Maximum Likelihood Decoding of M-ary Orthogonal Modulated Signals for Multi-Carrier Spread-Spectrum Systems

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ABSTRACT

For a *multi-carrier spread-spectrum* (MC-SS) system, the inner spreading can be optimized by applying M-ary orthogonal modulation. In this paper, we investigate the concatenation of *maximum likelihood* (ML) Viterbi decoding with M-ary orthogonal demodulation in a MC-SS system. The system operates over a frequency-selective Rayleigh fading indoor channel in the uplink. We first evaluate the *bit specific* log-likelihood ratio for ML Viterbi decoding and present an estimation of the ratio exploiting the MC technique. Furthermore, the trade-off between channel coding, M-ary orthogonal modulation and simple spreading will be considered by Monte-Carlo simulations. The results are always compared with BPSK performance and they emphasize for the concerned indoor transmission scenario that a moderate bit-error-rate can only be achieved if M-ary orthogonal modulation is employed.
All simulations are related to the European Hiperlan/2 standardization whereas the results are generally valid.

I. INTRODUCTION

Future wireless radio systems need to make efficient use of the frequency spectrum. One technique offering high spectrum efficiency is the multi-carrier-modulation technique *orthogonal frequency division multiplexing* (OFDM). With OFDM, the fading of each subcarrier can be assumed as frequency-nonselective. To achieve frequency diversity, OFDM can be combined with *code division multiple access* (CDMA), where the signal is spread over all subcarriers. This concept was first introduced as OFDM-CDMA or also as *multi-carrier* (MC-)CDMA by [1, 2] and is also called MC-*spread spectrum* (MC-SS).
In the recent few years, most publications dealing with MC-SS have been focused on classical modulation schemes such as BPSK [3, 4] and several *maximum likelihood* (ML) decoding strategies were proposed [5, 6].
Instead of using classical modulation schemes, SS techniques offer the possibility to apply M-ary orthogonal modulation [7]. With this modulation, the inner spreading of the MC-SS system is optimized by spreading in two successive steps: (*i*) spreading by the modulation itself and (*ii*) additional simple replication of the transmitted information into parallel subcarriers to perform SS technique. The optimization is carried out without changing the required bandwidth. Investigations without *forward error correction* (FEC) in [8] indicate a performance gain by applying M-ary orthogonal Walsh modulation in contrast to BPSK.
If we employ FEC, the combination of FEC with M-ary orthogonal Walsh modulation describes a concatenation of two codes. The channel code represents the outer one and the M-ary Walsh modulation can be interpreted as an inner (block-) code with rate $\log_2(M)/M$. For this combination, the derivation of the ML criterion yields a *bit specific* log-likelihood ratio supplied to the Viterbi decoder [9].
The bit-error-rate (BER) for SS systems highly depends on the trade-off between spreading and channel coding. Here, one can interchange the rate of the channel code, the spreading of the Walsh modulation and the simple replication. Besides the trade-off, inherently burst error structures caused by slowly fading channels lead to a significant performance degradation for systems with Viterbi decoding. In case of MC systems, one can combat bursty channels by sufficient frequency interleaving, i.e. one can avoid time-interleaving, which requires large size block-interleaver, especially in case of very slow fading.
The intention of this paper is to examine the concatenation of ML Viterbi decoding and M-ary orthogonal Walsh demodulation for a MC-SS system. The system operates in a frequency-selective and slowly Rayleigh fading indoor channel in the uplink. We first evaluate the *bit specific* log-likelihood ratio for ML Viterbi decoding assuming independent adjacent subcarriers and present an estimation of the ratio exploiting the MC technique. We will discuss the trade-off between channel coding, M-ary orthogonal modulation and simple spreading based on Monte-Carlo simulations in case of sufficient and insufficient frequency interleaving, respectively. Moreover, the results are always contrasted with BPSK performance.
The paper is organized as follows: In Section II, the MC-SS transmission system with FEC and M-ary orthogonal Walsh modulation is presented. The log-likelihood ratios for antipodal and M-ary orthogonal modulated signals are evaluated in section III. In section IV simulation results are shown for the uplink indoor transmission. Finally, section V summarizes the results.

