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Design of OFDM Systems for Frequency-Selective and Time-Variant Channels

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Abstract--We show that frequency-selectivity and time-variance represent adverse constraints for the design of wireless OFDM systems. This observation is important since OFDM is a promising candidate for the realisation of broadband wireless systems, which by definition transmit over frequency-selective and time-variant channels. In this contribution, we present a convenient design criterion for the OFDM symbol duration. The criterion is relating the two basic channel parameters, the maximum excess delay time and the maximum Doppler frequency, with two design parameters reflecting the bandwidth efficiency and the robustness of the system. The introduction of the latter two parameters allows the efficient design of OFDM systems by providing information on the system reserves. Furthermore, we show how the use of multiple antenna arrays exploiting spatial diversity can be incorporated in the system design.

Index Terms--OFDM, system design, broadband wireless channels, frequency-selective and time-variant fading

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is a suitable modulation technique for broadband systems. One of the first commercial systems utilizing OFDM was ADSL, a standard for high data-rate communication over twisted-pair wires. With DAB (digital audio broadcast) OFDM was used for the first time for wireless transmission. The decision to use OFDM for the digital video broadcasting standard DVB-T followed at the beginnings of the nineties. This system is a good example for a state-of-the-art broadband wireless system operating in up to 8MHz wide transmission channel. The latest and maybe most prominent system employing the OFDM modulation technique is the WLAN standard IEEE 802.11a.

All these systems have been tailored in a different fashion according to the respective system constraints. Especially the time-variance of the transmission channel was treated in different ways. In DAB it was proposed to use differential encoding to mitigate this effect, which obviously limits the use of higher order modulations. DVB-T and the WLAN system were designed for stationary and portable use, meaning low mobility only. Clearly, a WLAN system operating in the lower GHz bands will not experience a high degree of mobility due to the relatively low speeds involved. For DVB-T on the other hand, although originally not designed to sustain a high degree of mobility, it was felt later that such a feature offers a wide range of new and interesting applications, like e.g. for trains, cars or even airplanes. The same holds true for future OFDM-based systems. Such systems are likely to be deployed in higher frequency bands, e.g. between 20 and 70GHz. Since the Doppler frequency is proportional to the carrier frequency, a high degree of time-variance has to be expected. We here propose a design procedure that takes into account both the frequency-selective as well as the time-variant nature of the transmission channel. Furthermore, in order to mitigate the effect of the latter we propose the use of multiple antenna arrays and show how this can be incorporated in the system design.

The document is organized as follows: In section II we review the statistical characterisation of mobile transmission channels and present appropriate parameters describing the frequency-selective and time-variant behaviour. In section III we present and illustrate the proposed system design approach for such channels. We start with a review of design issues for time-invariant channels, introduce the system design parameters and finally present a design criterion for OFDM symbol duration. The criterion relates the system design parameters with channel parameters. In section IV the use of multiple antennas exploiting spatial diversity is discussed. Design examples are provided in the latter two sections in order to illustrate the design approach. A summary and conclusions are provided in section V.

II. CHARACTERISATION OF THE MOBILE TRANSMISSION CHANNEL

In what follows we consider a frequency-selective and time-variant transmission channel $h(f,t)$.

A. *Wide-Sense Stationary Uncorrelated Scattering*

Although not being exhaustive for all classes of channels, describing the statistics of wireless channel by means of first and second order moments gives already good insight on their behaviour [1]. Channels are said to be *wide-sense stationary* if the first order moments are independent of time, and the second order moments only depend on the time difference Δt. Similarly, channels that fulfill the same requirement with respect to the frequency variable are said to exhibit *uncorrelated scattering*. Finally, channels that combine both characteristics are known as *wide-sense stationary uncorrelated scattering* (WSSUS) channels. WSSUS channels build an important class of wireless channels that are useful for channel modelling. The autocorrelation function of channels with WSSUS characteristic only depends on the time difference Δt and the frequency difference Δf

$$\mathrm{E}\{h(f,t)h^*(f+\Delta f,t+\Delta t)\} = R_h(\Delta f,\Delta t) \qquad \textbf{(0.1)}$$

By applying a double Fourier transform we obtain the channel description in the time-delay Doppler domain. The corresponding autocorrelation function is referred to by *scattering function*

$$S(\tau,\nu) = \int\int R_h(\Delta f,\Delta t)e^{j2\pi\Delta f\tau}e^{-j2\pi\Delta t\nu}\, d\Delta f\, d\Delta t \qquad \textbf{(0.2)}$$

The scattering function is an important system function since it provides insight on the energy distribution along the time-delay variable τ and the Doppler frequency ν.

