OVERLAY SPECTRUM REUSE IN A MULTICARRIER BROADCAST NETWORK: SINGLE RECEIVER ANALYSIS

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ABSTRACT

The overlay cognitive radio paradigm presents a framework where a secondary user exploits the knowledge of the primary user's message to improve spectrum utilization. A multicarrier broadcast network is one of the scenarios where this knowledge is possible: the secondary user could join a single frequency network and, therefore, gain access to the primary message. However, if the primary signal is received with a strong line of sight component, its relaying from the secondary transmitter does not suffice to ensure the primary user quality of service. In this paper we study the scenario where a secondary transmitter maximizes its own transmission rate, keeping the quality of a primary receiver over a given threshold. The analytical results, based on bit error rate bounds, are verified by means of software simulations and hardware tests.

1. INTRODUCTION

During the past few years, the *overlay* approach [1] to the problem of a secondary user transmitting over a licensed band has been found of special interest: the knowledge of the primary message might potentially allow the secondary transmitter to use side information at the transmitter techniques such as *Dirty Paper Coding* (DPC) [2], and even to transmit the primary signal to reinforce the primary Quality of Service (QoS).

One scenario where it is possible to access the primary message is depicted in [3]: if the primary user is integrated in a broadcast Orthogonal Frequency Division Multiplexing (OFDM) Single Frequency Network (SFN), the signal is somehow delivered to the primary base stations (usually via satellite), so the secondary user could be able to take part of that SFN in order to gain access to the primary signal. Moreover, as the SFN operation requires synchronization between the primary transmitters, the secondary transmitter could also get synchronized by gaining access to the same synchronization source, usually the GPS signal.

However, the transmission of the primary message from the secondary transmitter does not immediately imply a better system performance, as the channel degradation due to the insertion of the secondary replica can be higher than the provided power gain, specially in those systems with a strong Line Of Sight (LOS) reception [4], with the Additive White Gaussian Noise (AWGN) channel as the worst case scenario. This problem is expected to appear in the SFN broadcast case, specially at those cases where rooftop antennas are used for fixed reception.

In [5] it was shown that the pre-filtering of the primary message at the secondary transmitter substantially improves the primary system performance, even for pure LOS channels. In that work, optimum transmission strategies were derived for a pure cooperative secondary user, i.e., a secondary user that tries to maximize the quality of the primary service. In this paper, we present a scenario where the secondary user transmits its own message overlaid on the primary one, while allocating a fraction of its available power to reinforce the primary signal in order to meet a given interference constraint. The study is extended to several users in the primary transmitter coverage area in the companion paper [6].

The structure of the paper is as follows: in Section 2 the system model and proposed metrics are introduced; in Section 3 a scenario where a secondary transmitter maximizes its own transmission rate subject to a QoS constraint on a single primary receiver is studied; in Section 4 the results are presented, and the proposed metrics are verified via software simulations and hardware measurements. Section 5 concludes the paper.

2. PROBLEM STATEMENT

Throughout the paper, we will study a multicarrier primary system, and assume that the links from both primary and secondary transmitters to a given primary receiver can be modeled as AWGN channels, so the equivalent baseband received signal after the Cyclic Prefix (CP) removal can be written as

$$y_n = \left(\delta_n + \gamma e^{-j\theta} f_{n-n_0}\right) \circledast x_n + \rho s_n + w_n \tag{1}$$

where the equivalent channel was normalized to set the channel from the primary transmitter to δ_n, while γ, θ and n_0 are the relative amplitude, phase and delay of the primary signal contribution sent from the secondary transmitter, $\circledast$ denotes the circular convolution operator, x_n denotes the n-th sample of the primary signal (normalized to have unit power), ρ denotes the relative amplitude of the secondary signal[1] $s_n \sim \mathcal{CN}(0, 1)$, assumed to be Gaussian, sent from the secondary transmitter, and $w_n \sim \mathcal{CN}\left(0, \sigma^2\right)$ is a sample of white Gaussian noise. As an additional degree of freedom, the secondary transmitter is allowed to (circularly) filter the primary signal with a transmit filter f_n. The convenience of this filtering was shown in [5]. This scenario is depicted in Figure 1.

[1] The phase and delay of the secondary signal have been omitted in (1), without loss of generality.

