# System Parameters Effect on DMT-Based Broadband Indoor Power Line Communications

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**ABSTRACT**

The main purpose of this paper is the determination of DMT (Discrete Multitone) modulation main parameters in a practical system for broadband indoor power line communications. Specifically, the influence of the length of the cyclic prefix and the number of carriers in the performance is investigated. The use of a time equalizer (TEQ), as a mean for reducing the cyclic prefix length, is also studied. Bit-rates attained with several parameter combinations are evaluated, so that, depending on the affordable complexity, one of them can be chosen.

## I. INTRODUCTION

DMT modulation characteristics make it especially suited for power line channels. Regardless of the well-known fact that multicarrier modulation (with infinite number of tones) is the optimum system for frequency shaped channels, in the sense of achieving the theoretical channel capacity, it has other interesting properties that allow to cope with the power line channel impairments. Due to the longer DMT symbols (compared to single-carrier systems), neither intersymbol interference (ISI) nor the impulsive noise introduced by the channel will be as destructive as in single-carrier systems. The existence of transmission gaps (non-permitted bands), because of electromagnetic compatibility problems, is easily overcome by DMT systems, which can fully exploit the spectral resources without the necessity of large continuous bands. Additionally, an adaptive DMT system can be used to minimise the effect of the channel variations in the performance.

It is very important to select an adequate number of carriers and cyclic prefix length. If the number of tones is too small, the above benefits will vanish. On the contrary, a very large number of carriers will result in very low distortion but may lead to an impractical implementation complexity and excessive delay. The cyclic prefix has a twofold impact on the bit-rate. On one hand, increasing its length reduces DMT symbol rate but, on the other hand, it improves the signal to noise and distortion ratio (SINAD), within a certain range, and consequently the number of bits per tone.

To perform this study, a wide range of indoor power line channels, in the frequency range up to 30MHz, have been generated using a software model [1] based on the interconnection of multiple transmission lines in a tree-form structure. In section II, this model and the channels selected for the analysis are described. A basic DMT system, along with a more complex one that includes a TEQ to shorten the effective impulse response of the channel, are presented in section III. The influence of the cyclic prefix length and the number of carriers in the signal to distortion ratio (SDR) is assessed in section IV, where SINAD and performance, in terms of attainable bit-rate, are also evaluated for each system in different noise scenarios. The reduction in the cyclic prefix length achieved by means of the TEQ is investigated in section IV as well. Finally, in section V, main conclusions from the above analysis are summarised.

## II. CHANNEL MODEL

The impulse responses used in the subsequent study have been obtained using a simulation program in which the indoor power line network is considered as a set of multiple transmission lines, terminated in loads of diverse nature, and interconnected to produce a tree-like structure, as presented in [1]. Every device connected to the power grid represents a load to the network, and is modelled with a complex impedance value and a noise source. Both time variant (depending on whether the device is in working state or not) values are taken from the characterisation of typical electrical appliances. Regarding the noise power spectral density (PSD), in addition to the ones given by the software model, results based on measurements [2] have been utilised.

### *A. Channel frequency response*

A great number of indoor residential channels have been generated and classified according to their attenuation profiles. Representative elements from the groups *worst* and *intermediate-to-good* have been chosen for presenting the simulations results. The one from the first set corresponds to a link of an about 40 m length (from now on, channel configuration 1), and the other to a link of approximately 20 m length (from now on, channel configuration 2). Fig.1 shows their amplitude frequency responses.

Main characteristics of the power line channels can be observed, namely: presence of deep notches, great variation of the attenuation with frequency and high dispersion of the amplitude values among channels (it must be emphasised that both responses correspond to a unique residential configuration).

