Cognitive Radio Overlay Paradigm Towards Satellite Communications

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***Abstract*—In this paper, we investigate the application of cognitive radio techniques within the overlay paradigm toward satellite communications, where the primary and cognitive users transmit concurrently at the same frequency, time and space. By means of dirty paper coding (DPC) and superposition techniques, a low complexity scheme is proposed when compared to classical precoding for interference pre-subtraction. The system performance is evaluated in different realistic scenarios, resulting in no degradation for primary user and a significant recovery of the shaping loss for the cognitive user.**

***Index Terms*—Satellite Communications, Cognitive Radio, Overlay Paradigm, Dirty Paper Coding, Trellis Shaping, Flexible Payloads.**

I. Introduction and Motivation of the Work

Satellite communication systems span across multiple applications, such as earth remote sensing, broadcast transmission of images and data, emergence backups, connecting remote regions etc. At time of writing, it is estimated over 900 satellites orbiting the earth, in an increasingly demanding and competitive market, which in turns generates new technical challenges in order to accommodate the rising new services [1]. As an example of particular relevance, we could point out the use of satellite systems for the support of machine-to-machine (M2M) communications, especially in remote areas. M2M communications are one of the central use cases in the upcoming new fifth generation (5G) standard [2]. It is predicted that the deployment of around 1 million device/km^2 that generate sporadic and low data rate packets (sensors, vehicules, factores machines,...) [3]. In this rising service, one faces three challenges: (i) battery power devices should consume as low energy as possible, (ii) use of cheap electronic components (to allow massive deployment) limits the onboard processing resources, (iii) overall spectrum scarcity requires a new optimization of the medium access and sharing.

In order to meet this demand, the communication system design for a particular mission shall be conceived basically under the following threefold factor: power efficiency, bandwidth efficiency and low complexity. However, due to the existing tradeoffs, equating these three factors often becomes a major challenge. On the other hand and by means of current measurements over the planet, one realizes that the average occupation of the spectrum as well as the actual static system for spectrum allocation should be improved [4].

In this context, with the premise of alleviating the increasingly spectral scarcity, the cognitive radio techniques have also become attractive for space applications [4], supported by the new trends in flexible payloads [5], [6] and software defined radio technology [7]. As a consequence, this research has acted as context for the development of spectral awareness and spectral exploitation techniques.

Generally, three paradigms emerge in order to classify the cognitive user (CU) (unlicensed user) communication strategies: (i) interweave, rubricated by [8] as *white* spectrum space, where the CU transmits opportunistically into the spaces not allocated by the primary user (PU) (licensed user); (ii) underlay, denominated as *gray* spectrum space, where the CU adjusts its parameters according to the PU signal characteristics and transmits, concurrently, below an interference power threshold; and (iii) overlay, where the CU, from the noncausal knowledge of the PU signal, uses advanced coding and modulation strategies to transmit simultaneously while mitigating the interference. The spectrum space in this last paradigm is called *black*, due to the fact that it is occupied by the interfering signals and noise, without power limitation.

Recently, [9], [10] and [11] addressed the previously mentioned paradigms (i) and (ii) for satellite communication systems, contributing mainly with respect to satellite-terrestrial integrated systems and dual satellite systems.

This paper explores the overlay scheme for satellite communication where both PU and CU transmit concurrently towards two different terminals, at the same frequency, time and space. To the best of the authors' knowledge, no previous works have been conducted on the overlay paradigm in the context of satellite communications. In this paper, we show that this novel approach is well suitable for low data rate M2M applications, which, in this case, plays the role of CU and has noncausal knowledge about PU. Basically, the proposed CU encoding strategy is comprised of superposition technique, in order to protect the PU, and dirty paper coding (DPC) [12], to adapts the cognitive signal to the direction of the interference. Contributing to reinforce this application, this work aims to present directions towards research to be explored, provides tools and key levers to design such systems, and evaluate the preliminary performance in a satellite communications context.

