Green Hybrid FMT for WLAN Applications

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***Abstract*—In this paper we propose a novel multicarrier (MC) architecture to address the energy saving problem in WLAN applications. We refer to this scheme as hybrid filtered multitone (H-FMT). Depending on the channel condition, H-FMT wisely switches between Filtered Multitone Modulation (FMT) with orthogonal pulses and adaptive orthogonal frequency division multiplexing (A-OFDM). We show that H-FMT provides a considerable gain in terms of energy efficiency compared to conventional OFDM, yet achieving the same rate. Furthermore, its additional computational complexity is negligible, making this scheme an attractive solution for WLAN applications.**

I. Introduction

In recent years the energy consumption problem of communication devices has had much consideration. Some standards have been developed to cope with the need of energy saving, e.g., the IEEE 802.3az and the HomePlug Green Phy that are respectively specified for Ethernet and power line communications. Power saving also plays a fundamental role in wireless applications since the communication devices can be battery driven. In this paper we consider a WLAN application scenario and we investigate the energy efficiency of a novel hybrid multicarrier (MC) modulation scheme [1]. The scheme is referred to as hybrid filtered multitone (H-FMT). Depending on the channel condition, H-FMT switches between orthogonal filtered multitone (FMT) modulation and adaptive orthogonal frequency division multiplexing (A-OFDM).

The FMT system is a discrete-time implementation of MC modulation that uses uniformly spaced sub-carriers and identical sub-channel pulses. OFDM can be viewed as an FMT scheme that deploys rectangular time domain filters. FMT originally was proposed for application over broadband wireline channels [2], and subsequently it was investigated for applications over wireless channels [3]. Short orthogonal FMT (SO-FMT) is an FMT that deploys orthogonal filters with minimal length equal to the duration of one transmitted symbol. This renders the system interference free in ideal conditions and makes the filters maximally confined in the frequency domain. Consequently, both ICI and ISI are well mitigated when signaling over a frequency selective channel. Furthermore, the computational complexity is comparable to that of OFDM [4].

As mentioned above, the OFDM system uses a rectangular impulse response prototype pulse. In presence of frequency selective channel, the cyclic prefix (CP) is needed to mitigate the interference components. If the CP is longer than the channel impulse response, the system maintains its perfect reconstruction property, thus, no interference is present at the decision stage. Clearly, when the CP is shorter than the channel duration, ISI and ICI terms arise [5]. In a previous work [6], we have found that over typical WLAN channels, the CP has not to be necessarily as long as the channel duration to maximize the achievable rate. Furthermore, for each channel class of the IEEE 802.11n WLAN channel model [7], we have found a nearly optimal value of CP designed according to the statistics of the capacity optimal CP duration. We refer to the OFDM that adapts the CP to the channel condition as A-ODFM. Thanks to the similarities of the efficient implementation of both SO-FMT and A-OFDM [4], the realization of H-FMT introduces a marginal increase in computational complexity w.r.t. the conventional OFDM.

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In this paper we exploit the peculiarities of H-FMT to cope with the problem of the energy saving in WLAN applications. Since the H-FMT system is based on A-OFDM and SO-FMT, it generally suffers from interference making the problem of power allocation not convex. Therefore, it cannot be solved by conventional methods such as Lagrange multipliers. For this purpose, we present an algorithm that significantly simplifies the power minimization problem. This algorithm allows us to assume convex the power allocation problem assuming that H-FMT suffers of negligible interference.

We compare H-FMT with the OFDM used in the 802.11 standard and we show that H-FMT affords a considerable gain in terms of energy efficiency compared to OFDM, yet reaching the same rate.

The paper is organized as follows. In Section II, we present the system model and we recall the fundamentals of A-OFDM, SO-FMT, and H-FMT. In Section III, we focus on the power minimization problem for H-FMT. In Section IV, we present numerical results. Finally, in Section V, the conclusions follow.

