M-ARY ORTHOGONAL MODULATION FOR MULTI-CARRIER SPREAD-SPECTRUM UPLINK TRANSMISSION

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Abstract: In this paper, we investigate the application of M-ary orthogonal modulation for *multicarrier spread spectrum* (MC-SS) uplink transmission over a Rayleigh fading indoor channel. Different coherent detection strategies with perfectly known channel coefficients are analyzed. Furthermore, we present a decision-directed estimation receiver to apply channel phase estimation. With decision-directed estimation no redundancy like training data is required to be transmitted. Simulation results show no performance loss for low $\bar{E}_b/N_0$ and only a moderate degradation for high $\bar{E}_b/N_0$ in comparison to ideal phase equalization. Moreover, the results are always contrasted with BPSK performance. Applying M-ary orthogonal modulation reveals to outperfom BPSK with respect to bit error rate and spectral efficiency, even if ideal equalization is considered for BPSK and decision-directed estimation for M-ary orthogonal modulation.

The results are generally valid, whereas in this paper, they are based on the European HIPERLAN/2-standardization.

1. INTRODUCTION

As the radio frequency spectrum is a scarce resource, future wireless radio systems need to make efficient use of the frequency spectrum. One possible 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 non-selective and the frequency selectivity of the channel over all subcarriers can be combatted by a simple one-tap equalizer. But the strong attenuations of some carriers in a multipath environment lead to relatively high bit error rates. To overcome this effect, OFDM can be combined with *code division multiple access* (CDMA), where the signal is spread over all subcarriers to achieve frequency diversity. This concept was first introduced as OFDM-CDMA or also as *multi-carrier* (MC-)CDMA by [1, 2, 3] 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 or QPSK. Detection strategies were proposed [4, 5, 6] assuming perfectly known channel coefficients. Furthermore, novel channel estimation techniques for MC-SS have been analyzed [7, 8, 9] mainly based on using redundancy like pilot-symbols.

Instead of using classical modulation schemes, SS techniques offer the possibility to apply M-ary orthogonal modulation. One well-known scheme is the so-called QUALCOMM system (IS-95) combining *direct sequence*-SS (DS-SS) with M-ary orthogonal Walsh modulation [10, 11].

As well as for DS-SS, M-ary orthogonal Walsh modulation can also be combined with the MC-SS technique [12]. Since a MC-SS scheme yields a frequency resolution by dividing the available bandwidth into a finite number of subbands, the idea behind the application of M-ary orthogonal Walsh modulation is to further increase this resolution in order to distinguish transmitted symbols without raising the required bandwidth. Therefore, instead of spreading one data symbol over all subcarriers, e.g. like for BPSK, spreading is achieved in two successive steps. By employing Walsh modulation, we further use orthogonality, which renders good performance in general.

Since this orthogonality is extremely sensitive to phase rotation, e.g. caused by the channel, the equalization scheme used should preferably correct these phase deviations. For example, this can be done by pilot-symbol-aided channel estimation. Due to necessary additional information, i.e. training data, pilot-aided channel estimation requires a certain signalling overhead and therefore results in less spectral efficiency. To overcome this disadvantage, one can use a phase estimation concept based on the so-called *decision-directed* (DD) loop technique [13]. For DD parameter estimation the incoming signal is quantized. Conceptually, the decision is assumed to be correct and the decided information symbols have been estimated over a certain observation time-interval. In this case, the phase can be estimated by comparing the incoming signal with the decided information symbol for a given time-instant.

The intention of this paper is to examine the application of the M-ary orthogonal Walsh modulation for MC-SS systems operating in frequency-selective, slowly fading channels in the uplink. First, different detection strategies for the M-ary orthogonal Walsh modulation are analyzed assuming perfectly known channel coefficients. Furthermore, a receiver structure is presented employing decision-directed estimation of the channel phase. With decision-directed estimation no redundancy like training data is required to be transmitted. The simulation results are also contrasted with BPSK performance.

The paper is organized as follows: In Section 2, the MC-SS

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transmitter with M-ary orthogonal Walsh modulation is presented and important properties of the Walsh modulation are introduced. The coherent receiver assuming perfectly known channel coefficients as well as the main aspects of the DD receiver are described in Section 3. In Section 4, simulation results are shown for the uplink bit error rate of an indoor Rayleigh fading channel. Finally, Section 5 summarizes the results.

