

Multicarrier COFDM scheme in high bitrate radio local area networks

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Abstract: After a short description of the principles of a multicarrier COFDM (Coded Orthogonal Frequency Division Multiplex) scheme some simulation results concerning bit/packet error rates gained with three different characteristic multipath channels are presented and discussed. Aspects of COFDM design principles and complexity of the modulation and demodulation part of the system, as well as effects of the non-linearity of amplifiers will be considered.

I. INTRODUCTION

The wideband radio channel is known to be frequency selective, which leads to intersymbol interferences in the time domain and to partially wide and deep notches in the signal transfer function in the frequency domain. Proceeding on that assumption the aim is to develop and design suitable modulation/demodulation schemes for high bit rate data transmission. To increase the data rate in serial transmission one method is to decrease the symbol period. A second way is the use of bandwidth efficient multilevel modulation schemes. Both solutions show serious drawbacks in a multipath environment. As symbol period is getting shorter than the root mean square delay spread of the radio channel the intersymbol interference is getting a problem. Equalisation in this case is complicated, complex and expensive. But the other way, where the number of bits per symbol is increased by using higher level modulation schemes decreases the distance of the symbols in the signal space and hence the negative influence of noise is getting more and more disastrous and leads to intolerable performance degradation of a communication system.

A different way to increase the symbol period in order to combat with ISI coming from multipath propagation is the multicarrier approach, where the given system bandwidth is divided into an appropriate number of subbands each of which is commissioned with a low data rate, corresponding to a long symbol period, and the sum of the low data rates correspond to the required (demanded) gross bit rates of the communication system. With harmless channels both approaches (parallel and serial methods) are equivalent.

II. PRINCIPLES OF COFDM

In recent time multicarrier transmission schemes such as COFDM have been proposed for radio LANs [1] and also for digital audio broadcast systems [2]. With multicarrier transmission schemes the nominal frequency band is splitted up into a suitable high number of subcarriers each modulated with a low data rate. The spacing of the band is managed by fast Fourier transformation (FFT and IFFT) processes in the baseband. In principle COFDM signals are easy to separate in the receiver because of their orthogonal quality. One of the main advantages of COFDM is its very flexible working range in behalf of rms delay and Doppler shift. Given an assessed maximum multipath delay τ [sec], the symbol time of the system should be chosen by a rule of thumb to be $T \geq 4\tau$. On the other hand the requirements of a maximum Dopplershift ν [Hz] which has to be mastered by the system, call for a guard time $T \leq 0.03 \cdot 1/\nu$. Therefore the relation to define a proper T is $4\tau \leq T \leq 0.03 \cdot 1/\nu$. This feature permits a certain flexibility in tailoring the system (tradeoff between max. possible delay and max. possible Dopplershift) to fit a particular requirement.

When assuming a conventional linear modulation scheme, the complex envelope of the burst waveforms at the modulator output, with respect to the selected carrier, can be given [3] by:

$$\sum_{n=0}^{N-1} c_n x(t - nT) \quad (1)$$

where $x(t)$ is the modulation pulse, and every complex coefficient c_n ($n = 0, 1, \dots, N-1$) is taken from a finite alphabet, at rate $1/T$, according to data at the modulator input. With standard QAM (Quadrature Amplitude Modulation) schemes, for example c_n is equal to $a_n + jb_n$, where $a_n, b_n = \pm 1, \pm 3, \dots, \pm (2^m - 1)$. Each modulation "symbol" is defined by an (a_n, b_n) pair, hence T is the symbol duration and $1/T$ symbols/s is the gross symbol rate. An interesting alternative to the classical modulation approach is to transmit burst waveforms whose complex envelopes can be written as

$$\sum_{n=0}^{N-1} c_n x(t) e^{j2\pi n F t} \quad (2)$$

with the same sequence $\{c_n\}$ and appropriately selected $x(t)$ and F . This means to use the frequency domain just in the same manner as the classical modulations employ the time domain. In fact, the Fourier transform of Eq.2 is

