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## To cite this version:

Sumit J. Darak, Sumedh Dhabu, Christophe Moy, Honggang Zhang, Jacques Palicot, et al.. Low complexity and efficient dynamic spectrum learning and tunable bandwidth access for heterogeneous decentralized cognitive radio networks. Digital Signal Processing, 2015, 37, pp.13-23. 10.1016/j.dsp.2014.12.001 . hal-01136014

## HAL Id: hal-01136014

## https://hal.science/hal-01136014

Submitted on 4 Nov 2015

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## Highlights

- Low complexity variable bandpass filter (VDF) design
- Tunable dynamic spectrum learning and access (DSLA) policy for heterogeneous cognitive radio networks (CRNs)
- System level approach to integrate VDF with the DSLA policy to support tunable bandwidth access in heterogeneous CRNs
- Higher spectrum efficiency and lower subband switching cost


# Low Complexity and Efficient Dynamic Spectrum Learning and Tunable Bandwidth Access for Heterogeneous Decentralized Cognitive Radio Networks 

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#### Abstract

This paper deals with the design of the low complexity and efficient dynamic spectrum learning and access (DSLA) scheme for next-generation heterogeneous decentralized Cognitive Radio Networks (CRNs) such as Long Term EvolutionAdvanced and 5G. Existing DSLA schemes for decentralized CRNs are focused predominantly on the decision making policies which perform the task of orthogonalization of secondary users to optimum vacant subbands of fixed bandwidth. The focus of this paper is the design of DSLA scheme for decentralized CRNs to support the tunable vacant bandwidth requirements of the secondary users while minimizing the computationally intensive subband switchings. We first propose a new low complexity VDF which is designed by modifying second order frequency transformation and subsequently combining it with the interpolation technique. It is referred to as Interpolation and Modified Frequency Transformation based VDF (IMFT-VDF) and it provides tunable bandpass responses anywhere over Nyquist band with complete control over the bandwidth as well as the center frequency. Second, we propose a tunable decision making policy, $\rho^{t-r a n d}$, consisting of learning and access unit, and is designed to take full advantage of exclusive frequency response control offered by IMFT-VDF. The simulation results verify the superiority of the proposed DSLA scheme over the existing DSLA schemes while complexity comparisons indicate total gate count savings from $11 \%$ to as high as $87 \%$ over various existing schemes. Also, lower number of subband switchings make the proposed scheme power-efficient and suitable for battery-operated cognitive radio terminals.


[^0]Keywords: Decentralized cognitive radio networks, dynamic spectrum learning and access, frequency transformation, multi-armed bandit, variable digital filter.

## 1. Introduction

Dynamic spectrum learning and access (DSLA) based Cognitive Radio Networks (CRNs) paradigm has been suggested to improve the overall utilization of the spectrum, which is crowded yet underutilized mainly due to static spectrum allocation policies [1, 2]. In CRNs, multiple secondary (unlicensed) users can access the vacant frequency subbands in an opportunistic way with minimal interference to the primary (licensed) users [1, 2]. Recently, the term "Green CRN" has been coined, which refers to the CRN which will minimize the $\mathrm{CO}_{2}$ emissions (by using area and energy efficient radio terminals) along with improving the spectrum utilization $[3,4]$. These CRNs can be deployed for smart grid networks, public safety networks, health-care networks, cellular networks [5] etc.

The two representative DSLA approaches are: 1) DSLA for centralized CRNs (centralized DSLA): A central unit is responsible for orthogonalization of $M$ secondary users to optimum subbands (i.e., the subset of $M$ subbands which are identified as superior, in terms of their probability of being vacant, lower interference and noise levels etc., amongst all available subbands in wideband input signal) $[6,7], 2)$ DSLA for decentralized CRNs (decentralized DSLA): Each secondary user independently learns, senses and accesses the vacant subbands without any explicit information exchanges or pre-agreements with other users [6-14]. The centralized DSLA schemes using game theory approach, auction based approach, multi-armed bandit based approach etc. have been proposed and more details can be found in $[6,7]$. Various decentralized DSLA schemes are discussed in detail in Section 2.2. In summary, the decentralized DSLA has the advantages such as ease of implementation, robustness to link or node failures, no communication overhead and lower delay over the centralized DSLA [6-14]. As a result, decentralized DSLA is preferred choice for public safety networks and proximity-aware social networking services. However, decentralized DSLA suffers from lower utilization of optimum subbands due to the potentially higher number of collisions between the secondary users. This is because of the lack of coordination among secondary users as well as hardware and signal processing constraints at each secondary user terminal.

Existing centralized as well as decentralized DSLA schemes [6-11, 13, 14] assume that the vacant bandwidth requirements of all secondary users are fixed and equal. Such assumption may not be true in the next-generation heterogeneous CRNs such as 5G where multiple wireless communication standards with distinct channel bandwidths co-exist. In order to extend the capabilities of such decentralized CRNs beyond voice to video, data, location and Internet based services demanding distinct bandwidths and to seamlessly integrate multiple

Another constraint for efficient DSLA scheme is the total penalty incurred in terms of delay, energy consumption, hardware reconfiguration and protocol overhead when secondary user switches from one frequency subband to another. Hereinafter, this penalty is referred to as subband switching cost (SSC) $[6,7]$.
${ }_{45}$ The total number of subband switchings and hence, SSC should be as small as possible, especially for the resource-constrained battery-operated secondary user terminals. Though the challenges of minimizing the collisions among secondary users are attracting much attention in the research community today, the design of decentralized DSLA scheme with additional constraints of tunable vacant bandwidth and minimum SSC for heterogeneous CRNs has yet to be systematically addressed. The stringent hardware specifications of minimum area and power make the design even more challenging and are the motivations behind the work presented in this paper. The proposed decentralized DSLA scheme consists of: scheme is not just a straightforward integration of the VDF and existing policy $\rho^{r a n d}$. In fact, it is designed by exploiting the exclusive unabridged center frequency control offered by the IMFT-VDF and adapting the $\rho$ rand to efficiently work with the IMFT-VDF.

The work presented in this paper is a significant extension of our preliminary works in $[12,15]$. For instance, the proposed IMFT-VDF offers complete control over the bandwidth as well as center frequency of bandpass responses compared to the coarse and limited control offered by our VDF in [15]. Furthermore,
the design, architecture as well as gate count complexity are discussed in detail in this paper. Our DSLA scheme in [12] was designed for CRNs consisting of single secondary user, which is now extended for more practical and challenging scenarios of multi-standard multi-user CRNs.

The rest of the paper is organized as follows. The details of network model and the literature review of different stages of DSLA scheme are presented in Section 2. In Section 3, the design and architecture of the proposed IMFT-VDF are discussed along with suitable design examples and gate count complexity comparisons. The proposed policy, $\rho^{t-r a n d}$, is presented in Section 4 followed by the simulation results and complexity comparisons in Section 5 . Section 6 concludes the paper.

