Parameter Optimization for OFDM Systems in Doubly-Selective Fading Channels with Line-of-Sight Components

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***Abstract*— In wireless communications, the channel can often be modelled as a time-selective frequency-selective fading channel with line-of-sight (LOS) component. Although orthogonal frequency division multiplexing (OFDM) systems are developed to cope with the frequency selectivity of the channel, they will suffer from interference caused by the time-varying character of the channel, present in particular in the LOS component, when the number of carriers increases. A proper selection of the system parameters, such as the number of carriers and the length of the cyclic prefix, can alleviate the effect of the doubly-selective channel on the system performance. In this paper, we investigate analytically the effect of the aforementioned system parameters on the performance, and determine the optimum values of the system parameters. Further, we study the effect of deviations from the optimum parameters on the system performance.**

***Index Terms*— Multicarrier systems, fading channel, parameter optimization.**

I. INTRODUCTION

MULTICARRIER (MC) techniques have received considerable attention for high data rate communications [1]-[5] as they combine bandwidth efficiency with an immunity to the channel dispersiveness and time-selectivity. The selection of the system parameters, i.e. the number of carriers and the cyclic prefix length, however is crucial. When the cyclic prefix is too short, the MC technique will suffer from interference caused by the channel dispersion. As the use of a cyclic prefix reduces the power efficiency and the effective throughput, the cyclic prefix must be kept small as compared to the number of carriers; hence the cyclic prefix length may not be too long and the number of carriers not too small. However, increasing the number of carriers, and hence the MC symbol duration, will make the MC system more sensitive to the time-selectivity of the channel.

In this paper, we study the effect of the system parameters on the performance of OFDM in a doubly selective fading channel with LOS component, and determine the optimum number of carriers and cyclic prefix length. This paper extends the results obtained in [6], where the optimum parameters have been determined for fading channels without LOS component. Further, in this paper, we investigate the effect of deviations from the optimum parameters on the system performance, which has not been considered in [6].

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II. SYSTEM DESCRIPTION

The data stream to be transmitted is serial-to-parallel converted into N lower rate data streams $\{a_{i,n}\}$, where $a_{i,n}$ is the ith symbol of the nth data stream, $n = 0, \ldots, N-1$. The data symbols $a_{i,n}$ are transmitted on the different carriers of the OFDM system using an inverse fast Fourier transform (IFFT). To cope with the presence of a dispersive channel, a cyclic prefix of length ν samples is inserted, resulting in the time-domain samples $s_i(m) = \sqrt{\frac{E_s}{N+\nu}} \sum_{n=0}^{N-1} a_{i,n} e^{j2\pi \frac{nm}{N}}, m = -\nu, \ldots, N-1$, where the data symbols are assumed to be statistically independent and have unit average energy, i.e. $E[a_{i,n} a^*_{i',n'}] = \delta_{i,i'} \delta_{n,n'}$, and the transmitted energy per symbol is equal to E_s.

The time-domain samples $s_i(m)$ are transmitted over a doubly-selective fading channel. The channel is modelled as a tapped delay line with channel coefficients $h(k;\ell)$, where ℓ is the time index and k is the index of the multipath delay. We assume that the channel consists of a line-of-sight (LOS) component $h_{LOS}(k;\ell)$ and a zero-mean multipath fading contribution $h_{MP}(k;\ell)$, i.e. $h(k;\ell) = h_{LOS}(k;\ell) + h_{MP}(k;\ell)$. The coefficients of the LOS component are given by $h_{LOS}(k;\ell) = \alpha e^{-j\phi(\ell)} \delta(k)$, where the quasi-static amplitude α is assumed to be constant over a number of OFDM symbols, and $\phi(\ell)$ is a time-varying phase that depends on the time-selectivity of the channel. The channel coefficients $h_{MP}(k;\ell)$ of the multipath component are assumed to be zero-mean Gaussian distributed according to the wide-sense stationary uncorrelated scattering (WSSUS) model of Bello [7] with autocorrelation function $E[h_{MP}(k_1;\ell_1) h^*_{MP}(k_2;\ell_2)] = \delta(k_1 - k_2) R_{MP}(k_1; \ell_1 - \ell_2)$. Let us define P_{LOS} and P_{MP} as the energy contained in the impulse response corresponding to the LOS and multipath component, respectively:

$$\begin{aligned} P_{LOS} &= |\alpha|^2 \\ P_{MP} &= \sum_{k=-\infty}^{+\infty} R_{MP}(k;0). \end{aligned} \qquad (1)$$