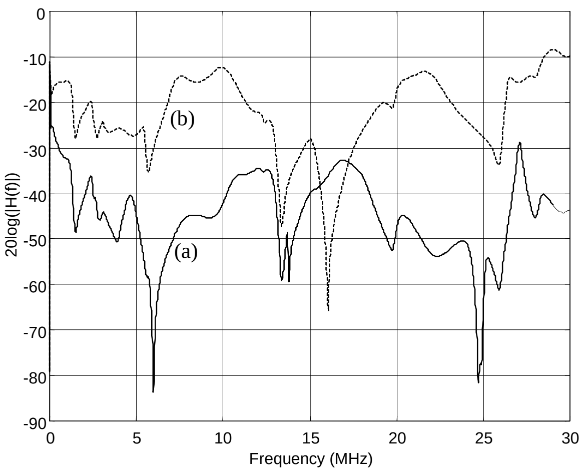


Figure 1. Amplitude frequency responses of the selected channels. (a) config.1, (b) config.2

### *B. Noise PSD*

Two noise PSD have been considered for each of the above responses: the one given by the simulator, shown in fig.2, and an Additive White Gaussian Noise (AWGN) with a PSD of –65 dBm/kHz. The latter is an upper bound of the worst noise scenario measured in a residential environment [2], and will be used to obtain lower limits for the bit-rate.

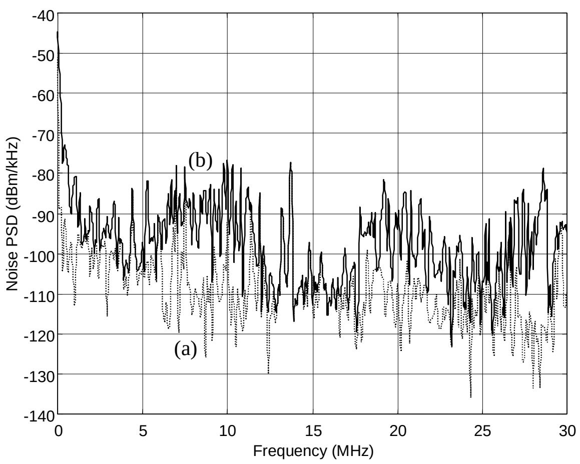


Figure 2. Modelled noise PSD (dBm/kHz) for channel config. 1 (a) and channel config. 2 (b)

## III. DMT SYSTEM MODEL

DMT is a particular case of multicarrier modulation. The basic principle of this technique is to divide the incoming data flow into several streams, each of them used to modulate different carriers that are transmitted simultaneously. Therefore, it can be seen as a combination of a generalised pulse amplitude modulation (PAM) and a multipulse modulation.

Denoting by $X_k$, k=0,1…N-1, the complex subsymbols to be transmitted, the discrete-time expression of a DMT signal (sampled at the Nyquist frequency) with M carriers can be written as a sequence of consecutive DMT symbols of the form

$$x[n]=\sum_{k=0}^{N-1} X_k e^{j\frac{2\pi}{N}\cdot k\cdot n}, \qquad 0\le n\le N-1 \qquad (1)$$

with N=2M. To ensure real values for the transmitted samples it is necessary to impose the conjugate symmetry constraint

$$X_k = X^*_{N-k}, \qquad k=1\ldots N-1 \qquad (2)$$

with $X_0$ and $X_{N/2}$ real valued.

Expression (1) can be seen as the IDFT (Inverse Discrete Fourier Transform) of N symbols $X_k$, k=0...N-1, and hence, the demodulation process can be done by computing the DFT of each DMT symbol.

With $N\to\infty$, this modulation is optimum in the sense of achieving the theoretical channel capacity, but because of the finite number of carriers, ISI (Intersymbol Interference) and ICI (Intercarrier Interference) appear. When $v$+1 is the length of the channel impulse response, these effects can be eliminated using a cyclic prefix (CP), in the form of the last $v$ samples of the DMT symbol, x[N-1-$v$]...x[N-1], that is transmitted at the beginning of each symbol and is eliminated at the receiver before the DFT. Hence, the longer the cyclic prefix, the greater the part of the transmitted energy that is not used in the detection process.