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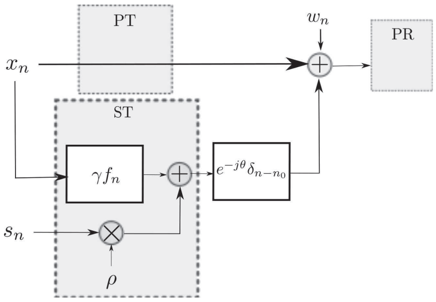

Fig. 1. System model: the Secondary Transmitter (ST) knows the message x_n of the Primary Transmitter (PT). The ST filters the primary signal with the filter γf_n, and scales the secondary message s_n with ρ. The signal received by the Primary Receiver (PR) can be seen to follow equation (1).

In the Discrete Fourier Transform (DFT) domain, the previous relation (1) reads for a given carrier k as

$$Y_k = \left(1 + \gamma e^{-j(2\pi k n_0/N+\theta)} F_k\right) X_k + \rho S_k + W_k,\ k = 1, ..., N \tag{2}$$

where X_k, S_k, F_k and W_k denote the N-DFT of x_n, s_n, f_n and w_n, respectively, and N is the number of carriers. Without loss of generality, we will restrict our filter F_k to be real in the DFT domain[2].

For the sake of simplicity, we will consider a Quadrature Phase Shift Keying (QPSK) constellation in the primary system, as the derived analytical bounds are easier to deal with, and we will assume perfect channel estimation and frequency synchronization in the analytical derivations, and an overall channel length shorter than the CP. However, these results will be extended to practical synchronization schemes by means of hardware measurements.

We propose to analyze the performance of the primary system by means of the Chernoff Bound (CB) for the uncoded Bit Error Rate (BER) or, equivalently, by the Exponential Effective SNR Metric (EESM), one of the Physical Layer abstraction metrics for multicarrier systems proposed in IEEE 802.16 [7]. The expression for the effective Signal to Noise Ratio (SNR) using the EESM metric is[3] $\Upsilon_{eff} = -2\log(\eta)$, where

$$\eta = \frac{1}{N}\sum_{k=1}^{N} e^{-\Upsilon |H_k|^2/2} \tag{3}$$

is the expression for the CB. In the previous expression $H_k = 1 + \gamma e^{-j(2\pi k n_0/N+\theta)} F_k$ denotes the equivalent channel seen by the k-th carrier at a given receiver, so

$$|H_k|^2 = 1 + \gamma_k^2 + 2\gamma_k \cos(\theta + 2\pi k n_0/N) \tag{4}$$

where $\gamma_k \doteq \gamma F_k$, and $\Upsilon \doteq \frac{1}{\sigma^2+\rho^2}$ denotes the SNR of the system in absence of the secondary transmitter conveying the primary message, which is constant along all the carriers due to the AWGN assumption.

[2] As θ is unknown and will be modeled as an uniform random variable, the insertion of a phase term in F_k does not affect the final result, thus allowing us to assume a real filter.

[3] The general expression for the EESM is $\Upsilon_{eff} = -\lambda \log\left(\frac{1}{N}\sum_{k=1}^{N} e^{-\Upsilon |H_k|^2/\lambda}\right)$, λ being a degree of freedom that depends on the particular modulation and coding scheme [8]. In this paper we will set $\lambda = 2$, as it is the value for the CB of the BER of a QPSK, although results can be easily extended to other values of λ.

In the following, we will assume that the value of γ and ρ are deterministic, as they can be obtained by means of a propagation model or by measurements, and model θ as a uniform Random Variable (RV) $\theta \sim U(0, 2\pi]$ as it is impossible to determine the exact phase difference between echoes θ. Therefore, the final expression for the CB is obtained by substituting (4) in (3), and averaging over θ,

$$\begin{aligned} \eta(\boldsymbol{\gamma}, \rho) &= \frac{1}{N}\sum_{k=1}^{N} E_\theta\left\{e^{-\frac{\Upsilon}{2}\left(1+\gamma_k^2+2\gamma_k\cos(\theta+2\pi k n_0/N)\right)}\right\} \\ &= \frac{e^{-\frac{\Upsilon}{2}}}{N}\sum_{k=1}^{N} e^{-\frac{\Upsilon}{2}\gamma_k^2}\frac{1}{2\pi}\int_0^{2\pi} e^{-\Upsilon\gamma_k cos(\theta)} \mathrm{d}\theta \\ &= \frac{e^{-\frac{\Upsilon}{2}}}{N}\sum_{k=1}^{N} e^{-\frac{\Upsilon}{2}\gamma_k^2} I_0(\Upsilon\gamma_k) \end{aligned} \tag{5}$$

