Bit-loading algorithms and SNR estimate for HomePlug AV

E. Guerrini, G. Dell'Amico, P. Bisaglia, L. Guerrieri
Dora Spa, STMicroelectronics Group, Via Lavoratori Vittime del Col du Mont 24, 11100 Aosta, Italy
Email: eleonora-dora-spa.guerrini@st.com

Abstract— **In this paper two different bit-loading algorithms with uniform power allocation are compared within the framework of the HomePlug AV system. Both algorithms try to maximize the overall throughput: the first one guarantees a target mean BER, while the second guarantees a target BER per sub-carrier. By means of simulations, we compare the performance of the two algorithms, in terms of throughput and BER versus average SNR under a realistic power-line channel model. Results show that, when ideal SNR knowledge is assumed, the first and more complex algorithm offers a greater throughput; however, when SNR estimate is included, the performance of the two algorithms gets closer.**

Keywords— **Bit-loading, SNR estimate, OFDM, HomePlug AV.**

I. Introduction

HOMEPLUG, an industrial organization that now comprises more than 70 companies, was formed in 2000 to develop and standardize specifications for home networking technology using existing power-line wiring. The first release of HomePlug in June 2001, called HomePlug 1.0, with raw data rates up to 14 Mbit/s, was followed in July 2005 by a second release, called HomePlug AV [1].

The HomePlug AV (HPAV) system, supporting raw data rates up to 150 Mbit/s, employs adaptive orthogonal frequency division multiplexing (OFDM) over a bandwidth from 1.8 to 30 MHz, using 917 useful sub-carriers. The possible modulations that can be employed per sub-carrier are: BPSK, 4-QAM, 8-QAM, 16-QAM, 64-QAM, 256-QAM and 1024-QAM. The choice of the modulation to use over each sub-carrier should be made by a proper bit-loading algorithm, on the basis of the estimated per sub-carrier signal-to noise ratio (SNR); the algorithm is not however specified in [1]. In [2], [3] an extensive overview of the HomePlug AV is given; however, to our knowledge, no previous results have been published proposing possible bit-loading algorithms for the system and evaluating their performance under a realistic environment.

In this paper the problem of bit-loading, i.e. of properly allocating the bits over the different sub-carriers, is tackled. Furthermore, an algorithm for per sub-carrier SNR estimation is also investigated and the results presented. We note that a similar problem was already faced in [4] for a HomePlug-like system; unfortunately, the bit-loading algorithm proposed in [4] cannot be straightforwardly applied to HomePlug AV. In addition, the results presented in [4] refer to a completely different OFDM setting and a direct derivative of the HomePlug AV system performance appears to be difficult.

The first bit-loading algorithm that we investigate is an iterative discrete bit-loading algorithm with non-adaptive power allocation, as recently proposed in [5]. The algorithm tries to maximize the overall throughput of the system, guaranteeing a target mean bit error rate (BER). We note that this algorithm ideally suits the HomePlug AV specification requirements, that do not allow power re-allocation over sub-carriers. Conversely, the more common rate-adaptive algorithms [6] - as employed for instance in [4] - or margin-adaptive loading algorithms [7], [8], all re-distribute the power over the different OFDM sub-carriers depending on the SNR, hence they cannot be applied to the HomePlug AV system. In [5] two different algorithms are proposed that iteratively try to converge to an almost optimum allocation; the number of iterations, that needs to be minimized in order to limit the computational complexity, is obviously a crucial point of the algorithms. In this paper we merge the two algorithms of [5] in order to limit the computational complexity and remove the dependence on the SNR value of the minimization of the number of iterations.

The second algorithm we consider tries to maximize the overall throughput of the system with a uniform power allocation, similarly to the first one. In this case, however, the target BER is guaranteed per sub-carrier. We highlight that this second algorithm is much simpler to implement, since no convergence iterations are required, but simply a look-up table in order to store, for each allowable constellation, the required SNR to guarantee a target BER.

In order to include the effect of the SNR estimate on the two algorithms' performance, we also develop a data-aided per sub-carrier SNR estimator [9].

The remainder of this paper is organized as follows: in Section II the physical layer of the HomePlug AV system is reviewed, while details on the investigated bit-loading algorithms are illustrated in Section III; in Section IV the considered SNR estimate algorithm is presented; and numerical results are reported in Section V with conclusions given in Section VI.