We define the ratio $\kappa = P_{LOS}/P_{MP}$ as the ratio of the power contained in the impulse response of the LOS component to the one of the multipath component. Without loss of generality, we can assume that $P_{LOS} + P_{MP} = 1$.[1] The output of the channel is disturbed by complex-valued additive white Gaussian noise $w_N(m)$, with zero mean and power spectral density N_0.

Without loss of generality, we concentrate on the detection of the data symbols transmitted during the OFDM block with time index $i = 0$. From the $N+\nu$ samples belonging to the

[1] Under this assumption, it follows that $P_{LOS} = \frac{\kappa}{1+\kappa}$ and $P_{MP} = \frac{1}{1+\kappa}$.

considered OFDM block, the receiver removes the ν samples corresponding to the cyclic prefix. The remaining N samples are then applied to a fast Fourier transform (FFT). The nth output of the FFT can be written as, assuming that all carriers are modulated,

$$z_n = \sqrt{E_s}\sqrt{\frac{N}{N+\nu}}\sum_{i=-\infty}^{+\infty}\sum_{n'=0}^{N-1} a_{i,n'}\gamma_i(n,n') + W_n \quad (2)$$

where

$$\gamma_i(n,n') = \frac{1}{N}\sum_{m=-\nu}^{N-1}\sum_{k=0}^{N-1} h(k-m-i(N+\nu);k)e^{-j2\pi\frac{kn-mn'}{N}} \quad (3)$$

and the noise contribution W_n is the FFT of the noise samples $w_N(k)$. In the presence of a fading channel, in general $\gamma_i(n,n') \neq 0$ for $n' \neq n$ and/or $i \neq 0$. Hence, the fading channel will cause interference. The power at the nth FFT output can be decomposed in $P(n) = \frac{N}{N+\nu}E_s(P_U(n) + P_{ICI}(n) + P_{ISI}(n)) + N_0$. It can easily be verified that the useful power $P_U(n) = E[|\gamma_0(n,n)|^2]$ and interference powers $P_{ICI}(n) = \sum_{n'=0,n'\neq n}^{N-1} E[|\gamma_0(n,n')|^2]$ and $P_{ISI}(n) = \sum_{i=-\infty,i\neq 0}^{+\infty}\sum_{n'=0}^{N-1} E[|\gamma_i(n,n')|^2]$ are independent of the carrier index n. Hence, in the following, we drop the carrier index.

As a performance measure, we use the signal to interference and noise ratio (SINR) at the output of the FFT, defined as the ratio of the useful power to the sum of the powers of the interference and noise.

$$SINR = \frac{\frac{N}{N+\nu}E_s P_U}{\frac{N}{N+\nu}E_s(P_{ICI} + P_{ISI}) + N_0} \quad (4)$$

In the presence of a fading channel, the SINR is degraded as compared to the SINR in the case of an AWGN channel and in the absence of a cyclic prefix, which equals $SINR_{AWGN} = E_s/N_0$ (i.e. in an AWGN channel, $P_{ICI} = P_{ISI} = 0$, and because of the normalization of the powers $P_{LOS} + P_{MP} = 1$, it follows $P_U = 1$). The degradation (in dB) of the SINR, caused by the fading channel and the presence of a cyclic prefix, is given by $Deg = -10\log\left(\frac{N}{N+\nu}P_U\right) + 10\log\left(1 + \frac{E_s}{N_0}\frac{N}{N+\nu}(P_{ICI} + P_{ISI})\right)$.