2. MC-SS TRANSMITTER WITH M-ARY ORTHOGONAL WALSH MODULATION

With SS systems, the *multiple access interference* (MAI) is mainly determined by the codes implemented. For MC-SS transmission over a downlink channel, where we have a synchronous situation, optimized codes like Walsh codes can be used. Since every user-specific codeword is affected by the same channel transfer function, the deterioration of the correlation properties can be restored by using suitable equalization schemes. For an uplink transmission scenario, which is considered in this paper, every user-specific codeword is affected by different channel transfer functions. This fact leads to an asynchronous situation involving the use of PN-sequences, which results in higher MAI. Such a MC-SS transmitter (mobile station) applying Walsh modulation is illustrated in fig. 1. For simplicity, one of J active users is taken into account (subscripts are omitted). The

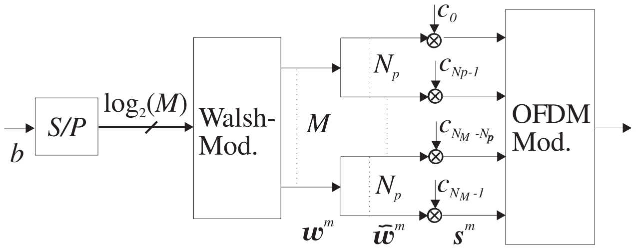

Figure 1: MC-SS transmitter with M-ary Walsh modulation

data bits, $b \in \{0, 1\}$ each of duration T_b, are serial-to-parallel converted to groups of $\log_2(M)$ data bits each. The Walsh modulation maps the $\log_2(M)$ data bits to one corresponding Walsh symbol $\boldsymbol{w}^m = [w_0^m, w_1^m, \ldots, w_{M-1}^m]^T$, $m \in \{0, \ldots, M-1\}$ including M Walsh chips $w_\mu^m \in \{\pm 1\}$, $\mu = 0 \ldots M-1$. Each of the M parallel Walsh chips has a symbol duration of $T = \log_2(M) T_b$. This modulation can also be interpreted as frequency spreading of value $M/\log_2(M)$. The set of M orthogonal Walsh symbols represents an orthogonal basis which can be evaluated recursively by applying

$$\boldsymbol{W}_{2M} = \begin{pmatrix} \boldsymbol{W}_M & \boldsymbol{W}_M \\ \boldsymbol{W}_M & -\boldsymbol{W}_M \end{pmatrix}, \quad \boldsymbol{W}_1 = (1), \tag{1}$$

where $\boldsymbol{W}_M$ is the $M \times M$ *Hadamard-Matrix* [10, 13]. The Euclidian distance is identical for all possible pairs of symbols and equals

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

To finally 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 each bandwidth of the M parallel subchannels decreases for larger values of M, the number of subcarriers N_M increases for the same available bandwidth B. In case of Walsh modulation, the number of subcarriers is determined by

$$N_M = N_2 \cdot \log_2(M) = M \cdot N_p, \tag{3}$$

where N_2 is the number of subcarriers used for 2-ary Walsh modulation or BPSK, respectively[1]. For the j^{th} user, this yields the transmission vector $\boldsymbol{s}^{m,j}$ before the OFDM modulation

$$\boldsymbol{s}^{m,j} = [s_0^{m,j}, \ldots, s_{N_M-1}^{m,j}]^T, \quad s_i^{m,j} = \tilde{w}_i^{m,j} \cdot c_i^j \tag{4}$$

with $\tilde{w}_i^{m,j} = w_\mu^{m,j}$, $\forall\ \mu = \lfloor i/N_p \rfloor$, m, j. OFDM modulation includes the IFFT (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. Since we focus on an indoor scenario, the channel can be regarded as slowly fading for the considered values M. Hence, $T \ll 1/f_{d,max}$ holds and $f_{d,max}$ indicates the maximum Doppler frequency. Due to the insertion of the guard interval, every subcarrier is affected by one single channel transfer coefficient. For the uplink transmission scenario assumed here, 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 the random amplitude and phase. Hence, the channel can be described for every user by an $N_M \times N_M$ diagonal matrix $\boldsymbol{H}^j$.

