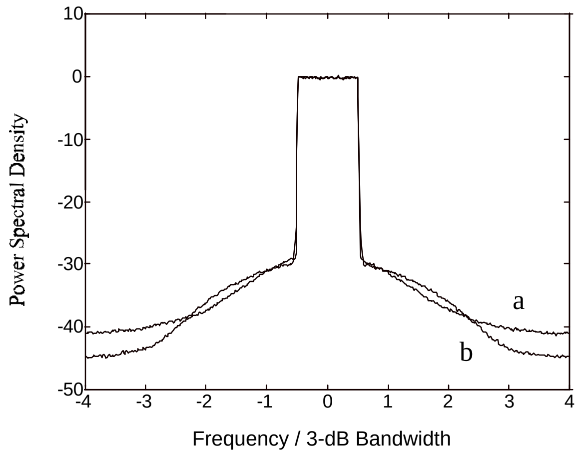

Figure 12: a) plain OFDM with 7.4 dB backoff, b) peak windowing with 5 dB backoff 1024 carriers, Rapp's model with p=3

To demonstrate the robustness of an OFDM link to nonlinear distortion, the figures 13 and 14 show bit and packet error curves with the signal being clipped to achieve a certain maximum peak-to-average power ratio. The simulated link has 48 subcarriers with 16-QAM and a rate ½ convolutional code with constraint length 7. The plots demonstrate that the peak-to-average power ratio can be reduced to 5 dB with a signal-to-noise degradation of about 0.3 dB. For a PAP ratio of 6 dB, the degradation is negligible.

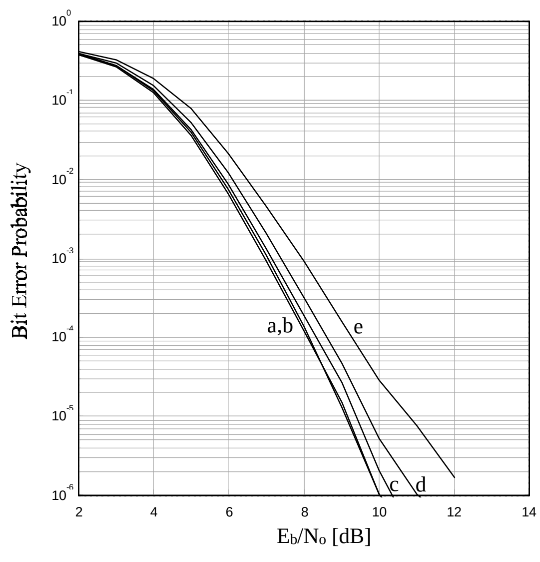

Figure 13: Bit error ratio in additive white Gaussian noise. Maximum PAP ratio is a) 16 dB (no clipping), b) 6 dB, c) 5 dB, d) 4 dB, e) 3 dB.

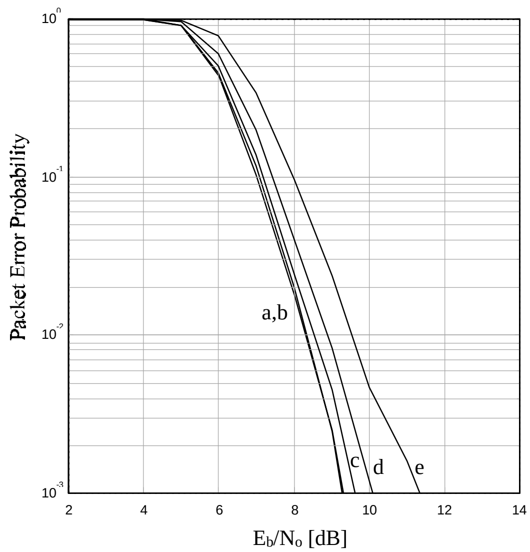

Figure 14: Packet error ratio for a packet length of 1024 bits in additive white Gaussian noise. Maximum PAP ratio is a) 16 dB (no clipping), b) 6 dB, c) 5 dB, d) 4 dB, e) 3 dB.

4.3 Training Overhead

Training overhead is necessary for training the automatic gain control (AGC) and for estimating symbol timing, frequency offset and channel estimates. In wireless packet transmissions, it is important to make this overhead as small as possible, to prevent a significant decrease in net throughput. Compared to single carrier systems, OFDM actually has an advantage in the sense that no equalizer training is required. However, it is sometimes argued that OFDM has the disadvantage that the training overhead increases as the number of subcarriers is increased. This would be the case if a fixed number of OFDM symbols is required for training. Increasing the number of subcarriers and the symbol duration then immediately increases the absolute amount of training overhead. However, this is not the best way to do training in OFDM. In the following, a practical training structure is given, and it is explained how the training overhead could be kept constant for larger number of subcarriers.