II. System Model

We consider a general multicarrier (MC) system with M sub-channels whose low rate data signals $a^{(k)}(Nn)$, with $k = 0, \ldots, M-1$, are upsampled by a factor $N = M + \beta$, where β denotes the overhead (OH) factor, and are filtered by the modulated pulses $g^{(k)}(n) = g(n)W_M^{-kn}$, with $g(n)$ being the prototype filter of the synthesis bank and $W_M^{kn} = e^{-j\frac{2\pi}{M}kn}$. Then, the sub-channel signals are summed and the transmitted

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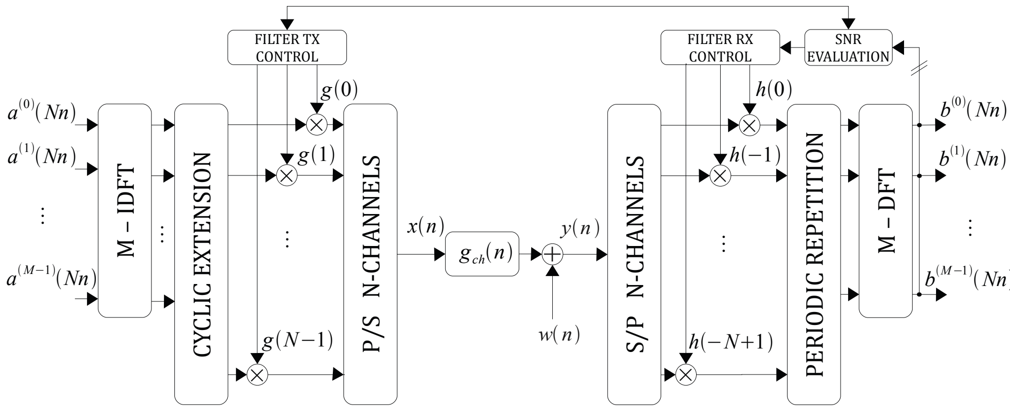

Fig. 1. Hybrid FMT scheme.

signal reads

$$x(n) = \sum_{k=1}^{M-1} \sum_{\ell \in \mathbb{Z}} a^{(k)}(N\ell) g^{(k)}(n - N\ell).$$

After channel propagation, the signal $y(n)$ is processed by the sub-channel modulated pulses $h^{(k)}(n) = h(n)W_M^{-kn}$, where $h(n)$ is the prototype pulse of the analysis bank. Therefore, the signal received in the k-th sub-channel, with sampling with period N, can be written as

$$b^{(k)}(Nn) = a^{(k)}(Nn)H^{(k)}(\beta) + I^{(k)}(Nn, \beta) + \eta^{(k)}(Nn, \beta), \quad (1)$$

where $H^{(k)}(\beta) = g^{(k)} * g_{ch} * h^{(k)}(0)$ and $g_{ch}(n)$ is the channel impulse response. With $I^{(k)}(Nn, \beta)$ and $\eta^{(k)}(Nn, \beta)$ we respectively denote the interference (ISI plus ICI) and the noise term in sub-channel k.

We use the IEEE 802.11 TGn [7] channel model. This model generates channels belonging to five classes labeled with B,C,D,E,F. Each class is representative of a certain environment, e.g., small office, large open space/office with line of sight (LOS) and non LOS (NLOS) propagation, and so on. Both small scale multipath fading and large scale path loss fading as a function of distance are taken into account. For a detailed description of the model, see [7].

In order to evaluate the system performances, we assume parallel Gaussian channels and independent and Gaussian distributed input signals, which render ISI and ICI also Gaussian (cf. e.g. [5]). Furthermore, assuming single tap zero forcing equalization, the achievable rate in bit/s for a given channel realization, can be computed as

$$C(\beta) = \frac{1}{(M+\beta)T} \sum_{k=0}^{M-1} \log_2 \left(1 + SINR^{(k)}(\beta)\right), \quad (2)$$

where T is the sampling period, and $SINR^{(k)} = (1/SNR^{(k)} + 1/SIR^{(k)})^{-1}$ denotes the signal over interference plus noise ratio experienced in sub-channel k. We denote with $SNR^{(k)}(\beta) = P_{\mathrm{U}}^{(k)}(\beta)/P_{\eta}^{(k)}$ and $SIR^{(k)}(\beta) = P_{\mathrm{U}}^{(k)}(\beta)/P_{\mathrm{I}}^{(k)}(\beta)$ the signal to noise and the signal to interference ratios. $P_{\mathrm{U}}^{(k)}(\beta)$, $P_{\mathrm{I}}^{(k)}(\beta)$ and $P_{\eta}^{(k)}$ respectively are the useful, the interference, and the noise power terms on sub-channel k. Details on their computation can be found in [8].