where $I_0(\cdot)$ is the zero-th order modified Bessel function of the first kind, $E_X\{\cdot\}$ denotes the expectation operator over the RV X, and $\boldsymbol{\gamma} = [\gamma_1, ..., \gamma_N]^T$. Note that the obtained expression does not depend on the time difference n_0, so there is no need to make any assumption about this value.

In order to obtain a relationship between the CB and the definition of the coverage zone, which is given by the coded BER, the performance of the primary system will be evaluated by using the following analytical bound for the BER after Viterbi decoding for DVB-T, taken from [9]:

$$BER \leq \frac{1}{4}\sum_{d=d_{min}}^{\infty} c_d \eta(\boldsymbol{\gamma}, \rho)^d \tag{6}$$

with d_{min} the minimum Hamming distance of the convolutional code, and c_d the total input weight due to an error event at distance d from the all-zero path.

3. SINGLE PRIMARY RECEIVER ANALYSIS

In this section, we will focus on the strategy the secondary user must use in order to maximize its own capacity subject to a controlled degradation of the primary service at a given receiver. We will assume that the secondary users are able to use some kind of interference mitigation techniques (DPC at the transmitter or Successive Interference Cancellation (SIC) at the receivers) so the capacity of the secondary link is equivalent to that in absence of the primary transmitter.

Thus, the channel model will be an interference Z channel [1], where the secondary message is treated as noise by the primary receivers. As the objective is to maximize the secondary user achievable rate, and the power distribution for the secondary message is uniform, this is equivalent to maximizing the power allocated to the secondary message ρ^2. Therefore, introducing constraints on the transmit power and the primary receiver CB, we can formulate the optimization problem as

$$\begin{aligned} &\text{minimize} && -\rho \\ &\text{subject to} && \eta(\boldsymbol{\gamma}, \rho) = \tfrac{1}{N}\sum_{k=1}^{N} e^{-\frac{1+\gamma_k^2}{\psi+2\rho^2}} I_0\left(\tfrac{2\gamma_k}{\psi+2\rho^2}\right) \leq \eta_0 \\ &&& \rho^2 + \tfrac{1}{N}\sum_{k=1}^{N}\gamma_k^2 \leq P \end{aligned} \tag{7}$$

where η_0 is the constraint on the CB, and $\psi \doteq 2\sigma^2$, leading to $\Upsilon = \frac{1}{\sigma^2+\rho^2} = \frac{2}{\psi+2\rho^2}$. We denote by P the secondary received power (normalized by the primary one) at a the primary receiver, that has to be shared between the primary prefiltered replica and the secondary signal, with respectively allocated powers $\frac{1}{N}\sum_{k=1}^{N}\gamma_k^2 = \gamma^2$ and ρ^2.

In [5] it was shown that for a pure cooperative secondary user, i.e., $P = \gamma^2$, the optimum power distribution concentrates the power in a fraction ϕ of carriers, leaving the remaining fraction $1-\phi$ set to zero. For a sufficiently large number of carriers, we can approximate the fraction ϕ by a real number in the interval $[0, 1]$, so problem (7) can be rephrased as

$$\begin{array}{ll} \text{min.} & -\rho \\ \text{s.t.} & e^{-\frac{1}{\psi+2\rho^2}}\left((1-\phi)+\phi e^{-\frac{\gamma^2/\phi}{\psi+2\rho^2}} I_0\left(\frac{2\gamma}{\sqrt{\phi}(\psi+2\rho^2)}\right)\right) \leq \eta_0 \\ & \rho^2+\gamma^2 \leq P \\ & 0 \leq \phi \leq 1. \end{array} \tag{8}$$

With this simplification we have reduced the number of variables from $N+1$ (ρ and the N variables γ_k to perform the primary signal power weighting) to 3. Furthermore, we can reduce the number of variables to 2 by approximating ϕ by its asymptotic optimum (and heuristic) value, $\phi = \frac{\gamma^2}{4}$, as it was shown in [5] that the use of this value instead of the optimum one suffers from small losses for a large range of SNR values. Note that this solution is only valid for $0 \leq \gamma^2 \leq 4$.