II. TRANSMISSION SYSTEM

A. Transmitter

With SS systems, the spreading codes implemented mainly determine the *multiple access interference* (MAI). For an uplink transmission scenario, we have an asynchronous reception using *Pseudo-Noise* (PN)-sequences, which results in higher MAI. A

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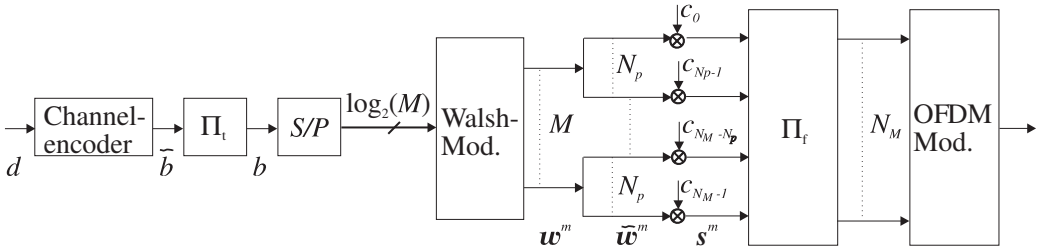

Figure 1: MC-SS transmitter

MC-SS transmitter (mobile station) with channel encoding and Walsh modulation is illustrated in fig. 1. For simplicity, one of J active users is shown and subscripts are omitted.

The data bits $d \in \{0,1\}$, each of duration T_d, are convolutionally encoded with code rate $R_c = K/N$. The input of the encoder is a sequence $\boldsymbol{d}$ of K subsequent data bits and the output is the encoded bit sequence $\tilde{\boldsymbol{b}}$ of N bits, each with duration $T_b = R_c\, T_d$. After block interleaving (Π_t) the encoded bits are serial/parallel converted to groups of $\log_2(M)$ bits each. The Walsh modulation maps the $\log_2(M)$ encoded bits to one corresponding Walsh symbol (vector) $\boldsymbol{w}^m = [w_0^m, w_1^m, \ldots, w_{M-1}^m]^T$, $m \in \{0, \ldots, M-1\}$ including M Walsh chips $w_\mu^m \in \{\pm 1\}$, $\mu = 0, \ldots, M-1$. Each of the M parallel Walsh chips has a duration of $T = \log_2(M)\, T_b$. This modulation can also be interpreted as a block code of rate $\log_2(M)/M$. The set of M orthogonal Walsh symbols describes the $M \times M$ *Hadamard-Matrix* [7, 10], and the Euclidian distance is identical for all possible pairs of symbols and equals

$$\varepsilon = \sum_{\mu=0}^{M-1} |w_\mu^m - w_\mu^n|^2 = 2M \quad \forall \quad m \neq n. \tag{1}$$

To obtain the transmitted symbols, the Walsh chips are replicated into N_p parallel copies where each branch of the parallel stream is multiplied with one chip of the user specific PN-code $c_i \in \{\pm 1/\sqrt{N_M}\}$, $i = 0, \ldots, N_M - 1$. Since the bandwidth for each of the M parallel subchannels decreases for larger values of M, the number of subcarriers $N_M = M \cdot N_p$ increases for the same bandwidth B.

For the j-th user, this yields the vector $\boldsymbol{s}^{m,j}$

$$\boldsymbol{s}^{m,j} = [s_0^{m,j}, \ldots, s_{N_M-1}^{m,j}]^T, \quad s_i^{m,j} = \tilde{w}_i^{m,j} \cdot c_i^j \tag{2}$$

with $\tilde{w}_i^{m,j} = w_\mu^{m,j}$, $\forall\ \mu = \lfloor i/N_p \rfloor$, m, j in front of the frequency interleaver (Π_f). The interleaver scrambles the outputs and passes them onto the OFDM modulator. Finally, OFDM modulation includes the IDFT and inserts the guard interval between adjacent OFDM symbols.