II. System model

In Figure 1 the HomePlug AV physical layer [1], [2] on which we base our analysis is shown[1]. At the transmitter, the

[1]As explained in [2] the input bits from the MAC are structured by the HPAV transmitter differently depending on whether they are HPAV data, HPAV control information or HomePlug 1.0 control information. In this paper, for the sake of simplicity, we will refer to the more common HPAV data format only.

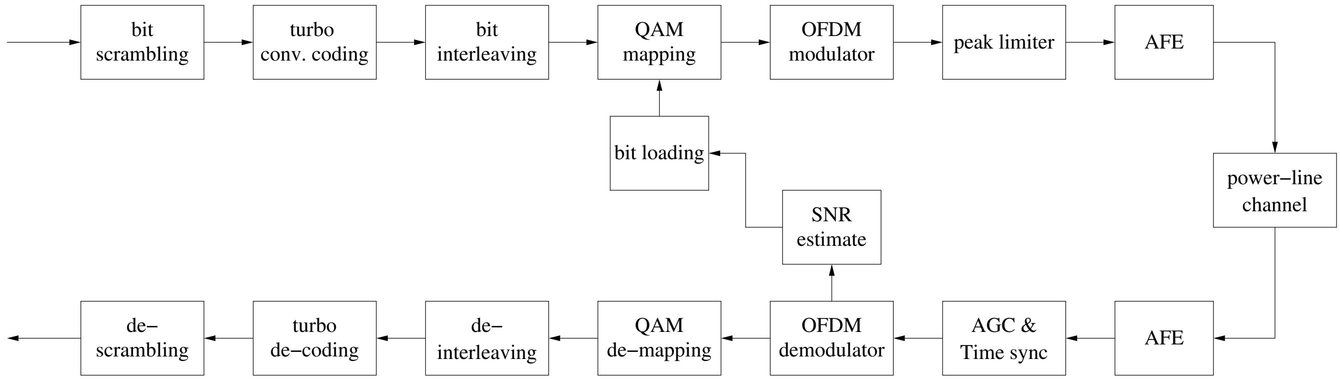

Fig. 1. Block diagram of the HomePlug AV physical layer for data transmission.

information bits for HPAV data transmission, after scrambling, are turbo convolutionally encoded, bit-by-bit interleaved and then converted into QAM symbols through a bit-mapper. The data symbols (belonging to unit power constellations) are serial-to-parallel-converted for OFDM modulation. Each OFDM sub-carrier can be differently loaded, depending on the estimated SNR per sub-carrier, with one of the following modulations: BPSK, 4-QAM, 8-QAM, 16-QAM, 64-QAM, 256-QAM and 1024-QAM. Both the bit-loading algorithms and the SNR estimate algorithm will be discussed in the following two sections. In HPAV the OFDM modulation is implemented by using a 3072-point inverse discrete Fourier transform (IDFT). Furthermore, to comply with frequency regulatory bodies, only 917 sub-carriers, out of the 1536 sub-carriers from DC to 37.5 MHz, are employed for useful transmission. To reduce the complexity of the receiver, a suitable cyclic prefix is used to remove both inter-symbol and inter-channel interference (ISI and ICI). Finally, before an analog front end (AFE) block, which sends the resulting signal to the power-line channel, a peak limiter block is inserted to minimize the peak-to-average power ratio (PAPR).

At the receiver the signal after the AFE block is fed to an automatic gain control (AGC) and time synchronism blocks. In this paper both blocks will be assumed to be ideal. After cyclic prefix removal and OFDM demodulation, which is implemented by the 3072-point discrete Fourier transform (DFT), and assuming the following: that the cyclic prefix completely eliminates ISI and ICI; that we can guarantee perfect synchronization; and that the channel is time invariant within each i-th OFDM symbol, the received signal over the generic sub-carrier k, for $k = 0, \ldots, N-1$, can be written as:

$$y_i[k] = G_i[k]a_i[k] + n_i[k] \tag{1}$$

$a_i[k]$, $G_i[k]$ and $n_i[k]$ denote the transmitted QAM symbol, the channel frequency response complex coefficient and the complex additive noise, over the generic k-th sub-carrier during the i-th OFDM symbol, respectively.