$$\sum_{n=0}^{N-1} c_n X(f - nF) \quad (3)$$

which can be directly compared to Eq.1 with f , F and $X(f)$ replacing t , T and $x(t)$, respectively where $X(f)$ is defined as the "Fourier transform of $x(t)$ ". In practice, this idea is used with a rectangular pulse $x(t)$ having a duration $T = 1/F$. Since each c_n is assigned to a particular subcarrier and the condition:

$$\int_{-\infty}^{+\infty} X(f - nF) X^*(f - mF) df = \int_{-\infty}^{+\infty} x(t) e^{j2\pi n F t} x^*(t) e^{-j2\pi m F t} dt = 0 \quad (4)$$

for $m \neq n$, is met with the above-mentioned rectangular pulse, this signal design method has been called OFDM (Orthogonal Frequency Division Multiplexing) [4]. In this case, a gross "symbol" rate of $1/T = F$ (i.e., corresponding to the subcarrier spacing) is usually defined, since there is a single pulse with duration T in the whole burst.

A. Multicarrier system design

When designing a multicarrier COFDM scheme for a high performance radio local area network, the following assumptions have been made: Nominal frequency band will be 25 MHz in the 5.1 GHz frequency range. The modulation on each subcarrier is assumed to be 4DPSK where a spectral efficiency of 1.6 bit/s/Hz is achieved, so the maximum gross bit rate will be 40 Mbit/s (ether bit rate). The COFDM efficiency is assumed to be $\eta = 0.8$ or -0.97 dB. Following a rule of thumb, the guard time should be at least more than 2 times the rms delay spread. As shown in Fig.1 the symbol period time T_s is per definition equivalent to the sum of FFT-window time t_{FFT} plus guard time Δ . From it follows that the symbol period in this case will be $T_s = 2400$ ns and therefore the FFT window time $t_{FFT} = 1920$ ns. The reciprocal of the window time is equal to the intercarrier distance F , which is therefore $1/1920\text{nsec} = 520.8$ kHz. With a nominal bandwidth of 25 MHz the number of subchannels will be 48. The dimension of the FFT field should be a power of 2, large enough to enable a hermitian symmetry on the signal to be processed. In that case 2^6 will be suffi-

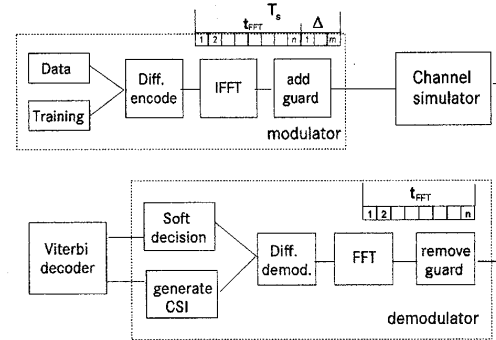


Fig. 1 COFDM simulation setup

cient. The data rate per carrier is equal to 0.833 Mbit/s (cross bit rate divided by the number of carriers). In a further version (Case II) the value of the guard time has been halved. Following the same design rules, the number of carriers is also halved whereas the bit rate per carrier is now 1.6 Mbit/s. Table 1 shows the assumptions as well as the derived COFDM parameters for the simulation in case I and II..

TABLE 1
COFDM PARAMETERS

Parameter	Symbol [unit]	Case I	Case II
Gross data rate	R [Mbit/s]	40	40
Nominal bandwidth	B_n [MHz]	25	25
Symbol period	T [nsec]	2400	1200
Guard interval	Δ [nsec]	480	240
FFT window time	F^{-1} [nsec]	1920	960
Inter-carrier distance	F [kHz]	521	1041
OFDM efficiency	η [dB]	-0.97	-0.97
Number of carriers	N_c	48	24
Number of FFT points	N_F	64	32
Modulation on Single carrier	M-(D)PSK	4	4
Data rate per carrier	R_c [Mbit/s]	0.833	1.6
Useful bits per packet	L_u	256	256
Bits per packet (incl FEC+Tail)	L_p	528	528