Figure 15 shows the proposed training structure at the beginning of a packet. Here, t_1 up to t_5 denote training symbols, while d_i are OFDM data symbols. The length of a training symbol is smaller than the duration of an OFDM data symbol because of two reasons. First, the training symbols do not have a guard time. This is done to benefit from the good cyclic autocorrelation properties of OFDM symbols. Second, the subcarrier spacing for training symbols may be larger than that of data symbols. This makes it possible to use the same training symbols in different applications for which the data symbols need to use different subcarrier spacings to meet delay spread requirements.

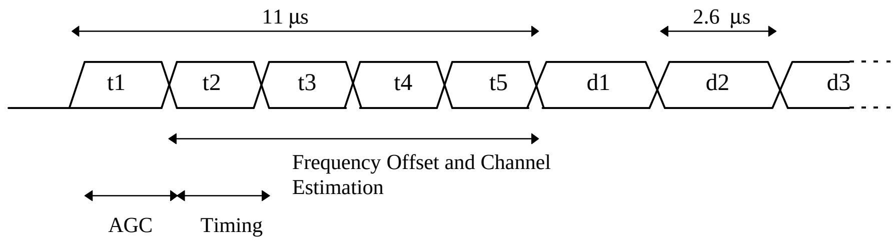

Figure 15: Training structure.

If we want to use a larger number of subcarriers, the training structure from figure 15 could still be used for timing and channel estimation. The number of subcarriers in the training symbols does not necessarily have to increased, which means that the amount of training overhead remains unchanged. However, for a larger number of subcarriers, it is necessary to improve the accuracy of the frequency offset estimate. One way of doing this without increasing the training overhead is to insert the first three data symbols after the third training symbol. By doing this, the time over which the frequency estimate can be estimated is proportional to the data symbol duration. Hence, if the symbol duration is increased, the accuracy of the frequency estimate is increased proportionally. So, the training overhead does not depend on the number of subcarriers, but rather on the tolerable delay spread, since that determines the minimum length of the training symbols.

5 SIMULATION RESULTS

Figure 16 and 17 show simulated bit error and packet error probabilities for a delay spread of 100 ns in a Rayleigh fading channel with an exponentially decaying power delay profile. The error ratios are plotted versus the average received E_b/N_o. The instantaneous E_b/N_o is different for each channel because of the Rayleigh fading. A comparison is made between QPSK with rate ½ and rate ¾ coding, and 16-QAM with rate ½ coding. Interestingly, 16-QAM performs better than QPSK with rate ¾ coding at low error rates, even though 16-QAM with rate ½ coding has a better spectral efficiency (2 bps/Hz versus 1.5 bps/Hz for QPSK with rate ¾). This is caused by the fact that a large Hamming distance is more important than a large Euclidean distance (which would favor QPSK), because in OFDM, the coding must be able to correct a certain amount of subcarriers which are lost in deep fades.

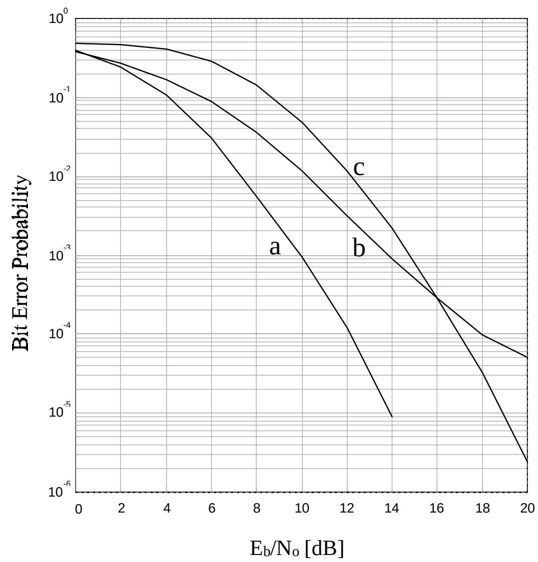

Figure 16: Bit error ratio for a) rate ½ coding with QPSK, b) rate ¾ coding with QPSK and c) rate ½ coding with 16-QAM. Results are obtained by simulating 4000 independent Rayleigh fading channels with an exponentially decaying power delay profile. Delay spread is 100 ns.