With this simplification, the CB constraint in (8) can be rewritten as $f(\gamma, \rho) \leq 0$, with

$$\begin{aligned} & f(\gamma,\rho) \doteq \\ & e^{-\frac{1}{\psi+2\rho^2}}\left(\left(1-\frac{\gamma^2}{4}\right)+\frac{\gamma^2}{4}e^{-\frac{4}{\psi+2\rho^2}} I_0\left(\frac{4}{(\psi+2\rho^2)}\right)\right)-\eta_0, \end{aligned} \tag{9}$$

that can be seen to be equivalent to

$$\gamma^2 \geq 4\frac{1-\eta_0 e^{\frac{1}{\psi+2\rho^2}}}{1-e^{-\frac{4}{\psi+2\rho^2}} I_0\left(\frac{4}{\psi+2\rho^2}\right)}. \tag{10}$$

The solution to this problem has a different behavior depending on the values of the SNR in absence of the secondary transmitter $\Upsilon_{NS} \doteq 2/\psi$ and the received power from the secondary transmitter P. This behavior is illustrated for different cases next.

3.1. Moderate values of P

For non-extreme values of P, if $\gamma^2 \leq 4$, the approximate optimum value of γ is the one that maximizes ρ while meeting constraint (10) and, therefore, is the value obtained from (10) with equality, so the BER restriction is active, and the remaining power is used to transmit the secondary information. From (10), it follows that the optimum value for the secondary signal received power ρ^2 can be obtained from the following implicit function:

$$\rho^2 = P - 4\frac{1-\eta_0 e^{\frac{1}{\psi+2\rho^2}}}{1-e^{-\frac{4}{\psi+2\rho^2}} I_0\left(\frac{4}{\psi+2\rho^2}\right)}. \tag{11}$$

If $\gamma^2 = P - \rho^2 > 4$, the optimum value of ϕ will be 1. Therefore, the solution can be obtained by forcing $\phi = 1$ and $\rho^2 + \gamma^2 = P$ in problem (8), so the desired value of ρ^2 can be obtained as the root of

$$e^{-\frac{1+P-\rho^2}{\psi+2\rho^2}} I_0\left(\frac{2\sqrt{P-\rho^2}}{\psi+2\rho^2}\right) - \eta_0 = 0. \tag{12}$$

3.2. $P \to 0$

For small values of P the solution will be strongly dependent on the value of Υ_{NS}. Let us define Υ_0 as the value of SNR such that the CB constraint is met with equality in absence of the secondary transmitter, i.e., $e^{-\frac{\Upsilon_0}{2}} = \eta_0$. Equivalently, we define $\psi_0 \doteq \frac{2}{\Upsilon_0} = \frac{-1}{\log(\eta_0)}$. We will restrict our analysis to those receivers in the original coverage region, i.e., $\Upsilon_{NS} \geq \Upsilon_0$.

3.2.1. $\Upsilon_{NS} > \Upsilon_0$

In this case, if the value of P is small enough, we have that $\eta(\mathbf{0}, \sqrt{P}) = e^{-\frac{1}{\psi+2P}} \leq \eta_0$, so the secondary transmitter can allocate all its available power to the secondary message without violating the CB constraint, i.e, its optimum allocated power is $\rho^2 = P$. This could be the case of a primary receiver operating at a very high SNR, or a low-power secondary user.