B. Effect of amplifier non-linearity

Multicarrier schemes belong to the category of so called linear or non-constant envelope schemes. In order to avoid or at least to reduce the non-linear distortion of the signal causing out-of-band and in-band interference a linear or at least a linearised amplifier must be used

when the signal is amplified in an HPA prior to transmission. The disadvantage is that linear amplifiers are inherently inefficient. Linear amplifiers at 5.1 GHz in the state of the art techniques are expensive, somewhat complex and operate with power efficiencies of around 5-10% compared to 30-50% efficiency for simple non-linear amplification [5]. This fact results in excessive power consumption of the modem. For desktop computers this is a minor problem but for battery powered lap-top computers it is not bearable. Another problem related to the signal envelope is the relative high crest factor of the scheme. As the number of carriers with essentially random data modulated on them is increased, the signal envelope tends to a Rayleigh distribution, that means that there is a small but finite possibility of multiple vector aligning and hence the peak to average power ratio may be very high. This fact is responsible for the high valued out-of-band radiation, which is normally intolerable. By sophisticated measures the dynamic of the envelope of the OFDM-signal may be reduced in such a way, that the following HPA is modulated only in its linearized range, so that any additional out-of-band radiation is avoided.

The maximum transmit power normally is specified in spectral regulations by the peak envelope power (PEP). To remain within the specified limits on the PEP the average power will be low, reducing the system range.

III SIMULATION

The transmission of signals through a mobile radio communication channel is known to be affected by multipath propagation and Doppler shift. Therefore an essential prerequisite with the judgement of qualified modulation/ demodulation schemes by simulation is the thorough knowledge of the characteristics of the broadband radio propagation channel and its adequate transformation into a suitable model.

Only if the software simulation package includes a proper model of the propagation channel, it will be able to produce characteristic performance figures and more knowledge about reliability, efficiency and bit error rate performance of the designed transmission system.

A. Channel model

Because of the randomly time-variant characteristics, correct and complete representation of the mobile radio channel is a very difficult task. Usually the interference processes and noise mechanisms are extremely complicated so that the analytical description of them is not possible in practice. Published channel parameters as well as the outputs from experimental propagation measurements serve to build up a characteristic multipath radio propagation channel model

Commonly modelling the multipath radio channel is performed by the classical tapped delay line approach, which has been thoroughly investigated by Arredondo [6] and many other scientists. In the tapped delay line channel model differently delayed copies of the input signal are individually multiplied with filtered Gaussian noise and are afterwards added; (known as Rayleigh channel). In that case, the channel behaviour is determined by the characteristics of the noise shaping filters. The drawback of this method is, that the exact influences of the filter parameters on the characteristics of the modelled channel are not easy to perceive. So it is often very cumbersome and requires extensive considerations to determine the adequate filter parameters for a given WSSUS channel. Another disadvantage is that there may exist a lot of special shapes that can only be realised by means of high-order filters of high numerical precision, which requires significant computation time in a software simulation.

A novel technique of channel simulation with the aid of Monte Carlo Method is in many respects superior to the tapped delay line channel simulation approach, because this simulation method is suited to cope with statistical, ray tracing and even hybrid channel data.

B. Monte Carlo Method

In document [7] A. Müller suggested the Monte Carlo method as an attractive approach to multipath channel simulation. This method is more general and in many respects superior to the tapped delay line approach, because its parameters are easily configured and the simulation requires less computer operations than the comparable tapped delay line models. The N-th order Monte Carlo model is entirely determined by N triplets $(\tau_n, \nu_n, \alpha_n)$, describing delay, Doppler shift and phase of path n. Thus the channel modelling problem has been reduced to the much simpler problem of determining the parameters τ_n, ν_n and α_n corresponding to the realistic distributions of these random variables, where $0 < n < N$, with N typically in the order of several hundreds. All paths which are used in the Monte Carlo simulation may be illustrated in a momentum scattering diagramme where each path is characterised by its delay (in nsec) and its Doppler shift (in Hz) relative to the carrier frequency. The phases of each path are randomly chosen out of the interval $[0, 2\pi]$. The maximal Doppler shift in this case was 250 Hz which corresponds to a 15 m/s velocity of the receiver and to a 5 GHz carrier frequency.