## 2. Network Model and Literature Review

Consider the slotted CRN consisting of $M$ secondary users. The desired vacant bandwidth of any $k^{t h}$ secondary user is denoted by $B_{v}(k), k \in\{1,2, . ., M\}$. The wideband input signal consists of at most $N_{p}$ active primary users with channel bandwidths, $B_{p}(j), j \in\left\{1,2, . ., N_{p}\right\}$, which are integer multiples of the smallest channel bandwidth, $B_{\text {cmin }}$ such that $N_{c}=\left(1 / B_{\text {cmin }}\right), \sum_{j} B_{p}(j) \leq 1$ and $M<N_{c}$. Note that all the frequency specifications are normalized with respect to half the sampling frequency. Let $P_{\text {vac }}(i), i \in\left\{1,2, . ., N_{c}\right\}$ be the probability of vacancy for each subband of bandwidth $B_{\text {cmin }}$. It is assumed to evolve as an i.i.d (independent and identically distributed) Bernoulli random process with time, $t$, stationary and unknown to secondary users [8-14].

Within each secondary user, DSLA scheme can be implemented through filtering stage, detection stage and decision making stage as illustrated in Fig. 1 [12]. Depending on the desired vacant bandwidth, $B_{v}(k)$, the decision making stage of corresponding $k^{\text {th }}$ secondary user instructs the filtering stage to extract the desired frequency subband of interest via parameters, $B_{v}(k)$ and $f_{\text {center }}(k)$ (for center frequency). The extracted subband is then sensed using an imperfect detector i.e., sensing errors may occur. Due to resource and hardware constraints of secondary user terminal, it is assumed that each secondary user can sense only one subband at a time.


Figure 1: Decentralized DSLA scheme in $k^{t h}$ secondary user terminal.
When $B_{v}(k)=B_{\text {cmin }}$, then the corresponding $k^{t h}$ secondary user can choose any one of $N_{r}(k)=N_{c}$ orthogonal subbands. However, when $B_{v}(k)>B_{c m i n}$,
the number of subbands choices for the corresponding $k^{\text {th }}$ secondary user are $N_{r}(k)=\left[N_{c}+1-\left\lceil\left(B_{v}(k) / B_{c m i n}\right)\right\rceil\right]$ non-orthogonal subbands. When multiple secondary users transmit on the non-orthogonal subbands, collision occurs and
dins can afford only one detector due to the imposed power constraints for longer battery life and hence, only one subband or multiple contiguous subbands
in conjunction can be sensed at a time. Since, the other extracted subbands are not used, filter bank is not utilized to its full potential. Also, filter banks are not efficient for wideband input signal with non-contiguous bandwidth. In such cases, VDFs offer an efficient alternative.

### 2.1.2. Variable Digital Filters

In memory based digital FIR filters, filter coefficients corresponding to all desired bandpass response specifications are stored in memory beforehand. Such filter based scheme is area and power efficient for secondary user terminals due to fewer number of multipliers and adders. However, filter coefficients need to be updated frequently whenever $B_{v}(k)$ and/or $f_{\text {center }}(k)$ changes, which incurs huge penalty in terms of sensing time due to multiple memory read/write operations. Higher sensing time means less time for data communication and hence, lower spectrum throughput. Furthermore, memory size increases with the increase in the granularity of $B_{v}(k)$ and $f_{\text {center }}(k)$ and is very large. As a result, DSLA using memory based FIR filter is suitable for CRN supporting single communication standard, i.e., $B_{v}(k)$ is fixed and $f_{\text {center }}(k)$ is limited to a set of fewer discrete values.

The VDF provides on-the-fly control over $B_{v}(k)$ and $f_{\text {center }}(k)$ of the bandpass response through a small number of parameters [15-26]. The design of linear phase bandpass VDF for decentralized DSLA is a challenging research problem and highly nontrivial because of the fixed transition bandwidth (TBW) requirement as well as area, power and delay constraints.

A number of bandpass VDF designs are available in the literature [15-26]. The allpass transformation (APT) based VDF [23] offers complete control over the center frequency of bandpass responses but the phase is non-linear which limits its usefulness in DSLA scheme. The frequency response masking based VDFs [16] offer complete control over the cut-off frequency of lowpass responses and are suitable for applications requiring sharp TBW. However, their usefulness in the case of bandpass responses with complete control over $B_{v}(k)$ and $f_{\text {center }}(k)$ has not been explored and the group delay of the [16] is very high. The first and second order frequency transformation based VDFs [15, 17, 18] and coefficient decimation method (CDM) based VDFs [24] offer only coarse control over the $B_{v}(k)$ and $f_{\text {center }}(k)$. The spectral parameter approximation based VDFs (SPA-VDFs) [19-21] have the advantages of fixed TBW, fewer variable multipliers and lower group delay. However, the gate count complexity of these VDFs is very high and large dynamic range of filter coefficients may impose constraints when fixed-point implementation is desired. Though the SPA-VDF in [19] and our SPA-MCDM-VDF [20] overcome some of these drawbacks, their gate count complexity is still high.

### 2.2. Decision Making Stage

Various decision making policies have been proposed for the decentralized DSLA in single user as well as multi-user CRNs [6-14]. These policies have shown to outperform the conventional random selection based policies under
self-play i.e. when implemented at all secondary users [6-14]. They mainly differ in the type of online learning algorithms they employ for decision making such as $\varepsilon$-greedy algorithm, Multi-Armed Bandit based upper confidence bound (UCB) algorithms such as $U C B_{1}, U C B_{H}, U C B_{t}, U C B_{v}$ etc. [12-14] or combination of these algorithms [10] as well as different strategies for orthogonalization of secondary users [8-11]. Analytically, the UCB based algorithms [13, 14] are mathematically proved to be optimal with logarithmic regret bound, which is the best we can expect when secondary users do not have any information about $P_{v a c}(i) \forall i$. Hence, the discussion in this paper is limited to the UCB algorithm.