On the role of superposition, the CU shares part of its power to relay the PU signal [13]. Concerning the DPC, in a brief historical perspective, the first idea of practical scheme was proposed by Erez, Shamai and Zamir [14]. It pointed out the Tomlinson-Harashima precoding (THP) for intersymbol interference (ISI) canceling, which can be seen as a DPC application for frequency selective channel. In this technique, the modulo operation is used to pre-subtracted the interference with a mini-

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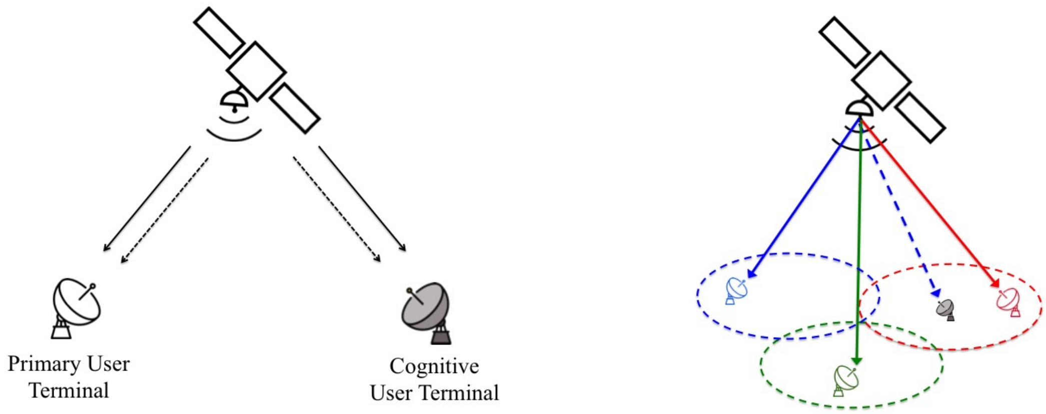


(a) LEO/MEO satellite scenario

(b) Multibeam satellite scenario

Fig. 1: Examples of Satellites Scenarios: PU - solid line; CU - dashed line

mum power increase. Also in this work, the precoding losses, i.e. shaping loss at high signal-to-noise ratio (SNR) regime and the combined modulo and power losses at intermediate/low SNR regimes, was well characterized. Moreover, Eyuboglu and Forney in their seminal paper [15] generalized the combination of the trellis shaping (TS) [16] with THP for Gaussian ISI channels. The so-called Trellis Precoding (TP) allows the recovery of the shaping loss. Likewise, a little bit more closer of our application, an extension of TP for multiuser interference was proposed to recover the shaping loss with sufficient high constellation expansion in [17] and [18], where the TS technique acts as a vetorial quantization, replacing the modulo operator. In addition to this discussion, however, a special attention must be paid in the use of output modulo signal as it incurs into nonlinear distortion. This consideration is of high interest, especially in the context of satellite communications, when the embedded High Power Amplifier (HPA) has to work near of saturation for power efficiency. The mitigation of this effect was studied in [19] and some countermeasures were disposed for satellite channels over ISI. However, this topic remains an open problem for system with modulo processing.

In view of the above, a scheme involving TS with slight constellation expansion combined with THP is proposed for DPC. This leads to a good tradeoff between power efficiency and complexity, key levers for embedded satellite processing and cheap M2M terminals.

The rest of the paper is organized as follows. The Section II presents a system model, exposing the main channel concepts. The system design, comprised by the superposition and DPC techniques are depicted in Section III. Subsequently, the simulation results are presented and discussed in Section IV and the conclusions with suggestions for future works are exposed in Section V.

II. System Model

The Fig. 1a and Fig. 1b illustrate two possible satellite scenarios in which overlay paradigm could be applied. In the first one, a LEO/MEO satellite transmits the primary and cognitive signals simultaneously towards its respective terminals from two independent antennas. The interference is present in both terminals and different techniques must be employed to protect both links. In the second scenario, we consider a multibeam satellite transmitting by three colors of frequency reuse. In this assumption, the CU transmitter is also able to transmit at the same spot of PU receiver with another color assigned and, hence, the interference among adjacent beams of same color shall be resolved.