C. Simulation results

Fig. 1 shows a block diagramme of the COFDM simulation setup. The simulation has been made in three consecutive steps which consume different amounts of computer time. First the modulator output signal has been

computed and the sample values written to a working file. This process takes nearly no time for computation and has to be done only once. In a second step the multipath channel with a fixed S/N and the demodulation procedure is simulated, which takes about 0.5 h per simulation point (per fixed S/N). The last step is the simulation of the error correction procedure with the aid of the Vitebi-algorithm which takes about 1-1.5 h per simulation point. In each simulation run about 1 million bits are transmitted and computed, which corresponds to about 2000 packets of a length of 528 bits.

Fig. 2 shows the exponential power delay profiles of the multipath channels, with three values of rms-delay spread (50, 100, 150 nsec). In that representation the power density in dB of the exponential profiles is drawn versus the delay, so the curves are straight lines in that diagramme. Fig. 3 shows the COFDM bit error probability in the case of AWGN without FEC together with the theoretical curve of the DPSK scheme. The curves are shown to demonstrate the correctness of the COFDM modulation and demodulation modules of the simulation tool.

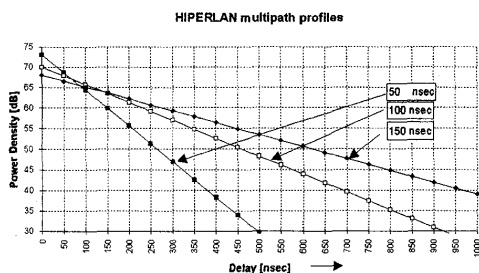


Fig. 2 HIPERLAN profiles

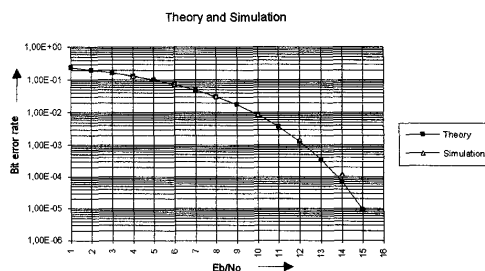


Fig. 3 COFDM bit error probability (AWGN channel)

Fig. 4 is a first result from simulation of a COFDM bit error probability in a multipath channel with exponential power delay profiles. As expected the 150 nsec profile without error correction shows worst results in the higher region of $\gamma = \text{S/N-ratio}$. In a typical

operation range of γ between 12 and 18 dB there is no visible difference within the three bit error rate curves. That means that the additional interferences caused by paths with long delays are small compared to the AWGN influence. This is because of the proper chosen

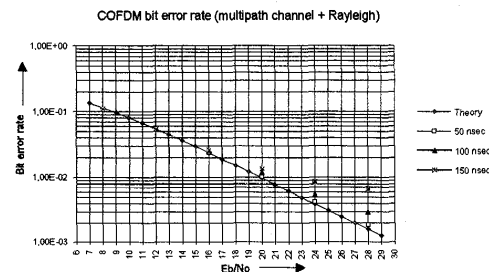


Fig. 4 COFDM bit error probability (multipath channel)

guard time of the COFDM scheme, which in this case is large when compared to the rms-delay value. (rms=150 nsec, $\Delta = 480$ nsec). The splitting-up of the curves begins at a S/N which is greater than 18 dB's. For comparison the theoretical Rayleigh-curve is also shown in Fig. 4.

Fig. 5, here the simulated COFDM packet error rate is shown for three different exponential power delay profiles after FEC. As the picture shows, the 50 nsec rms delay profile has worst results. This "inversion" of performance compared to the previous Fig. 4 is to explain by the benefits of the inherent frequency diversity of the COFDM scheme.