3.2.2. $\Upsilon_{NS} = \Upsilon_0$

In this case, we have that $\eta(\mathbf{0}, \sqrt{P}) = e^{-\frac{1}{\psi_0+2P}} > \eta_0 \,\forall\, P > 0$, so the CB constraint is not fulfilled if all the power P is allocated to the secondary message. Following (9) and from the definition of ψ_0, we have $f(0,0) = 0$. For small values of P, we have $\rho \to 0$, $\gamma \to 0$. Therefore, and as $\nabla_{\gamma,\rho} f(0,0) = \mathbf{0}$, we can approximate $f(\gamma,\rho)$ by its second order Taylor polynomial around the point $(0,0)$ $f(\gamma,\rho) \approx \frac{1}{2}[\gamma\ \rho]\nabla^2_{\gamma,\rho} f(0,0)\,[\gamma\ \rho]^T$, where $\nabla_{\gamma,\rho} f(\gamma_0,\rho_0)$ denotes the gradient of the function f evaluated in (γ_0,ρ_0), and $\nabla^2_{\gamma,\rho} f(\gamma_0,\rho_0)$ denotes the Hessian matrix evaluated in the same point. In this case, the Hessian evaluated in $(0,0)$ is a diagonal matrix with entries $\frac{\partial^2 f}{\partial\gamma^2}(0,0) = \frac{1}{2}e^{-5/\psi_0}\left(I_0\left(\frac{4}{\psi_0}\right) - e^{4/\psi_0}\right)$, and $\frac{\partial^2 f}{\partial\rho^2}(0,0) = \frac{4e^{-1/\psi_0}}{\psi_0^2}$.

The maximum value of ρ will be obtained when both the CB constraint and the power constraint are met with equality, i.e, $f(\gamma,\rho) = 0$ and $P = \gamma^2 + \rho^2$, so

$$\frac{\rho^2}{P} = \frac{\frac{\partial^2 f}{\partial\gamma^2}(0,0)}{\frac{\partial^2 f}{\partial\gamma^2}(0,0) - \frac{\partial^2 f}{\partial\rho^2}(0,0)} = \frac{\psi_0^2\left(e^{4/\psi_0} - I_0\left(\frac{4}{\psi_0}\right)\right)}{e^{4/\psi_0}\left(\psi_0^2+8\right) - \psi_0^2 I_0\left(\frac{4}{\psi_0}\right)}. \tag{13}$$

3.3. $P \to \infty$

For high values of P, the high power coming from the secondary transmitter makes the primary contribution negligible. If this is the case, we can write the CB constraint as

$$\eta(\gamma,\rho) = e^{-\frac{\gamma^2}{2\rho^2}} \leq \eta_0, \tag{14}$$

so the optimum value of ρ^2 will be obtained when (14) and the power constraint are met with equality, so we arrive to

$$\frac{\rho^2}{P} = \frac{1}{1-2\log(\eta_0)} = \frac{1}{1+\Upsilon_0}. \tag{15}$$

Note that this is the case when the primary user power is negligible, so the constraint for the secondary user is to keep the ratio between primary and secondary messages over the limit SNR value, $\frac{\gamma^2}{\rho^2} = \Upsilon_0$.

The analytical power allocation results are summarized in Table 1 for the different cases.

Table 1. Values for the design parameters ϕ, γ^2 and ρ^2 for the different cases under study.

Case	ϕ	γ^2	ρ^2
P Moderate, $\gamma^2 < 4$	$\gamma^2/4$	$P-\rho^2$	Root of (11)
P Moderate, $\gamma^2 \geq 4$	1	$P-\rho^2$	Root of (12)
$P \to 0, \Upsilon_{NS} > \Upsilon_0$	N/A	0	P
$P \to 0, \Upsilon_{NS} = \Upsilon_0$	$\gamma^2/4$	$P-\rho^2$	(13)
$P \to \infty$	1	$\frac{P\Upsilon_0}{1+\Upsilon_0}$	$\frac{P}{1+\Upsilon_0}$

4. RESULTS

We have obtained the power ρ^2 which can be reserved for the secondary signal under different reception conditions of the primary receiver, and for different fractions ϕ of reinforced primary carriers, namely the approximation $\phi = \gamma^2/4$, $\phi = 1$ and the optimum value of $\phi = \phi_{opt}$. These two latter solutions are obtained by MATLAB `fmincon` applied to the problem (8), whereas the value of ρ with the approximation $\phi = \gamma^2/4$ is obtained following the analytical derivations of the previous section: first, we calculate $\hat{\rho}$ as the root of equation (11), and then set $\rho = \min\left\{\sqrt{P}, \hat{\rho}\right\}$. This $\min\{\cdot\}$ operation is necessary in case all the secondary available power can be allocated to the secondary message, i.e., $P = \rho^2$, as previously discussed. If $P - \hat{\rho}^2 > 4$, ρ is obtained as the root of (12).