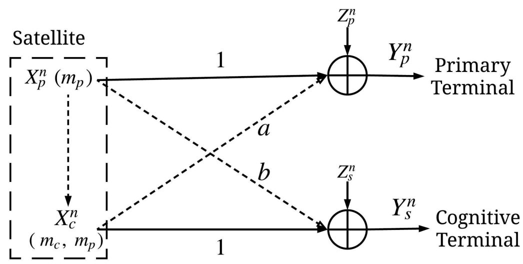


Fig. 2: Channel Model

In this context, the standard interference model with noncausal knowledge, presented at Fig. 2 and adapted from [13], can be applied. In our case, the primary and cognitive transmitters are placed onboard satellite and transmit both signals concurrently at the same frequency, time and polarization to its respective terminals. The primary and cognitive messages are defined as m_p and m_c, respectively. Without loss of generality, it is assumed that direct links have unitary gains and the parameters a and b represent the attenuation factors due to different impairments (earth station locations, antenna patterns, mismatches etc.). The transmitted signals of primary and cognitive users are defined as X_p^n and X_c^n. The power constraints to be satisfied are $E[|X_p^n|^2] = P_p$ and $E[|X_c^n|^2] \leq P_c$, respectively. At the receiver side, the received signals are described by:

$$Y_p^n = X_p^n + aX_c^n + Z_p^n \tag{1}$$

$$Y_s^n = X_c^n + bX_p^n + Z_s^n, \tag{2}$$

where the components Z_p^n and Z_s^n are assumed as a circular Gaussian noise with zero mean and powers $E[|Z_p^n|^2] = N_p$ and $E[|Z_s^n|^2] = N_s$, respectively.

III. System Design

A. Superposition strategy

Firstly, it is necessary to ensure that PU SNR remains the same even with the presence of the CU interfering signal. To solve this problem, since the cognitive encoder knows the primary message m_p as well as its modulation and coding, X_p^n can be formed by superposition coding. Thus, the CU transmitted signal can be expressed as:

$$X_c^n = \hat{X}_c^n + \sqrt{\alpha \frac{P_c}{P_p}} X_p^n, \tag{3}$$

where $\alpha \in [0,1]$ is the fraction of power P_c shared by CU to relay PU.

It is also worth highlighting from Eq. (3) that the two components $\hat{X}_c^n$ and X_p^n shall be statistically independent, otherwise the correlation may cause an additional interference to the PU link. Due to the statistical independence, the new power constraint is defined as $E[|\hat{X}_c^n|^2] \leq (1-\alpha)P_c$ and, under these circumstances, we define the signal-to-interference-plus-noise ratio (SINR) at the primary receiver by:

$$SINR_P = \frac{E[|(1 + a\sqrt{\alpha^* \frac{P_c}{P_p}})X_p^n|^2]}{E[|a\hat{X}_c^n|^2] + E[|Z_p^n|^2]}, \tag{4}$$

where $\alpha^* \in [0,1]$ is the superposition factor that should be optimized. By convenience, the noise power N_p is assumed unitary.

Observe in Eq. (4) that for no interference condition $(a = 0)$, any choices for α^* will satisfy the AWGN capacity, therefore, $\alpha^* = 0$. However, when the interference is present $(a > 0)$, the optimized superposition factor α^* that maximize Eq. (4), is given by [13]:

$$\alpha^* = \left(\frac{\sqrt{P_p}(\sqrt{1 + a^2(1+P_p)P_c} - 1}{a\sqrt{P_c}(1+P_p)} \right)^2, \tag{5}$$

It is important to note that the superposition factor α^* does not depend on b. It means that, even if the interference is high at the PU receiver, the CU is always able to reach the AWGN capacity as far as the decrease of SNR (due to power sharing) is considered.

B. Dirty Paper Coding

1- Main Concepts

The main concept behind the DPC technique is to, instead of cancel, adapt the CU signal to the interference. By rearranging different terms in Eq. (2) and combining with Eq. (3), we obtain:

$$Y_s^n = \hat{X}_c^n + \underbrace{\left(b + \sqrt{\alpha \frac{P_c}{P_p}}\right) X_p^n}_{S^n} + Z_s^n, \tag{6}$$

where S^n is the interference at the CU receiver.