For the following treatment, it is convenient to express the SINR as $SINR^{(k)}(\beta) = P_a^{(k)}\gamma^{(k)}(\beta)$, where $P_a^{(k)}$ is the transmitted signal power in the k-th sub-channel, and $\gamma^{(k)}(\beta) = |H^{(k)}(\beta)|^2 / \left(P_{\mathrm{I}}^{(k)}(\beta) + P_{\eta}^{(k)}\right)$.

In the next sub-sections we first briefly recall A-OFDM and SO-FMT, and then we present H-FMT.

A. Adaptive OFDM (A-OFDM)

A-OFDM is an OFDM system which adapts the cyclic prefix duration to the channel condition. The OFDM scheme can be obtained setting the coefficients $g(n)$ and $h(n)$ respectively equal to

$$g(n) = \frac{1}{\sqrt{N}}\, rect\left(\frac{n}{N}\right), \quad h(n) = \frac{1}{\sqrt{M}}\, rect\left(-\frac{n}{M}\right), \quad (3)$$

where $rect\,(n/A) = 1$ for $n = \{0, 1, \ldots, A-1\}$ and zero otherwise. Note that $N = M + \beta$, and β is the duration of the CP in samples. It is well known that when the CP lasts more than the channel duration, the received signal is neither affected by ISI nor by ICI [5]. Nevertheless, in our previous work [6], we have shown that the CP has not to be necessarily as long as the channel duration to maximize the achievable rate. The rationale behind is that the level of interference can be raised in noise-limited systems without loosing in achievable rate (see (2)).

From the statistics of the optimal CP duration, in [6], we have found a limited set of CP values over which to adapt the OFDM system to the channel condition. We have denoted this set as $\mathbb{P}_{OFDM} = \left\{\beta_B^{(99\%)}, \beta_C^{(99\%)}, \beta_D^{(99\%)}, \beta_E^{(99\%)}, \beta_F^{(99\%)}\right\}$. That is, depending on the experienced channel class, the A-OFDM picks the corresponding CP value from the set $\mathbb{P}_{OFDM}$.

B. Short Orthogonal FMT (SO-FMT)

Differently from OFDM that privileges time confinement using rectangular pulses, in general, FMT schemes privilege the frequency confinement. The SO-FMT scheme can be obtained from the general system model simply substituting the prototype pulse $g(n) = h^*(-n)$ with an FMT orthogonal pulse having minimal length. These pulses satisfy the orthogonality conditions given by the following system of equations

$$\left[g^{(k)} * h^{(i)}\right](Nn) = \delta_n \, \delta_{i-k}, \qquad (4)$$
$$\forall (k,i) \in \{0, \cdots, M-1\}, \; \forall n \in \mathbb{Z},$$

where we denote with δ_n the Kronecker delta. The solution of this system is not unique, and thus we parameterize the filter coefficients with a minimal set of parameters θ. In order to have maximally frequency confined pulses, we choose those parameters θ that satisfy the minimum square error from a target pulse $H(f)$.

$$\arg\min_{\theta} \int_{-.5}^{.5} |G(f, \theta) - H(f)|^2 df, \qquad (5)$$

where $G(f, \theta)$ is the frequency response of the prototype pulse $g(n)$ as function of the parameters θ. It is worth noting that the choice of the filter determines the achievable rate (2) inasmuch it is proportional to the OH, therefore β can be adapted to the channel condition to maximize the achievable rate. Nevertheless, in order to keep low the system implementation complexity, the family of filters adopted in this paper has minimal length equal to one symbol duration, i.e., $L_g = N$. Furthermore, numerical results have shown that for SO-FMT the adaptation of β to the channel conditions does not significantly improve the achievable rate, and in general the best performer results to be the filter whose $N = (9/8)M$, i.e., a filter with $\beta = 8$ samples. Therefore, when showing numerical results for SO-FMT, we set the OH β equal to 8 samples. More details regarding the filter design of the SO-FMT scheme are reported in [4].

C. Hybrid FMT

We have numerically found that depending on the channel condition it is convenient to use A-OFDM or SO-FMT to maximize the achievable rate (2) (see numerical results section).

This can be simply justified observing the following. Firstly, the interference experienced by A-OFDM is always lower than that of SO-FMT if a further equalization stage is not deployed. Secondly, OFDM suffers of a loss of SNR of a factor equal to M/N because it does not use a matched filter. Hence, we can understand that, when looking at the SINR of SO-FMT, if it results noise limited, namely if the SNR is lower than the SIR, it is convenient to use SO-FMT. Vice versa, it is convenient to use A-OFDM.