However, the actual impulse response of a power line channel is of significant duration. This means that very long CP's would be needed and, since it conveys no information, important energy loss would occur. In an attempt to shorten this length a TEQ is used. Denoting the impulse response of the channel by h[n], and the equivalent one of the TEQ by w[n], the employed design criterion [3] has been to obtain an effective response $h_{eff}[n]=h[n]*w[n]$ (where $*$ is the convolution operation) with just $v+1$ non zero samples. This can be expressed in the form

$$\sum_{k=0}^{N_w-1} h[n-k]\cdot w[k]=0, \quad \begin{matrix} n=0\ldots D-1 \\ n\ge D+v+1 \end{matrix} \qquad (3)$$

where $N_w$ is the length of w[n], and D is the time index of the first non zero sample of $h_{eff}[n]$. To prevent the all zero trivial solution, the unit-tap constraint (4) is applied.

$$\sum_{k=0}^{N_w} h[n-k]\cdot w[k]=1, \quad n=D \qquad (4)$$

Nevertheless, in practice is not possible to achieve this objective, and some residual energy will remain out of the $v+1$ samples window after the equalization process, distorting, in the form of ISI and ICI, the received symbols, $Y_k$.

After the IDFT, the different attenuation of the carriers is compensated for by a zero-forzing frequency equalizer (FEQ). In this way, the block diagram of the whole system is depicted in fig.3.

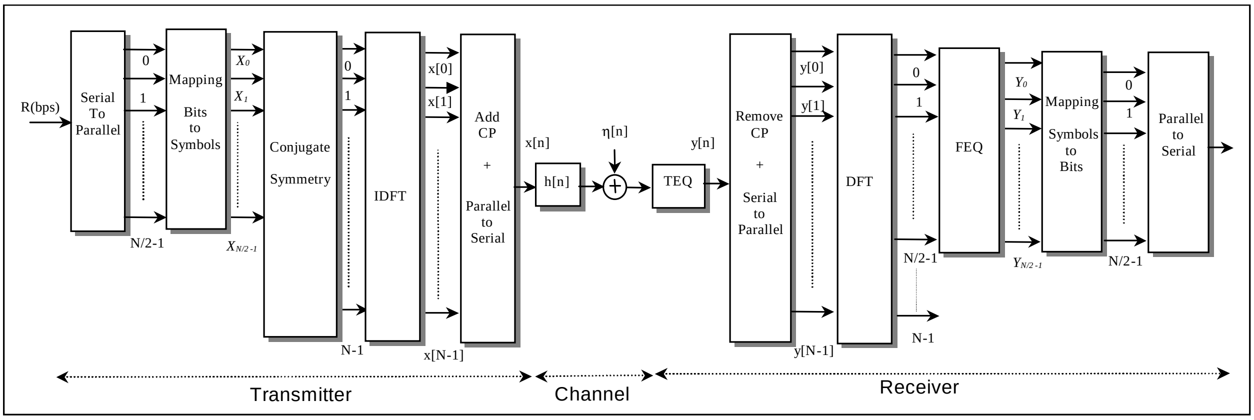


Figure 3. Discrete-time DMT system model

## IV. SYSTEM PARAMETRISATION

Parametrisation is assessed in two phases. In a first instance, the performance of a system without TEQ, as a function of the cyclic prefix length, $v$, and the number of carriers, M, is analysed. Once this has been done, the reduction in the cyclic prefix length obtained by means of a TEQ, with the number of taps, $N_w$, as a parameter, is studied.

Performance is measured in achievable bit-rate. Ideally, without any constraint in the alphabet and assuming a large number of carriers, it can be approximated by

$$R = \frac{f_s}{N+v} \sum_{k=0}^{N/2-1} \log_2\left(1 + SINAD_k\right) \;\; bits/s \qquad (5)$$

where $f_s$ is the sampling frequency, 60MHz in our case, and $SINAD_k$ is the signal to noise and distortion ratio (SINAD) on the $k^{th}$ carrier. Denoting by $S_k$, $D_k$ and $N_k$ the power of the signal, distortion (caused by ISI and ICI) and noise, respectively, at the output of subchannel k (after the FEQ), it may be written

$$SINAD_k = \frac{S_k}{D_k + N_k}, \quad SDR_k = \frac{S_k}{D_k} \qquad (6)$$

where $SDR_k$ is the signal to distortion ratio on the $k^{th}$ carrier.

### *A. System parametrisation without TEQ*

In order to assess the impact of the distortion on the bit-rate, the SDR is investigated at first. Simulations with 512, 1024, 2048 and 4096 carriers have been performed. For each of these systems, the relative length of the cyclic prefix with respect of the total length of the DMT symbol (N+$v$) is varied from 5% to 20%. The PSD of the transmitted signal is fixed to –20dBm/kHz [2]. Results for channel config.1 (continuous line) and channel config.2 (dashed line) are presented in fig.4.