The output from the OFDM demodulator is then sent to a QAM de-mapper, a de-interleaver, a turbo convolutional decoder and a de-scrambler to reconstruct and estimate the transmitted bits.

III. Bit-loading algorithms

In the literature the bit-loading issue has been studied using two different main optimization approaches, which led to two families of algorithms, namely: *rate-adaptive* and *margin-adaptive* algorithms. The rate-adaptive algorithms [6] try to maximize the overall throughput of the system with a constraint on the transmit power. Whereas, the margin-adaptive algorithms [7], try to minimize the transmit power with a constraint on the overall throughput of the system. Both families are based on the possibility of re-distributing the power and the bits over the different OFDM sub-carriers depending on the estimated SNR. Unfortunately, regulatory constraints of HomePlug AV do not allow the exploitation of power re-allocation. Hence, the above two families of algorithms cannot be applied to the HomePlug AV system.

Recently, two iterative discrete rate-adaptive bit-loading algorithms with non-adaptive power allocation have been proposed in [5]; indeed, these nearly optimum algorithms ideally suit the HomePlug AV specification requirements. We recall that the first algorithm of [5] has a simpler structure but slower convergence, while the second has a faster convergence but it strongly depends on a parameter δ, whose optimum value is a function of the SNR.

In the following sections an alternative algorithm derived from [5] is presented; furthermore, another non-adaptive power allocation algorithm with a lower complexity is considered.

A. Decreasing BER-constrained (DBC) allocation algorithm

The adaptive bit-loading algorithms proposed in [5] try to solve the following problem[2]:

[2]In the following expressions, to simplify the notation, the suffix i denoting belonging to the i-th OFDM symbol has been omitted.

$$\begin{cases} \max\limits_{b[k]} \sum\limits_{k=0}^{N-1} b[k] \\ \overline{P} = \dfrac{\sum\limits_{k=0}^{N-1} b[k]P[k]}{\sum\limits_{k=0}^{N-1} b[k]} \leq P_T \end{cases} \tag{2}$$

where $b[k]$ is the number of bits over the k-th sub-carrier, $\overline{P}$ is the bit error rate (BER) of the system, P_T is the specified BER threshold, and $P[k]$ is the BER for sub-carrier k, which is determined from the sub-carrier SNR value $\Gamma[k]$. To compute the probability of bit error, closed-form expressions of all the modulation schemes that the system employs can be used [10]. The BER-constrained allocation algorithm is reported in Table I:

1) Best case: set the modulation scheme of all the sub-carriers to 1024-QAM.
   a) Given the SNR value $\Gamma[k]$ for each sub-carrier, determine $P[k]$ and compute $\overline{P}$.
   b) Compare $\overline{P}$ with P_T. If $\overline{P}$ is less than P_T, the current configuration is kept and the algorithm ends.
2) Worst case: search for the k-th sub-carrier with the highest SNR. Set the BPSK modulation and compute $P[k]$.
   a) Compare $P[k]$ with P_T. If $P[k]$ is greater than P_T, the target BER P_T cannot be reached and the algorithm ends.
3) Compute the value of $\hat{P}$, where $\hat{P}$ is the initial peak BER threshold (see Table II).
4) Set the modulation scheme of all sub-carriers to the maximum constellation that satisfies $P[k] < \hat{P}$, $\forall\ k = 0, 1, \ldots, N-1$.
5) Compute the current value of $\overline{P}$.
6) Compare $\overline{P}$ with P_T. If $\overline{P}$ is lower than P_T, the current configuration is kept and the algorithm ends.
7) Search for the sub-carrier with the worst $P[k]$ and reduce the constellation size. If $b[k] = 1$, null the sub-carrier (i.e., set $b[k] = 0$).
8) Recompute $P[k]$ and $\overline{P}$ with changed allocation and return to step 6.

TABLE I
BER-CONSTRAINED ALLOCATION ALGORITHM.

In order to reduce the complexity, instead of computing the values of $P[k]$ for each sub-carrier in real-time using close-form expressions, the algorithm employs a *memory look-up table*, which contains the P values at different SNR values[3] (Γ) for each allowable modulation.