To prevent *intersymbol interference* (ISI), the guard time T_g is chosen such that $T_g \geq \tau_{max}$, where τ_{max} is the maximum delay spread of the channel. Due to the insertion of the guard interval, every subcarrier is affected by only one channel transfer coefficient. For an uplink transmission scenario, the i-th coefficient for every user j is given by $h_i^j = \rho_i^j \cdot \exp(\phi_i^j)$, where ρ_i^j and ϕ_i^j are random amplitudes and phases, respectively. Hence, the channel can be described for each user by an $N_M \times N_M$ diagonal matrix $\boldsymbol{H}^j$.

Applying SS technique, the BER can be reduced by an optimized compound design of channel coding and spreading. Here, the product between bandwidth and bit duration is given by

$$B\, T_d = \underbrace{\frac{M}{\log_2(M)} \cdot \frac{N_p}{R_c}}_{P_G} \cdot \frac{T}{T - T_g}. \tag{3}$$

The first term of (3) describes the spreading of the Walsh modulation, the second one the ratio of simple spreading (N_p) over the code rate R_c of the channel code, whereas the third term takes into account the guard time T_g. For our comparisons, the product $B\, T_d$ remains unchanged due to the exploitation of the inherently overall spreading factor[1] P_G. Decreasing the BER can be achieved by sharing P_G between channel coding, M-ary orthogonal Walsh modulation and simple spreading. The data rate $R = 1/T_d$ is based on binary modulation without channel coding ($R_c = 1$) and the bandwidth B is generally given.

B. Coherent receiver

The MC-SS receiver for M-ary orthogonal Walsh modulation is shown in fig. 2. Paying attention to J active users, the received signal after OFDM demodulation and deinterleaving (Π_f^{-1}) can be written as a sum of vectors $\boldsymbol{r}^{m,j}$

$$\boldsymbol{r} = \sum_{j=0}^{J-1} \boldsymbol{r}^{m,j} + \boldsymbol{n} = \sum_{j=0}^{J-1} \boldsymbol{H}^j \boldsymbol{s}^{m,j} + \boldsymbol{n} \tag{4}$$

with elements $r_i^{m,j} = s_i^{m,j} \cdot h_i^j$, $i = 0, \ldots, N_M - 1$ and $m \in \{0, \ldots, M-1\}$. The vector $\boldsymbol{n}$ represents *Additive White Gaussian Noise* AWGN. After multiplication with the user specific code and equalization, the first part of despreading is obtained by subcorrelating N_p subcarriers. Reception for the user $j = 0$ is assumed (subscripts are omitted). The components of $\boldsymbol{v}$ are given by

$$v_\mu = \sum_{i=\mu N_p}^{(\mu+1)N_p-1} \mathrm{Re}\left\{ c_i e_i r_i^{m,0} + \sum_{j=1}^{J-1} c_i e_i r_i^{m,j} + c_i e_i n_i \right\}, \quad \mu, m \in \{0, \ldots, M-1\} \tag{5}$$

where e_i indicates the equalization coefficient of the i-th subcarrier. We take into account the well-known equalization scheme *maximum ratio combining* (MRC) which shows the best performance for an uplink transmission [8]. If perfectly known channel coefficients are assumed[2], then $e_i = h_i^*$. To enable *maximum*

[1] Being aware of a changed symbol duration T

[2] Channel phase estimation based on decision-directed estimation for the M-ary orthogonal modulation is investigated in [11]