Clearly, the optimal criterion for switching between A-OFDM and SO-FMT should target the maximization of the achievable rate for each channel realization. Nevertheless, this criterion is not desirable for practical systems where the channel conditions can quickly change. Therefore, in our previous work [1] we have proposed a metric that, based on the estimation of the mean SNR, chooses the scheme to be used. More precisely, numerical results have shown that the average level of the SIR experienced by SO-FMT over WLAN channels is equal to 38 dB (this value is obtained averaging across sub-channels and channel realizations). Thus, if the mean SNR (SNR averaged across sub-channels) is below this value the system uses SO-FMT, otherwise it switches to A-OFDM.

As it will be shown in the numerical results section, the criterion based on the estimation of the mean SNR gives results that are close to that given by the optimal one.

To implement H-FMT we make use of the efficient implementation of a DFT modulated filter bank presented in [4]. Fig. 1 shows the H-FMT architecture. As we can see, at the transmitter the M data signals $a^{(k)}(Nn)$, with $k = 0, \ldots, M-1$, are processed by an M points IDFT. The N output samples from the cyclically extension, are multiplied by the coefficients $g(n)$, with $n = 0, \ldots, N-1$ which correspond to the prototype synthesis pulse coefficients of SO-FMT or A-OFDM. Finally, the outputs are parallel-to-serial converted and transmitted over the channel.

The receiver consists in a serial-to-parallel conversion with a converter of size N. Depending on the modulation scheme used at the transmitter, namely SO-FMT or A-OFDM, the output signals are multiplied by the corresponding prototype analysis pulse coefficients $h(-n)$ with $n = 0, \ldots, N-1$. Then, the periodic repetition with period M of the block of coefficients of size N is applied. Finally, the M-point DFT is performed.

The block SNR Evaluation estimates the mean SNR and feedbacks it to the transmitter. Note that the SNR evaluation can be done sending a known robust training sequence of symbols using OFDM with a conservative long CP. In this case the received signal results free of interference and the SINR corresponds to the SNR. Furthermore, to adapt the system to the channel condition, the SNR evaluation is done periodically.

Since both SO-FMT and OFDM use filter with almost the same length, the implementation complexity of H-FMT is basically the same of OFDM (see [4] for more details).

III. Power Allocation

In this section we propose a practical algorithm to solve the *power minimization* (PM) problem using H-FMT when a constraint on the power spectral density (PSD) mask is imposed.

The PM problem can be formulated as

$$\begin{aligned} \arg\min_{\mathbf{P}_a} \quad & \sum_{k=0}^{M-1} P_a^{(k)}, \qquad (6) \\ \text{s.t.} \quad & 0 \leq P_a^{(k)} \leq \overline{P}, \quad k = 0, \ldots, M-1. \\ & \frac{1}{NT} \sum_{k=0}^{M-1} \log_2\left(1 + P_a^{(k)} \gamma^{(k)}(\beta)\right) = R, \end{aligned}$$

where R denotes the target achievable rate in [bit/s], $\mathbf{P}_a = \{P_a^{(k)}$, with $k = 0, \ldots, M-1\}$ is the vector of the transmitted powers, and $\overline{P}$ is the power constraint given by the PSD limit level.

In general, problem (6) is not convex because the interference terms that appear in $\gamma^{(k)}(\beta)$ are dependent on the transmitted power distribution. A solution to (6) could be found doing and exhaustive search on $\mathbf{P}_a$. However, this procedure is practically unfeasible. This is true even supposing to quantize the sub-channel power in a finite number L of levels. In fact, the exhaustive search would have a complexity $O(L^M)$. Nevertheless, if we consider the interference term independent from the distribution of the transmitted power, the problem becomes convex with constraints satisfying the Slater's condition [9]. Therefore, if exists, the optimal solution satisfies the Karush Kuhn Tucker (KKT) conditions, and it is given by [10]

$$P_a^{(k)} = P_a^{(k)}(\nu) = \left[\nu - 1/\gamma^{(k)}(\beta)\right]_0^{\overline{P}}, \tag{7}$$

where

$$[x]_a^b = \begin{cases} b, & x \geq b \\ x, & a < x < b \\ a, & x \leq a, \end{cases} \tag{8}$$

and ν is given by the solution of the equation

$$\frac{1}{NT}\sum_{k=0}^{M-1}\left(1 + P_a^{(k)}(\nu)\gamma^{(k)}(\beta)\right) = R. \tag{9}$$

To solve (7),(9) we make use of the iterative algorithm presented in [10].