[3] Γ and P denote the SNR value and the BER value, over a generic sub-carrier, respectively.

1) Initial peak BER calculation: The speed at which the iterative algorithm reaches its final allocation depends on the initial $\hat{P}$ chosen. $\hat{P}$ determines how much any given sub-carrier can individually exceed P_T, while $\overline{P}$ remains below it. The algorithm for finding the initial peak BER $\hat{P}$ estimate is as summarized in Table II.

The proposed algorithm is based on the simplicity of the first algorithm of [5], using, however, the starting-point-condition of the second of [5] to speed up convergence. The interested reader should refer to [5] for a thorough analysis of the other two algorithms.

We re-name the BER-constrained allocation with this new starting-point-condition: decreasing BER-constrained (DBC) allocation algorithm.

1) Given the sub-carrier SNR values $\Gamma[k]$, calculate $P[k]$ for all the different modulation schemes that could potentially be employed in the system.
2) Find $\beta[k]$, the largest $P[k]$ that does not exceed P_T.
3) Find $\alpha[k]$, the smallest $P[k]$ that exceeds P_T.
4) Determine ΔP to have several sub-carriers exceeding P_T while having $\overline{P}$ below P_T. In this case, we have

$$P_T = \frac{\sum \beta[k]b[k] + \Delta P}{\sum b[k]} \tag{3}$$

or equivalently

$$\begin{aligned} \Delta P &= P_T \sum b[k] - \sum b[k]\beta[k] \\ &= \sum b[k](P_T - \beta[k]) \end{aligned} \tag{4}$$

5) Sort the values of $\alpha[k]$ (and $b[k]$) in increasing order. Find the largest value of I for which

$$\Delta P \geq \sum_{i=0}^{I-1} b[i](\alpha[i] - P_T) \tag{5}$$

is true, and set $\alpha[I-1]$ as the initial $\hat{P}$ for the algorithm.

TABLE II
INITIAL PEAK BER CALCULATION FOR THE BER-CONSTRAINED ALLOCATION ALGORITHM.

B. BER threshold constant (BTC) for all sub-carriers algorithm

The second algorithm we consider, tries to maximize the overall throughput of the system with a uniform power allocation, similarly to the first one. In this case, however, the target BER is guaranteed per sub-carrier. Hence, the problem to solve is:

$$\begin{cases} \max\limits_{b[k]} \sum\limits_{k=0}^{N-1} b[k] \\ P[k] \leq P_T\ , \qquad k = 0, \ldots, N-1 \end{cases} \tag{6}$$

We highlight that this second algorithm is much simpler to implement, since no convergence iterations are required, but simply a look-up table. Hereafter, we will refer to this algorithm as the BER threshold constant (BTC) for all sub-carriers algorithm.

IV. Sub-carrier SNR estimate

The SNR estimate is a critical and essential element in every bit-loading algorithm. In the explanation that follows, we propose a data-aided per sub-carrier SNR estimator, which makes use of known 4-QAM symbols transmitted during special packets, called *packet sounds*. These are available in the HomePlug AV system, for the purpose of estimating, for each sub-carrier k, the SNR value $\Gamma[k]$, before starting the useful data transmission. We first evaluate

$$\begin{aligned}\frac{1}{S}\sum_{i=0}^{S-1}\frac{y_i[k]}{a_i[k]} &= \frac{1}{S}\sum_{i=0}^{S-1}\frac{G_i[k]a_i[k]+n_i[k]}{a_i[k]} \\ &= \frac{1}{S}\sum_{i=0}^{S-1}G_i[k]+\frac{1}{S}\sum_{i=0}^{S-1}\frac{n_i[k]}{a_i[k]} \qquad (7)\end{aligned}$$

where $y_i[k]$ was given in (1); $a_i[k]$ are the known 4-QAM symbols transmitted over the k-th sub-carrier during the i-th OFDM symbol; and S represents the number of known OFDM symbols transmitted during the packet sounds to estimate the SNRs. Assuming that the channel is time invariant over S OFDM symbols, i.e. $G_0[k] = G_1[k] = \ldots, G_{S-1}[k] = G[k]$, and S is sufficiently large enough to average out the noise (which is assumed to be zero mean), it follows that (7) represents a least squares (LS) estimate, averaged over S samples, of the channel frequency response over sub-carrier k. Hence, we can write [11]:

$$\hat{G}[k] = \frac{1}{S}\sum_{i=0}^{S-1}\frac{y_i[k]}{a_i[k]} \qquad (8)$$

Furthermore, let calculate

$$\begin{aligned}\frac{1}{S}\sum_{i=0}^{S-1}|y_i[k]|^2 &= \frac{1}{S}\sum_{i=0}^{S-1}|G_i[k]a_i[k]|^2+\frac{1}{S}\sum_{i=0}^{S-1}|n_i[k]|^2 \\ &+ \frac{1}{S}\sum_{i=0}^{S-1}G_i[k]a_i[k]n_i^*[k]+\frac{1}{S}\sum_{i=0}^{S-1}G_i^*[k]a_i^*[k]n_i[k] \qquad (9)\end{aligned}$$

Again, assuming that the channel is time invariant over S OFDM symbols and S is sufficiently big to average out the noise, exploiting the fact that $|a_i[k]|^2 = 1$ for unit power 4-QAM symbols, (9) can be assumed to be an estimate of the square amplitude frequency response of the channel plus the noise variance, hence we can pose:

$$|\hat{G}[k]|^2+\hat{\sigma}_n^2[k] = \frac{1}{S}\sum_{i=0}^{S-1}|y_i[k]|^2 \qquad (10)$$

where $\hat{\sigma}_n^2[k]$ is the estimate noise variance over the k-th sub-carrier. Hence, from (8) and (10) it turns out that an estimate of the sub-carrier SNR value $\Gamma[k]$ can be determined by:

$$\hat{\Gamma}[k] = \frac{\frac{1}{S^2}\left|\sum_{i=0}^{S-1}\frac{y_i[k]}{a_i[k]}\right|^2}{\frac{1}{S}\sum_{i=0}^{S-1}|y_i[k]|^2-\frac{1}{S^2}\left|\sum_{i=0}^{S-1}\frac{y_i[k]}{a_i[k]}\right|^2} \qquad (11)$$

We note that the estimator considered is reminiscent of the squared signal-to-noise variance (SNV) presented in [9] in the framework of single-carrier BPSK or QPSK systems corrupted only by AWGN.

V. Numerical results

In order to investigate the considered DBC and BTC algorithms under a realistic environment, we have developed a power-line channel model. One characteristic of the power-line channel is its cyclic variation [12] with the phase of the underlying AC line cycle. To achieve high data rate, power-line systems should exploit this cyclic variation by dividing the AC line cycle into multiple independent regions where the channel can be considered linear time invariant [13]. In each region an adaptive bit loading should be performed in order to determine the modulation to use over each sub-carrier. Based on this consideration, we have developed a linear time invariant multi-path channel model, as described in [14]. Furthermore, accordingly to [15], [16], two kinds of noise are also considered, namely: additive colored Gaussian noise and narrow-band interference. We note that in order to see only the effect of the bit-loading algorithms, in the following we will refer to an un-coded HomePlug AV system[4]. Hence, the impulsive noise has not been included in the numerical results, since it will have a massive detrimental effect on the considered system.

First of all, in Figure 2 we compare the performance of the two proposed algorithms (DBC with continuous lines and BTC with dashed lines) in terms of maximum average system throughput versus the average received SNR of the HomePlug AV system. The comparison is made in the case of ideal sub-carrier signal-to noise ratio knowledge Γ. The throughput is calculated by averaging the allocated bits of the bit-loading algorithms over many realizations of the considered power-line channel model. For a fixed target BER of $P_T = 10^{-3}$, results show that, the DBC algorithm, as expected, outperforms the BTC algorithm at all SNR values; in particular, the difference in throughput is of approximately 11 Mbit/s at 20 dB and reduces to 4.5 Mbit/s at 40 dB. As far as the number of iterations for the DBC convergence is concerned, results, not reported here, have shown that the number of iterations is comparable to the second algorithm of [5]. Furthermore, in Figure 2 we report the performance of the two algorithms

[4] We observe that for an un-coded HomePlug AV system the maximum raw data rate of 150 Mbit/s would increase up to 189 Mbit/s.