B. Time-Variations

The so-called *mean Doppler power spectral density* provides a description of the integral effect of the time-variant nature of the channel

$$P(\nu) = \int_{-\infty}^{\infty} S(\tau,\nu)d\tau \qquad \textbf{(0.3)}$$

The most commonly used Doppler power spectral density is the so-called *Jakes* spectrum, which is based on the assumption of uniformly distributed incident waves

$$P(\nu) = \begin{cases} \dfrac{1}{\pi\nu_{\max}}\dfrac{1}{\sqrt{1-(\nu/\nu_{\max})^2}} & \text{if } |\nu| < \nu_{\max} \\ 0 & \text{otherwise} \end{cases} \qquad \textbf{(0.4)}$$

where $\nu_{\max}$ is the maximum Doppler frequency. Two extreme cases of Doppler spectral densities are the uniform and the two-paths models, which have been studied in [3]. The Doppler power spectral density for these two models writes

$$P(\nu) = \begin{cases} 0.5/\nu_{\max} & \text{if } |\nu| < \nu_{\max} \\ 0 & \text{otherwise} \end{cases} \qquad \textbf{(0.5)}$$

and

$$P(\nu) = 0.5\cdot[\delta(\nu+\nu_{\max})+\delta(\nu-\nu_{\max})], \qquad \textbf{(0.6)}$$

respectively. The *mean time correlation function* $R(\Delta t)$ describes the time-variant behaviour of the channel in the time domain. $R(\Delta t)$ is obtained as the Fourier transform of $P(\nu)$ and consequently we find $R(\Delta t) = R_h(0,\Delta t)$. Based on the time correlation function an important parameter, namely the coherence time t_{coh} can be derived. We define the coherence time as the time difference for which the mean time correlation function attains 50% of its zero-lag value. Note that the coherence time t_{coh} is not only a convenient but also an important parameter providing insight in the time-variant behaviour of the channel.

The coherence time for the above-mentioned Doppler spectral densities can be evaluated numerically and the mathematical expressions are summarized in the following table.

Table 1 - Coherence time for different Doppler spectra

	Jakes	*Uniform*	*Two-Path*
t_{coh}	$0.242/\nu_{\max}$	$0.6/\nu_{\max}$	$0.16/\nu_{\max}$

We observe that the coherence time is inversely proportional to the maximum Doppler frequency. Following this observation and examining the values in Table 1, a simple rule-of-thumb can be used if the actual Doppler spectral density is unknown

$$t_{coh} \approx 0.25/\nu_{\max} \qquad \textbf{(0.7)}$$

C. Frequency-Selectivity

The so-called *power-delay profile* describes the average frequency-selective behaviour of the channel

$$P(\tau) = \int_{-\infty}^{\infty} P(\tau,\nu)d\nu \qquad \textbf{(0.8)}$$

Wireless channels most often exhibit an exponential decay $P(\tau) \propto e^{-\tau/\tau_0}$ [1], where τ_0 is the decay constant. Parameters describing the power-delay profile are the first and second order moments, also known as mean time $\bar{\tau}$ and delay spread σ_τ. It is easy to show that for exponential decaying power-delay profiles both parameters equal to the decay constant $\bar{\tau} = \sigma_\tau = \tau_0$.