### 2.2.1. CRN with Single Secondary User ( $M=1$ )

The direct application of UCB algorithm for single user CRN $(M=k=1)$ is discussed in $[6,7,12-14]$. In brief, secondary user sequentially but randomly senses all $N_{r}(1)$ subbands once at the beginning. Then, for $t>N_{r}(1)$, the subband, $I_{t}(1)$, with the maximum value of g -statistic, $B_{l_{1}, 1}(t)$, is chosen [13]. The $g$-statistic, $B_{l_{1}, 1}(t)$, for channel $l_{1}$ and time $t$ is the sum of the learned mean availability of the subband, $\bar{X}_{l_{1}, 1}(t)$ (exploitation factor) and UCB-bias, $A_{l_{1}, 1}(t)$ (exploration factor) [13]. Mathematically [13],

$$
\begin{equation*}
I_{t}(1)=\underset{l_{1}}{\arg \max } B_{l_{1}, 1}(t) \tag{2}
\end{equation*}
$$

where

$$
\begin{equation*}
B_{l_{1}, 1}(t)=\underbrace{\frac{X_{l_{1}, 1}(t)}{T_{l_{1}, 1}(t)}}_{\bar{X}_{l_{1}, 1}(t)}+\underbrace{\sqrt{\frac{2 \cdot \ln (t)}{T_{l_{1}, 1}(t)}}}_{A_{l_{1}, 1}(t)} \forall l_{1} \tag{3}
\end{equation*}
$$

In [12], we proposed DSLA scheme using UCB to support the tunable vacant bandwidth requirements for CRNs with single secondary user.

### 2.2.2. $C R N$ with Multiple Secondary Users $(M>1)$

For practical scenario where CRN consists of $M(>1)$ secondary users, notable research on the integration of learning algorithms such as UCB algorithm with different access strategies for orthogonalization of secondary users to optimum subbands has been pursued in [8-11].

In particular, the $\rho^{\text {rand }}$ policy in [8], achieves orthogonalization by randomly and independently assigning rank to each secondary user. For example, $k^{t h}$ secondary user with the rank, $p_{k} \in\{1,2, . . M\}$, selects the subband having $p_{k}^{t h}$ maximum value of $g$-statistic, $B_{l_{k}, k}(t)$. When collision occurs, the corresponding secondary users randomly and independently change their rank. Another access policy in [9] follows time division fare share approach where the rank of each secondary user is not fixed but is updated in round robin fashion to allow equal access to optimum subbands among all secondary users. Analytically, both the policies are simple, order-optimal and their regret, $U_{t}$, is logarithmic with time $t$. However, the SSC of the policy in [9] increases linearly with $t$ since each secondary user switches the subband in each time slot compared to the policy

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in [8] where subband switching occurs only when secondary user experiences collision. Thus, policy in [8] is preferred over the policy in [9].

The mixed strategy based policy in [10] employs soft-max action selection method that uses the Boltzmann distribution. The performance of this method is similar to [8] but its SSC is high and convergence has not been proved analytically. The policy in [11] is designed by combining access strategies in [8, 9] with two learning algorithms to take into account different $P_{\text {vac }}$ distributions. However, the SSC of [11] is high for the $P_{v a c}$ distributions where it performs better than [8]. The proposed policy is discussed later in Section 4.

## 3. Proposed IMFT-VDF

The IMFT-VDF is the first contribution of this paper and constitutes the filtering stage of the proposed decentralized DSLA scheme shown in Fig. 1.

### 3.1. IMFT-VDF Design

Consider a linear-phase prototype lowpass filter, $H(Z)$, of order $2 N$ and symmetric filter coefficients $h_{n}, n \in\{0,1, . .2 N\}$, i.e., $h_{n}=h_{2 N-n} \forall n$. The transfer function of the prototype filter, $H(Z)$, in Taylor form is given as [17],

$$
\begin{equation*}
H(Z)=\sum_{n=0}^{N} a_{n} Z^{-N}\left[\frac{Z+Z^{-1}}{2}\right] \tag{4}
\end{equation*}
$$

where the coefficients $a_{n}$ are related to the impulse response coefficients $h_{n}$ through the Chebyshev polynomials [17]. The second order frequency transformation based VDF, $H_{2}(z)$, is then given by [17],

$$
\begin{equation*}
H_{2}(z)=\sum_{n=0}^{N} a_{n} Z^{-N} D(z) \tag{5}
\end{equation*}
$$

where $Z \rightarrow z$ represents the frequency transformation [17]. The term $D(z)$ denotes the second order frequency transformation and is given by [17],

$$
\begin{equation*}
D(z)=\frac{Z+Z^{-1}}{2}=\sum_{v=0}^{2} A_{v}\left(\frac{z+z^{-1}}{2}\right)^{v} \tag{6}
\end{equation*}
$$

where $A_{v}, v \in\{0,1,2\}$, are the controlling parameters of $D(z)$ and they decide the cut-off frequency and the TBW of the $H_{2}(z)$ [17]. Substituting Eq. 6 into Eq. 5, we get

$$
\begin{equation*}
H_{2}(Z)=\sum_{n=0}^{N} a_{n} z^{-2(N-n)}\left[\sum_{v=0}^{2} A_{v} z^{v-2}\left(\frac{1+z^{-2}}{2}\right)^{v}\right]^{n} \tag{7}
\end{equation*}
$$

Let $\omega_{c}$ and $\Omega_{c}$ be the cut-off frequencies of the $H(Z)$ and $H_{2}(z)$, respectively. Similarly, $T B W_{p}$ and $T B W_{2 f t}$ are the TBW of the $H(Z)$ and $H_{2}(z)$, respectively. Then, $\Omega_{c}$ and $T B W_{2 f t}$ are given by [17],

$$
\begin{align*}
& \Omega_{c}=\cos ^{-1}\left(\frac{-A_{1} \pm \sqrt{A_{1}^{2}-4 A_{2}\left(A_{0}-\cos \omega_{c}\right)}}{2 A_{2}}\right)  \tag{8}\\
& T B W_{2 f t}=T B W_{p} \cdot\left(\frac{A_{1} \sin \Omega_{c}+A_{2} \sin \left(2 \Omega_{c}\right)}{\sin \omega_{c}}\right) \tag{9}
\end{align*}
$$

if the following constraints are met [17]:

$$
\begin{align*}
A_{0}+A_{1}+A_{2} & =1  \tag{10a}\\
0 \leq A_{1} & \leq 1  \tag{10b}\\
A_{1}^{2}-4 A_{2}\left(1-A_{1}-A_{2}-\cos \omega_{c}\right) & \geq 0 \tag{10c}
\end{align*}
$$

In [17], $A_{1}=1$ and hence, according to Eq. 10(a), $A_{2}=-A_{0}$, which eventually leads to reduction in the number of variable multipliers from 3 to 1. However, the detailed analysis of Eq. 8 and Eq. 10 indicate that, by restricting $A_{1}=1$, the cut-off frequency range of $H_{2}(z)$ is limited to approximately $25 \%$ of the Nyquist band. This in turn narrows the bandwidth and the center frequency range of the bandpass responses. Our modified second order frequency transformation based VDF in [15] allows $A_{1}$ to take any reciprocal of power-oftwo value instead of unity. The resultant VDF is then combined with the CDM to obtain variable lowpass, highpass, bandpass and bandstop responses from a fixed-coefficient lowpass prototype filter. However, it can not provide complete control over the center frequency and the bandwidth of the bandpass response over the entire Nyquist band. Furthermore, the TBW deteriorates significantly at higher cut-off frequencies. To overcome these drawbacks, especially for the DSLA application, a new IMFT-VDF based on the modified frequency transformation and interpolation technique is proposed. The main differences between IMFT-VDF and our VDF in [15] are :

1. The parameter $A_{1}$ is tunable as against to fixed $A_{1}$ in [15] but restricted to the reciprocal of power-of-two values and zero.
2. The architecture consists of two branches of $D(z)$ (discussed in detail later in Section 3.3) compared to single branch of $D(z)$ in [15].
3. The VDF in [15] is combined with the CDM to obtain variable bandpass responses from variable lowpass responses. In IMFT-VDF, the interpolation and conventional lowpass to highpass conversion by reversing the sign of every alternate coefficient are used to increase the tunable cut-off frequency range of lowpass responses.

