Cognitive GSM OpenBTS

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Declaration

I declare that this written submission represents my ideas in my own words, and where ideas or words of others have been included, I have adequately cited and referenced the original sources. I also declare that I have adhered to all principles of academic honesty and integrity and have not misrepresented or fabricated or falsified any idea/data/fact/source in my submission. I understand that any violation of the above will be a cause for disciplinary action by the Institute and can also evoke penal action from the sources that have thus not been properly cited, or from whom proper permission has not been taken when needed.

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Approval Sheet

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Dedication

Dedicated to my beloved family and adviser.

Abstract

GSM technology is one of the wireless communication networks which is the easiest and most viable solution for robust voice communication. No wonder it is the most successful communication standard in the world covering over 90% of globe population. Inspite of this, it has been repeatedly observed that utilization of GSM spectrum is poor. The resources available to support the users in wireless communication networks are very limited. In particular spectrum that are used as a path between users are scarce and underutilized. For example in New York and Chicago, GSM cell phone frequencies are occupied 45% and 55% of the time.

We focus mainly on how to overcome the congestion and underutiliztion of the spectrum using cognitive radio technology. We concentrate on building of a CR testbed using existing GSM protocol stack and that aims to provide QoS for both PU and SU.

We integrate cognitive radio capabilities, on the fly, with GSM base station (BTS) using OpenBTS architecture. The implementation has following novel features:

- Cognitive sensing of the channel, in use by secondary user, while transmitting,
- On the fly shifting to free GSM subchannel without call drop,
- Works with existing GSM phones without any modification in hardware and software.

Trials at IIT Hyderabad over the last one year has validated the viability of the system. We have implemented FFT based energy detection for spectrum sensing in testbed. We also worked with improving cyclostationary based spectrum sensing algorithm for implementing in the testbed.

Cyclostationary detection is considered as one of the major methods for spectrum sensing in cognitive radio. But due to cyclic frequency mismatch the performance degrades as the sensing time increases. In this paper we propose block based cyclostationary detection which overcomes the sensitivity of tra- ditional cyclostationary detection to cyclic frequency mismatch. Theoretical analysis of block based cyclostationary detection is presented which shows that there is an optimal block length for which the performance is best. Simulation results corroborating the theoretical analysis are then presented.

Spectrum has also been underutilized in the TV UHF and VHF bands. TV 'White Space' spectrum is characterized by large amounts of geographically available spectrum with excellent propagation characteristics, offering long range and exceptional building penetration compared to the spectrum used by Wi-Fi and Bluetooth equipment. As a result TV white space spectrum is ideal for providing fixed broadband internet services to locations where the routing of cables or optical fibre is neither practical nor economical; and hence important from a developing world perspective. We have presented the study on availability of Digital TV White Space in India.

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Chapter 1

Outline

In the present work , we discuss the cognitive radio testbed which could be used for testing the real time working of cognitive radio algorithms. A brief outline of the thesis is presented in Chapter 1.

In Chapter 2, we present the ideation of the Cognitive GSM OpenBTS (CGBTS). Testbed setup, features and results out of CGBTS are discussed.

In Chapter 3, we present the method to improve cyclostationary detection for spectrum sensing in cognitive radio. We discuss the impact of frequency offset in the existing cyclostationary detection. We develop a block based cyclostationary detection which is robust for frequency offset.

In Chapter 4, we study the availability of Digital TV white space.

In Chapter 5, we give the conclusion for the thesis.

Chapter 2

Cognitive GSM OpenBTS

2.1 Introduction

Over past few decades increase in the wireless technologies and applications like GSM, Bluetooth, Wi-Fi, broadcast TV, etc., has found increase in the usage of available spectrum which has created spectrum scarcity. Future wireless communication focuses on delivering maximum throughput, better coverage, high data rate, high spectral efficiency by dynamic spectrum access. These deliverables could be achieved by effectively using cognitive radio technology which enables sensing of free licensed spectrum belonging to primary user (PU) and utilization of the spectrum efficiently by assigning or allowing 'free spectrum' usage by secondary user (SU).

In order to implement cognitive radio (CR) communication technologies, it is necessary to evaluate the practicability of algorithms in real time. It also makes sense to use existing architectures to reduce the implementation time. Thanks to the availability of Software Defined Radio (SDR), it is now easy to modify radio capabilities via software and can be made as potential testbed. Open source hardware platforms like Ettus USRP, Range Networks RAD-1, Rice WARP, Fairwaves UmTRX can be used as off-the-shelf SDR.