To simplify the expressions for P_U, P_{ICI} and P_{ISI}, we define similarly as in [6] the following two-dimensional weight function

$$w(q;r) = \frac{1}{N}\begin{cases} N-|r| & 0 \le q \le \nu \\ & 0 \le |r| \le N \\ N-q+\nu-|r| & \nu \le q \le N+\nu \\ & 0 \le |r| \le N-q+\nu \\ N+q-|r| & -N \le q \le 0 \\ & 0 \le |r| \le N+q \\ 0 & \text{otherwise.} \end{cases} \quad (5)$$

Using (5) and following the same method as in [6], we find after tedious manipulations the following simple expressions:

$$P_U = |\alpha|^2|\Phi(0)|^2 + \frac{1}{N}\sum_{k=-\infty}^{+\infty}\sum_{\ell=-\infty}^{+\infty} w(k;\ell)R_{MP}(k;\ell) \quad (6)$$

$$P_{ICI} = |\alpha|^2 + \sum_{k=-\infty}^{+\infty} w(k;0)R_{MP}(k;0) - P_U \quad (7)$$

$$P_{ISI} = \sum_{k=-\infty}^{+\infty}(1-w(k;0))R_{MP}(k;0) \quad (8)$$

where $\Phi(0)$ is the discrete Fourier transform of length N of the LOS phase $\phi(\ell)$ at frequency 0. For example, assuming that the phase rotation of the LOS component is caused by a Doppler shift f_D, i.e. $\phi(\ell) = \phi(0) - 2\pi f_D \ell T$, with $1/(NT)$ the carrier spacing, $|\Phi(0)|^2$ reduces to

$$|\Phi(0)|^2 = \left|\frac{\sin(\pi N f_D T)}{N\sin(\pi f_D T)}\right|^2. \quad (9)$$

Note that, because of the assumption that $P_{LOS} + P_{MP} = 1$, the sum of the powers of the useful component, the ICI and ISI equals one: $P_U + P_{ICI} + P_{ISI} = 1$. In the absence of a LOS component (i.e. $\alpha = 0$), the equations (6)–(8) reduce to the expressions in [6].

As can be observed in (6)–(8), the useful power and the ICI power depend on both the LOS component and the multipath component, whereas the ISI power only depends on the multipath component. This can easily be explained, as the channel $h_{LOS}(k;\ell)$ consists of a single tap and hence will not cause interference between successive OFDM blocks. Following a similar reasoning as in [6], it can be shown that an optimum set of system parameters (N_{opt}, ν_{opt}) can be found. Further, for a given shape of the LOS and multipath channel impulse responses, the ratio κ of the powers of the LOS contribution to the multipath contribution will have an effect on the optimal parameters. For increasing κ, i.e. for channels with stronger LOS component, the optimum parameters will move to $N_{opt} = 1$ and $\nu_{opt} = 0$, which are the optimum system parameters when there is no multipath component.

III. Limiting Cases

As observed in (6), the useful power (and thus also the ICI power in (7)) corresponding to the multipath component consists of a double sum, which requires a high computational complexity, especially when N is large. To reduce the computational complexity, we have shown in [6] for multipath Rayleigh fading channels that the useful power and interference powers can be well approximated by considering two limiting cases, i.e. the time-flat and frequency-flat fading Rayleigh fading channels derived from the considered Rayleigh fading channel, and take the sum of the powers of the limiting cases to obtain the total useful power and interference powers.

The same reasoning can be followed for the channel considered in this paper. As we observe in (6)–(8), the powers of the useful component, the ICI and ISI consist of the sum of the powers originating from the LOS component and the multipath component. Further, taking into account the results from [6], the powers corresponding to the multipath component can be further split into the sum of the powers corresponding to the limiting cases of a time-flat channel with autocorrelation function $R_{MP,TF}(k;\ell) = R_{MP}(k;0)$ and a frequency-flat channel with autocorrelation function

$R_{MP,FF}(k;\ell) = \tilde{R}_{MP,FF}(\ell)\delta(k)$, where $\tilde{R}_{MP,FF}(\ell) = \sum_{k=-\infty}^{+\infty} R_{MP}(k;\ell)$. Hence, we approximate the powers in (6)–(8) using the following decomposition: $P_X = P_{X,LOS} + P_{X,MP,TF} + P_{X,MP,FF}$, with $X = U, ICI$ and ISI. The powers corresponding to the LOS component are given by: $P_{U,LOS} = |\alpha|^2|\Phi(0)|^2$, $P_{ICI,LOS} = |\alpha|^2(1 - |\Phi(0)|^2)$ and $P_{ISI,LOS} = 0$. The useful powers corresponding to the two limiting cases of the multipath component are given by [6]:

$$P_{U,MP,TF} = \sum_{k=-\infty}^{+\infty} w^2(k;0) R_{MP}(k;0) \qquad (10)$$

$$P_{U,MP,FF} = \frac{1}{N} \sum_{\ell=-N}^{N} \left(1 - \frac{|\ell|}{N}\right) \tilde{R}_{MP,FF}(\ell) \qquad (11)$$

and the derivation of the corresponding interference powers follow straightforward from (6)–(8) and the definition of $R_{MP,TF}(k;\ell)$ and $R_{MP,FF}(k;\ell)$. In the next section, we will show the validity of the used approximations by means of numerical evaluation.

IV. Numerical Results

For the numerical evaluation of the obtained analytical expressions, we consider a channel bandwidth $B = 1$ MHz and a center frequency of $f_c = 1$ GHz. Hence, the duration of a sample equals $T = 1\mu$s. For the LOS component of the channel, we assume that the channel has a fixed amplitude α and a phase rotation $\phi(\ell)$ caused by a Doppler shift f_D, i.e. $\phi(\ell) = \phi(0) - 2\pi f_D \ell T$. Assuming that the receiver is moving at a speed of $v = 135$ km/hr, the Doppler shift is given by $f_D = (v/c)f_c = 125$ Hz, where c is the velocity of light. For the multipath component of the channel, we use an impulse response with a delay spread of 10μs. The coherence time of the multipath component is obtained using the rule of thumb $T_0 = 0.5/f_D$ [9]-[10], where f_D is the Doppler frequency of the LOS component, resulting in the coherence time $T_0 = 4$ ms. We assume that the autocorrelation function for the multipath component has an exponentially decaying multipath intensity profile and a Gaussian time correlation profile

$$R_{MP}(k;\ell) = C \exp\left(-\frac{k}{\tau_0}\right) \exp\left(-\frac{\ell^2}{2\sigma_0^2}\right) \qquad k \geq 0, -\infty \leq \ell \leq +\infty \qquad (12)$$

Defining the delay spread as the time at which the multipath intensity profile falls 10 dB below the level of the strongest component, the parameter τ_0 is found to be about five samples. Further, we fix the coherence time T_0 to twice the duration of the spreading of the Gaussian time correlation profile, i.e. $\sigma_0 = 2000$ samples. The constants α and C are normalization constants.

In Fig. 1, we show the total interference power $P_{ICI} + P_{ISI}$, obtained with (7) and (8), together with the interference power of the limiting cases and the sum of the interference powers of the limiting cases as function of N, for $\kappa = 0$ dB and $\nu = 10$. As expected, the interference power corresponding to the LOS component increases with increasing N. For $Nf_DT \ll 1$, i.e. when the Doppler shift is sufficiently smaller than the carrier spacing, the approximation $|\frac{\sin(\pi N f_D T)}{N \sin(\pi f_D T)}|^2 \approx 1 - \frac{1}{3}(\pi N f_D T)^2$ can be used, such that the interference power for the LOS component, given by $|\alpha|^2(1-|\Phi(0)|^2)$, increases quadratically with N. In [6], it was shown that for a multipath fading channel the total interference power for the time-flat channel with autocorrelation function $R_{MP,TF}(k;\ell) = R_{MP}(k;0)$ is proportional to $1/N$. Further, in the case of the frequency-flat limiting case of the multipath channel, when the autocorrelation function $R_{MP,FF}(k;\ell) = \tilde{R}_{MP,FF}(\ell)\delta(k)$ can be approximated[2] by $\tilde{R}_{MP,FF}(\ell) \approx \tilde{R}_{MP,FF}(0)(1 - \beta\ell^2)$ for $\ell \ll T_0$, it can easily be shown that the total interference power increases with N^2. This can be observed in Fig. 1. Figure 1 also shows the total interference power of the multipath component (indicated in the figure as MP, doubly selective), obtained with the contributions of $R_{MP}(k;\ell)$ in (7) and (8). This interference power is well approximated by the sum of the interference powers (indicated in the figure as MP, TF+FF) corresponding to the time-flat and frequency-flat limiting cases [6]. Further, we show in Fig. 1 the sum of the interference powers of the LOS component and the limiting cases of the multipath component (indicated as LOS + MP,FF + MP,TF), and the total interference power obtained with (7) and (8). As we observe, the sum of the interference powers of the limiting cases well approximate the total interference power of the actual doubly selective channel with LOS component. For small values of N, the interference power of the actual channel converges to the interference power of the time-flat limit of the multipath component, whereas for large N, it converges to the sum of the interference powers of the LOS component and the frequency-flat limit of the multipath component (indicated as LOS + MP,FF). As the computational complexity for the interference powers of the limiting cases is much lower than that for the actual channel, the excellent agreement demonstrates the practical relevance of the method of the limiting cases to assess the performance.