The goal is then to design $\hat{X}_c^n$ in such way to pre-subtract S^n and respect the power constraint $E[|\hat{X}_c^n|^2] \leq (1-\alpha)P_c$.

According to the lattice strategy adopted in [14], at low and intermediate SNR regimes, which is the case of our scenario, a partial interference pre-subtraction (PIP) scheme must be considered. The main idea is to presubtracted λS^n, instead of S^n, with $\lambda \in [0,1]$. In this case the output cognitive signal $\hat{X}_c^n$ is formed as following:

$$\hat{X}_c^n = \left[X_{cc}^n - \lambda S^n \right] MOD_\Delta \ , \tag{7}$$

where X_{cc}^n is the coded signal, detailed in the next section, and MOD_Δ is the complex-valued modulo operation. By assuming a Gaussian distributed interference, $\hat{X}_c^n$ becomes, firstly, independent with X_{cc}^n, to satisfy DPC theory condition, and, secondly, independent with primary signal X_p^n. Otherwise Eq. (4) will not be valid and the superposition will not work properly due to the correlation. In order to achieve this statistical independence, a dither may be added at both transmitter and receiver. Finally, the modulo function (MOD_Δ) has amplitude $\Delta = \sqrt{M}d_{min}$, where M is the number of points of the expanded square QAM constellation and d_{min} is the minimum intersymbol distance.

2- Trellis Shaping Based DPC Scheme

The TS technique was introduced by Forney in [16] and could be seen as a method for sequence optimization to achieve any desired signal property. In our case, the objective is to design X_{cc}^n to be as close as possible to the scaled interference λS^n, without sacrificing the shaping gain as it is the case for THP [17]. The optimal shaping gain is knows to be 1.53 dB.

The Fig. 3 presents the implemented encoder for cognitive user. Three gains can be achieved by this systems: code gain, represented by the upper part of the diagram; shaping gain, achieved by the trellis shaping code in the lower part, and precoding gain, achieved by the modulo operation jointly with shaping code.

The two codes work independently. The input bit sequence is split into two parts. In the upper part, the k_c-bits message is encoded into a n_c-bits coded sequence by a channel code C_c. In the lower part, the r_s-bits syndrome sequence passes through an inverse syndrome former H_s^{-T} for the shaping code C_s. This initial sequence t jointly with the channel coded sequence and the scaled noncausal interference λS are fed into the Viterbi decoder. This later selects, according to a well chosen branch metric, the shaping coded sequence y_s. After that, the shaped sequence z_s is obtained by the XOR operation between t and y_s. Note that z_s and t are within the same coset, according to the trellis shaping on regions strategy, detailed in [16]. Finally, the output shaped sequence X_{cc}^n is obtained by mapping the d symbols as a function of w and the sign mapped bits z. The Viterbi decoder branch metric chosen is defined as:

$$\| \left[X_{cc}^n - \lambda S^n \right] MOD_\Delta \|^2 \tag{8}$$

The key difference between our implementation and [17] and [18] is the combination of a less expanded constellation with a modulo operation over the expanded constellation. Despite of the fact that only partial shaping loss is recovered, this practice reduces the complexity avoiding exhaustive comparisons in the shaping Viterbi decoder. In addition, the use of the modulo guarantees that the transmitted power is limited even in high interference scenarios [14].

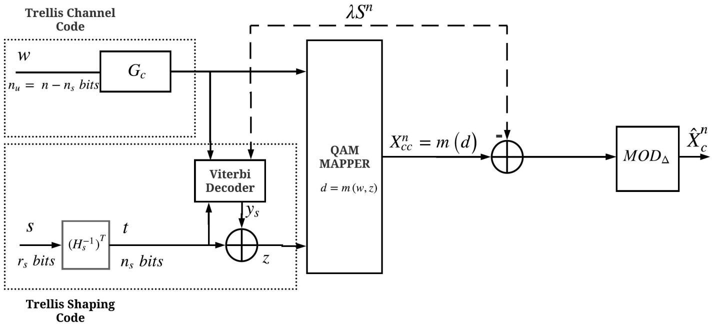


Fig. 3: Proposed DPC Encoder

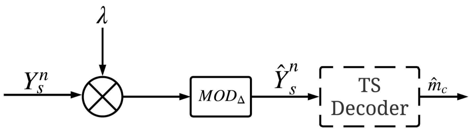


Fig. 4: DPC Receiver

3- DPC Receiver

The Fig. 4 presents the DPC receiver. Basically, the reverse chain is implemented: firstly, Y_s^n is multiplied by the factor λ. Before enter in the DPC decoder, the signal is modulo operated again. At the decoder input, the signal is:

$$\hat{Y}_s^n = \Big[(\hat{X}_c^n + S^n + Z^n)\lambda\Big] MOD_\Delta \tag{9}$$

$$= \Big[\lambda \hat{X}_c^n + (X_{cc}^n - \hat{X}_c^n) MOD_\Delta + \lambda Z_s^n\Big] MOD_\Delta \tag{10}$$

$$= \Big[X_{cc}^n - (1-\lambda)\hat{X}_c^n + \lambda Z_s^n\Big] MOD_\Delta, \tag{11}$$

where in Eq. (11) the following property was utilized $[(a)MOD_\Delta + b]MOD_\Delta = \ (a+b)MOD_\Delta$.

The value of λ that minimizes the effective noise $(1-\lambda)\hat{X}_c^n + \lambda Z_s^n$ is obtained by [12]:

$$\lambda = \frac{(1-\alpha^*)P_c}{(1-\alpha^*)P_c + E[|Z_s^n|^2]}. \tag{12}$$

It is worth noting that in case of full recovery of the shaping loss by $\hat{X}_c^n$, the capacity of the AWGN channel is achieved and the factor λ presented in Eq. (12) is optimum, since the effective noise has, in fact, Gaussian distribution.

Finally, the decoder for the signal $\hat{Y}_s$ is identical to referenced in [15], which is comprised by a Trellis decoder for C_c and the syndrome mapper for C_s.

IV. Simulation and Results

In this section, without loss of generality, we assume that a and b are real values and equal to 0.2.

For the PU, a 4-QAM transmitted signal encoded by a 16-state, rate $1/2$, convolutional code C_c specified in octal notation by generators $g_1(D) = 31$ and $g_2(D) = 33$ was implemented. The transmitted power values P_p of 3, 7 and 15 dB were considered in accordance with the ranges of DVB-S2 standard [1].

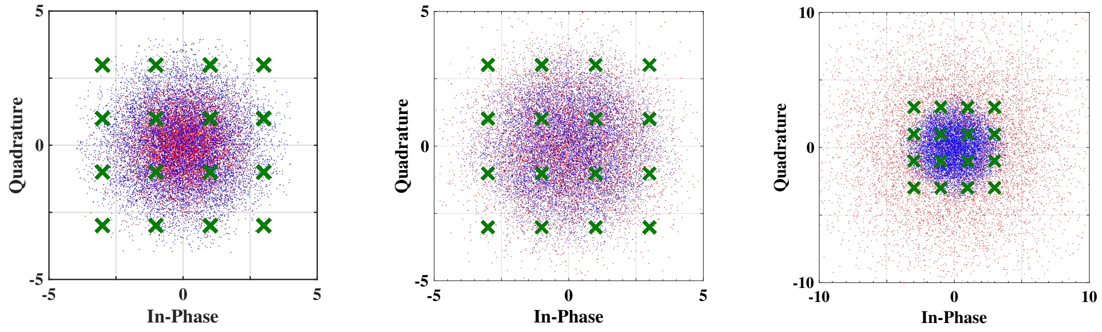
For the CU, the available remaining symbol power (after the superposition) was assumed as 10 dB, with a transmission rate $R_{cu} = 2$ bits/symbol. We utilized a slight expanded constellation of $n_s = 2$, in such way to respect the maximum 16-QAM standardized constellation in DVB-S2. Further, a systematic 64-state, rate $1/2$, convolutional code specified in octal notation by the feedforward polynomial $h_1(D) = 54$ and the feedback $h_0(D) = 161$ was assumed jointly with the 4-state, rate $1/2$, *Ungerboeck* code for C_s, specified by generators $g_{s,1}(D) = 7$ and $g_{s,2}(D) = 5$.