Now the problem is: how does the H-FMT select the modulation to be used? For each channel realization, the optimal approach should target the PM objective. We can have two cases. The first is when the solution to (6) exists for both SO-FMT and A-OFDM. In such a case, we should choose the system that transmits less power. The second is when only one system admits a valid solution to (6). In this case the choice is trivial. In both cases, the systems in general can be interference limited (i.e., the interference level is higher than that of noise). Thus, (6) has to be solved exhaustively and this renders this approach unfeasible.

To diminish the complexity we use the two switching criteria that have been proposed in [1] and summarized in Section II-C. They have been derived considering the dual problem of (6), i.e., the *achievable rate maximization* (RM) under a PSD mask constraint. The first one is optimal for the RM problem in the sense that for each channel realization, it first computes the achievable rate of both SO-FMT and A-OFDM, and consequently chooses the system that maximizes it. The second is sub-optimal, indeed it selects the modulation to be used depending on the experienced mean SNR.

In this way the H-FMT works in a condition where it is always noise limited (i.e., it uses SO-FMT if its interference is lower than the noise, otherwise uses A-OFDM that, thanks to the CP, renders the interference negligible).

It is worth noting that, supposing the interference term independent from the transmitted power distribution, the optimal power allocation that is the solution of the RM problem is a uniform distribution at the PSD limit level [10]. Therefore, when showing numerical results for the achievable rate obtained using RM power allocation we assume the previous hypothesis verified.

The practical algorithm for the PM of H-FMT can be summarized as follows.

1) *Select the system to be used (SO-FMT or A-OFDM) with one of the criteria proposed for the RM problem.* Namely, the optimal or the mean SNR based criterion.
2) *Compute the coefficients $\gamma^{(k)}(\beta)$ for $k = 0, \ldots, M-1$, fixing the transmitted power at the PSD limit level.* Note that in such case the interference is maximal and the corresponding $\gamma^{(k)}(\beta)$ are the smallest. Also compute the rate achievable with this power distribution. It is worth noting that it also equals the maximum achievable rate when the interference is negligible.
3) *Set the value for the target rate R.*
4) *Compute the optimal power allocation (7).*

IV. Numerical Results

To obtain numerical results, we have chosen the following system parameters that essentially are those of the IEEE 802.11 standard [11]. The MC system uses $M = 64$ sub-channels with a transmission bandwidth of 20 MHz. The signal is transmitted with a constraint on the PSD limit level of -53 dBm/Hz. At the receiver side, we add white Gaussian noise with PSD equal to -168 dBm/Hz. To show the performance of H-FMT, we use an OFDM baseline system which deploys a fixed CP of 0.8 μs ($\beta = 16$ samples), that is the value of CP employed by the IEEE 802.11 standard [11].

The H-FMT deploys SO-FMT with fixed OH of 0.4 μs ($\beta = 8$ samples, $N = (9/8)M$), and A-OFDM with values of CP equal to $\mathbb{P}_{OFDM} = \{\beta_B^{(99\%)} = 0.4\ \mu s,\ \beta_C^{(99\%)} = 0.5\ \mu s,\ \beta_D^{(99\%)} = 0.6\ \mu s,\ \beta_E^{(99\%)} = 0.9\ \mu s,\ \beta_F^{(99\%)} = 1.1\ \mu s\}$ [6]. Both schemes use single tap zero forcing sub-channel equalization.

Fig. 2 shows the complementary cumulative distribution function (CCDF) of the achievable rate obtained using the RM power allocation for SO-FMT, A-OFDM, and the baseline system. For the sake of readability, we only show results for channel classes B, C, and D and for distances between transmitter and receiver equal to 10 m, 30 m, and 60 m. More results for the baseline OFDM and for A-OFDM can be found in [6].

From Fig. 2 we can observe how for high values of distance $d = \{30, 60\}$ m, and thus in the low SNR region, the SO-FMT shows better performance than both A-OFDM and baseline OFDM. On the contrary, for a short distance of 10 m, A-OFDM is able to achieve higher rate than SO-FMT. Although not shown, the results obtained with a distance of 3 m behave as the ones obtained with a distance of 10 m.