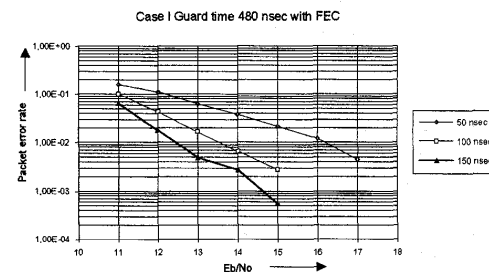


Fig. 5 Simulated COFDM packet error rate

In single carrier schemes the bit error probability gets worsen and worsen with increasing delay spread. In COFDM schemes the reverse is true due to the inherent frequency diversity. The rms delay spread is in case of equalisation an adequate criteria to decide whether a radio channel is severe to handle or not. In schemes with inherent frequency diversity the coherence bandwidth, the minimum frequency separation required for independent fading characteristics, which is defined

as the reciprocal of the rms-delay spread ($W_c=1/2\pi\sigma$) is a proper measure to assess in which degree the subchannels are statistically independent from each other. With 150 nsec rms delay the coherence bandwidth is about 1 MHz and it can be seen that the correlation of the fading characteristics between adjacent subchannels is not so strong as in the case of a 50 nsec delay spread, where the coherence bandwidth is 3 MHz. Similar results have been reported by van Driest [8]. The profits are visible as long as the relevant paths of a profile lie within the guard time of a COFDM-symbol. The 100 and 150 nsec curves in Fig. 5 end with the S/N=15 dB value because there was no error stated in that simulation run (no error in 10^6 bit).

Fig.6 shows the relations when the guard time was halved to $\Delta=240$ nsec. In this case the twofold maximum rms delay value lies outside the guard interval with the result, that this curve shows no longer best performance as shown in Fig. 5. The curves are registered versus the signal-to-noise ratio. With 4DPSK, which is used in each subchannel as modulation method, the correspondent E_b/N_0 values lie typically 3 dB to the left in that diagramme. As with COFDM not the full

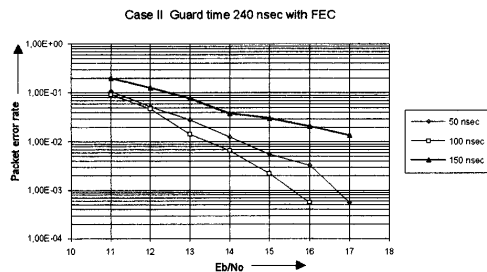


Fig. 6 Simulated COFDM packet error rate (tg=240 nsec)

bitlength is used but only a part of it, the 3 dB have to be reduced by the efficiency of the system, which is $\eta = -0.97$ dB. That means that the corresponding E_b/N_0 values lie about 2 dB to the left compared to the S/N-curves shown in the graphs. Table II shows a table of used coding parameters.

TABLE II
CODING PARAMETERS

Coding type	convolutional
Code rate	1/2
Memory	6
Polynomials	133, 171 (oct.)
Decoder	Viterbi
Decision	soft
Channel state	2 bit CSI
Trellis states	64

Time interleaving	convolutional
Modulus	32
Spread	60
Delay	$60 \times 32 \times 31 = 59520$ bit

The maximum Doppler frequency value corresponds to a car velocity of 15 m/sec. The wavelength at 5 GHz is $\lambda = 60$ mm and the time to pass through that distance with a 15 m/sec speed takes 4 msec. Within 4 msec the number of transmitted bits is 1.6×10^5 (40 Mbit/s), so a 10^6 bit transmission corresponds to a distance of 6.25λ i.e. during the simulation of one point the $10\lambda \times 10\lambda$ area, where the WSSUS assumption is valid, is not left.

IV. CONCLUSION

Non-constant envelope modulation schemes require some degree of linearity in power amplification because in case of non-achievement of that demand the phenomenon of inter-modulation will efficiently block adjacent channels and therefore reduce system capacity. There are several advantages of COFDM schemes, the most important being the excellent performance in time dispersive multipath channels. This arises because, firstly the data rate on each carrier is low so that the signal is not affected by time dispersion and secondly, the scheme has frequency diversity which can be exploited with the application of coding to combat frequency selective fading. Generally multicarrier systems have excellent potential but several problems of RF-stage implementation have not been solved. Techniques for reducing the envelope variation of COFDM to reduce the peak to mean envelope power ratio and consequently to reduce the linearity problem have to be investigated.

V. REFERENCES

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