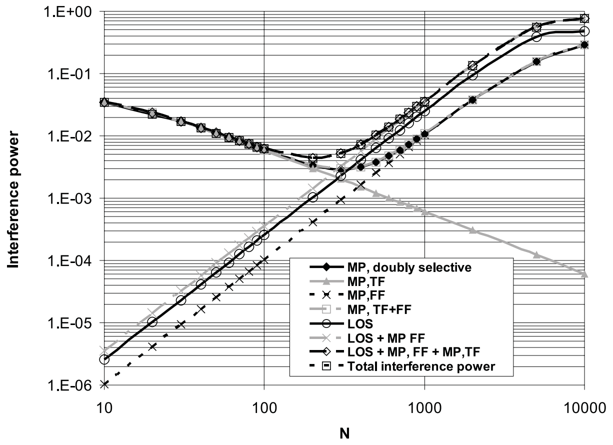

Fig. 1. Interference power as function of FFT length, $\kappa = 0$ dB, $\nu = 10$.

In Fig. 2, the interference power of the actual channel, the limiting cases and the sum of the interference powers of the limiting cases are shown as function of ν, for $\kappa = 0$ dB and $N = 256$. Similarly as in Fig. 1, the sum of the

[2] In most practical channels, this approximation is valid.

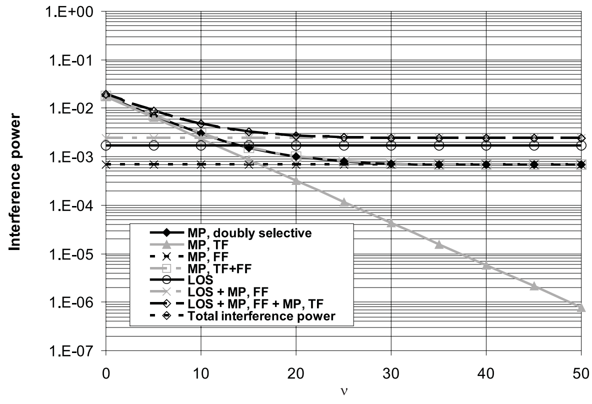

Fig. 2. Interference power as function of cyclic prefix length, $\kappa = 0$ dB, $N = 256$.

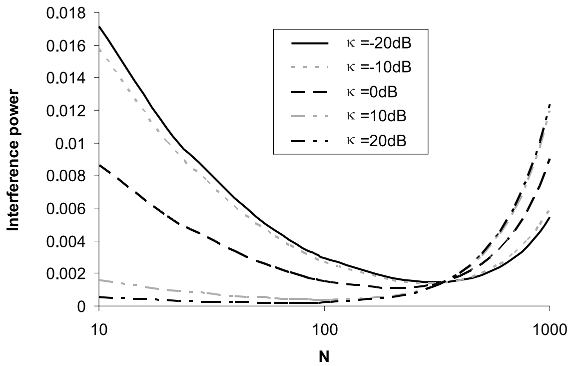

Fig. 3. Influence of N and κ on interference power, $\nu = 10$.