### 3.2. IMFT-VDF Design Procedure

Consider the design of the IMFT-VDF which provides complete control over the cut-off frequency of lowpass response from $\Omega_{c 1}$ to $\Omega_{c 2}$ with the desired TBW
of $T B W_{d}$, desired passband and stopband ripples of $\delta_{p d}$ and $\delta_{s d}$, respectively. The desired bandpass response with lower and upper cut-off frequencies $\Omega_{b p c 1}$ and $\Omega_{b p c 2}$, respectively, is obtained by subtracting a lowpass response with cutoff frequency of $\Omega_{b p c 1}$ from another lowpass response with cut-off frequency of $\Omega_{b p c 2}$ using the same prototype filter and two branches of $D(z)$. Without loss of generality, we have $\Omega_{b p c 1} \geq \Omega_{c 1}, \Omega_{b p c 2} \leq \Omega_{c 2}, \Omega_{c 1} \approx 0$ and $\Omega_{c 2} \approx 1$. The design steps of IMFT-VDF are:

1. The $\hat{\Omega}_{c 1}$ is chosen such that $\hat{\Omega}_{c 1}=\min \left\{\Omega_{c 1},\left(1-\Omega_{c 2}\right)\right\}$ and $\hat{\Omega}_{c 2}=\left(1-\hat{\Omega}_{c 1}\right)$ to maintain the symmetry with respect to center of the Nyquist band.
2. The lowpass cut-off frequency range from $\hat{\Omega}_{c 1}$ to $\hat{\Omega}_{c 2}$ is divided into two equal parts : a) $\hat{\Omega}_{c 1}$ to 0.5 , b) 0.5 to $\hat{\Omega}_{c 2}$.
3. For the desired frequency specifications of $T B W_{d}, \delta_{p d}$ and $\delta_{s d}$, the cut-off frequency of the prototype filter, $\omega_{c}$, the parameters $A_{1}$ and $A_{2}$ are chosen using the Eq. 8-10 such that lowpass responses with desired frequency specifications and cut-off frequency range from $\bar{\Omega}_{c 1}$ to 0.5 are obtained. This is a simple optimization problem to obtain $\bar{\Omega}_{c 1}$ to be as close to $\hat{\Omega}_{c 1}$ as possible with the constraints of meeting the desired frequency specifications.
4. When $\bar{\Omega}_{c 1} \leq \hat{\Omega}_{c 1}$, interpolation is not required. Otherwise, lowpass responses with cut-off frequency $\Omega_{c}$ where $\hat{\Omega}_{c 1} \leq \Omega_{c} \leq \bar{\Omega}_{c 1}$ or $\bar{\Omega}_{c 2}(=$ $\left.1-\bar{\Omega}_{c 1}\right) \leq \Omega_{c} \leq \hat{\Omega}_{c 2}$ are obtained using conventional interpolation technique. Corresponding interpolation factor is $p \approx\left(\bar{\Omega}_{c 1} / \hat{\Omega}_{c 1}\right)$ such that $p \in\{2,4,8 \ldots\}$ and masking filter, $H_{m}(z)$, is required to mask the undesired subbands.
5. Finally, the lowpass response with $0.5 \leq \Omega_{c} \leq \bar{\Omega}_{c 2}$ is obtained from the corresponding lowpass response with cut-off frequency of $\left(1-\Omega_{c}\right)$ by changing the sign of every alternate filter coefficients and then taking the complementary response. Note that $\bar{\Omega}_{c 1} \leq\left(1-\Omega_{c}\right) \leq 0.5$.

The IMFT-VDF is not just a straightforward integration of the frequency transformation and interpolation. In fact, the IMFT-VDF is carefully designed by exploiting the architectural advantages such as lower deterioration in TBW and lower multiplier complexity of the proposed second order modified frequency transformation as well as exclusive multiband response capability of the conventional interpolation technique, without compromising on the total gate counts. The architecture of the IMFT-VDF is discussed in the next section.

### 3.3. IMFT-VDF Architecture

The architecture of the IMFT-VDF is shown in Fig. 2 where the modified second order frequency transformation is realized through $D(z)$ shown in Fig. 3. The IMFT-VDF architecture consists of prototype filter of order $2 N$ with coefficients, $a_{0}, a_{1}, . ., a_{N}$, obtained from the original filter coefficients, $h_{0}, h_{1}, . . h_{2 N}$, as discussed in Section 3.2. These coefficients are fixed and can be hardwired. In the IMFT-VDF architecture, the Taylor structure of filter implementation is redesigned by shifting each $D(z)$ block adjacent to the corresponding delay block.

Even though the number of delay blocks are higher in the transposed Taylor structure compared to conventional Taylor structure [17], the total group delay of both structures is equal for a given $N$. Moreover, in the FPGA implementation of the VDF, delays can be implemented without significant hardware cost.

* $D(z)=$ Modified second order frequency transformation, ${ }^{*} p=$ Interpolation factor.

Figure 2: Architecture of the IMFT-VDF using transposed Taylor structure.
The $D(z)$ is given by,

$$
\begin{equation*}
D(z)=A_{0} z^{-2}+A_{1}\left(\frac{1+z^{-2}}{2}\right)+A_{2}\left(\frac{1+z^{-2}}{2}\right)^{2} \tag{11}
\end{equation*}
$$

From Eq. 10(a), we have $A_{0}=1-A_{1}-A_{2}$. Substituting this relationship into Eq. (11) and simplifying, we obtain

$$
\begin{equation*}
D(z)=z^{-2}+A_{1}\left[\left(\frac{1+z^{-2}}{2}\right) z^{-1}-z^{-2}\right]+A_{2}\left[\left(\frac{1+z^{-2}}{2}\right)^{2}-z^{-2}\right] \tag{12}
\end{equation*}
$$

Using Eq. 12 and interpolation factor $p, D(z)$ can be efficiently implemented as shown in Fig. 3.