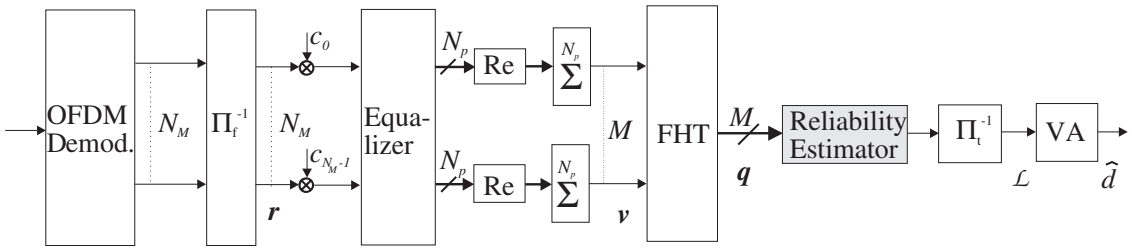

Figure 2: MC-SS receiver

likelihood detection (MLD), the signal is correlated with all possible Walsh symbols $\boldsymbol{w}^m, m \in \{0, \ldots, M-1\}$. The MLD can be realized by the *Fast Hadamard Transform* (FHT), where the output vector $\boldsymbol{q}$ contains M decision variables. In case of MLD, optimum decoding, e.g. of a convolutional code, is performed by the application of a soft input Viterbi algorithm (VA) where channel state information is exploited. Therefore, the reliability estimator exploits the M decision variables to calculate the assigned *bit specific* log-likelihood ratio $\mathcal{L}$ which is fed to the Viterbi channel decoder after deinterleaving (Π_t^{-1}). Afterwards, we obtain the estimated data stream with bits $\hat{d}$.

III. ML DECISION DECODING

In the following, we will derive the *bit specific* log-likelihood criterion for ML decoding of M-ary orthogonally modulated signals. We consider coherent MC-SS demodulation with MRC equalization. The MAI is taken into account as AWGN and independent adjacent fading subcarriers ensured by perfect frequency interleaving (Π_f) are assumed.

A. Antipodal Transmission

Let us first assume binary antipodal transmission on a single carrier system with $b_\nu \in \{-1, +1\}$ denoting the ν-th transmitted encoded bit. Hard decision at the receiver leads to the binary encoded bit $z_\nu \in \{-1, +1\}$. For a memoryless discrete binary symmetric channel the likelihood criterion for ML channel decoding is given by [12]

$$\Phi_{ML}^{BSC} = \sum_\nu b_\nu \underbrace{z_\nu \ln \frac{P_\nu}{1-P_\nu}}_{\triangleq\ \mathcal{L}_\nu} \stackrel{!}{=} \max\,, \tag{6}$$

where $P_\nu \stackrel{\triangle}{=} P(z_\nu = b_\nu)$ denotes the probability for correct decision. The likelihood criterion can efficiently be solved by the Viterbi algorithm (VA) where the log-likelihood ratio $\mathcal{L}_\nu = z_\nu\, L_\nu$ serves as a soft input signal, and, in this context, the absolute value $L_\nu = \ln \frac{P_\nu}{1-P_\nu}$ of the log-likelihood ratio can be interpreted as a reliability indicator for the hard decided code bit z_ν. Due to the fact that $\mathcal{L}_\nu$ involves channel state information the application of eq. (6) results in optimum performance under time-variant channels. For example, if the transmitted signal b_ν suffers from fading h_ν and is disturbed by AWGN, the log-likelihood ratio results in

$$\mathcal{L}_\nu = h_\nu \frac{4E_s}{N_0} r_\nu, \tag{7}$$

where r_ν denotes the received signal after matched-filtering, E_s is the symbol energy and $N_0/2$ indicates the Gaussian noise spectral density [12].
In case of a MC-SS with antipodal transmission, the derivation of the log-likelihood ratio yields an equivalent expression as (7), and several approaches are given for a downlink transmission in [5, 13]. Derivations for the uplink result in approximately the same approaches except of individual channel transfer functions for each specific user.

B. M-ary orthogonal Walsh modulation

Let us now return to the problem of M-ary orthogonal modulation with MRC detection. Here, it is also possible to describe the reliability estimator based on hard symbol decision. Fig. (3)

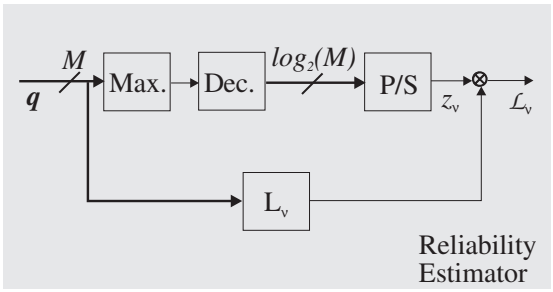

Figure 3: Reliability estimator model

shows the corresponding reliability estimator model. The choice of the maximum value of vector $\boldsymbol{q}$ results in symbol decision. Decoding of the decided symbol to $\log_2(M)$ data bits and parallel/serial conversion leads to the estimated bits z_ν. Subsequently, the hard decided bits are multiplied with the reliability indicator L_ν to form the bit specific soft input signal $\mathcal{L}_\nu$ for the VA.
To derive the bit specific reliability estimator[3] $\mathcal{L}_\nu$, we start with the calculation of the conditional probability that $\boldsymbol{\omega}^n$ is the transmitted symbol under the observation of $\boldsymbol{q}$. Furthermore, one time instant is taken into account (subscripts are omitted). Applying Bayes law results in