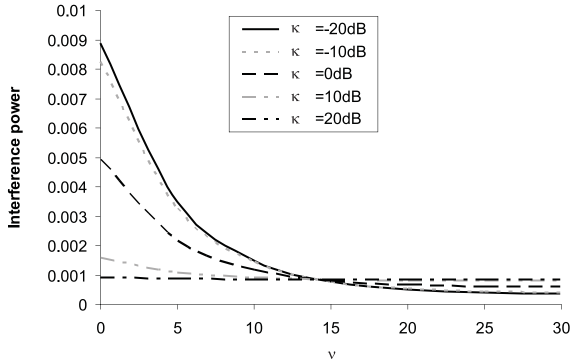

Fig. 4. Influence of ν and κ on interference power, $N = 256$.

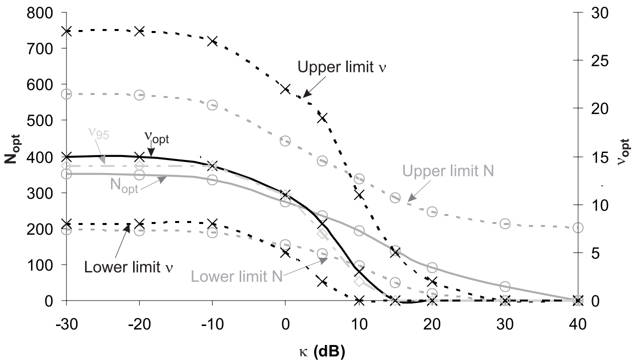

Fig. 5. Optimal N and ν and range for N and ν for which degradation $\leq 0.1dB$ larger than minimum degradation, $E_s/N_0 = 10$ dB.

interference powers of the limiting cases well approximate the interference power of the actual channel. As expected, the interference powers of the frequency-flat limit of the multipath component and the LOS component are independent of the length of the cyclic prefix. The interference power of the time-flat limit of the multipath component is a decreasing function of ν as the effect of the channel dispersion reduces for increasing ν; the slope of the decreasing function will depend on the shape of the multipath intensity profile. For small ν, the interference power of the actual channel converges to the interference power of the time-flat limit of the multipath component, whereas for large ν, the interference power of the actual channel no longer decreases but converges to the sum of the interference powers of the LOS component and the frequency-flat limit of the multipath component; the resulting interference is caused by the time-selectivity of the channel and cannot be combatted by a further increase of the cyclic prefix length.

Figures 3 and 4 show the effect of κ, indicating the relative amount of power contained in the LOS component and the multipath component, on the interference power. As can be observed in Fig. 3, the minimum of the interference power moves to lower values of N when κ increases. This can be explained as follows. When κ increases, the LOS component becomes the dominating component. As in this case the interference is caused by the time-selectivity of the LOS component, the interference will increase when N increases, resulting in a reduction of the optimum value of N. The effect of κ on the dependency of the interference power on ν is shown in Fig. 4. As for large κ, the channel consists of mainly LOS, and the relative importance of the frequency selectivity decreases, the interference power becomes essentially independent of ν, whereas for small values of κ, when the multipath component is dominating, the interference power strongly depends on the value of ν.

These effects can also be observed in Fig. 5, where the optimum value of N and ν that minimize the degradation are shown as function of κ. Similarly as in Fig. 3 we observe that N_{opt} decreases for increasing κ. At the same time, ν_{opt} decreases for increasing κ; as the interference power for large κ becomes essentially independent of ν, this reduction of ν_{opt} will be caused by the power efficiency factor $\frac{N}{N+\nu}$. Further, in Fig. 5, the range over which N (ν) given ν_{opt} (N_{opt}) may vary such that the degradation is smaller than the minimum degradation plus 0.1 dB, is shown (dashed curves). The range over which N or ν may vary is large for small κ and reduces when κ increases. In Fig. 5, we also have shown the value of ν for which $|\alpha|^2 + \sum_{k=0}^{\nu} R_{MP}(k;0) = 0.95$, i.e. 95% of the power of the channel impulse response falls within the guard interval. As can be observed, this value ν_{95} well corresponds to ν_{opt}. Hence, as a rule of thumb, ν_{opt} can be approximated by ν_{95}. Further, defining $\tau_{max} = \nu_{95}T$ as the delay spread of the total channel (LOS+MP), the coherence bandwidth of the