The IMFT-VDF comprises of two parallel branches of $D(z)$, controlled via signals $S_{x p}, C_{x n}, A_{1 x}$ and $A_{2 x}$ where $x \in\{1,2\}$. Each branch provides either lowpass response (when $p=1$ ) or multiband response (when $p>1$ ) depending on the value of corresponding $S_{x p} . A_{1 x}$ and $A_{2 x}$ control the values of parameters $A_{1}$ and $A_{2}$, respectively. $C_{x n}$ controls adder-subtractor block where $C_{x n}=1$ corresponds to the addition and $C_{x n}=-1$ corresponds to the subtraction. In general, $C_{x n}=(-1)^{n}$ when the desired cut-off frequency lies in the range 0.5 to $\bar{\Omega}_{c 2}$. Otherwise, $C_{x n}=1 \forall n$. The outputs of these two branches, $y_{1}$ and $y_{2}$,
 are given to the output logic unit.

This is done to obtain multiple lowpass responses, each with distinct cut-off frequency, using single fixed-coefficient prototype filter and multiple branches of $D(z)$. The proposed structure is referred to as transposed Taylor structure.


Figure 3: Modified second order frequency transformation, $D(z)$.

The output logic unit is depicted in Fig. 4. It consists of $N_{m}^{t h}\left(N_{m} \ll\right.$ $N$ ) order fixed-coefficient masking filter, $H_{m}(z)$, to mask the undesired subbands in multiband response when $p>1$ as discussed in Section 3.2. The cut-off frequency, TBW, passband and stopband ripples of $H_{m}(z)$ are $(1 / p)$, layed version of input signal). In Fig. 4, $y_{1 c}$ is a complementary response of $y_{1}$ obtained by subtracting $y_{1}$ from an appropriately delayed version of the input signal.


Figure 4: Output logic unit.
The multiplexer control signal, sel_y, is a 6 bit signal which controls two 2:1 multiplexers and two $4: 1$ multiplexers as shown in Fig. 4. The IMFTVDF architecture provides variable lowpass, highpass, bandpass and bandstop responses using fixed-coefficient lowpass filter. The control signals to obtain variable bandpass responses are given in Table 1 where " X " denotes "don't care"condition. Similarly, variable lowpass responses can be obtained using
e $\{101100,101111,101111,101110\}$ for cut-off frequency lying in the range $\hat{\Omega}_{c 1}$ to $\bar{\Omega}_{c 1}, \bar{\Omega}_{c 1}$ to $0.5,0.5$ to $\bar{\Omega}_{c 2}$ and $\bar{\Omega}_{c 2}$ to $\hat{\Omega}_{c 2}$, respectively. The bandstop and highpass responses can be obtained as the complementary responses of the corresponding bandpass and lowpass responses.

Table 1: Control Singals for Bandpass Response

| $\Omega_{\text {bpc1 }}$ | $\Omega_{\mathrm{bpc} 2}$ | $\mathrm{S}_{1 \mathrm{p}}$ | $\mathbf{S}_{2 \mathrm{p}}$ | $\mathrm{C}_{1 \mathrm{n}}$ | $\mathrm{C}_{2 \mathrm{n}}$ | sel_y |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| $\hat{\Omega}_{c 1}$ to $\bar{\Omega}_{c 1}$ | $\hat{\Omega}_{c 1}$ to $\bar{\Omega}_{c 1}$ | $p$ | $p$ | $1 \forall n$ | $1 \forall n$ | X11100 |
|  | $\bar{\Omega}_{c 1}$ to 0.5 | $p$ | 1 | $1 \forall n$ | $1 \forall n$ | 001000 |
|  | 0.5 to $\bar{\Omega}_{c 2}$ | $p$ | 1 | $1 \forall n$ | $(-1)^{n}$ | 001000 |
|  | $\bar{\Omega}_{c 2}$ to $\hat{\Omega}_{c 2}$ | $p$ | $p$ | $1 \forall n$ | $1 \forall n$ | X10101 |
| $\bar{\Omega}_{c 1}$ to 0.5 | $\bar{\Omega}_{c 1}$ to 0.5 | 1 | 1 | $1 \forall n$ | $1 \forall n$ | X11111 |
|  | 0.5 to $\bar{\Omega}_{c 2}$ | 1 | 1 | $1 \forall n$ | $(-1)^{n}$ | X11111 |
|  | $\bar{\Omega}_{c 2}$ to $\hat{\Omega}_{c 2}$ | 1 | $p$ | $1 \forall n$ | $1 \forall n$ | 100010 |
| 0.5 to $\bar{\Omega}_{c 2}$ | 0.5 to $\bar{\Omega}_{c 2}$ | 1 | 1 | $(-1)^{n}$ | $(-1)^{n}$ | X11111 |
|  | $\bar{\Omega}_{c 2}$ to $\hat{\Omega}_{c 2}$ | 1 | $p$ | $(-1)^{n}$ | $1 \forall n$ | 100010 |
| $\bar{\Omega}_{c 2}$ to $\hat{\Omega}_{c 2}$ | $\bar{\Omega}_{c 2}$ to $\hat{\Omega}_{c 2}$ | $p$ | $p$ | $1 \forall n$ | $1 \forall n$ | X11101 |

### 3.4. Design Example and Complexity Comparison

Consider the design of the VDF with $T B W_{d} \leq 0.06 \pi, \delta_{s d}=-50 \mathrm{~dB}$ and $\delta_{p d}=0.1 \mathrm{~dB}$ that provides complete control over the bandwidth as well as the center frequency of bandpass responses over the desired range from $\Omega_{c 1}=0.05 \pi$ to $\Omega_{c 2}=0.95 \pi$. For these frequency specifications, the IMFT-VDF is designed with $N=42, \omega_{c}=0.25 \pi, A_{1} \in\{0,0.25,0.5,1\}$ and $p \in\{1,2\}$ where $\bar{\Omega}_{c 1}=0.1 \pi$ and $\bar{\Omega}_{c 2}=0.9 \pi$. The illustrative lowpass responses obtained using the IMFTVDF are shown in Fig. 5 via different colors. The responses shown by black color are obtained with $A_{1}=1$, i.e. conventional second order frequency transformation [17]. The responses shown in blue color are obtained using $A_{1} \neq 1$ i.e., the proposed second order modified frequency transformation. The frequency resposes in red and green colors are obtained using interpolation technique. Since the bandpass responses can be easily obtained by arithmatic subtraction of lowpass responses as discussed in Section 3.3, we can say that IMFT-VDF offers complete control over the bandwidth as well as the center frequency of bandpass responses.