$$P(\boldsymbol{\omega}^n|\boldsymbol{q}) \;=\; \frac{P(\boldsymbol{\omega}^n)}{p(\boldsymbol{q})} \cdot p(\boldsymbol{q}|\boldsymbol{\omega}^n)\,. \tag{8}$$

If equiprobable transmitted symbols $\boldsymbol{\omega}^n$ are assumed, we get $P(\boldsymbol{\omega}^n) = 1/M$. Due to the statistical independence of the components of $\boldsymbol{q}$, the conditional probability density $p(\boldsymbol{q}|\boldsymbol{\omega}^n)$ can be

[3] We take pattern from the derivation given in [9], where noncoherent reception in a *direct-sequence* SS system is considered.

written as a product

$$p(\boldsymbol{q}|\boldsymbol{\omega}^n) = r(q_0) \cdot \ldots \cdot R(q_j) \cdot \ldots \cdot r(q_{M-1}), \tag{9}$$

where $r(q_m) = p(q_m|\boldsymbol{\omega}^n \neq \boldsymbol{\omega}^m)$ and $R(q_n) = p(q_n|\boldsymbol{\omega}^n)$. The terms $r(q_m)$ and $R(q_n)$ can be calculated uniquely. In case of the considered coherent MRC detection one obtain a zero mean Gaussian distribution with variance σ_q^2 for the m-th component where the hypothesis $\boldsymbol{\omega}^n$ does not hold, i.e. $m \neq n$:

$$r(q_m) = \frac{1}{\sqrt{2\pi\sigma_q^2}} \exp\left(-\frac{q_m^2}{2\sigma_q^2}\right). \tag{10}$$

In contrast, for the n-th component corresponding to the current hypothesis $\boldsymbol{\omega}^n$ the FHT results in a Gaussian distribution

$$R(q_n) = \frac{1}{\sqrt{2\pi\sigma_q^2}} \exp\left(-\frac{(q_n - E_h)^2}{2\sigma_q^2}\right) \tag{11}$$

with mean $E_h = E\left\{\frac{1}{N_M}\sum_{i=0}^{N_M-1}|h_i|^2\right\}$, where $E\{\cdot\}$ denotes the expected value.
With this, eq. (8) can be rewritten in the form

$$P(\boldsymbol{\omega}^n|\boldsymbol{q}) = \underbrace{\frac{1/M}{p(\boldsymbol{q})}\prod_{m=0}^{M-1} r(q_m)}_{\triangleq\, \alpha} \cdot \frac{R(q_n)}{r(q_n)},$$
$$n \in \{0, \ldots, M-1\}. \tag{12}$$

The unknown factor α can be determined indirectly by the exploitation of the relationship

$$\sum_{n=0}^{M-1} P(\boldsymbol{\omega}^n|\boldsymbol{q}) = \alpha \cdot \sum_{n=0}^{M-1} \frac{R(q_n)}{r(q_n)} = 1 \tag{13}$$

(the union of all elementary events has probability one). The calculation of

$$\frac{R(q_n)}{r(q_n)} = \exp\left(\frac{2\,q_n\,E_h - E_h^2}{2\sigma_q^2}\right) \tag{14}$$

requires the knowledge of the noise variance σ_q^2 and the expected value of the average channel energy over all subcarriers E_h.
These parameters can be estimated as follows:
(*i*) *noise variance estimation*
Assuming that the decision variable with maximum magnitude, denoted by q_l, represents the true hypothesis, we obviously obtain an estimate of the noise power $\hat{\sigma}_q^2$ by the mean over $M-1$ components $q_m, m \neq l$, containing the false hypotheses:

$$\hat{\sigma}_q^2 = \frac{1}{M-1} \cdot \sum_{\substack{m=0 \\ m\neq l}}^{M-1} q_m^2\,. \tag{15}$$

