NEW ADAPTIVE BIT AND POWER LOADING POLICIES FOR GENERALIZED MULTICARRIER TRANSMISSION

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ABSTRACT

In this paper, the adaptive modulation for the Generalized MultiCarrier (GMC) transmission is described, showing that non-orthogonality of the subcarriers affects the complexity of the adaptive techniques and the system performance. The mathematical derivation for the GMC channel capacity is presented. Since the straightforward formula is impractical from the implementation point of view, the efficient modifications of the Hughes-Hartogs algorithm are discussed. Simulation results are also included.

1. INTRODUCTION

Adaptive techniques are being applied in wireless communications to improve the broadly-understood efficiency of a telecommunication system. They serve better utilization of available resources (time, frequency or power) while guaranteeing the Quality-of-Service. Some of these techniques are successively applied in wireless systems, e.g. Hybrid Automatic Repeat-Request (HARQ) in HSDPA or Adaptive Modulation and Coding (AMC) in IEEE 802.11 standards, WiMAX, or TETRA2.

Adaptive modulation for frequency-selective channels, originally proposed for and applied in Orthogonal Frequency Division Multiplexing (OFDM) systems, bases on the optimal assignment of a number of bits to be transmitted at distinct subcarriers given the instantaneous channel characteristic and the total transmit power constraint. This optimal bit assignment should maximize the channel capacity. The solution of this classical constraint-optimization problem for independent subbands defined within a given frequency band (e.g. for the OFDM subcarriers) can be found in many books on the information theory, e.g. in [1].

Within this paper, we take the so-called Generalized MultiCarrier (GMC) transmission into account (described in [2] in detail), and consider adaptive bit and power loading for the GMC channel capacity maximization. It can be shown, that GMC transmission is very effective with respect to spectral efficiency since it does not require the cyclic prefix (necessary in case of the OFDM), can be easily parametrized, and displays flexibility, which makes it suitable for the application in multi-standard transceivers [2]. As it will be shown in Section II, our GMC transmission is based on non-orthogonal subcarriers and filtered subband signals, what results in overlapping of these signals both in time and in the frequency domain. In face of such overlapping and mutual dependence of baseband filtered and nonorthogonal signals the classical solution for bit and power loading presented in the literature for OFDM is no longer appropriate for our GMC system.

The reminder of the paper is organized as follows. First, the basics of the GMC system are presented in Section II. The modifications of the well known *water-filling* principle [1] for application in two-dimensional (time-frequency represented) signals are described in Section III. In Section IV the idea of bit-and-power loading for GMC is presented, whereas the adjustment of the Hughes-Hartogs algorithm for GMC systems is explained in Section V. The simulation results are given in Section VI and the work is concluded in Section VII.

2. THE GENERALIZED MULTICARRIER SIGNAL

Our GMC representation of signals is based on the Gabor signal expansion using the non-orthogonal basis functions [3], [4]. The basis functions $g_{l,m}(t)= g(t-lT)\exp(2\pi jmFt)$ in our case are constructed from the so-called synthesis window $g(t)$ equally shifted by lT in time and mF in frequency, where l and m are the indices of the time-domain and frequency-domain intervals respectively, T is the time-distance between consecutive shifted versions of $g(t)$, and F is the distance between adjacent subcarriers in the frequency domain. The transmit signal can be represented as:

$$s(t)=\sum_{l,m} c_{l,m}\cdot g_{l,m}(t), \qquad (1)$$

where $c_{l,m}$ are the so-called the Gabor coefficients, which represent the transmit signal $s(t)$ on the TF plane and can be obtained using the Gabor transform of the transmit signal and the so-called analysis window $\gamma(t)$ [2].