A 16x16 bit multiplier, a 4:1 multiplexer, a 2:1 multiplexer, 32 bit adder and a word of memory were synthesized on a TSMC 65 nm process. The Synopsys Design Compiler was used to estimate the cell area. The area in terms of gate count is obtained by normalizing the cell area values by that of a two input NAND gate from the same library. The total gate count is the sum of gate counts of all the components. For the design examples with $T B W_{d}=0.06 \pi$


Figure 5: Variable lowpass responses using IMFT-VDF with $\Omega_{c 1}=0.05 \pi, \Omega_{c 2}=0.95 \pi$, $T B W_{d} \leq 0.06 \pi, \delta_{s d}=-50 \mathrm{~dB}$ and $\delta_{p d}=0.1 \mathrm{~dB}$.

Table 2: Gate Count Complexity Comparison

| VDFs | $T B W_{d}=0.06 \pi$ |  |  |  |  |  | $T B W_{d}=0.01 \pi$ |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
|  | No. of |  |  |  | GroupTotal Delay Gate Count |  | Group <br> Delay | Total Gate Count |
|  |  | Adders | Mux's | Memory <br> Words |  |  |  |  |
| FIRFilter | 46 | 90 | 0 | 23000 | 45 | 668850 | 250 | 3651600 |
| $\begin{aligned} & \text { MCDM- } \\ & \text { VDF }[24] \end{aligned}$ | 1003 | 8003 | $8002$ | 0 | 1000 | 3685545 | 3500 | 12885545 |
| APT- <br> CDM- <br> VDF [23] | 241 | 640 | 480 | 810 | NLP | 566650 | NLP | 3178850 |
| $\begin{aligned} & \hline \text { SPA- } \\ & \text { VDF [19] } \end{aligned}$ | 562 | 1093 | 830 | 0 | 45 | 1174175 | 250 | 6346325 |
| $\begin{aligned} & \text { SPA- } \\ & \text { MCDM- } \\ & \text { VDF }[20] \\ & \hline \end{aligned}$ | 294 |  | $564$ | 0 | 52 | 739665 | 256 | 3898715 |
| $\begin{aligned} & \text { IMFT- } \\ & \text { VDF } \end{aligned}$ | 155 | 903 | 406 | 0 | 104 | 465385 | 622 | 2835085 |

and $T B W_{d}=0.01 \pi$ with $\delta_{s d}=-50 \mathrm{~dB}$ and $\delta_{p d}=0.1 \mathrm{~dB}$, the gate count requirements of memory based FIR filter, MCDM-VDF [24], non-linear phase APT-CDM-VDF [23], SPA-VDF [19], SPA-MCDM-VDF [20] and IMFT-VDF are given in Table 2. The term "NLP"in Table 2 stands for non-linear phase.

For the first design example, the order of an FIR filter is 90 and coefficients corresponding to different bandpass responses are stored in memory beforehand. The MCDM-VDF consists of prototype filter of order 2000 and decimation factor, $D$, varies from 1 to 25 . Our APT-CDM-VDF consists of prototype filter

## $B_{v}(k)$ indicates large $N_{c}(\gg 16)$.

When the filtering stage of each secondary user is implemented using filter bank, the desired resolution (i.e., number of subbands) of the filter bank is equal to $\left(1 / B_{v}(k)\right)$. For example, if $N_{c}=16$ and $B_{v}(k)=B_{\text {cmin }}$ (or $B_{v}(k)=$ $2 \cdot B_{\text {cmin }}$, respectively), then 16 subband (or 8 subband, respectively) filter bank APT with the reduced second order APT. The SPA-VDF consists of 2 subsets of 6 sub-filters, each of order 90 . Our SPA-MCDM-VDF consists of 6 sub-filters each of order 90 and a masking filter of order 14.

The IMFT-VDF offers total gate count savings of $\{30 \%, 87 \%, 18 \%, 60 \%$, $37 \%\}$ and $\{22 \%, 78 \%, 11 \%, 55 \%, 27 \%\}$ over FIR filter, MCDM-VDF [24], APT-CDM-VDF [23], SPA-VDF [19] and SPA-MCDM-VDF [20] for the first and second design examples, respectively. The group delay of the IMFT-VDF is slightly higher than other VDFs. However, it must be taken into account that the total latency of the physical layer in each secondary user terminal is small compared to the latency of the CRNs [6, 7]. This means that little penalty in group delay is acceptable for more than $27 \%$ savings in total gate count. For delay sensitive applications, our SPA-MCDM-VDF [20] offers good alternative at the cost of higher gate count.

## 4. Proposed Decision Making Policy, $\rho^{\text {t_rand }}$

The second contribution of this paper is a tunable decision making policy, $\rho^{\text {t_rand }}$, designed to take full advantage of exclusive bandwidth and center frequency control offered by IMFT-VDF. The objective is to strike a balance between the number of collisions among secondary users and the total time spent on the non-optimum subbands, in order to achieve lower regret, $S_{t}$ as well as lower SSC for different distributions of $P_{v a c}(i) \forall i$ and distinct $B_{v}(k) \forall k$ requirements in heterogeneous decentralized CRNs. It is assumed that all secondary users employ the same policy but do not exchange any information with other users.

The advantages and disadvantages of different policies are discussed in Section 2.2. Accordingly, the proposed policy, $\rho^{t-r a n d}$, is designed via three refinements to existing policy, $\rho^{\text {rand }}[8]$, which are: 1) Low complexity tunable filtering stage (discussed in Section 4.1), 2) An efficient approach to identify the optimum subset of contiguous orthogonal subbands (discussed in Section 4.2), 3) Tunable subset size (discussed in Section 4.3).

### 4.1. Low Complexity Tunable Filtering Stage

In the literature, the input signal scenarios with $N_{c} \leq 16$ are considered. As per the assumption of stationary vacant subband statistics, the subband bandwidth must be narrow which means that the total bandwidth of the input signal, when $N_{c} \leq 16$, is not sufficiently wide. On the other hand, the practical scenarios usually involve wideband input signal and the requirement of tunable is required. To support tunable bandwidth requirement, an efficient approach
of order 160 and variable responses are obtained by integrating the first order design example considered here. Thus, timely selection of the second subset will definitely lead to the lower regret, $U_{t}$.


Figure 6: Input signal of bandwidth $B$ with $P_{v a c}$ distribution.
The IMFT-VDF, due to its finer granularity of center frequency of bandpass responses, offers large number of subsets of orthogonal contiguous subbands.
would be to implement $N_{c}$ subband filter bank and then, $\left(N_{c} / 2\right),\left(N_{c} / 4\right)$ or ( $N_{c} / 8$ ) subband filter bank responses can be obtained by combining adjacent subbands. However, higher the value of $N_{c}$, higher the resolution and hence the gate count, power consumption as well as group delay of the filter bank are. ance, $N_{r}(k)=N_{c} k$, i.e., all the suban that the performance of existing schemes $[8-11,13,14]$ under tunable vacant bandwidth requirements $\left(B_{v}(k)>B_{c m i n}\right)$ and wideband input signal scenarios ( $N_{c} \gg 16$ ) has not been analyzed yet.