(*ii*) *average channel energy estimation*
If we assume statistically independent fading subcarriers and the same average energy on all subcarriers, one can estimate E_h by the mean over all subcarriers:

$$\hat{E}_h = \frac{1}{N_M}\sum_{i=0}^{N_M-1}|h_i|^2. \tag{16}$$

With (13)-(16), all parameters needed for the calculation of the conditional symbol probability (12) are known. The results can be summarized for all M hypotheses: If the symbols $\boldsymbol{\omega}^n$ are replaced by their corresponding bit patterns $\{b_0, \ldots, b_{\log_2(M)-1}\}$ in antipodal representation, we obtain (e.g. for $M = 16$)

$$\begin{aligned}
P(-1,-1,-1,-1|\boldsymbol{q}) &= \alpha \cdot R(q_0)/r(q_0) \\
P(-1,-1,-1,+1|\boldsymbol{q}) &= \alpha \cdot R(q_1)/r(q_1) \\
&\vdots \\
P(+1,+1,+1,+1|\boldsymbol{q}) &= \alpha \cdot R(q_{M-1})/r(q_{M-1}).
\end{aligned} \tag{17}$$

The application of eq. (6) requires knowledge of the probabilities of individual bit-decisions which can be gained by a columnwise evaluation of (17). Note, at first, that eq. (6) does not depend on the specific choice of the decision for bit z_ν, since

$$z_\nu = +1 \;\rightarrow\; z_\nu \cdot L_\nu = +\ln\frac{P(b_\nu = +1|\boldsymbol{q})}{1 - P(b_\nu = +1|\boldsymbol{q})}$$

and

$$\begin{aligned}
z_\nu = -1 \;\rightarrow\; z_\nu \cdot L_\nu &= -\ln\frac{P(b_\nu = -1|\boldsymbol{q})}{1 - P(b_\nu = -1|\boldsymbol{q})} \\
&= +\ln\frac{P(b_\nu = +1|\boldsymbol{q})}{1 - P(b_\nu = +1|\boldsymbol{q})}
\end{aligned}$$

yield the same expression.
Now consider the calculation of soft information, e.g. for the first bit b_0 of the $\log_2(M)$-bit groups (most left bit-"column" in (17)): Adding all probabilities belonging to '$b_0 = +1$' results in

$$P(b_0 = +1|\boldsymbol{q}) = \sum_{n=M/2}^{M-1} P(\boldsymbol{\omega}^n|\boldsymbol{q}).$$

This leads to the reliability information for the first bit of the Walsh symbol being '$b_0 = +1$'

$$L_{b_0} = \ln\frac{P(b_0 = +1|\boldsymbol{q})}{1 - P(b_0 = +1|\boldsymbol{q})}$$

and serves as a soft input for the Viterbi channel decoder. The other bits $b_1 \ldots b_{\log_2(M)-1}$ are treated equivalently.

IV. SYSTEM COMPARISON

A. System description

Related to the European HIPERLAN/2 standardization, the results followed are given for a transmission over an indoor Rayleigh fading channel. The bandwidth is $B = 25\,\mathrm{MHz}$ in the $5.2\,\mathrm{GHz}$ range. A velocity of $v_0 = 0.5$ m/s results in a very low maximum Doppler frequency $f_{d,max}$ of about $9\,\mathrm{Hz}$. Hence, long symbol

durations T are possible and thus high level M-ary orthogonal modulation can be applied. Furthermore, due to the very low maximum Doppler frequency, the fading on each subcarrier is highly statistically dependent. To avoid bursty errors, sufficient frequency interleaving (Π_f) is required. First, the indoor channel is modeled assuming sufficient frequency interleaving. Subsequently, we take into account an exponential delay profile with a maximum delay spread of $\tau_{max} \approx 350\,\mathrm{ns}$ and a coherent bandwidth of approximately $2.8\,\mathrm{MHz}$. I.e. the number of independent subcarriers is reduced dramatically.