The influence of the number of carriers is, in general, quite straightforward and independent of the considered channel, each time the number of tones doubles the averaged SDR increases around 11dB.

Regarding the cyclic prefix, the greatest gains are attained when the relative length is augmented from 5% to 10%. From this point on, each time it is augmented a 5%, smaller improvements, which also depend on the number of carriers of the system, are obtained.

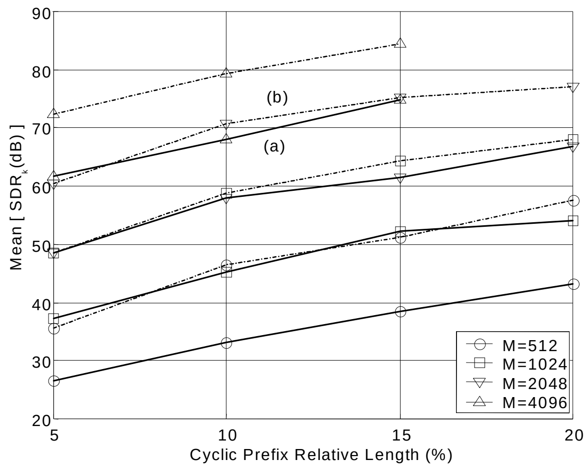


Figure 4. Averaged SDR for different number of carriers and cyclic prefix lengths for channel config.1 (a) and channel config.2 (b)

Once the effect of the system parameters in the SDR has been evaluated, it is time to take into account the noise introduced by the channel. This will avoid selecting values that, although lead to lower distortion figures (with the cost of a very high implementation complexity), may have a negligible effect on the SINAD and hence, in the bit-rate.

To this end, each channel configuration has been initially analysed in its corresponding noise scenario from fig.2. Results are shown in fig.5. As seen, except for the case of 512 tones, SINAD remains nearly constant when lengthen the cyclic prefix from 15% to 20%. Equivalently, for relative lengths equal or greater than 10%, minimum gains are experienced when the number of carriers goes from 2048 to 4096.

Secondly, the study has been repeated for the -65dBm/kHz AWGN case, from now on referred to as *worst case.* In this situation, it has been found that SINAD is dominated by the noise, even for 512 tones and a 5% of relative length.

After assessing the contributions of noise and distortion to the SINAD, the definitive step for selecting the most appropriate parameters is to evaluate the bit-rate achieved by each system in the two noise scenarios previously examined. When doing this for the ones depicted in fig.2, the curves displayed in fig.6 are obtained, where for 4096 and 2048 tones, lengths shorter than 5% have also been included to clearly observe the trends in the curves.

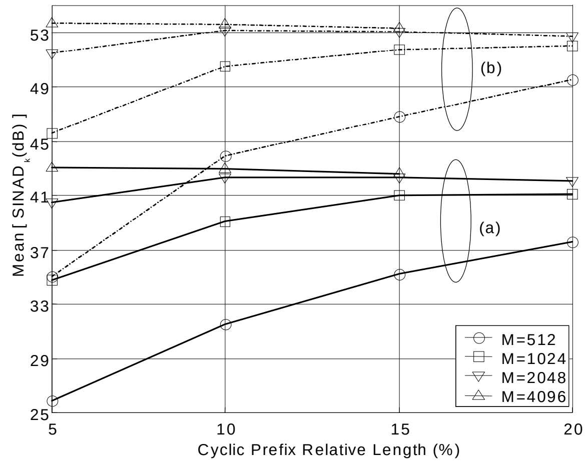


Figure 5. Averaged SINAD for channel config.1 (a) and channel config.2 (b) in its corresponding noise scenario from Figure 2.

As seen, there is an optimum relative cyclic prefix length that depends on the number of carriers (although when translated to absolute duration, nearly the same value –around 3.83 μs, 230 samples at a sampling rate of 60MHz- results for all of them). The reason is that for lower cyclic prefixes SINAD is dominated by SDR. In these circumstances, when *v* is augmented, the increment experienced by the numerator in (5) (i.e. the number of bits per carrier) is greater than the one of the denominator (i.e. the DMT symbol length in samples), resulting in higher bit-rates. This process continues until SINAD starts to be dominated by noise. From this point on, lengthen the cyclic prefix reduces the frequency of the DMT symbols but produces no increment in the number of bits per tone, so the bit-rate decreases.