The *mean frequency correlation function* $R(\Delta f)$ describes the frequency-selective behaviour of the channel in the frequency domain. Following similar reasoning as for the time correlation function we find $R(\Delta f) = R_h(\Delta f,0)$. We define the coherence bandwidth B_{coh} as the frequency where frequency correlation function attains 50% of its zero-lag value. It can be shown that for exponentially decaying power delay profiles the coherence bandwidth is given by

$$B_{coh} = 1.1/\tau_{\max} \approx 1/\tau_{\max} \qquad \textbf{(0.9)}$$

where $\tau_{\max}$ is the maximum excess time delay, which we define as the time delay where the power delay profile attains -30dB of its zero-lag value. For exponential power delay profiles $\tau_{\max}$ is given by $\tau_{\max} = 6.9\tau_0$.

Note that the coherence bandwidth B_{coh} and consequently also the maximum excess delay $\tau_{\max}$ are important parameters that describe the frequency-selective behaviour of the channel.

III. OFDM System Design Aspects

In this section we first review the design of OFDM systems operating in stationary channels, then illustrate issues related to the time-variant nature of wireless channels and finally develop the OFDM system design approach for frequency-selective and time-variant channels.

A. OFDM Design for Stationary Channels

The OFDM modulation technique divides the total available bandwidth B into N_{mod} subchannels. Each individual subchannel occupies the bandwidth Δf_c and we exhibit $\Delta f_c << B_{coh}$. Therefore, the individual subchannels can be considered to show frequency-flat transmission characteristic, which in turn greatly simplifies channel equalisation. The actual OFDM modulation is performed by means of inverse discrete Fourier transform (IDFT). For implementational simplicity fast Fourier transform (FFT) algorithms are used, which consequently puts constraints on the DFT-size N. For this reason the actual number of subchannels is likely to exceed the number of modulated carriers $N \geq N_{\text{mod}}$. The duration of an OFDM symbol is determined, however, by the time-domain sequence and equals to $T_S = NT$, with T being the sampling period. Due to the properties of the DFT, T_S is related to the carrier separation Δf_c by $T_S = 1/\Delta f_c$. Consequently, by using the approximation in (0.9) the condition to obtain frequency-flat subchannels translates to

$$T_S >> \tau_{\max} \qquad \textbf{(0.10)}$$

In order to eliminate interference due to the excess delay components, a guard interval of length $T_G \geq \tau_{\max}$ is sent prior to each OFDM symbol. Clearly, this reduces the bandwidth efficiency. Let $G = T_G / T$ be the number of guard samples, then the spectral efficiency for a constellation size of C is given by

$$\eta = \frac{N_{\text{mod}}}{N+G} \cdot \frac{\log_2 C}{BT} \qquad \textbf{(0.11)}$$

We introduce the bandwidth coefficient $\alpha_B = N_{\text{mod}} \Delta f_c / B$ that relates the actual modulated bandwidth to the total available bandwidth. Note that α_B is an important system design parameter that determines the implementational complexity of the channel selection filters at the input of the receiver (IF or baseband). High-performance systems exhibit bandwidth coefficients close to one ($\alpha_B = 0.95$ for DVB-T) whereas low-cost systems operate with smaller values: α_B equals to 0.8125 for HiperLAN2. Using $\Delta f_c = 1/T_S$ and the definition for the bandwidth coefficient, the spectral efficiency can be expressed as

$$\eta = \frac{\alpha_B}{1+\zeta_f} \cdot \log_2 C \qquad \textbf{(0.12)}$$

In (0.12) we introduced the system design parameter ζ_f, which is given by $\zeta_f = G/N = T_G/T_S$. We deduce that ζ_f has to be small in order to obtain a high spectral efficiency. Consequently, by setting $T_G = \tau_{\max}$ we obtain a design criterion for the OFDM symbol duration that guarantees a high spectral efficiency

$$T_S = \tau_{\max} / \zeta_f \qquad \zeta_f \leq 0.25 \qquad \textbf{(0.13)}$$

Note that by limiting ζ_f to be smaller than 0.25 we also satisfy (0.10) thus assuring frequency-flat subchannel.