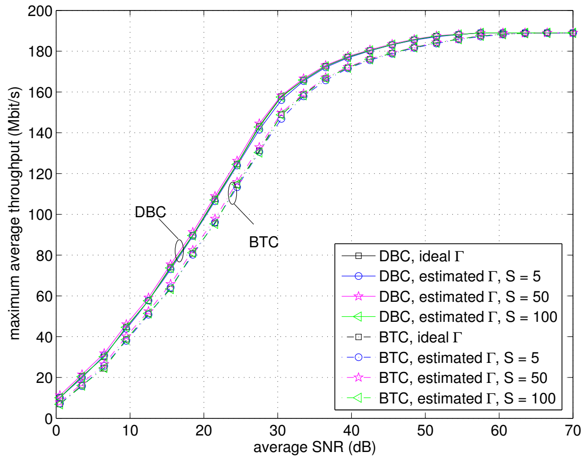

Fig. 2. Comparison of DBC (continuous lines) and BTC (dashed lines) algorithms with a target $P_T = 10^{-3}$. Maximum average HomePlug AV throughput versus average SNR, with ideal and estimated Γ, for $S = 5$, 50 and 100.

when sub-carrier SNR estimation is included for three different values of S, namely: 5, 50 and 100. The effect of imperfect SNR estimate, in the case of the first considered algorithm, was already investigated in [5], [17]; we note, however, that the sub-carrier channel estimation error was only approximated by a Normal distribution, but an SNR estimation algorithm was not implemented. Results, in terms of maximum average throughput, show that the curves are almost invariant when the SNR estimation effect is included. This can be explained by the fact that the estimated $\hat{\Gamma}[k]$, over a sub-carrier k, can either cause a lower bit allocation (when $\hat{\Gamma}[k] < \Gamma[k]$) or a higher bit allocation (when $\hat{\Gamma}[k] > \Gamma[k]$) on that sub-carrier, respectively. On average these effects on different sub-carriers balance out each other, and the resulting aggregate throughput is almost the same as for the ideal SNR case. Unfortunately, when $\Gamma[k]$ is overestimated over a sub-carrier k, i.e. $\hat{\Gamma}[k] > \Gamma[k]$, the bit error rate $P[k]$ can increase significantly over that sub-carrier and this effect is not balanced by a lower $P[l]$ due to an underestimated $\Gamma[l]$ over a sub-carrier l. To see these effects, in Figure 3 we compare the two algorithms (DBC on the left and BTC on the right) in terms of average BER versus average received SNR for a fixed target BER of $P_T = 10^{-3}$. Comparing the DBC and the BTC algorithms in the case of ideal Γ, we note that the DBC algorithm guarantees an average BER very close to the target, while the BTC is much further away. This greater margin of the BTC from the target, however, causes a loss in throughput, as shown in Figure 2. When sub-carrier SNR estimation is included (investigated cases are for $S = 5$, 50 and 100) for both algorithms the average BER always increases, as explained above. In particular, estimating Γ causes the BER of the DBC algorithm to exceed the target BER, even when the most accurate estimation is performed ($S = 100$). Whereas, the BER of the BTC algorithm, when a coarse estimation is performed ($S = 5$), is below the target for SNR values greater than 30 dB, while it is always below the target when the most accurate estimations are performed ($S = 50$ and 100). This is due to the intrinsic margin of the BTC algorithm.

In order to increase the robustness of the DBC algorithm and meet the bit error rate requirements, we investigate the solution of introducing a margin, against error estimation, in the signal-to-noise ratios at the receiver. That is to say, if Γ is the minimum SNR that guarantees a certain P, when we introduce a margin Δ, the bit-loading is done assuming $\Gamma + \Delta$ as the new minimum SNR. Obviously, Δ is a system parameter that needs to be optimized.