Notice that fig.6 is a useful tool to identify some parameter combinations that report insignificant improvements but involve additional efforts in complexity.

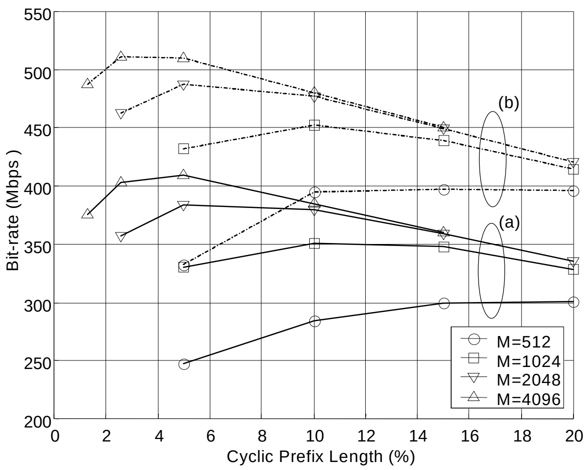


Figure 6. Bit-rates for channel config.1 (a) and channel config.2 (b) in its corresponding noise scenario from Figure 2.

In the *worst case*, due to the previously mentioned behaviour of the SINAD in this situation, the bit-rate curves in the two channel configurations are monotonically decreasing with the cyclic prefix length. Maximum values are around 40 Mbps for config.1 and 210 Mbps for config.2. However, it must be taken into account that this is not, by far, the most common noise scenario. In this sense, even if the selection is done based on the combined bit-rate from both cases, optimum systems in fig.6 are still the most appropriate ones.

### B. *Worst case analysis*

A noise PSD of–65dBm/kHz is one of the worst situations that can be found in a residential environment. In addition, since config.1 and config.2 can be taken as representative elements of the *worst* and *intermediate-to-good* ensembles of possible channels, performances obtained in these circumstances represent lower bounds for indoor power line communications. To better approximate the bit-rate that a real system would attain in these conditions, the signal constellations have been constrained to MQAM (M=0,2,4,16,...). That is, each carrier is assigned a certain number of bits depending on its SINAD, with the objective of maintaining an error probability lower than $10^{-6}$. Results, with the PSD of the transmitted signal as a parameter, are presented in Table 1.

| | Channel config.1 | | | Channel config.2 | | |
|---|---|---|---|---|---|---|
| | PSD of the transmitted signal (dBm/kHz) | | | | | |
| System Parameters | -22 | -20 | -18 | -22 | -20 | -18 |
| M=512, cp=15% | 0.95 | 2.30 | 4.33 | 73.40 | 91.57 | 99.19 |
| M=1024, cp=10% | 1.19 | 2.82 | 5.35 | 83.70 | 99.46 | 114.1 |
| M=2048, cp=5% | 1.40 | 3.35 | 6.94 | 93.02 | 106.7 | 120.6 |
| M=4096, cp=2.6% | 1.58 | 3.64 | 7.77 | 98.16 | 110.2 | 121.8 |

Table 1. Bit-rates (Mbps) for channel config.1and channel config.2 in the *worst case* noise scenario with uncoded BPSK and square QAM.

From Table 1 it can be inferred that when practical constellations are used, the transmission PSD has a decisive influence on the bit-rate of bad channels, e.g. in config.1, a 2 dB increment multiplies the bit-rate by a factor greater than 2. Nevertheless, it must be noted that no coding is used. It would not be difficult to gain 4dB by means of a trellis code, and thus, reduce this effect.

### C. *System parametrisation with TEQ*

In this subsection, the use of the aforementioned TEQ (section III), as a mean for reducing the cyclic prefix length, is studied.

Since the TEQ involves additional computational load, to assess its influence only the system with 512 tones has been analysed. Results from simulations performed for both channel configurations in the noise scenarios of fig.2 are depicted in fig.7 and fig.8. For reference, the maximum bit-rates that can be attained with 512 and 1024 carriers but without a TEQ are also shown (straight lines).