The discrete representation of (1) can be written as:

$$s(n)=\sum_{l=-\infty}^{\infty}\sum_{m=-\infty}^{\infty} c_{l,m}\, g_{l,m}(n),$$

where

$$g_{l,m}(n)=g(n-lN)\exp(j2\pi mn/M)$$

is the discrete form of the synthesis window shifted in time by lT (note that $T = N\cdot T_s$, where T_s is the sampling interval, and N is the distance between neighboring $g_{l,m}(n)$ windows in samples) and in frequency by mF, $0 \le m \le M-1$, and M is the total number of subcarriers. The Time-Frequency (TF) shifted versions of the synthesis window $g(n)$, i.e. $g_{l,m}(n)$ are called the *Gabor atoms* since they represent elementary pulses, which are used by the information-carrying data symbols $c_{l,m}$. The transformation from the two-dimensional (2D) TF domain signal representation to the time domain and backwards is possible only when the analysis and synthesis windows are biorthogonal [3]. Any signal can be represented in the way of GMC representation by choosing the appropriate synthesis window shape, length N_g, the distance between consecutive atoms N (in samples) and number of subcarriers M [2]. For example, when the window shape is selected to be rectangular, N and N_g are equal to M, and $s(n)$ represents the OFDM signal (before adding the cyclic-prefix).

Let us stress again the differences between OFDM and GMC transmission. In the case of non-orthogonal modulation, the spectral efficiency can be higher due to the possibilities of decreasing of adjacent subcarriers spacing and discarding of the cyclic prefix. Moreover, GMC signal can be very easily parametrized, what makes it suitable for next generation, software defined transceivers, as well as for opportunistic radio [5]. On the other side, the lack of orthogonality causes that typical algorithms (suitable for OFDM) cannot be applied in GMC case, since the information-bearing pulses suffers from self-interference. This phenomena has to be considered in the transmitter as well, e.g. when the adaptive techniques are applied, because by assigning some amount of power to some pulse affects significantly the neighboring pulses, degrading the overall channel quality, estimated earlier for these TF instants (TF pulses).

3. TIME-FREQUENCY POWER LOADING IN THE GMC TRANSMITTER

In case of the TF representation of signals the problem of adaptive power loading (PL) becomes two-dimensional. Let us notice, that we assume the perfect knowledge of the channel gains or, if these values change within the frame, that an appropriate channel prediction has been applied. We start with dividing the channel frequency band B into infinitesimal subbands df in such a way, that the channel characteristic could be considered flat in frequency and invariable in the time instant t. The channel capacity can be calculated using the Shannon formula (based on the derivations in [1]):

$$C = \int_B \int_{T_B} \frac{1}{T_B} \log\left(1 + \frac{P(f,t)\left|H(f,t)\right|^2}{G(f,t)}\right) dtdf \,, \qquad (2)$$

where $G(f,t)$, $H(f,t)$ denote the noise power spectral density and the channel characteristic respectively at the time instant t, $P(f,t)$ denotes the power assigned to the signal localized at the frequency f and the time instant t, and T_B is the frame time duration. As a result of the maximization of C using the Lagrange multipliers, the function $P(f,t)$ that maximizes formula (2) has the following form:

$$P(f,t) = W - \frac{G(f,t)}{\left|H(f,t)\right|^2} \,, \text{ for } \frac{G(f,t)}{\left|H(f,t)\right|^2} \le W$$

and $P(f,t)=0$ otherwise. The 2D *water-surface* W (we use the phrase *water-surface* instead of *water-line* to distinguish between one- and two-dimensional scenario; however, in both cases the phrase *water-level* can be used) can be computed using the power constraint, what results in:

$$W \cdot B \cdot T_B = P + \int_B \int_{T_B} \frac{G(f,t)}{\left|H(f,t)\right|^2} dfdt \,.$$

Let us now consider, how this 2D water-filling principle relates to our GMC signal and the channel model. As mentioned above, the subsequent atoms of the GMC signal can overlap the neighboring atoms both in time and in frequency domain. Thus, any change of the power assigned to one atom has its repercussions on the power of the neighboring atoms. This can be understood as TF self-interference. To include the window shape and overlapping phenomena into the calculation of the optimal power allocation, let us represent the transmit signal power as a function of the power assigned to the time and frequency-shifted synthesis window $g_{l,m}(t)$ and the power assigned to all respective data symbols $c_{l,m}$. Let's compute the TF distribution of the transmit signal:

$$S(f,t) = \sum_{l=0}^{L-1}\sum_{m=0}^{M-1} c_{l,m} STFT\left(g_{l,m}(t)\right) \,,$$