For robust voice communication, GSM technology is still the easiest and most viable solution. No wonder it is the most successful communication standard in the world covering over 90% of globe population. Inspite of this, it has been repeatedly observed that utilization of GSM spectrum is poor. For example in New York and Chicago, GSM cell phone frequencies are occupied 45% and 55% of the time, respectively [3]. In this chapter we concentrate on building of a CR testbed using existing GSM protocol stack and that aims to provide QoS for both PU and SU.

2.2 Related Work

After the emergence of SDR, open source based application based on SDR has seen tremendous growth. Especially, open-source software for telephony like OpenBTS [4], OpenBSC [5], Osmocom BB [6], Openmoko [7] has proved that real time voice or data communication can be done at low cost, providing a viable solution for less dense populated areas like rural and underdeveloped countries.

Figure 2.1: Cognitive GSM BTS

In particular, OpenBTS has revolutionized the GSM telephony making GSM BTS setup possible using SDRs like Range Networks RAD-1, USRP, UmTRX, etc..,

Efforts to use OpenBTS as cognitive radio testbed has been reported in [1] and [2]. In [1], cognitive capability is incorporated indirectly by monitoring frame error rate (FER). If the FER is above a certain threshold, it is assumed that the PU is active and the call is abruptly dropped. This scheme suffers from high false alarm rate (because of obvious reasons) and abrupt call termination. In case of [2] frequency selection is done based on GSM whitespace database from PU which doesn't involve spectrum sensing and on the fly detection and shifting of PU frequency. In this chapter, our contribution is focused on making OpenBTS a cognitive BTS (CGBTS) making it smart to detect the PU on the fly and assign the available free frequency to SU without degrading QoS of both PU and SU.

2.3 Cognitive GSM BTS

CGBTS allows secondary user to make GSM based calls using OpenBTS without call drop and degradation of QoS for primary as well as secondary user. As mentioned earlier, we need to detect the presence of PU in a reliable manner while transmitting on the same frequency. We achieve this by placing the sensing radio just outside the cell coverage of our SU pico cell, so that it can monitor PU activity, as shown in 2.1. In order to provide subchannel (frequency) change, on the fly, and a handover is initiated as soon as the PU is detected. The proposed implementation does not require hardware, software changes to the existing GSM phones or GSM standards.

2.3.1 Testbed Setup

A Block diagram of the testbed is showed in 2.2. The setup is explained as follows.

- 2 RAD-1 devices are connected to a PC with two instances of modified OpenBTS software which acts as SU and is integrated with spectrum sensing application.
- 1 USRP is connected to a PC with spectrum sensing application , and the radio is placed outside the cell coverage of SU.
- 1 PC with VoIP soft client is connected with SU using wireless backhaul.

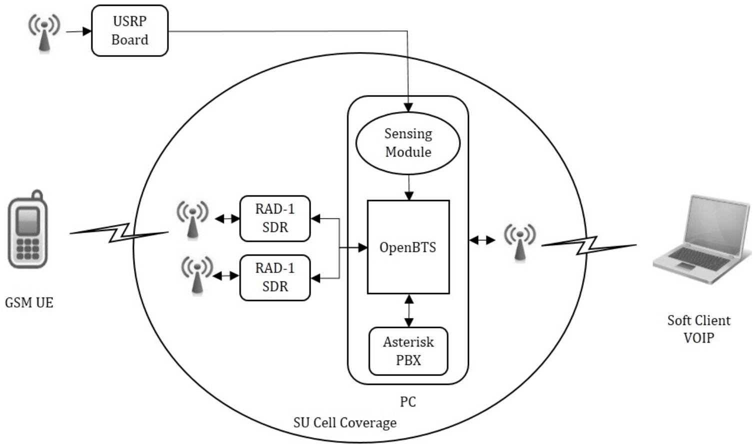

Figure 2.2: Cognitive GSM BTS Block Diagram

2.3.2 Spectrum Sensing

Spectrum sensing is done using USRP device. USRP device has capability to stream baseband IQ signals at 25MS/s in 16 bit mode or 50MS/s in 8 bit mode to host application for processing. We use 25 MS/s 16 bit mode in USRP device to stream IQ samples for processing. This allows to sense spectrum of bandwidth 12.5 MHz, simultaneously. We use FFT based energy detection for sensing GSM subchannels.