The trade-off between coding and spreading is carried out for an unchanged product $B\,T_d$ (3) and is based on binary modulation without FEC with $N_2 = 64$ subcarriers. The guard time is chosen to be $T_g = 640\,\mathrm{ns}$. This system design results in a data rate of $R = 312.5\,\mathrm{kbit/s}$ for all cases of spreading and coding. A block interleaver (Π_t) with 32 rows and 18 columns is used to separate the channel code and the M-ary Walsh modulation. We employed two convolutional codes of rate $R_c = 1/2$ and $R_c = 1/3$, both with constraint length 7 [10]. For all cases, $J = 8$ active users are taken into account and perfect channel estimation as well as perfect synchronization is assumed for the considered user.

B. Simulation Results

Fig. 4 shows the BER of the M-ary Walsh modulation for $M = 16,\ 32$ by sharing P_G between simple spreading N_p and FEC of rates $R_c = 1,\ 1/2,\ 1/3$; see (3). Independent of the M-ary modulation the results indicate that choosing a higher amount of simple spreading N_p combined with less powerful FEC ($R_c = 1/2$) outperforms the scheme with lower amount N_p combined with more powerful FEC ($R_c = 1/3$). Note the gain of approximately $1\,\mathrm{dB}$ in $\bar{E}_b/N_0$ at a BER of 10^{-3} for the 32-ary modulation. This fact can be explained for $R_c = 1/3$ by

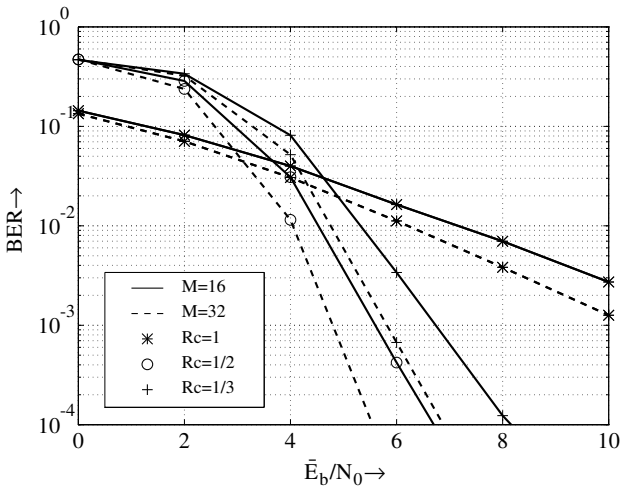

Figure 4: Sufficient frequency interleaving: Trade-off between channel coding, M-ary Walsh modulation and simple spreading for $M = 16,\ 32$ and codes of rate $R_c = 1,\ 1/2,\ 1/3$

the higher loss due to an unchanged guard time T_g but a shorter symbol duration T and an additional lower energy of the coded bit in front of the Viterbi decoder[4]. Moreover, less number of subcarriers ($N_M = M \cdot N_p$) results in higher MAI in case of the used PN-sequences. Hence, the more powerful FEC scheme is not able to compensate for the summarized losses.

We now focus on the trade-off between M-ary Walsh modulation and FEC (N_p remains unchanged due to an unchanged $B\,T_d$, see (3)). For example, 32-ary modulated signals with rate of code $R_c = 1/2$ yield a gain of $\bar{E}_b/N_0 \approx 2\,\mathrm{dB}$ in comparison with 16-ary Walsh modulation with $R_c = 1/3$ at a BER of 10^{-3}. This results from an increasing Euclidian distance (1) for higher values M and less MAI (higher number of subcarriers). I.e., the quality of the FHT output is decisive for the overall system performance.