The numerical evaluations were performed for a DVB-T system with a 2/3 convolutional code rate, for which the bound (6) predicts a value of $\Upsilon_{0,dB} = 10\log_{10}(\Upsilon_0) \approx 5.6dB$ for a BER of $2 \cdot 10^{-4}$, with $\eta_0 \approx 0.16$.

In Figure 2 it is shown the evolution of ρ with the total available power for $0 \leq P \leq 4$ for three different values of Υ_{NS}. Obviously, as all the three cases have the same BER restriction, the one with the higher $\Upsilon_{NS,dB} = 10\log_{10}(\Upsilon_{NS})$ will require a lighter *support* from the secondary transmitter and, therefore, the available power for the secondary message ρ^2 will be higher. It is also noticeable that the evolution of ρ^2 (in both the $\phi = \gamma^2/4$ and ϕ_{opt} cases) has two differentiated regimes: the *low power* regime, where all the secondary power can be allocated to the secondary message without breaking the BER constraint and, therefore, in this region $\rho^2 = P$; and the *moderate power* regime. Note also that the case of $\Upsilon_{NS} = \Upsilon_0$ does only admit the *moderate power* regime, as the BER constraint is met with equality even in absence of the secondary transmitter. The solution for $\phi = 1$ has a slightly different behavior:

- For those cases for which all the power can be allocated to the secondary message without breaking the BER constraint, the solution is the same as in the other approximations. If this region does not exist (for $\Upsilon_{NS} = \Upsilon_0$) the optimum value of ρ is zero for a large range of values of P.
- For moderate values of P, an increment on the value of P is not reflected in the value of ρ, as allocating some power to the primary message would increase the BER bound.
- For high values of P, the value of ρ increases with P. In this region, the value of ρ is obtained as the root of (12), and approximates the optimum solution as P increases.

It is also noticeable that the solution with $\phi = \gamma^2/4$ offers very little degradation with respect to the optimum value of ϕ for small values of P, while the solution for $\phi = 1$ offers a good performance for larger values. Therefore, a near-optimum solution could be obtained just by solving the $\phi = 1$ and $\phi = \gamma^2/4$ problems, and choosing the one whose performance is better.

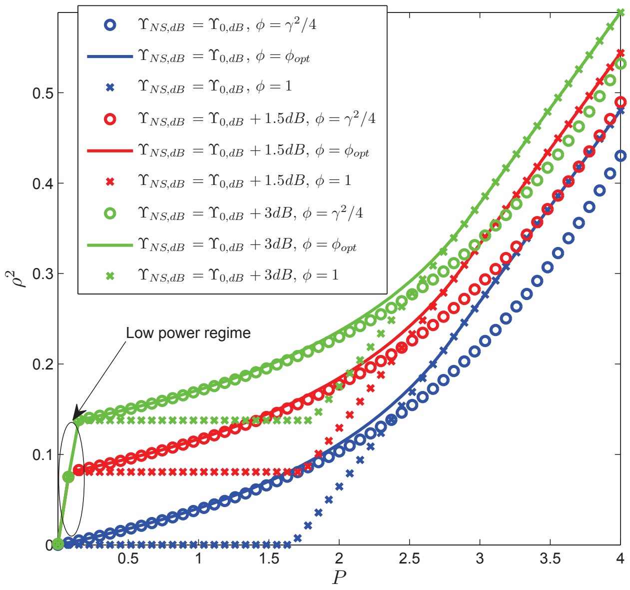

Fig. 2. Power (seen at reception) allocated at the secondary transmitter to the secondary signal as a function of the received secondary power.

In Figure 3 the accuracy of the $P \to \infty$ (14) and $P \to 0, \Upsilon_{NS} = \Upsilon_0$ (13) expressions for ρ^2/P is shown. For moderate values of P, it is also shown that if the target receiver has $\Upsilon_{NS} > \Upsilon_0$, then the fraction of available power used for the secondary transmission can be quite high for low values of P and then it has to decrease. In fact, in the low power regime, all the available power can be allocated to the secondary message without breaking the BER constraint, as previously stated. It can be also seen that the family of curves for $\Upsilon_{NS} > \Upsilon_0$ tend to approach the $\Upsilon_{NS} = \Upsilon_0$ curve as Υ_{NS} approaches Υ_0.