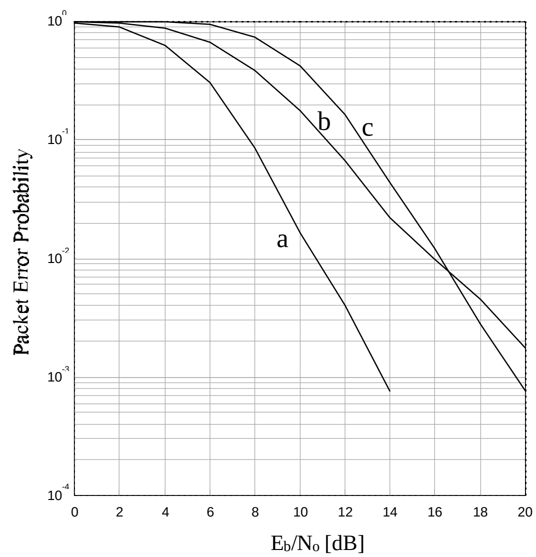

Figure 17: Packet error ratio for a) rate ½ coding with QPSK, b) rate ¾ coding with QPSK and c) rate ½ coding with 16-QAM. Packet size is 1024 bits.

6 CONCLUSIONS

OFDM is an efficient way to provide high data rates in wireless channels. Compared to single carrier modulation with equalizers, OFDM has the following advantages:

- Processing load is significantly lower; this advantage grows as bit rates and/or delay spreads increase
- Feasibility of OFDM is already proven by existing chips [6]
- Spectral efficiency of OFDM is better compared to GMSK
- No equalizer training is needed, saving extra complexity and training overhead

It was shown in this paper that the ‘traditional disadvantages’ of OFDM, being the peak-to-average power ratio, phase noise and frequency offset, do not cause significant problems for the selected parameters. Further, it was demonstrated that the training overhead can be kept relatively small, and that the overhead is in fact depending on the tolerable delay spread rather than on the number of subcarriers. Finally, some simulation results showed that 16-QAM in combination with a rate ½ code gives better results in terms of E_b/N_o and spectral efficiency than QPSK with rate ¾ coding.

REFERENCES

[1] J. Tellado-Mourelo, E.K.Wesel, J.M.Cioffi, ‘Adaptive DFE for GMSK in Indoor Radio Channels’, IEEE Trans. On Sel. Areas in Comm., vol. 14, no. 3, April 1996, pp. 492-501.

[2] S.W. Wales, ‘Modulation and Equalization Techniques for HIPERLAN’, Proceedings of PIMRC/WCN, The Hague, September 21-23, 1994, pp. 959-963.

[3] Lucent Technologies, ‘Scaleable OFDM Radio Parameters’, IEEE P 802.11-97/92, September 1997.

[4] W. Eberle et al., ‘Design Aspects of an OFDM Based Wireless LAN with Regard to ASIC Integration’, OFDM-Fachgespräch, Braunschweig, Germany, September 16-17, 1997.

[5] U. Reimers, ‘DVB-T: the COFDM-based system for terrestrial television’, Electronics & Communications Engineering Journal, February 1997, pp. 28-32.

[6] http://www.lsilogic.com/plweb-cgi/idoc.pl?176+unix+_free_user_+www.lsilogic.com..80+LSI+LSI+Products+Products++OFDM (alternative is to use the LSI search engine at http://www.lsilogic.com/footer/search.html)

[7] T. Pollet, M. van Bladel and M. Moeneclaey, ‘BER Sensitivity of OFDM Systems to Carrier Frequency Offset and Wiener Phase Noise’, IEEE Trans. On Comm., Vol. 43, No. 2/3/4, February-April 1995, pp. 191-193.

[8] J. Kivinen and P. Vainikainen, ‘Phase Noise in a Direct Sequence Based Channel Sounder’, Proceedings of IEEE PIMRC ’97, Helsinki, September 1-4, 1997, pp. 1115-1119.

[9] M. Pauli and H.P. Kuchenbecker, ‘Minimization of the Intermodulation Distortion of a Nonlinearly Amplified OFDM Signal’, Wireless Personal Communications 4, Kluwer Academic Publishers, 1997.

[10] C. Rapp, ‘Effects of HPA-Nonlinearity on a 4-DPSK/OFDM Signal for a Digital Sound Broadcasting System’, Proceedings of the Second European Conference on Satellite Communications, Liège, Belgium, October 22-24, 1991, pp.179-184.