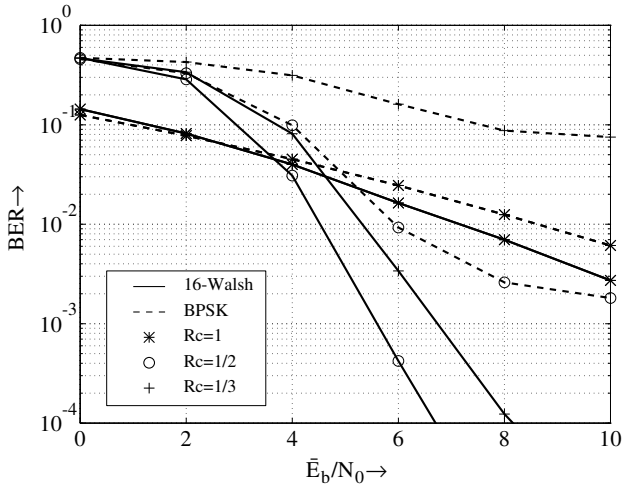

Figure 5: Sufficient frequency interleaving: Comparison of 16-ary Walsh modulation with BPSK, $R_c = 1,\ 1/2,\ 1/3$

In fig. 5, the 16-ary Walsh modulation is compared to BPSK for rates of code $R_c = 1,\ 1/2,\ 1/3$. For BPSK, the results show the same behavior concerning the trade-off between simple spreading and FEC as the Walsh modulation. Comparing the two modulation schemes indicates that already 16-ary Walsh modulation outperforms BPSK for all considered trade-offs, even if Walsh modulation is combined with a code of rate $R_c = 1/3$ and BPSK with a code of rate $R_c = 1/2$.

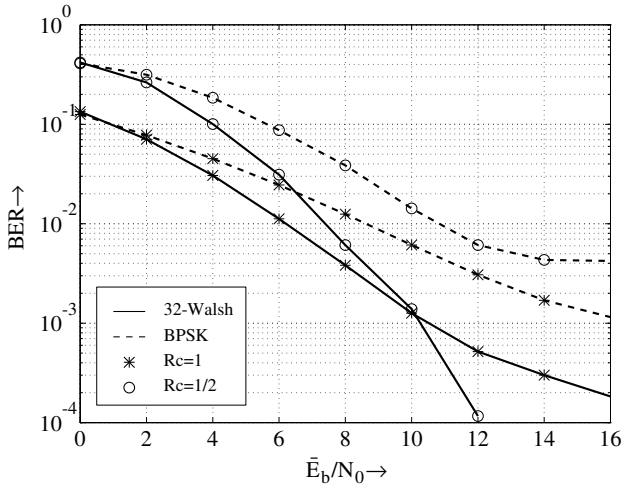

Figure 6: Insufficient frequency interleaving: Comparison of 32-ary Walsh modulation and BPSK, $R_c = 1,\ 1/2$

[4] E_b denotes the bit energy at the output of the VA decoder

The simulation results in fig. 6 show the BER for the transmission over an indoor channel with coherence bandwidth of approximately $2.8\,\mathrm{MHz}$. BPSK and 32-ary Walsh modulation are considered. In contrast to fig. 5, there exists a significant loss due to the statistical dependence of the adjacent subcarriers. The results emphasize for an MC-SS uplink transmission over an indoor Rayleigh fading channel that BPSK even combined with FEC leads to an unacceptable BER. Moderate performance can only be achieved if M-ary orthogonal Walsh modulation is applied. One has to mention that the degradation is mainly caused by the insufficient frequency interleaving. There exists only a slight degradation due to the assumption of independent subcarriers for the derivations of the log-likelihood ratios.

V. CONCLUSION

In this paper, we have investigated the concatenation of M-ary orthogonal Walsh modulation and ML Viterbi decoding for the MC-SS transmission over a frequency-selective slowly Rayleigh fading channel in the uplink. In particular, we have first evaluated the bit specific log-likelihood ratio of the M-ary orthogonal modulated signal for the ML Viterbi decoding. Assuming an unchanged bandwidth and data rate, we further analyzed the trade-off between coding and spreading to optimize the BER. Monte-Carlo simulation results indicate to better strengthen the inner spreading consisting of M-ary orthogonal modulation and simple spreading with an inherently less powerful channel code. The results are always compared with BPSK performance and it is shown that already 16-ary orthogonal Walsh modulation outperforms BPSK. Moreover, they emphasize that for an indoor scenario a moderate BER can only be achieved if the inner spreading is optimized, for example, by applying M-ary orthogonal Walsh modulation. BPSK even combined with channel coding leads to an unacceptable BER.

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