Consider an input signal bandwidth $B$ as shown in Fig. 6. It is divided into $16\left(=N_{c}\right)$ uniform subbands, each with bandwidth $B_{c m i n}=B / 16$ and $P_{\text {vac }}(i), i \in\{1,2, . .16\}$. When $B_{v}(k)>B_{c m i n}$, then the corresponding subbands of bandwidth $B_{v}(k)$ are non-orthogonal. For example, if $B_{v}(k)=2 \cdot B_{c m i n} \forall k$, then there are 15 non-orthogonal subbands of bandwidth $B_{v}(k)$. Since each subband overlaps in frequency with at most $\left[2 \cdot\left(\left\lceil B_{v}(k) / B_{c m i n}\right\rceil-1\right)\right]$ number of other subbands of bandwidth $B_{v}(k)$, clustering of non-orthogonal subbands into different subsets is necessary to reduce the number of collisions among secondary users. Furthermore, quick and accurate selection of the optimum subset will lead to smaller total regret, $U_{t}$, and the SSC. For example, let us consider the two possible subsets of orthogonal subbands when $B_{v}(k)=2 \cdot B_{c m i n}$ : $\{(1-2),(3-4), \ldots,(15-16)\}$ or $\{(2-3),(4-5), \ldots,(14-15)\}$. Using the relation $P(A B)=$ $P(A)+P(B)-P(A) \cdot P(B)$, corresponding combined probability of vacancies are obtained as $\{0.39,0.11,0.19,0.52,0.62,0.36,0.87,0.28\}$ and $\{0.14,0.03$, $0.21,0.8,0.06,0.59,0.84\}$, respectively. It can be easily observed that the second subset has higher number of optimum subbands than the first for this Higher the number of subsets, higher is the time spent in the non-optimum subbands and hence, higher is the regret, $U_{t}$. Thus, the number of subbands choices are limited to two in the proposed policy where the first subset starts
with the subband numbered as 1 while the second subset starts with the subband numbered as $\left(\left\lceil B_{v}(k) / 2 \cdot B_{c m i n}\right\rceil+1\right)$, except the case $B_{v}(k)=B_{c m i n}$ where only one subset is available. The idea behind such subset division is to keep less than $50 \%$ overlapping between the corresponding subbands of two subsets so that the subsets are sufficiently apart in terms of their combined vacancy statistics which in turn leads to fewer errors in the selection of optimal subset by secondary users.

In the proposed $\rho^{t_{-} \text {rand }}$ policy, all $M$ secondary users independently learn the vacancy statistics of these two subsets during initial phase spanning up to $t_{s}(k)$ time slots. This means that the secondary user spends half the number of time slots, i.e., $\left(t_{s}(k) / 2\right)$ slots, in the first subset and remaining half in other subset. Then, at $t=t_{s}(k)+1$, secondary user independently choses the optimum subset based on the past sensing events. The number of time slots, $t_{s}(k)$, needs to be chosen carefully since very small or very large value of $t_{s}(k)$ would lead to the higher $U_{t}$ as well as higher SSC as discussed before. Based on empirical observations, we have,

$$
\begin{equation*}
t_{s}(k)=2 \lambda \cdot L_{t}(k), \forall k \tag{13}
\end{equation*}
$$

where $L_{t}(k)$ is the tunable subset size for $k^{\text {th }}$ user (discussed in Section 4.3) and $\lambda$ is a constant between 5 to 10 .

### 4.3. Tunable Subset Size, $L_{t}$

In the existing policies for multi-user CRNs [8-11], the size of the subset, comprising of optimum subbands, is fixed, i.e., $L_{t}(k)=M \forall k$. This means that the distinct and tunable vacant bandwidth requirements of each secondary user as well as the different types of distributions of $P_{v a c}$ are not taken into consideration while choosing $L_{t}(k)$. Also, the number of collisions and hence, SSC are bound to be large in the beginning as well as for higher $M$ since it is unlikely that secondary users will always agree on the order of the optimum subbands $[8,9]$. The solution to these issues can be provided by making the subset size, $L_{t}(k) \forall k$, tunable as discussed below.

The minimum and maximum subset sizes, $L_{t \min }(k)$ and $L_{t \max }(k)$, respectively, for $k^{t h}$ secondary user are given by,

$$
\begin{align*}
L_{t \min }(k) & =\left(\frac{\sum_{u=1}^{M}\left\lceil N_{c} \cdot B_{v}(u)\right\rceil}{\left\lceil N_{c} \cdot B_{v}(k)\right\rceil}\right)  \tag{14}\\
L_{t \max }(k) & =\left(\frac{N_{c} \cdot B_{c \min }}{B_{v}(k)}-e\right) \tag{15}
\end{align*}
$$

where $e=1$ if the first subset is chosen, i.e. $\operatorname{sub}(k)=1$, otherwise $e=0$. It can be observed that $L_{\text {tmin }}(k)=L_{\text {tmax }}(k)=M$ when $B_{v}(k)=B_{\text {cmin }} \forall k$, same as that in [8-11].

If $L_{t}(k) \approx L_{t \min }(k)$, then the number of collisions among secondary users, $U_{t}$ and SSC are all high. On the other hand, very large value of $L_{t}(k) \approx L_{\text {tmax }}(k)$,

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offers lesser number of collisions and lower SSC , but leads to higher regret $U_{t}$ due to the selection of the sub-optimum subbands i.e. the subbands with lower $P_{v a c}$. By taking into account the fact that the $U_{t}$ and SSC due to the collisions are much worse than the same due to the selection of the sub-optimum subbands in the DSLA scenario compared to conventional machine learning scenarios, a new method to make the value of $L_{t}(k) \forall k$ tunable is proposed as follows:

1. For each integer value from $L_{t \min }(k)$ to $L_{t \max }(k)$, the total predicted reward, $\tilde{S}_{t}\left(k, q_{k}\right) \forall k$ and $q_{k} \in\left\{1,2, . .,\left(L_{\text {tmax }}(k)-L_{\text {tmin }}(k)+1\right)\right\}$, at any given time $t$, are calculated as,

$$
\begin{equation*}
\tilde{S}_{t}\left(k, q_{k}\right)=\sum_{u=1}^{L_{t m i n}(k)+q_{k}-1} \frac{\bar{P}_{v a c}\left(k, j_{k, u}\right) \cdot L_{t \min }(k)}{L_{t m i n}(k)+q_{k}-1} \tag{16}
\end{equation*}
$$

where

$$
\begin{equation*}
j_{k, u}=\left(\operatorname{sub}(k)+(u-1) \cdot\left\lceil\frac{B_{v}(k)}{B_{\text {cmin }}}\right\rceil\right) \tag{17}
\end{equation*}
$$