FFT based sensing algorithm involves the following steps. Sensing is done over the downlink of GSM-900 enabling us to see the presence of primary GSM base station. USRP N210 streams 25MS/s rate raw data to the spectrum sensing module in the host. Sensing Module then uses complex FFT over the received data, resulting in double sided spectrum in complex form. Double sided spectrum is then converted into single sided spectrum and power spectral density (PSD) of 62 subchannels, spaced 200 KHz apart is found and then compared with the threshold. The free channels are stored for further use.

2.3.3 The Cognitive Functionality

CGBTS will start the spectrum sensing procedure. Once it has list of free channels it assigns best out of the available channel to OpenBTS. GSM user equipment (UE) is camped into one (say first) instance of the OpenBTS. Once UE is camped into BTS, call connection is established between UE and VoiP client using OpenBTS. Spectrum sensing is done simultaneously by keeping it outside the range of secondary BTS, making it possible to detect PU even when SU call is in progress.

If a PU has been detected by the sensing module, forced handover procedure is triggered by OpenBTS, enabling UE to handover to another (second) instance of OpenBTS working on a frequency that is not occupied. This procedure is repeated as and when the PU is detected on the subchannel being used by the SU. QoS of the ongoing call is calculated using frame error rate (FER). FER experienced by the SU when a high power PU with received signal strength 39 dBm is transmitting is 38.05% and the same reduces to 3.8% when the received signal strength is 35 dBm . Thus

Table 2.1: Cognitive BTS parameters during collision with PU with received power of 35 dBm.

Time taken for spectrum sensing	10 ms
Time taken for completion of handover	500 ms
FER during detection and handover	3.8%

for 510 ms PU will be interfering with SU and vice versa. SU is 'handovered' to free frequency within 510 ms thus maintaing QoS of both PU and SU.

2.4 Conclusion

The testbed has presented the opportunity for realizing the next generation wireless system without changing the existing standards of GSM by extending the capability of OpenBTS as CGBTS by integrating FFT based spectrum sensing algorithm module which allows OpenBTS to use free spectrum and handover UE without degrading PU and SU QoS.

Chapter 3

Robust Block Based Cyclostationary Detection

3.1 Introduction

Cognitive radio is widely believed to be the next boom in wireless communication due to the promise it holds in exploiting the under-utilized spectrum [9]. Accordingly, there has been tremendous academic and industrial interest on cognitive radios and its applications.

The idea behind cognitive radio is spectrum sharing, that allows the secondary users to communicate over the spectrum allocated to the primary users while primary users are not using it. To make it possible, the secondary users must sense spectrum of primary users frequently for detecting the presence.It is must that whenever the primary users become active while secondary users are using allocated spectrum, they must be capable of detecting primary users with high probability of detection and vacate or reduce transmit power within certain amount of time [15, 14].

Spectrum detection is great challenge in cognitive radio due to

1. the low SNR required for detection,
2. multipath fading and time dispersion [10] and
3. variability of the noise/interference level with time and location[15].

As a result spectrum sensing has reborn as a very active area of research despite its long history.

Cyclostationary detection(CSD) [11, 12, 13] is one of the methods that has attracted renewed interest due to robustness to uncertainty in noise power and propagation channel. CSD takes advantage of cyclostationary property of a signal for distinguishing it from noise by extracting cyclostationary property of signal either by finding its cyclic autocorrelation(CAC) or spectral correlation density(SCD). However it suffers due to high computational complexity required and cyclic frequency mismatch (due to the sampling time error and frequency offset) that affect its performance [15, 16, 17, 18].

In general without cyclic frequency mismatch CSD shows increased performance with increased sensing time, but in the presence of cyclic frequency mismatch, as sensing time increases, performance degrades [16, 17, 18, 19, 20]. In this paper we propose block based cyclostationary detection to overcome this performance degradation.

The rest of the paper is organized as follows. Review of conventional CSD is presented in section 3.2. Method of block based CSD in presented in section 3.3. Robustness of block based CSD by theoretical analysis is presented in section 3.4, Simulation results are presented in section 3.5 and Conclusions are given in section 3.6.