3. MC-SS RECEIVER FOR M-ARY ORTHOGONAL WALSH MODULATION

3.1. Ideal coherent receiver

The MC-SS receiver for M-ary orthogonal Walsh modulation assuming perfectly known channel coefficients, called ideal coherent receiver, is presented in fig. 2. Paying attention to J

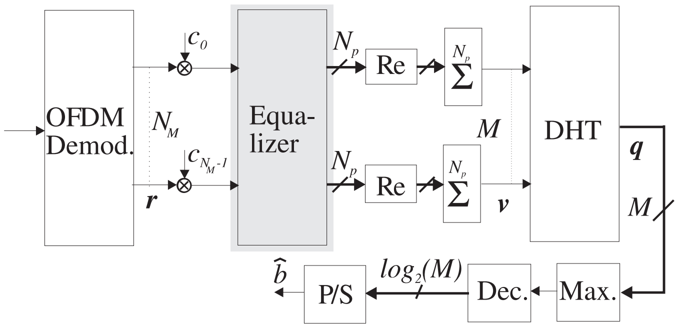

Figure 2: Ideal coherent MC-SS receiver for M-ary Walsh modulation

active users, the received signal after OFDM demodulation can be written as a sum of vectors $\boldsymbol{r}^{m,j}$

$$\boldsymbol{r} = \sum_{j=0}^{J-1} \boldsymbol{r}^{m,j} + \boldsymbol{n} = \sum_{j=0}^{J-1} \boldsymbol{H}^j \boldsymbol{s}^{m,j} + \boldsymbol{n} \tag{5}$$

[1] I.e. the system design is based on 2-ary modulation for comparison.

with elements $r_i^{m,j} = s_i^{m,j} \cdot h_i^j, \quad i = 0 \ldots N_M - 1$ and $m \in \{0, \ldots, M-1\}$. The vector $\boldsymbol{n}$ represents 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

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

$$\mu, m \in \{0, \ldots, M-1\} \qquad (6)$$

where e_i indicates the equalization coefficient of the i^{th} subcarrier. For our purpose, we take into account the well-known equalization schemes *equal gain combining* (EGC) and *maximum ratio combining* (MRC) [5]. If perfectly known channel coefficients are assumed, then $e_i = h_i^*/|h_i|$ for EGC and $e_i = h_i^*$ for MRC. To obtain *maximum 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 *Discrete Hadamard Transform* (DHT) [11], using the Hadamard matrix (1) according to

$$\boldsymbol{q} = \boldsymbol{W}_M \cdot \boldsymbol{v}. \qquad (7)$$

Vector $\boldsymbol{q}$ includes M decision variables where the choice of the maximum value results in a symbol decision. Decoding of the decided symbol to $\log_2(M)$ data bits and parallel-to-serial conversion finally leads to the estimated data stream with bits $\hat{b}$. Furthermore, it should be mentioned that using a guard interval results in mismatching

$$\frac{S}{N} = \frac{E_s}{N_0} \cdot \left(1 - \frac{T_g}{T}\right) \quad \text{with} \quad \frac{T_g}{T} = \frac{B\,T_g}{N_M + B\,T_g}, \qquad (8)$$

where E_s is the symbol energy, $N_0/2$ indicates the Gaussian noise spectral density and S/N the signal-to-noise ratio used for MLD. The amount of mismatching is determined by the ratio T_g/T of the guard time T_g to the overall symbol duration T. With (3) and (8) it can be seen, that mismatching is reduced by raising M if the product $B\,T_g$ is assumed to be constant.

3.2. Decision Directed (DD) receiver

As already mentioned, one can estimate the channel coefficients by pilot-symbol techniques with the drawback of signalling overhead. Applying M-ary orthogonal Walsh modulation, the deterioration of the orthogonality is mainly determined by phase rotation. Therefore, the estimation of the phase should be focused on, which leads to consider a phase-decision-directed estimation technique. Using this technique, the phase is estimated by comparing the received signal with the decided and fed back symbols under the assumption of correct decisions[2].

Fig. 3 illustrates the DD receiver for M-ary orthogonal Walsh modulation. In contrast to the ideal receiver shown in fig. 2, the equalization block is expanded by the phase-estimation and an additional feedback with modulation and spreading (see transmitter, fig. 1) of the decided group of $\log_2(M)$ bits is employed. Hence, (6) still holds whereas the equalization coefficient is here given by $e_i = \exp(-j\hat{\phi}_i)$, $i = 0 \ldots N_M - 1$, where $\hat{\phi}_i$ indicates the estimated phase for the i^{th} subcarrier.