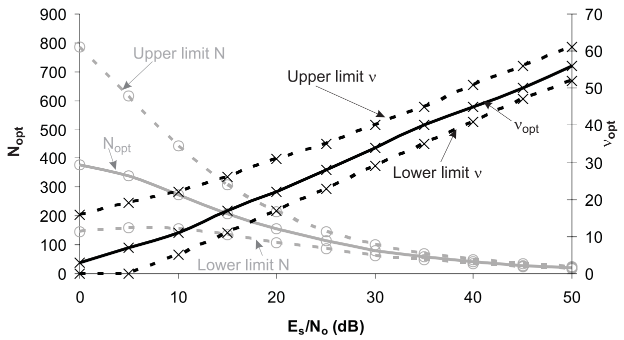

Fig. 6. Optimal N and ν and range for N and ν for which degradation ≤ 0.1 dB larger than minimum degradation, $\kappa = 0$ dB.

channel can be defined as $B_c = 1/(2\pi\tau_{max})$. Note that the carrier spacing $1/N_{opt}T$ is located in the interval determined by the Doppler frequency and the coherence bandwidth: $f_D \ll 1/NT \ll B_c$. Hence, N can be bounded by $2\pi\tau_{max}/T \ll N << 1/(f_D T)$, resulting in $88 \ll N \ll 8000$ for $\kappa = -30$ dB and $0 \ll N \ll 8000$ for $\kappa = 30$ dB.

In Fig. 6, the optimum value of N and ν that minimize the degradation are shown as function of E_s/N_0. As can be observed, N_{opt} decreases and ν_{opt} increases when E_s/N_0 increases. This can be explained as follows. For the optimum values of N and ν, the interference power $P_{ICI} + P_{ISI}$ is small, hence $P_U \approx 1$. For small E_s/N_0, the degradation therefore can be approximated by $Deg \approx -10\log\frac{N}{N+\nu}$; this becomes minimal for large N and small ν. For large E_s/N_0, the degradation can be approximated by $Deg \approx 10\log\frac{E_s}{N_0}P_I$. The interference power caused by the frequency selectivity will reduce for increasing ν while the interference power corresponding to the time selectivity will reduce for decreasing N. Hence, the resulting optimum N will be small while ν_{opt} will be large. This can also be observed in Figs. 3 and 4. Further, Fig. 6 shows the range over which N (ν) given ν_{opt} (N_{opt}) may vary such that the degradation is maximally 0.1 dB larger than the minimum degradation (dashed curves). It follows from the figure that the range over which N may vary decreases for increasing E_s/N_0, whereas the range for ν essentially does not change. Hence, the larger E_s/N_0, the more critical is the selection of the number of carriers N.

V. Conclusions

In this paper, we have investigated the effect of the system parameters, i.e. the cyclic prefix length and the number of carriers, on the performance of an OFDM system in a doubly selective fading channel with LOS component. We have derived analytical expressions for the degradation. Further, we have found simple but accurate approximations for the power of the useful component and the interference, by decomposing the doubly selective fading channel with LOS component into three limiting cases, i.e. a time-flat and a frequency-flat Rayleigh fading channel and a channel with only LOS, and taking the sum of the powers of the three limiting cases as approximation for the powers of the actual channel. Using these analytical expressions, the optimal system parameters (N_{opt}, ν_{opt}) that minimize the degradation, are obtained. Further, the range over which the system parameters N and ν may vary when a small degradation (0.1 dB) as compared to the minimum degradation is allowed has been determined. Moreover, the influence of the channel parameters κ, i.e. the ratio of the powers contained in the LOS component and the multipath component of the channel, and E_s/N_0 on the optimum parameters and deviations from the optimum parameters has been studied. The results can be summarized as follows.

- When κ increases, i.e. when the LOS component is the dominating component, the values of N_{opt} and ν_{opt} decrease. Further, as the range over which the system parameters may vary when a small degradation is allowed as compared to the minimum degradation decreases, the selection of the system parameters N and ν becomes more critical when κ increases.
- For increasing E_s/N_0, N_{opt} decreases whereas ν_{opt} increases. The range for N for which the degradation is less than 0.1 dB larger than the minimum degradation decreases when E_s/N_0 increases, such that the choice of N is more critical than for lower E_s/N_0. On the other hand, the range for ν is essentially independent of E_s/N_0.

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