Here $\bar{P}_{v a c}\left(k, j_{k, u}\right) \forall u$ denote the learned subband availability statistics arranged in the descending order, i.e., $\bar{P}_{v a c}(k, 1) \geq \bar{P}_{v a c}(k, 2) \geq \cdots \geq$ $\bar{P}_{\text {vac }}\left(k, L_{\text {tmax }}(k)\right)$.
2. For each integer value from $L_{\text {tmin }}(k)$ to $L_{\text {tmax }}(k)$, the total predicted number of collisions, $\tilde{C}_{t}\left(k, q_{k}\right) \forall k$ and $\forall q_{k}$, at any given time $t$, are calculated as

$$
\begin{equation*}
\tilde{C}_{t}\left(k, q_{k}\right)=1-\left[\frac{\left(L_{t \min }(k)+q_{k}-1\right)!}{\left(q_{k}-1\right)!\left(L_{t m i n}(k)+q_{k}-1\right)^{L_{t \text { min }}(k)}}\right] \tag{18}
\end{equation*}
$$

3. Finally, the optimum value of $L_{t}(k)$ is given by

$$
\begin{equation*}
L_{t}(k)=L_{\text {tmin }}(k)+s-1 \tag{19}
\end{equation*}
$$

where

$$
\begin{equation*}
s=\underset{q_{k}}{\arg \max }\left[\tilde{S}_{t}\left(k, q_{k}\right) \cdot t-\tilde{C}_{t}\left(k, q_{k}\right) \cdot t\right] \tag{20}
\end{equation*}
$$

Each secondary user independently calculates the value of $L_{t}(k)$. To minimize the computational complexity, the value of $L_{t}(k)$ is updated only when corresponding secondary user experiences collision with other secondary users. The proposed learning policy, $\rho^{t-r a n d}$, is described in detail in Algorithm 2.

## 5. Simulation Results and Complexity Analysis

In this section, we presents the simulation results and compare the performances of the proposed decentralized DSLA scheme (IMFT-VDF and $\rho^{t-\text { rand }}$ ) with the existing decentralized schemes (DFTFB and $\rho^{\text {rand }}[8]$ ) in terms of total vacant spectrum utilization (average reward), SSC and gate count complexity. The wideband input signal consists of $128\left(=N_{c}\right)$ subbands i.e., $B_{c m i n}=$

```
Algorithm 1: Proposed tunable DSLA policy, \rho}\mp@subsup{\rho}{}{t-rand
1) Parameters
    t: Current time slot, }t\in{1,2,\ldots,T
    B _ { \text { cmin } } \text { : Smallest primary channel bandwidth}
    N
    P
    M: Number of secondary users
    B
        secondary user, }k\in{1,2,\ldots,M
```



```
        Ltmin}(k)\leq\mp@subsup{L}{t}{}(k)\leq\mp@subsup{L}{t\mathrm{ tmax }}{}(k
```

2) Initialization: Sense each of the $N_{r}(k)$ subband once.
$L_{t}(k) \leftarrow L_{\text {tmin }}(k), t \leftarrow N_{r}(k)+1 \quad \forall k$
Calculate $t_{s}(k) \forall k$, using Eq. (11).
3) Transit phase: Optimum subset selection.
I. Identify the two subsets of orthogonal subbands.
II. Select the optimum subset, independently at all secondary users in time, $t_{s}(k)$ using the procedure in Section IV-A. $t \leftarrow t+t_{s}(k) \quad \forall k$
4) Start Loop: $t \leftarrow t+1$
I. Run $\rho^{\text {rand }}$ policy (randomized access over $L_{t}(k)$ subbands of chosen subset) independently at each user.
II. When collision occurs, calculate $\bar{S}_{t}\left(k, q_{k}\right), \bar{C}_{t}\left(k, q_{k}\right)$ and update $L_{t}(k) \forall k$ using the steps given in Section IV-B.
End Loop when $t=T$
(1/128). Three different resolutions of DFTFB, $\{128,64,32\}$ are considered. Each numerical result reported hereafter is the average of the values obtained over 50 independent experiments and the simulations consider a time horizon of 50000 iterations.

Consider $B_{v}(k)>B_{c m i n} \forall k$ and $B_{v}(1)=\ldots=B_{v}(M)$. This scenario is similar to the one considered in the existing literature in the sense that the vacant bandwidth requirement of all users is identical. In Fig. 7a and Fig. 7b, the plots of average reward, $S_{t}$, in percentage vs. time $t$, are shown for two different types of input distributions, case $1(\mathrm{C} 1)$ and case $2(\mathrm{C} 2)$, respectively. Here, $M=$ $25, B_{v}(k)=0.012 \forall k$ and $S_{t} \%=\left[100 \cdot\left(S_{t}^{*}-S_{t}\right) / S_{t}^{*}\right]$ where $S_{t}^{*}$ is the average reward of the genie-aided policy (i.e., centralized DSLA). It can be observed that the total reward (vacant spectrum utilization) of the proposed scheme is higher than other schemes. In Fig. 7e, the plot of average reward at $t=20000$ is plotted for different values of $M$. It can be observed that the proposed scheme outperforms other schemes for all $M$. Furthermore, the proposed scheme offers $15 \%$ and $26 \%$ lower average SSC than 128-DFTFB based scheme for case 1 and research direction includes the efficient implementation of the proposed scheme on the USRP-GNU-FPGA platform and analyzing the performance using real radio signals.

## Acknowledgment

This work has received a French state support granted to the CominLabs excellence laboratory and managed by the National Research Agency in the "Investing for the Future" program under reference No. ANR-10-LABX-07-01. The authors would also like to thank Region Bretagne, France, for its support of this work.


Figure 7: (a) Case 1: Reward $S_{t}$ in $\%$ vs. time $t$ for $M=25$ and $B_{v}(k)$ $=0.012 \forall k$, (b) Case 2: Reward $S_{t}$ in $\%$ vs. time $t$ for $M=25$ and $B_{v}(k)$ $=0.012 \forall k$, (c) Case 1: Reward $S_{t}$ in $\%$ vs. time $t$ for $M=20$ with $B_{v}(k)$ $=\{(1 / 256),(1 / 256),(1 / 256),(1 / 256),(2 / 256),(2 / 256),(3 / 256),(3 / 256),(3 / 256),(3 / 256),(3 / 256)$, (3/256), (3/256),(3/256),(5/256),(5/256), (5/256),(6/256)\}, (d) Case 2: Reward $S_{t}$ in \% vs. time $t$ for $M=20$ and same $B_{v}(k)$ as in (d), (e) Reward $S_{t}$ in \% vs. number of secondary users, $M$ for $t=20000$ and $B_{v}(k)=0.012 \forall k$ where C1 and C2 denote case 1 and case 2 , respectively, (f) Reward $S_{t}$ in \% vs. number of secondary users, $M$ for $t=20000$ and different $B_{v}(k)$.

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