It is worth noting that also if not highlighted, in Fig. 2 the CCDFs of the achievable rate of H-FMT would correspond to the most right curves.

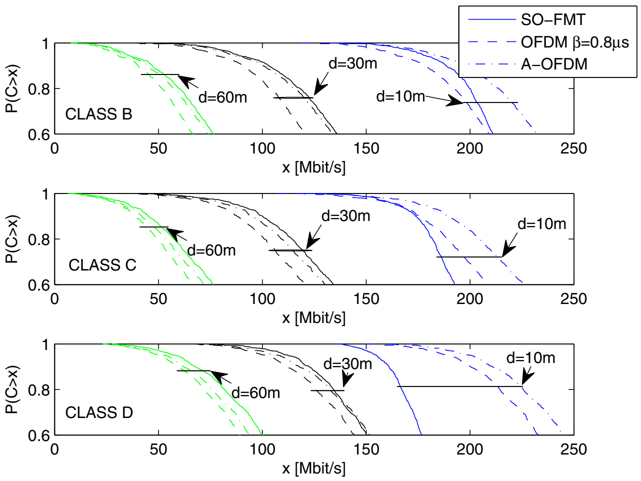

Fig. 2. Achievable rate CCDF obtained using SO-FMT, A-OFDM, and OFDM with a fixed CP of 0.8 μs. The employed channel classes are the B,C, and D. The distance between transmitter and receiver is set to 10 m, 30 m, and 60 m.

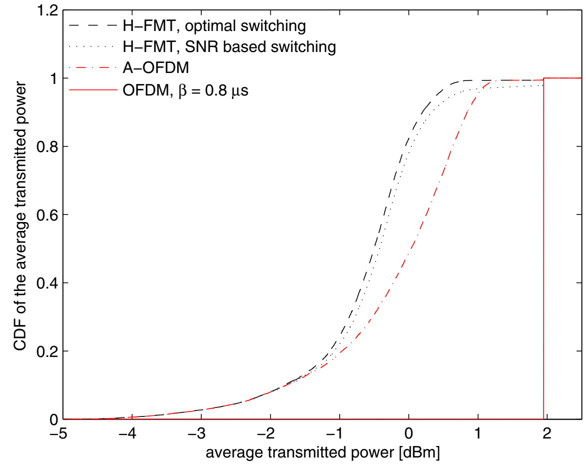

Fig. 3. CDF of the average transmitted power obtained using PM algorithm. The comparison is among H-FMT that adopts the optimal and the sub-optimal switching criterion, OFDM with CP equal to 0.8μs, and A-OFDM.

We now turn our attention to the PM problem (6). Fig. 3 shows the cumulative distribution functions (CDFs) of the average transmitted power. They have been obtained solving the PM problem (6) for H-FMT with the proposed PM algorithm using both the optimal and the mean SNR based switching criteria, and for the baseline system. Also the results obtained with A-OFDM are shown. These are meant to show the best performance obtainable using OFDM. The results are obtained drawing the channels randomly among the classes $\{B, C, D\}$, and for distances in $[1, 60]$ m. Furthermore, for each channel realization we set the target rate R equal to the achievable rate of the baseline system (OFDM with fixed CP of 0.8 μs) whose transmitted powers are given by the PSD limit level. This choice is justified observing that the baseline system shows worse performances than both H-FMT and A-OFDM fixing the transmitted power at the PSD level (see Fig. 2). Therefore, assuming this value as the target rate, we can first compare the proposed system with the baseline, and further we ensure that a solution to the PM problem exists for both H-FMT and A-OFDM.

From Fig. 3 we observe the followings.

- The switching criterion based on the mean SNR estimation works close to the optimal one.
- With probability equal to 0.9, H-FMT and A-OFDM respectively transmit with power of about 1.9 and 1.4 dB less than the baseline system, yet achieving the same rate. This translate in a power saving of about 36% and 30% w.r.t. the baseline system.
- There are a 20% of channels where the power saving given by H-FMT equals 3 dB.

V. Conclusions

We have investigated the use of a hybrid architecture that wisely exploits the peculiarities of adaptive OFDM and FMT with minimal length pulses to reduce the power transmitted by the WLAN devices. Numerical results have shown that hybrid FMT provides significant power saving gains compared to an OFDM scheme which deploys a fixed CP, while assuring the same achievable rate. It also achieves a marginal increment of computational complexity.

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