Figures 4 and 5 depict the maximum average throughput and the average BER versus the average received SNR for the DBC algorithm, when a margin is introduced. For each S value, the margin Δ has been optimized by simulations, i.e., the value has been selected that guarantees the maximum throughput, still satisfying the target P_T. In order to make a comparison, the DBC and the BTC algorithms with ideal Γ are also reported in Figure 4. We notice that the introduction of the margin causes a degradation in terms of throughput, although it allows meeting the target $P_T = 10^{-3}$, as shown in Figure 5. For instance, at an SNR of 20 dB the throughput with the DBC algorithm and estimated Γ with $S = 5$ losses 19 Mbit/s from the ideal DBC, and 8 Mbit/s from the ideal BTC. When Γ is estimated with $S = 50$ the loss from the ideal DBC reduces to 6 Mbit/s and the performance with respect to the ideal BTC algorithm now results 5 Mbit/s better. We note that (result not shown here) with $S = 100$ the loss in throughput, with respect to the ideal DBC, would be even lower.

To summarize, when a coarse estimation ($S = 5$) is performed, the DBC and BTC algorithms would require a margin to meet the BER requirements with a significant degradation

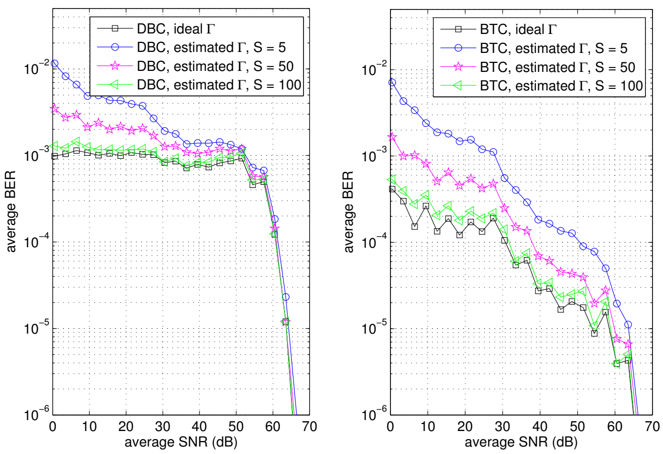

Fig. 3. Comparison of DBC (on the left) and BTC (on the right) algorithms with a target $P_T = 10^{-3}$. Average HomePlug AV BER versus average SNR, with ideal and estimated Γ, for $S = 5$, 50 and 100.

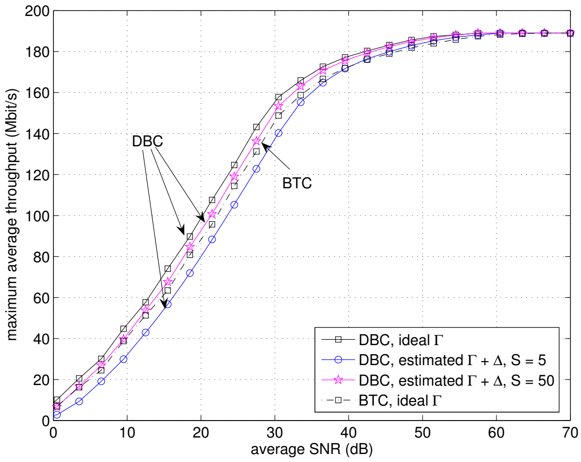

Fig. 4. DBC algorithm with a target $P_T = 10^{-3}$. Maximum average HomePlug AV throughput versus average SNR, with estimated $\Gamma + \Delta$, for $S = 5$ and $S = 50$.

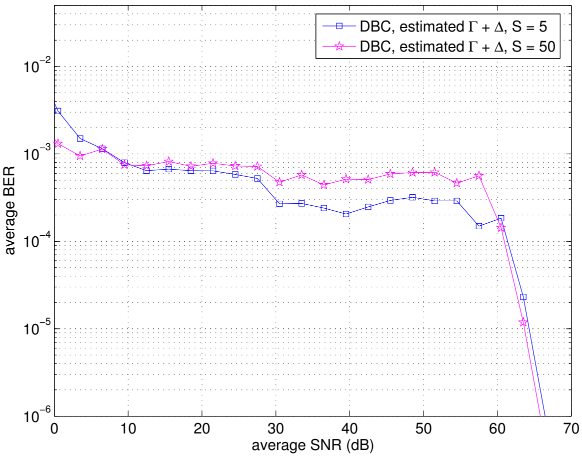

Fig. 5. DBC algorithm with a target $P_T = 10^{-3}$. Average HomePlug AV BER versus average SNR, with estimated $\Gamma + \Delta$, for $S = 5$ and $S = 50$.

in both throughputs. When more accurate estimations are performed ($S = 50$ or 100), the BTC already meets the target BER, while the DBC still requires a margin, which has, however, the following drawbacks: i) it reduces the gain of the DBC in terms of throughput with respect to the BTC, ii) it increases the design complexity of the DBC with respect to the BTC even further because of the margin.