where $STFT(\cdot)$ denotes the Short Time Fourier Transform [6], [7] used for our TF analysis. One can observe, that the TF representation of the signal $s(t)$ is equivalent to the sum of weighted TF representations of the original synthesis window $g(t)$. Now, the power distribution of the signal $s(t)$ on the TF plane can be calculated in the following way:

$$P(f,t) = E\left\{\left|S(f,t)\right|^2\right\} = E\left\{\left|\sum_{l=0}^{L-1}\sum_{m=0}^{M-1} c_{l,m} STFT\left(g_{l,m}(t)\right)\right|^2\right\}$$

or in equivalent form,

$$P(f,t) = \sum_{l=0}^{L-1}\sum_{m=0}^{M-1} E\left\{\left|c_{l,m}\right|^2\right\} E\left\{\left|\Gamma_{l,m}(f,t)\right|^2\right\}, \qquad (3)$$

where $E(\cdot)$ is the expected value, and $\Gamma_{l,m}(f,t)$ is the STFT of $g_{l,m}(t)$. The above relation shows, that based on the independency of the user data $c_{l,m}$ the power distribution on TF plane can be computed as the weighted sum of the spectrograms of the synthesis window $g(t)$, shifted in time and in frequency. Moreover, the formula (3) can be rewritten as:

$$P(f,t) = \sum_{l=0}^{L-1}\sum_{m=0}^{M-1} P_{c_{l,m}} P_{g_{l,m}}(f,t) \,,$$

where $P_{c_{l,m}}$ is the power of the data symbol $c_{l,m}$ and $P_{g_{l,m}}(f,t)$ is the power TF density of the synthesis window function $g(t)$ shifted to the (l, m) location in the TF plane.

Let us consider the discrete representation of $s(t)$. In such a case, the signal power density $P_{g_{l,m}}(f,t) = P_{g_{l,m}}(k,n)$ for $t = lT + nT_s$ and $f = mF + kF_s$, where T and F are the distances between adjacent atoms in time and in frequency, and T_s and F_s are the sample intervals in time and frequency reflecting the resolution on our TF plane. Thus, assuming that $G_{l,m}(k,n)$ and $H_{l,m}(k,n)$ denote $G(f,t)$ and $H(f,t)$ respectively for $t = lT + nT_s$ and $f = mF + kF_s$ our 2D water-filling principle for discrete signals is as in (4). Let us integrate the relation (4) over both time and frequency domain within the rectangular area of TF grid.

$$\sum_{l'=0}^{L-1}\sum_{m'=0}^{M-1} P_{c_{l',m'}} P_{g_{(l'-l),(m'-m)}}(k,n) = \tag{4}$$
$$= \begin{cases} W - \dfrac{G_{l,m}(k,n)}{\left|H_{l,m}(k,n)\right|^2} & \text{if } W > \dfrac{G_{l,m}(k,n)}{\left|H_{l,m}(k,n)\right|^2}, \\ 0 & \text{otherwise} \end{cases}$$

Such rectangular areas are edged with each other, and their centers are the locations of the atoms $g_{l,m}(t)$. This integration of discrete values is naturally equivalent to summation for TF indices $k = 0,\ldots, K-1$ and $n = 0,\ldots, N-1$. Thus,

$$\sum_{l'=0}^{L-1}\sum_{m'=0}^{M-1} P_{c_{l',m'}} P_{g_{(l'-l),(m'-m)}} = \begin{cases} W - \dfrac{G_{l,m}}{\left|H_{l,m}\right|^2} & \text{for } W > \dfrac{G_{l,m}}{\left|H_{l,m}\right|^2}, \\ 0 & \text{otherwise} \end{cases}$$

where $P_{g_{(l'-l),(m'-m)}} = \frac{1}{KN}\sum_{k=0}^{K-1}\sum_{n=0}^{N-1} P_{g_{(l'-l),(m'-m)}}(k,n)$ and $G_{l,m} = G_{lm}(k,n)$, as well as $H_{l,m} = H_{lm}(k,n)$ for any values of k and n within the abovementioned range, because we consider these values invariant within an atom rectangular area in the TF grid. Thus, we obtain the close-form formula for PL in the GMC transmitter as follows:

$$P_{c_{l,m}} = \frac{W}{P_{g_{0,0}}} - \frac{G_{l,m} + \left|H_{l,m}\right|^2 \sum_{(m',l')\in D} P_{c_{l',m'}} P_{g_{(l'-l),(m'-m)}}}{P_{g_{0,0}}\left|H_{l,m}\right|^2}, \tag{5}$$

for $W > \left(G_{l,m} + \left|H_{l,m}\right|^2 \sum_{(m',l')\in D} P_{c_{l',m'}} P_{g_{(l'-l),(m'-m)}}\right)\left|H_{l,m}\right|^{-2}$

and $P_{c_{l,m}} = 0$ otherwise. In (5), the set D is defined as: $D = \{(m',l') \in Z^2 \wedge 0 < l' < L-1 \wedge 0 < m' < M-1 \wedge l' \neq l \wedge m' \neq m\}$

The formula (5) shows, that the optimal power allocation to a particular atom representing the $c_{l,m}$ data symbol depends on the power allocated to all other atoms. To find a joint solution to this problem we need to solve a set of LM equations with LM unknown variables, namely the powers allocated to LM atoms. Equation (5) can be rewritten in the matrix form: $\mathbf{P}_g\mathbf{P}_c = \mathbf{X}$, where the matrices are defined as $\mathbf{P}_g = \{P_{g_{(l'-l,m'-m)}}\}$, $\mathbf{P}_c = \{P_{c_{l',m'}}\}^{T}$ and $\mathbf{X} = \{X_{l,m}\}^{T}$. The elements of $\mathbf{X}$ are de-fined as: $X_{l,m} = W - G_{l,m} \cdot |H_{l,m}|^{-2}$. The solution of the proposed matrix equation will be $\mathbf{P_c} = \mathbf{P_g^{-1}} \cdot \mathbf{X}$. The calculated power values cannot be negative, and thus, apart from solving the set of equations we must solve the set of inequalities:

$$\frac{W}{P_{g_{0,0}}} \geq \frac{G_{l,m} + \left|H_{l,m}\right|^2 \sum_{(m',l')\in D} P_{c_{l',m'}} P_{g_{(l'-l),(m'-m)}}}{P_{g_{0,0}}\left|H_{l,m}\right|^2}.$$

The parameter W has to be calculated from the initial condition on the total power constraint:

$$W \cdot M \cdot L = P + \sum_{m=0}^{M-1}\sum_{l=0}^{L-1} G_{l,m}\left|H_{l,m}\right|^{-2}.$$

4. BIT AND POWER LOADING IN THE GMC TRANSMITTER

As we have seen, the problem of power loading in GMC environment is already very complex. For the bit-and power loading (BPL) purpose, the function from (2) has to be complemented by the α parameter ($\alpha = -1.5/\ln(5 \cdot \mathrm{Pr}_b)$ for the QAM order higher than 2 and 0<SNR<30 dB, where Pr_b is the Bit-Error Probability (BEP) [1].):

$$C = \int_B \int_{T_B} \frac{1}{T_B}\log\left(1 + \frac{\alpha \cdot P(f,t)\left|H(f,t)\right|^2}{G(f,t)}\right) dt df.$$

Maximization of C as our objective function using the Lagrange multipliers results in the following solution for PL:

$$P_{c_{l,m}} = \frac{W}{P_{g_{0,0}}} - \frac{G_{l,m} + \left|H_{l,m}\right|^2 \sum_{(m',l')\in D} P_{c_{l',m'}} P_{g_{(l'-l),(m'-m)}}}{\alpha\left|H_{l,m}\right|^2 P_{g_{0,0}}}, \tag{6}$$

for
$$\frac{G_{l,m} + \left|H_{l,m}\right|^2 \sum_{(m',l')\in D} P_{c_{l',m'}} P_{g_{(l'-l),(m'-m)}}}{\left|H_{l,m}\right|^2} \leq W \tag{7}$$