As expected, lower gains are obtained in channel config.2 (where the impulse response is inherently shorter) than in config.1. It is interesting to note that, if a TEQ is used, it is worthless to use cyclic prefix longer than 10%. In relation to the number of taps, the most appropriate one depends on the relative cyclic prefix length and on the channel configuration. However, no less than 15 taps (approximately) should be used if we want to get appreciable benefits in both channels. At this stage, it is of particular interest to compare the computational requirements for the different systems. To this end, only the major functions of the transmitter (i.e. the IFFT) and receiver (i.e. the FFT and the FEQ) have been considered. Thus, the implementation of the optimum system with 1024 tones needs approximately 17% more instructions per second (IPS) than the optimum with 512 carriers. A two taps TEQ involves 40% more IPS. With these values, it can be concluded that a TEQ is not a good option for power line channels.

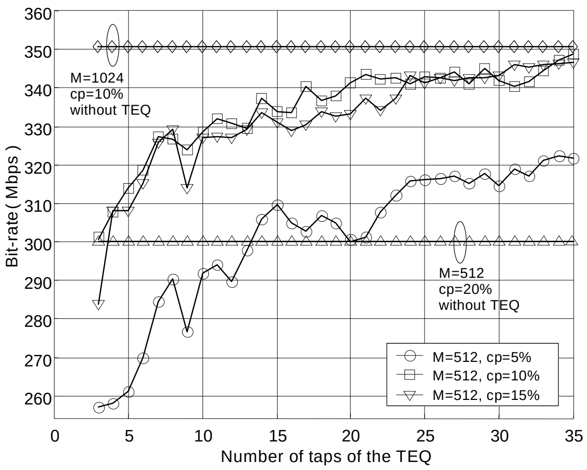


Figure 7. Bit-rates versus the number of taps of the TEQ for channel config.1. A system with 512 tones and different relative cyclic prefix lengths is used.

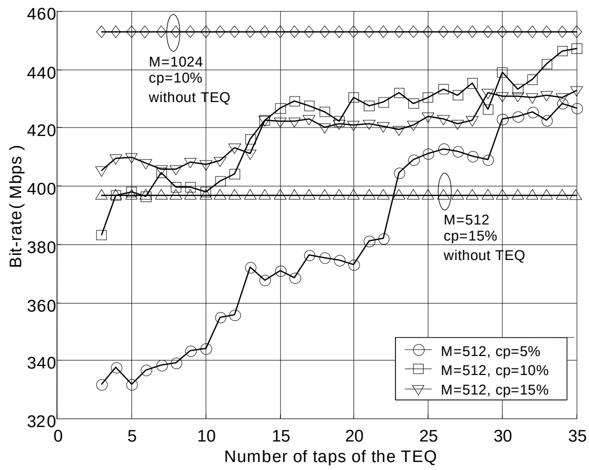


Figure 8. Bit-rates versus the number of taps of the TEQ for channel config.2. A system with 512 tones and different relative cyclic prefix lengths is used.

In the *worst case*, as SINAD is dominated by the noise, the use of a TEQ provides no improvement on the bit-rate.

## V. CONCLUSIONS

Parametrisation of the DMT modulation for indoor power line communications in the frequency range up to 30MHz has been investigated. The performance, in terms of bit-rate, of systems with 512 to 4096 carriers, and relative cyclic prefix lengths from 5% to 20%, have been analysed in a *worst case* channel and in an *intermediate to good* one in different noise scenarios. It has been found that, once the number of tones is fixed, there is an optimum relative length that maximises the bit-rate in both channel configurations. Furthermore, when translated to absolute duration, nearly the same value results for all of them. Sets of parameters that produce negligible improvements but may involve increased efforts in complexity or energy efficiency have been identified.

In addition, lower bounds for the performance of practical systems have been presented, assessing the influence of the transmission PSD.

Finally, in order to shorten the cyclic prefix, the use of a TEQ has been studied by evaluating the bit-rates as a function of the number of taps. It has been found that the improvements obtained do not compensate, by far, the increments in the computational complexity.

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