In order to achieve a specified spectral efficiency, the system designer has the freedom to control the implementational complexity. Besides the constellation size C, the spectral efficiency is given by the factor $\alpha_B/(1+\zeta_f)$. To obtain a certain value for this factor, a tradeoff between ζ_f and α_B is possible as depicted in Figure 1. We observe that within certain limits the bandwidth coefficient α_B can be kept small by reducing the value for ζ_f, thus maintaining a low implementational complexity.

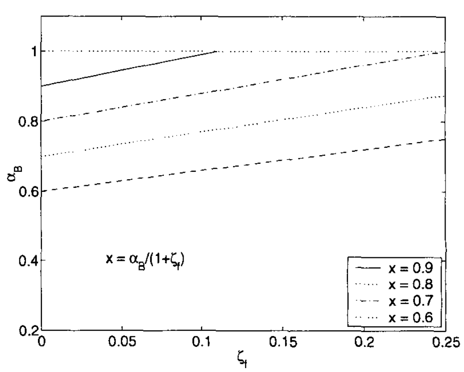

Figure 1 – Tradeoff between ζ_f and α_B

B. Issues Related to Time-Variations

If the channel transfer function is not constant by varies during the OFDM symbol period T_S the subcarriers are no longer orthogonal. This results in so-called intercarrier interference (ICI) at the output of the OFDM demodulator. The signal power of this additional interference depends on the Doppler power spectral density as well as the ratio of the maximum Doppler frequency $\nu_{\max}$ to the OFDM carrier spacing Δf_c. We recall that the OFDM carrier spacing is given as the inverse of the OFDM symbol duration and introduce the second design parameter $\zeta_t = \nu_{\max} T_S$ that reflects the extent of time-variations. In [3] tight upper bounds for the ICI power were derived and it was shown that the influence of the Doppler spectral density is only marginal. This can be seen from Table 2 where the upper bounds are summarized for the Doppler spectral densities presented in section II.B.

Table 2 - Exact upper bounds for ICI power, from [3]

Jakes	*Uniform*	*Two-Path*
$P_{ICI} < \frac{\pi^2}{6} \cdot \zeta_t^2$	$P_{ICI} < \frac{\pi^2}{9} \cdot \zeta_t^2$	$P_{ICI} < \frac{\pi^2}{3} \cdot \zeta_t^2$

In [3] it was also shown that a *universal* upper bound exists which is useful in cases where knowledge of the actual Doppler power spectral density is absent

$$P_{ICI} < \frac{\pi^2}{3} \cdot \zeta_t^2 \qquad (0.14)$$

If the ICI power exceeds the noise power $P_{ICI} > P_{th}$ the appearance of an error floor in the BER curves is obvious. In order to guarantee a quality of service with a particular BER performance it is necessary to assure $P_{ICI} << P_{th}$. This is especially important for the quality of channel estimation, which deteriorates with increased ICI power [4]. In practice the uncoded symbol error rate (SER) can be used to determine the required signal to noise ratio (SNR) and hence limits on the ICI power. The SER for transmission of QAM modulated symbols over time-invariant channels and additive white Gaussian noise (AWGN) is given by

$$P_e(\gamma) = 4a \cdot Q(\sqrt{b\gamma}) \cdot (1 - a \cdot Q(\sqrt{b\gamma})) \qquad (0.15)$$

The constants a and b are summarized for commonly used constellation alphabets in the following table.

Table 3 - Parameters a and b for different modulation sizes

	QPSK	*16-QAM*	*64-QAM*
a	$1/2$	$3/4$	$7/8$
b	1	$1/5$	$1/21$

The mean SER $P_e(\bar{\gamma})$ for time-variant channels can be calculated as

$$P_e(\bar{\gamma}) = \int_0^\infty p(\gamma) P_e(\gamma) d\gamma \qquad (0.16)$$

where $p(\gamma)$ is the probability density function of the amplitude distribution and $\bar{\gamma}$ is the mean signal to noise ratio (SNR). In our case (0.16) is applicable to each OFDM carrier since ICI is white and Gaussian [7]. In order to keep the deterioration due to the time-variations small, we solicit $P_{ICI} = 0.1 \cdot P_{th}$. Based on this constraint sensible values for the design parameter ζ_t can be obtained through numerical integration of (0.16). By assuming Rayleigh distributed amplitude variations by using the universal upper bound of (0.14) we find the representative values presented in Table 4.