The Fig. 5 presents the signals constellations involved by the described processing. The expanded constellation signal X_{cc}^n, from original 4-QAM to shaped 16-QAM, is shown in green "x". The Gaussian distributed version of the interference λS^n is superposed in red points and, concluding the chain, the transmitted signal $\hat{X}_c^n$, after presubtraction and modulo operation, is shown in blue dots.

We notice that in Fig. 5a, low interference condition, almost all interference is confined within the expanded constellation. In this case, a significant part of shaping loss is recovered. In Fig. 5b, we note that, even at higher interference, a considerable part remains inside the expanded constellation set, which led us to expect that some shaping loss could be recovered by the system. Finally, in Fig. 5c, we observe that the expansion is not sufficient for this strong amplitude of interference and a near-uniform distribution is expected after shaping operation, i.e. no shaping gain is achieved.

The Fig. 6 presents the bit error rate (BER) curves for the presented schemes. For reference, the curve of no interference condition is also plotted. As expected, for P_p of 3 and 7 dB, the shaping loss can be partially recovered, resulting in shaping gain around 1.0 and 0.5 dB respectively (considering a BER of 10^{-3}), when compared with P_p of 15 dB. The gap observed in relation to TS AWGN channel curve emphasizes the work [17], which recommends a higher constellation expansion (n_s about 5 or 6) to accommodate DPC application, that might not be reasonable for satellite and M2M applications due to high complexity and peak-to-average-power-ratio (PAPR).

Moreover, the Fig. 6 may suggest that DPC performs better than TS at low SNR regime. This is not the case, since the ratio here does not consider the previously defined effective noise. This can be explained by the effect



(a) $P_P = 3$ dB (b) $P_P = 7$ dB (c) $P_P = 15$ dB

Fig. 5: Scatter plots of signals constellations at different PU interfering powers

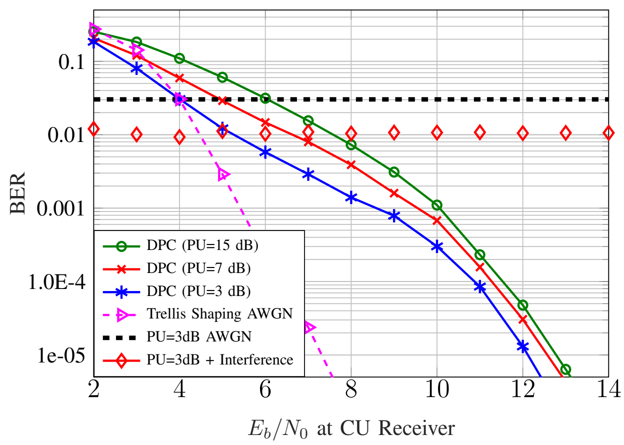


Fig. 6: BER Cognitive User

of λ on Z_s^n, especially at this regime (cf. Eq. (12)).

Furthermore, to verify that PU operates properly, the BER simulated for PU at P_p of 3 dB is depicted in presence and absence of CU interference. Thus, thanks to the superposition the BER PU is not degraded. In our case, the superposition even improves this BER. This can be explained by the fact that the superposition factor α^* was designed considering $E[|\hat{X}_c^n|^2] = (1-\alpha)P_c$. However, thanks to the trellis precoding, the transmitted power is actually $E[|\hat{X}_c^n|^2] \leq (1-\alpha)P_c$.

V. Conclusion

This paper proposes a low complexity transmission scheme using cognitive radio overlay paradigm towards satellite application, introducing some possible scenarios. The combination of trellis precoding and THP was implemented and the results presented no PU degradation and a partial shaping loss recovering for CU signal.

In the future research, we will investigate the effect of unwanted impairments typical on satellite communication, such as nonlinear distortions, jitter, amplitude/phase umbalances and phase noise.

Acknowledgment

This work was supported by National Council for Scientific and Technological Development (CNPq/Brazil) and by National Institute for Space Research (INPE/Brazil).

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