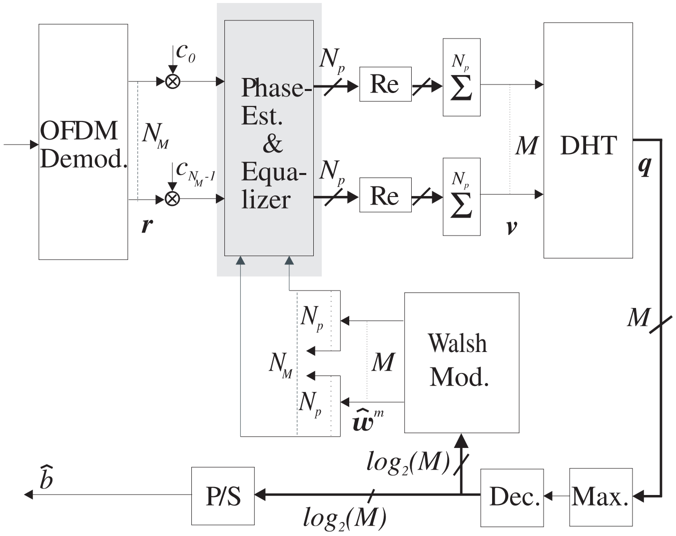

Figure 3: MC-SS DD receiver for M-ary Walsh modulation

In the following, we focus on the phase-estimation and equalization block. For simplicity, one subcarrier at one time-instant is considered (subscripts are omitted) and all interferences are taken into account as AWGN. Such a subsystem for one subcarrier of this block is shown in fig. 4. The received signal x and the output signal y are given by

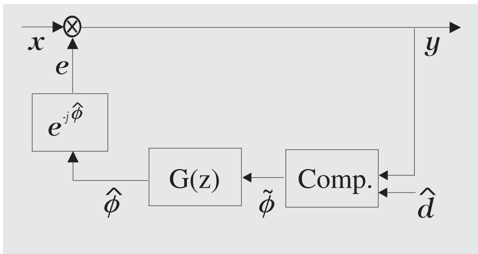

Figure 4: Phase-estimation and equalization subsystem

$$y = x \cdot e = |y| \cdot e^{j\,\Delta\phi}, \quad x = d \cdot h + n \qquad (9)$$

where $d \in \{\pm 1\}$ represents the transmitted information symbol, $h = \rho \cdot \exp(\phi)$ the subchannel coefficient and n the AWGN with variance σ_n^2. The received signal is phase corrected by the multiplication with $e = \exp(-j\hat{\phi})$, and $\hat{\phi}$ denotes the estimated phase. The comparator performs the operation $\tilde{\phi} = \mathrm{Im}\left\{\hat{d} \cdot y\right\}$. According to the assumption of correct estimation ($\hat{d} = d$), the comparator output signal can be written as

$$\tilde{\phi} = \mathrm{Im}\left\{\hat{d} \cdot y\right\}\Big|_{\hat{d}=d} = \rho \cdot \sin(\Delta\phi) + \phi_n, \qquad (10)$$

with $\Delta\phi = \phi - \hat{\phi}$ as the phase of the output signal y and $\phi_n = \mathrm{Im}\{d \cdot e \cdot n\}$ as phase noise. In case of proper phase tracking, $\Delta\phi$ is small and, hence, $\sin(\Delta\phi) \approx \Delta\phi$, which results

[2]Due to the fact that no additional information has to be transmitted, this method can also be interpreted as *noncoherent* reception.

to the comparator output signal

$$\tilde{\phi} \approx \rho \cdot \Delta\phi + \phi_n. \tag{11}$$

Assuming statistical independence between the data symbols and the noise, the variance of the phase noise ϕ_n is given by $\sigma^2_{\phi_n} = 0.5\sigma^2_n$. Finally, the phase-correction is carried out after filtering $\tilde{\phi}$ by a first order loop (IIR-filter)

$$G(z) = a_0 \frac{z^{-1}}{1 - z^{-1}}, \tag{12}$$

with gain parameter a_0. It is shown in [14] that the tracking performance and the noise interference of the system is mainly determined by a_0 (i.e. depending on the channel transmission conditions, one has to find a compromise between tracking performance and noise reduction ensuring system stability).