According to the definition of the design parameter ζ_t, the OFDM symbol duration has to be decreased as we want to employ higher order modulations or when a higher reliability is required.

Table 4 - Sensible settings for ζ_t for different modulation sizes

	QPSK	*16-QAM*	*64-QAM*
$SER \le 10^{-2}$	$\zeta_t \le 0.017$	$\zeta_t \le 0.007$	$\zeta_t \le 0.003$
$SER \le 10^{-3}$	$\zeta_t \le 0.005$	$\zeta_t \le 0.002$	$\zeta_t \le 0.001$

C. Design of OFDM systems for Frequency-Selective and Time-Variant Channels

From III.A we recall that a long OFDM symbol duration T_S is desired in order to achieve a high spectral efficiency. On the other hand, in III.B we showed that the ratio of the OFDM carrier spacing Δf_c and the maximum Doppler frequency should be kept small. Hence a short OFDM symbol duration T_S is desired in order to keep the impact of the time-variations small. Therefore, time-variance and frequency-selectivity put adverse design constraints on the OFDM symbol duration. This observation can be expressed mathematically by using the definitions of the system design parameters ζ_f and ζ_t

$$\frac{\tau_{max}}{\zeta_f} \le T_S \le \frac{\zeta_t}{\nu_{max}} \qquad (0.17)$$

Equation (0.17) provides a design criterion for the OFDM symbol duration by combining the channel parameters with the system design parameters. For a specific application the channel parameters τ_{max} and ν_{max} are known or can be deduced from the system specifications. Accordingly, the system design parameters are determined by system specification and appropriate values for ζ_f and ζ_t can be chosen as shown in III.A and by using Table 4, respectively.

Note that the OFDM symbol duration is an important system parameter since it determines other system parameters like e.g. the DFT size or the number of modulated carriers N_{mod}.

D. Example - Short-Range Communications

As an example we consider the design of a short-range communication system operating in the 60GHz bands. We assume that the delay spread equals to $\sigma_\tau = 50 = \tau_0$ resulting in a maximum excess delay time of approximately $\tau_{max} = 350\text{ns}$. The system should support low mobility users with a maximum speed of 5m/s. For the given frequency band this is equal to a maximum Doppler frequency of $\nu_{max} = 1\text{kHz}$. We target a low-performance system for the consumer market and therefore solicit a low-cost solution. Assuming high bandwidth availability in the 60GHz band, we consequently opt for QPSK modulation. Furthermore we choose $\zeta_f = 0.1$ and $\alpha_B = 0.8$ in order to kept implementational costs low. Similarly, with respect to time-variations we request $\zeta_t \le 0.005$, corresponding to transmission with a guaranteed SER of at least 10^{-3}. Given these settings, we find that the OFDM symbol duration should be in the tight range $3.5\mu s \le T_S \le 5\mu s$. We assume that the total available bandwidth is 10MHz and by using the lower limit for T_S, we find the number of modulated carriers $N_{mod} = 0.8 * 10 \cdot 10^6 * 3.5 \cdot 10^{-6} = 28$. A suitable DFT-size for e.g. a radix-2 algorithm is $N = 32$. We choose the number of guard samples to be equal to $G = 4$, which results in a modification of the design parameter $\zeta_f = 0.125$. Thus we obtain decreased bandwidth efficiency with respect to the original requirements: $\eta = 0.71 \cdot \log_2 C$ with $C = 2$ reflects QPSK modulation (for comparison, the bandwidth efficiency for the WLAN standard *IEEE 802.11a* is $\eta = 0.65 \cdot \log_2 C$). The sampling period evaluates to $T = 3.5/32\mu s$ corresponding to a clock rate of 9.1428MHz (which is, by coincidence, equal to that used for the digital video broadcast system DVB-T).