and $P_{c_{l,m}} = 0$ otherwise. The above formulas show that in the case of the GMC-based transmission with a particular BEP requirement, one has to consider distortion, which contains the noise and the interference originating from the neighboring atoms (self-interference). The water-surface value W has to be calculated from the power constraints as in the power loading approach. However, integrating both sides of (6) over time and frequency leads to the following relation:

$$W \cdot M \cdot L =$$
$$= P + \sum_{m=0}^{M-1}\sum_{l=0}^{L-1} \frac{G_{l,m} + (1-\alpha)\left|H_{l,m}\right|^2 \sum_{(m',l')\in D} P_{c_{l',m'}} P_{g_{(l'-l),(m'-m)}}}{\alpha\left|H_{l,m}\right|^2}.$$

One can observe, that now, the water-surface W depends on the power levels $P_{c_{l',m'}}$ assigned to the data symbols $c_{l',m'}$. In such a situation, these power values have to be calculated jointly with the water-surface value W, i.e. the set of $LM+1$ equations has to be solved. Moreover, the inequality conditions defined by (7) should be taken into account. The set of equations can be also represented in the matrix form as it was in a case of PL for the GMC transmitter. The matrices defined for the PL problem have to be complemented by the elements concerning the water-surface W parameter. In such a case we may be able to solve the problem of joint water-surface determination and optimal power allocation. Finally, once we determine the power allocation for a given BEP, the assignment of bits (limited to the permissible constellation size) to the respective TF atom locations can be done according to:

$$\tilde{M}_{l,m} = 1 + \alpha \cdot P_{c_{l,m}} \left|H_{l,m}\right|^2 \Big/ G_{l,m} \ .$$

5. MODIFICATION OF THE HUGHES-HARTOGS ALGORITHM FOR THE GMC TRANSMITTER

The theoretical approach, presented in the previous sections, requires solving of the large matrix equation what is impractical from the real-time implementation point of view. Moreover, a set of inequalities has to be solved as well. Some practical algorithms of lower complexity for BPL have already been proposed for OFDM e.g. [8], [9]. One of them is the Hughes-Hartogs (H-H) algorithm [10]. Its main idea bases on the iterative increase of the number of bits assigned to subcarriers (in the OFDM case). Additional bit is assigned to the subcarrier, for which the smallest amount of power is required to send one additional bit.

Below, the modification of the H-H algorithm for the TF-represented signals is shortly presented. First, we describe extension of the conventional algorithm to the TF represented signals, but we neglect the lack of orthogonality of subcarriers and overlapping of the atoms in both dimensions. This 2D H-H algorithm will serve us for further performance comparisons with the modified H-H algorithm, which does take this overlapping and lack of orthogonality into account. First, in the initialization phase the *incremental matrix* for each time and frequency location is defined. In this matrix, the rows relate to possible numbers of bits per symbol (constellation sizes), and the columns relate to all possible atoms TF locations in a considered frame. The element in the *b*th row and *(l,m)*th column (here the column index relates to the (*l*,*m*) atom location in TF plane) of this matrix denotes the amount of power required to transmit one additional bit by the atom $g_{l,m}(t)$: $\Delta P_{b,(l,m)} = P_{b,(l,m)} - P_{b-1,(l,m)}$, where $P_{b,(l,m)}$ is the transmit power needed at the atom location (*l*,*m*) to transfer *b* bits per symbol at a required BEP. Note, that $P_{0,(l,m)} = 0$.

In a simplified case, when overlapping of the Gabor atoms is neglected, the power at the (*l*,*m*) atom location:

$$P(f = mF, t = lN) = P_{c_{l,m}} = \frac{\tilde{M}_{l,m} - 1}{\alpha \left|H_{l,m}\right|^2} G_{l,m} \ . \qquad (8)$$

In case of the GMC system with overlapping atoms:

$$P(f = mM, t = lN + n) = P_{c_{l,m}} \left| g_{l,m}(n) \right|^2 = \frac{\tilde{M}_{l,m} - 1}{\alpha \left|H_{l,m}\right|^2} G'_{l,m} \ , \qquad (9)$$

where

$$G'_{l,m} = G_{l,m} + \left|H_{l,m}\right|^2 \sum_{(m',l') \in D} P_{c_{l',m'}} P_{g_{(l'-l),(m'-m)}} \ . \qquad (10)$$