4. SIMULATION RESULTS

4.1. System Descriptions

Based on the European HIPERLAN/2 standardization, the following results are given for an indoor Rayleigh fading channel. Assuming an exponential delay profile with a root mean square (RMS) delay of $\tau_{RMS} = 150$ ns and neglecting echoes less than 10% of the maximum power yield a maximum delay spread of $\tau_{max} \approx 350$ ns. The bandwidth is chosen to be $B = 25$ MHz in the 5.2 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 Hz. Hence, long symbol durations T are possible. Assuming a constant guard time T_g, higher values of M lead to less mismatching and better spectral efficiency, see (8). For example, if for BPSK $N_2 = 64$ subcarriers and $T_g/T = 0.2$ are considered ($T_g = 640$ ns $> \tau_{max}$), mismatching will be reduced from 1 dB to 0.2 dB and the bit rate will be increased from 312 kbit/s to 372 kbit/s by raising M from 2 to 32.
In case of DD reception, the gain parameter is chosen to be $a_0 = 1.5$. This choice leads to a good compromise between tracking performance and phase noise reduction for the given indoor transmission conditions ensuring system stability. Furthermore, for DD reception steady state is considered.
In all cases, $J = 8$ active users and $N_2 = 64$ subcarriers are taken into account.

4.2. System Comparisons

Fig. 5 presents the bit error rate (BER) of the 32-ary Walsh modulation and BPSK for MRC and EGC equalization assuming perfectly known channel coefficients. As can be seen, MRC outperforms EGC for BPSK because of less MAI. This is also true for Walsh modulation, although MRC leads to a higher deterioration of the orthogonality due to magnitude influence. This influence causes (self-) interference by all possible symbols applying MLD. Comparison of both modulation schemes results in a better performance for the 32-ary Walsh modulation, even in case where Walsh modulation is applied with EGC and

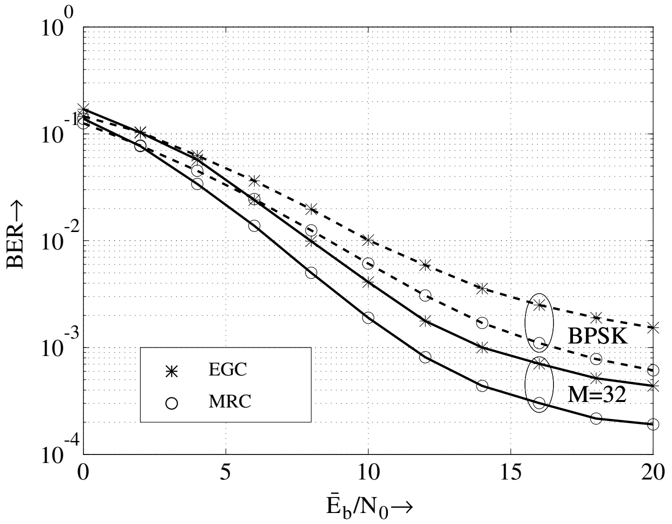

Figure 5: Comparison of ideal MRC and ideal EGC for 32-ary Walsh modulation and BPSK with $J = 8$ active users and $N_2 = 64$ subcarriers

BPSK with MRC. This means that only phase equalization combined with Walsh modulation is more powerful than complete channel equalization with BPSK.

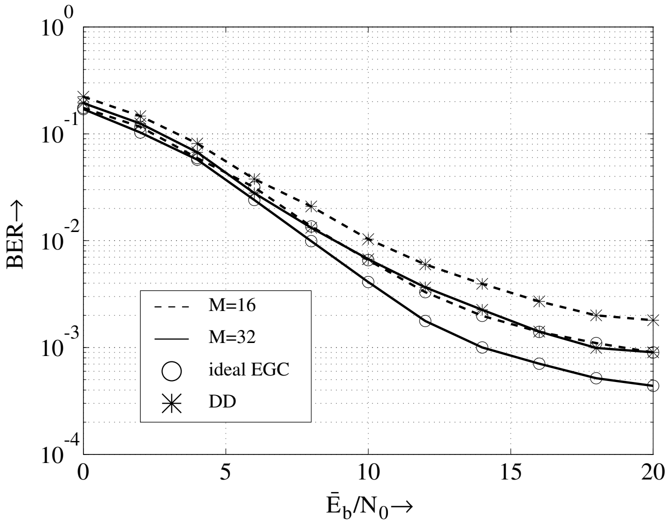

Figure 6: Comparison of DD and ideal EGC for 32-ary Walsh modulation with gain parameter $a_0 = 1.5$, $J = 8$ active users and $N_2 = 64$ subcarriers