IV. Multiple Antenna Systems

The above example illustrates that the quality of service requirements can no longer be met for higher order modulation schemes, since the upper limit for the OFDM symbol duration would be smaller than its lower limit. For systems that are already operating at their limits, we propose the use of multiple antenna systems in order to improve system performance, e.g. in terms of increased throughput and/or reliability. In what follows we focus on multiple antenna systems exploiting spatial diversity on either the transmitter or receiver.

A. Spatial Diversity

Multiple antennas can be used to reduce the time-variations of wireless communication channels. In [5] it was shown that for reception with M antennas, the distribution of the received power migrates from an exponential (Rayleigh fading) to a χ^2 distribution with $2M$ degrees of freedom. For illustration purposes the pdfs of the received signal energy r^2 are depicted in Figure 2. Plots are normalized by the variance of the received signal of a single antenna σ in antennas reception is two-fold: first, the mean signal is

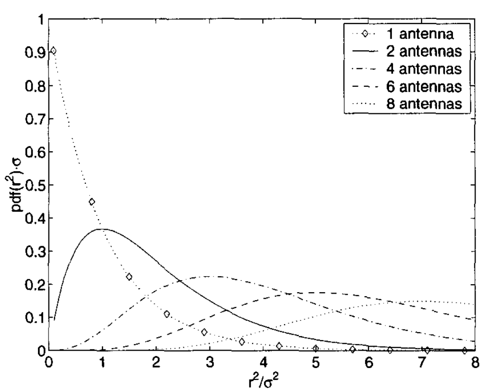

Figure 2 – Normalized envelope distribution for reception with multiple antennas

increased by a factor M and second, the variance of the amplitude variations is decreased. These two phenomena are referred to by the terms *antenna* and *diversity gains*, respectively [6].

Due to these effects, the required SNR for a specific quality of service can be reduced with respect to single antenna reception. Consequently, owing to the constraint $P_{ICI} = 0.1 \cdot P_{th}$ larger ICI powers can be tolerated. In the same way as previously in III.B, appropriate design values for the normalized Doppler frequency ζ_t can be obtained by using the χ^2 distribution for the numerical evaluation of (0.16). Suitable settings that assure $SER \le 10^{-2}$ are summarized in Table 5.

Table 5 - Settings for ζ_t and $SER \le 10^{-2}$

	QPSK	*16-QAM*	*64-QAM*
$M = 1$	$\zeta_t \le 0.017$	$\zeta_t \le 0.007$	$\zeta_t \le 0.003$
$M = 2$	$\zeta_t \le 0.069$	$\zeta_t \le 0.026$	$\zeta_t \le 0.012$
$M = 3$	$\zeta_t \le 0.110$	$\zeta_t \le 0.043$	$\zeta_t \le 0.020$
$M = 4$	$\zeta_t \le 0.146$	$\zeta_t \le 0.058$	$\zeta_t \le 0.024$

Note that the reduction of the time-variations can be seen as reducing the effective maximum Doppler frequency. Consequently, ζ_t can be increased with increasing number of antennas. With two antennas for example, ζ_t is approximately four-times higher compared to the single antenna system. Therefore, a four times longer OFDM symbol duration can now be tolerated. In principle, the number of antennas can be used to achieve any desired level of robustness against time-variations. It can be seen from Table 5, however, that the increase of ζ_t saturates as the number of receive antennas increases. From this viewpoint, the first additional antenna is the most advantageous, whereas the benefit coming from each further antenna will have less significant impact.

In order to illustrate the benefit of using multiple antennas simulation results are presented in Figure 3. Results for spatial diversity reception with carrier-based maximum ratio combining (MRC) [7] are depicted. We present BER results for 16-QAM modulated symbols and the maximum Doppler frequency ν_{max} equals to 5% of the carrier spacing Δf_c. It can be seen that, due to the errorfloor, transmission at a BER of 10^{-2} cannot be guaranteed for single antenna receivers. With two receive antennas the errorfloor is reduced sufficiently and the desired BER can be obtained with a mean SNR of approximately 16dB. The use of three antennas enables the same quality of service with an SNR advantage 7dB.