From (9) and (10) we obtain:

$$\sum_{l'=0}^{L-1} \sum_{m'=0}^{M-1} \beta_{l',l,m',m} P_{c_{l',m}} P_{g_{(l'-l),(m'-m)}} = \frac{\tilde{M}_{l,m} - 1}{\alpha \left|H_{l,m}\right|^2} G_{l,m} \ , \qquad (11)$$

where $\beta_{l',l,m',m} = 1$ for *l'=l* and *m'=m* and $\beta_{l',l,m',m} = -(\tilde{M}_{l,m} - 1)/\alpha$ otherwise. The solutions, i.e. $P_{c_{l,m}}$ for all values of *l* and *m* can be found by solving the set of linear equations defined above for a given $\tilde{M}_{l,m}$. Finally, the values of $P_{b,(l,m)}$ needed for the incremental-power matrix equal: $P_{b,(l,m)} = P_{c_{l,m}}$ for $b = \log(\tilde{M}_{l,m})$, while $P_{c_{l,m}}$ are calculated based either on (8) or (11) depending on which version of the H-H algorithm we want to apply (the simplified or the exact one). After the initialization phase the main loop of the algorithm is executed (see Tab. 1).

Tab.1. The modified Hughes-Hartogs algorithm

Initialization phase:

Fill in the incremental-power matrix and assign 0 bits to each atom

Main loop:

1) *Search the first row for the smallest* $\Delta P_{1,l \cdot m}$ *Result: column l'·m'*
2) *Assign q more bits* (*e.g. q=2 for QAM*) *to atom l',m'* ,
3) *Increment the total number of bits* L_b *and the total transmitted power* P_{tot}:

$$L_b := L_b + q \wedge P_{tot} := P_{tot} + \Delta P_{1,l' \cdot m'} \cdot \sum_{n=-N_g/2}^{N_g/2} \left| g_{0,0}(n) \right|^2$$

4) *Move all terms of column l'· m' one place up:* $\Delta P_{i,l' \cdot m'} = \Delta P_{i+1,l' \cdot m'}$
5) *If* $P_{tot} > P$ *or* $L_b > L_{b\,\text{assumed}}$ *finish, else go to* 2.

6. SIMULATION RESULTS

In our simulations, the following parameters have been adopted: the number of subcarriers M = 16, and the time frame length (in the intervals of T) L=32 (consequently, TF plane dimensions has been: 32 by 16). The transmit power constraint has been L times the normalized power of the signal in every time interval T. A fading channel with Ls=12 paths of exponentially decaying power has been selected, and the Doppler frequency fD = 10-2 (in samples-1). Let us stress, that we have to assume the perfect knowledge of the channel gains |Hl,m|. If these values change within the frame, we assume that an appropriate channel prediction has been applied. The assumed BEP: Prb = 10-3, and the maximum considered number of bits per symbol: 10. The exemplary channel realisation is shown in Fig. 1.

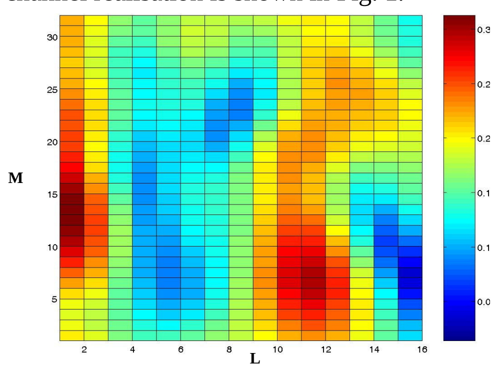

Fig. 1. Exemplary TF channel characteristic

Figs. 2 and 3 show the results obtained in the case when the original (but 2D) and modified H-H algorithms (described in the previous section) have been applied. In Fig. 4, achievable channel capacities versus SNR for the two variants of H-H algorithm described above.