If we now focus on the DD reception, fig. 6 shows the bit error rate compared to EGC (applying MRC requires the knowledge of the magnitudes and phases) with perfectly known phases for 16-ary and 32-ary Walsh modulation. For approximately $\bar{E}_b/N_0 < 8$ dB there is no loss in performance for the DD phase estimation. Due to the fact of MAI, there exists a moderate degradation for high ratios $\bar{E}_b/N_0$, i.e. the BER is not even doubled in these cases. Different values of M are considered for the Walsh modulation with DD reception in fig. 7. The results show that the BER is reduced for larger values M, which can be explained by the increasing Euclidian distance, see (2). Since more subcarriers are used ($N_M = 64 \cdot \log_2(M)$), the MAI is

also reduced. Furthermore, the results indicate that tracking is ensured independent of M.
For comparison, the BER of BPSK combined with the more powerful equalization technique MRC (BPSK-MRC) assuming perfectly known channel coefficients is also presented in fig. 7. It should be emphasized that M-ary orthogonal Walsh modulation with DD reception, and, therefore with not ideal phase equalization, outperforms the ideal BPSK-MRC scheme for values $M \geq 32$. The performance gain is even higher if channel estimation is considered for BPSK, e.g. pilot-symbol techniques. Furthermore, there is no redundancy required for

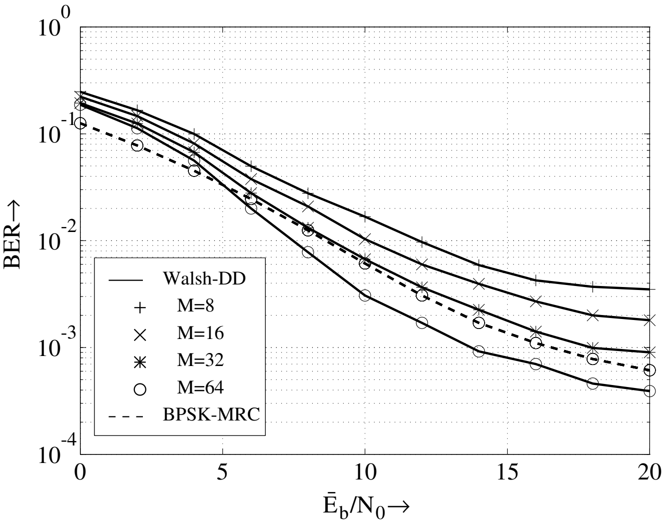

Figure 7: BER of M-ary Walsh modulation with DD reception for gain parameter $a_0 = 1.5$, $J = 8$ active users and $N_2 = 64$ subcarriers

the DD phase estimation which offers an additional coding gain in case of channel coding.

5. CONCLUSION

In this paper, we have investigated the application of M-ary orthogonal Walsh modulation for MC-SS uplink transmission over a Rayleigh fading indoor channel. The equalization techniques EGC and MRC have been analyzed assuming perfectly known channel coeffcients. Simulation results show that MRC outperforms EGC due to less MAI.
Furthermore, we have proposed a reception concept based on decision-directed (DD) phase estimation. Simulation results for the steady state have been presented. Compared to EGC with perfectly known channel phases, the DD reception leads to no loss in bit error rate performance for low $\bar{E}_b/N_0$. In case of high $\bar{E}_b/N_0$, there exists a moderate degradation caused by MAI whereat the BER is not even doubled.
The results are also always contrasted to BPSK performance. It was shown that Walsh modulation performs better in terms of bit error rate and spectral efficiency, even if ideal equalization is considered for BPSK and DD estimation for M-ary orthogonal Walsh modulation.
The concept of better using the available bandwidth by increasing the frequency resolution, combined with the advantages of raising Euclidian distance, less mismatching and thereby higher spectral efficiency for larger values M, appears to be an interesting modulation scheme for indoor mobile communication systems. The possibility to apply DD reception and, hence, to estimate the channel without using redundancy like pilot-symbols, enables a higher coding gain in case of channel coding. Additional analyses have therefore to be done by applying channel coding techniques.

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