VI. Conclusions

In this paper two uniform power allocation bit-loading algorithms have been compared using the HomePlug AV physical layer as a platform and a typical channel model for in-home power-line networks. Results have shown that, when ideal sub-carrier SNR knowledge is assumed, the first and more complex algorithm offers a greater throughput. However, when sub-carrier SNR estimate and, if necessary, a margin are included, performance of the two algorithms gets closer. Hence, the choice of the DBC algorithm over the BTC should consider whether or not its gain in throughput, with a reasonable length SNR estimator, justifies the extra cost.

References

[1] HomePlug PowerLine Alliance, "HomePlug AV baseline specification," Version 1.0.00, Dec. 2005.

[2] "HomePlug AV white paper," http://www.homeplug.org.

[3] K. H. Afkhamie, S. Katar, L. Yonge and R. Newman, "An overview of the upcoming HomePlug AV standard," *ISPLC 2005*, pp. 400-404, April 2005.

[4] S. Morosi, D. Marabissi, E. Del Re, R. Fantacci and N. Del Santo, "A rate adaptive bit-loading algorithm for in-building power-line communications based on DMT-modulated systems," *IEEE Trans. on Power Delivery*, vol. 21, pp. 1892-1897, Oct. 2006.

[5] A. M. Wyglinski, F. Labeau and P. Kabal, "Bit loading with BER-constraint for multicarrier systems," *IEEE Trans. on Wireless Comm.*, vol. 4, pp. 1383-1387, July 2005.

[6] B. S. Krongold, K. Ramchandran and D. L. Jones, "Computationally efficient optimal power allocation algorithms for multicarrier communication systems," *IEEE Trans. on Comm.*, vol. 48, pp. 23-27, Jan. 2000.

[7] J. Campello, "Practical bit loading for DMT," *ICC 1999*, pp. 801-805, June 1999.

[8] N. Papandreou and T. Antonakopoulos, "Dynamic bit-loading in pDSL communications systems," *ISPLC 2005*, pp. 356-360, April 2005.

[9] Pauluzzi and N. C. Beaulieu, "A comparison of SNR estimation techniques for the AWGN channel," *IEEE Trans. on Comm.*, vol. 4, pp. 1681-1691, Oct. 2000.

[10] J. G. Proakis, *Digital Communications*, McGraw-Hill, 3rd ed., 1995.

[11] J. van de Beek, O. Edfors, M. Sandell, S.K. Wilson and P.O. Börjesson, "On channel estimation in OFDM systems," *VTC 1995*, pp. 815-819, July 1995.

[12] F.J. Canete, J.A. Cortes, L. Diez, J.T. Entrambasaguas and J.L. Carmona, "Fundamentals of the cyclic short-time variation of the indoor power-line channels," *ISPLC 2005*, pp. 157-161, April 2005.

[13] S. Katar, B. Mashburn, K. Afkhamie, H. Latchman and R. Newman, "Channel adaptation based on cyclo-stationary noise characteristics in PLC systems," *ISPLC 2006*, pp. 16-21, March 2006.

[14] H. Phillips, "Modeling of powerline communication channels," *ISPLC 1996*, pp. 724-728, Nov. 1996.

[15] M. Zimmermann and K. Dostert, "Analysis and modeling of impulsive noise in broad-band powerline communications," *IEEE Trans. on Elect. Comp.*, vol. 44, pp. 249-258, Feb. 2002.

[16] R. Hormis, I. Berenguer and X. Wang, "A simple baseband transmission scheme for power line channels," *IEEE Journal on Select. Areas in Comm.*, vol. 24, pp. 1351-1363, July 2006.

[17] A. M. Wyglinski, F. Labeau and P. Kabal, "Effect of imperfect SNR information on adaptive bit loading algorithms for multicarrier systems," *IEEE Globecom 2004*, pp. 3835-3839, Nov. 2004.