The proposed quality metric, based on the Chernoff bound of the BER, was verified by means of software simulations and hardware tests for a DVB-T system. In both cases, the equivalent channel was averaged for 50 different values of relative delay n_0 between echoes and phase difference θ. As the secondary signal is assumed to be Gaussian, and, therefore, acts as an additional noise source, the system was simulated for different values of SNR, where the noise contribution takes into account both the thermal noise σ^2 and the secondary contribution, which is a function of ρ^2. In this case, the SNR (or, equivalently, the Carrier to Noise Ratio (CNR)) is calculated after the insertion of the secondary message, but does not take into account the transmission of the primary message from the secondary transmitter. The obtained BER results are shown in Figure 4, where the proposed filtering $\phi = \gamma^2/4$ is shown to outperform the non-filtered approach $\phi = 1$, and the bound, despite not being remarkably tight, leads to a system that offers a good performance and a security margin for the primary QoS.

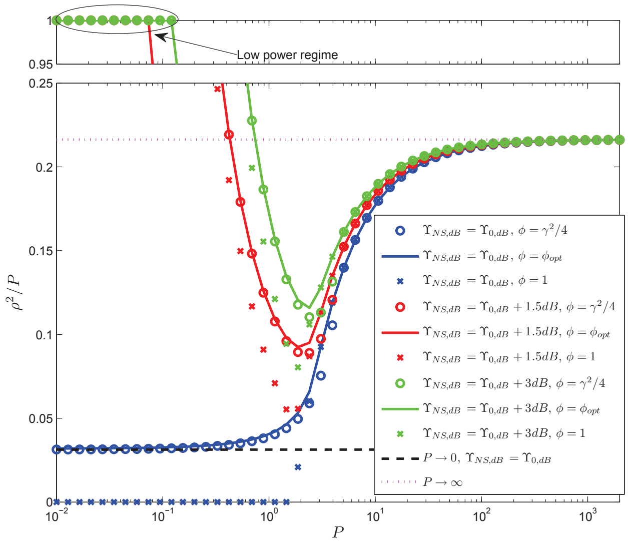

Fig. 3. Fraction of power used for the transmission of the secondary message as a function of the total received power from the secondary transmitter.

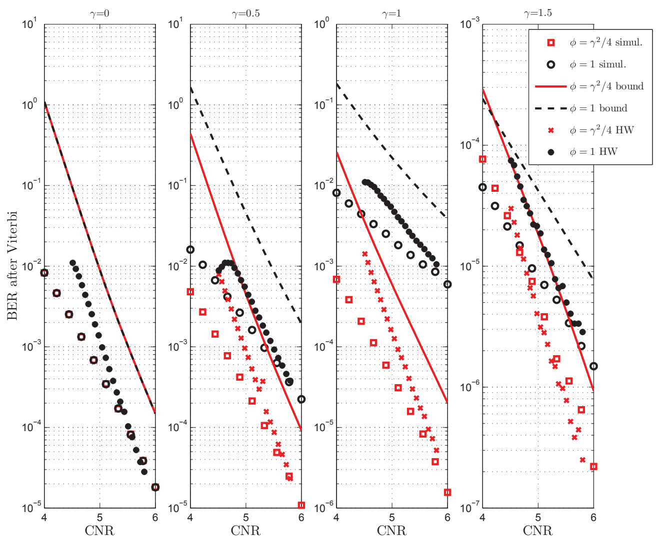

Fig. 4. Analytical bounds, simulation and Hardware (HW) results for multiple CNR and γ values. FFT Size: 8K. Constellation: QPSK. Code Rate: 2/3. The CNR is calculated prior to the transmission of the primary signal contribution coming from the secondary user, i.e., $\Upsilon = 10^{(CNR+0.33)/10}$[10].

5. CONCLUSIONS

A scenario where a secondary cognitive user is inserted in a multicarrier broadcast network has been studied. The objective of the secondary user is to maximize its own transmission rate while keeping the Quality of Service of a primary receiver over a given threshold. This constraint is fulfilled by allocating some of the secondary available power to the transmission of the primary message. The primary reception is assured by means of the Chernoff bound for the bit error rate, where the possible loss due to the insertion of an echo in a dominant line of sight environment is taken into account. The results are verified via software simulations and hardware measurements.

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