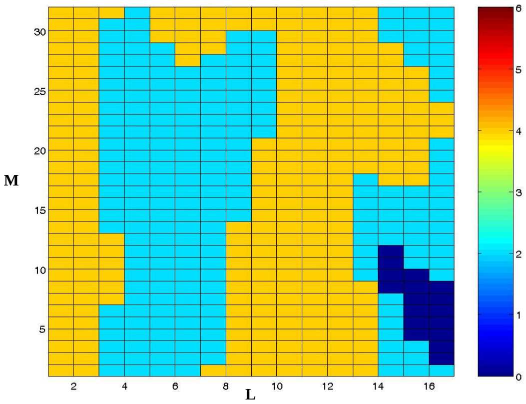

Fig. 2: Bit assignment using original 2D Hughes-Hartogs algorithm.

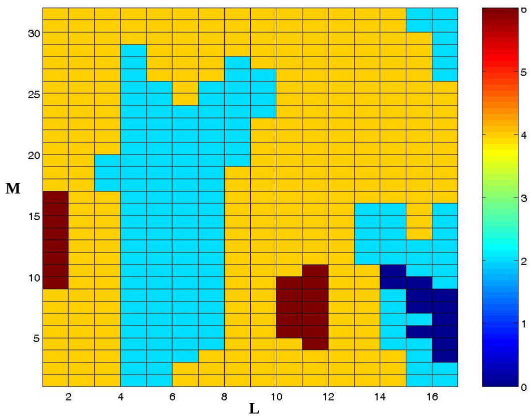

Fig. 3. Bit assignment using H-H algorithm modified for GMC transmission

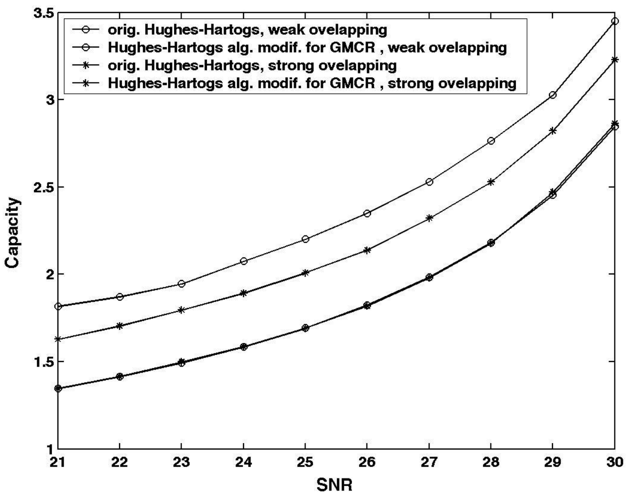

Fig. 4. Channel capacity vs. SNR obtained for original and modified H-H algorithm for two cases: strong and weak overlapping.

Two variants of atoms overlapping have been considered – strong overlapping (when N_g is 2.1 times higher than N) and weak overlapping (when $N_g = 1.2N$) One can observe that the in both cases the proposed modified algorithm ensures better power usage than the original one. The gain in average capacity is significant in the whole range of assumed values of SNR. Such observation allows to conclude, that the proposed modifications are important and by including them into the Hughes-Hartogs algorithm the comparable system capacity can be obtained as for the OFDM systems, where no overlapping is assumed at the transmitter. As a consequence, the modifications makes possible to implement the adaptive techniques into GMC systems.

7. CONCLOSIONS

We have shown that optimal bit and power loading for the GMC signaling maximizing the channel capacity is a complex problem. The TF representation of signals, the lack of orthogonality between the subcarriers and overlapping of signals has to be taken into account in the considered adaptive BPL technique. The mathematical analysis of the problem leads to the new formula (6) for optimal power allocation (and resulting bit allocation), which incorporates the interference between the Gabor atoms. This formula describes a large set of equations and the same large set of inequalities, which have to be solved jointly to obtain a vector of allocated power values. Moreover, for BPL the value of the water surface has to be calculated jointly with the power levels. Apart from the theoretical formulas a practical method has been proposed which modifies the Hughes-Hartogs algorithm for the GMC represented signals. The comparison of the effectiveness of the original and modified Hughes-Hartogs algorithms allows to conclude, that the inclusion of the atoms-interference in the computation of the power incremental matrix, leads to better spectrum usage.

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