In principle, the same performance gains can be achieved through transmit diversity, i.e. the deployment of multiple antennas at the transmitter. Transmit diversity is advan-

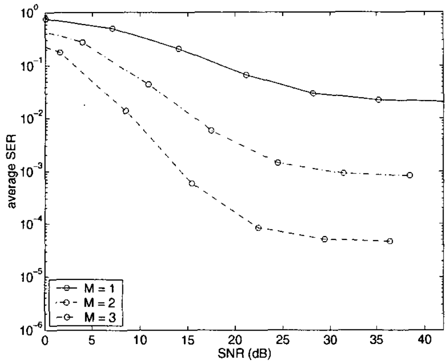

Figure 3 – Spatial receive diversity – results for maximum ratio combining

tageous for systems with asymmetric data traffic where high data-rates have to be delivered to the mobile terminals only. For such scenarios it is economically beneficial to install multiple antenna arrays at the base stations rather than at the far more numerous mobile terminals. However, a penalty in training-overhead has to be paid since the receiver has to estimate the channels associated with all transmit antennas.

B. Example – Wireless Local Loop (WLL)

We envision designing a WLL system operating at 28GHz. For maximum throughput 64-QAM symbol modulation is to be used. In order to obtain easy network deployment, no specific cell sites or directional transmit antennas (e.g. horn antennas) will be used. Therefore, non line-of-sight transmission has to be anticipated and the maximum excess delay time is expected to be $\tau_{\max} = 2\mu s$. Time-variations are mostly originating from street traffic with speeds of 10m/s for urban and rural areas. This results in a maximum Doppler frequency of $\nu_{\max} = 933\text{Hz}$. For single antenna systems with 64-QAM modulation, we find $\zeta_t = 0.003$ in Table 5. Therefore, the lower limit for the OFDM symbol duration for reception with one antenna evaluates to $3.21\mu s$. Obviously this is too small since T_S is in the order of $\tau_{\max}$ and the resulting spectral efficiency is intolerable. For this reason we choose to use three receive antennas resulting in an upper limit for the OFDM symbol durations of $T_S = 21.49\mu s$. We choose $\zeta_f = 0.125$ in order to yield a good spectral efficiency and since we are looking at a high performance system, a bandwidth coefficient $\alpha_B = 0.95$ seems adequate. With these settings, the lower limit for the OFDM symbol duration evaluates to $T_S = 16\mu s$. Now for $T_S = 16\mu s$ and by assuming a total available bandwidth of $B = 20\text{MHz}$, the number of modulated carriers evaluates to $N_{\text{mod}} = 304$ and we choose a DFT-size of $N = 512$. Owing to $T = T_S / N$, we obtain a sampling rate of 32MHz. If this sampling rate is too high with respect to the total bandwidth, one might reduce the signal bandwidth. For example, we could envision a system operating with a total available bandwidth of 15MHz. Then, the number of modulated carriers would be 228. Consequently a 256-point DFT operating at 16MHz could be employed. Obviously, the throughput would be reduced by ¾ with respect to the 20MHz system. To avoid this we could also choose the upper limit for the OFDM symbol duration. For $T_S = 22\mu s$ we require 418 modulated carriers and with $N = 512$ the sampling rate would be approximately 23MHz. Note that the system now operates at the upper bound for T_S and is therefore more vulnerable to exceedingly high time-variations.

V. CONCLUSIONS

The design of OFDM systems that operate in frequency-selective and time-variant channels was investigated. We showed that frequency-selectivity, which can be characterized by the maximum excess delay time, sets a lower limit for the OFDM symbol duration. On the other hand, the time-variant nature, which is characterized through the maximum Doppler frequency, dictates an upper limit for the OFDM symbol duration.

In this contribution we introduced two design parameters that reflect the bandwidth efficiency and the robustness of the OFDM system. We showed how these parameters are related to the system specifications and provided sensible settings to be used for system design. These settings together with the channel specifications allow the determination of the OFDM symbol duration. Based on this OFDM system parameter further parameters follow, like the DFT-size, the number of modulated carriers and the sampling rate.

Furthermore we showed the potential of multiple antenna systems and outlined how the use of multiple antennas can be incorporated in the system design.

VI. ACKNOWLEDGEMENTS

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