3.2 Cyclostationary Detection

In this section we review the results presented in [17, 18]. In general data, which is a wide-sense stationary random process, is modulated by coupling it with signals like pulse trains and hoping sequence which end up with building in periodicity resulting in cyclostationary as their statistics (mean and autocorreltion) exhibit periodicity. Cyclostationary features can be extracted by implementing cyclic autocorrelation (CAC) or spectral correlation density (SCD) of received signal. For a given α, called the cyclic frequency, and time lag, τ, the CAC of a signal $x(t)$ is defined as [11, 12, 13]

$$R_x^\alpha(\tau) = \lim_{\Delta\to\infty} \frac{1}{\Delta} \int_{-\frac{\Delta}{2}}^{\frac{\Delta}{2}} x(t+\frac{\tau}{2})x^*(t+\frac{\tau}{2})e^{-j2\pi\alpha t}dt. \tag{3.1}$$

If there exist at least one non-zero α such that $\max_\tau |R_{xx}^\alpha(\tau)| > 0$, we say that $x(t)$ exhibits cyclostationary. The value of such α depends on type of modulation, symbol duration,etc.. The SCD, the fourier transform of the CAC, is similarly defined as

$$S_x^\alpha(f) = \int_{-\infty}^{\infty} R_x^\alpha(\tau)e^{-j2\pi f\tau}d\tau. \tag{3.2}$$

3.2.1 System Model

Corresponding to the two hypotheses: $\mathcal{H}_0$; signal absent, and $\mathcal{H}_1$; signal present, the received signal can be written as

$$\begin{aligned} &\mathcal{H}_0 : y(t) = \eta(t) \\ &\mathcal{H}_1 : y(t) = hx(t) + \eta(t), \end{aligned} \tag{3.3}$$

where $x(t)$ is transmitted signal from the primary user, h is the channel response and $\eta(t)$ is the additive white Gaussian noise. Assuming that $x(t)$ has cyclostationary features in it, there exists at least one non-zero cyclic frequency α such that $R_x^\alpha(\tau) \neq 0$ for some τ, while for the noise, $R_\eta^\alpha(\tau) = 0, \forall\tau$, as it is a pure stationary process. Assuming signal and noise are mutually independent, we have for cyclic frequency, $\alpha_0 \neq 0$

$$\begin{aligned} &\mathcal{H}_0 : R_y^{\alpha_0}(\tau) = 0 \\ &\mathcal{H}_1 : R_y^{\alpha_0}(\tau) = R_x^{\alpha_0}(\tau) \neq 0, \text{ for some } \tau. \end{aligned} \tag{3.4}$$

Therefore $\mathcal{H}_0$ and $\mathcal{H}_1$ can be distinguished by generating a test statistic from the CAC of the received signal at cyclic frequency α_0 and comparing the test statistic with a threshold. In a digital system, received signal is sampled with sampling period, T_s, for a limited period of time that results in limited number of samples, N. Accordingly the discrete version of CAC is given by

$$R_y^{\alpha}(kT_s) = \frac{1}{N}\sum_{n=0}^{N-1} y((n+k)T_s)y^*(nT_s)e^{-j2\pi\alpha nT_s} \tag{3.5}$$

where the lag $k = 0, 1, ..., M-1$ with $M << N$ and the corresponding test static is given by [21, 22]

$$\mathcal{C}_1 = \sum_{k=0}^{M-1} |R_y^{\alpha_0}(kT_s)|^2. \tag{3.6}$$

The threshold is chosen as $\beta\hat{\sigma}^4$, where β is a scalar to meet the pre-defined probability of false alarm and $\hat{\sigma}^2$ is approximately the noise power at $\mathcal{H}_0$, but includes both signal power and noise power at $\mathcal{H}_1$ [17, 18].

3.2.2 Effect of Cyclic Frequency Mismatch

To bring out the effect of cyclic frequency mismatch, we consider two types of primary signals [17],[18]: a single carrier (SC) signal $x(t) = \cos(2\pi f_0 t)$ and the digitally modulated (DM) signal $x(t) = \sum_{n=-\infty}^{+\infty} a(n)p(t - nT_b)$, where T_b is the symbol duration, $a(n)$ is quantized and modulated signal at time nT_b, and $p(t)$ is the pulse shaping filter supported in the bit interval $[0, T_b)$. Note that the analysis of the SC signal neatly brings out the effect of the cyclic frequency mismatch; and the effect of cyclic frequency mismatch on DM signal is essentially similar [17, 18].

For SC Signal as shown in [18] the expectation of the test statistic $\mathcal{C}_1$ for hypothesis $\mathcal{H}_0$, is given by

$$\begin{aligned}\mathbf{E}(C_1) &= \sum_{k=0}^{M-1} E(|R_\eta^{\alpha}(kT_s)|^2)\\ &= \begin{cases} \frac{M+1}{N}\sigma_\eta^4 + \sigma_\eta^4 & , \alpha = 0 \\ \frac{M+1}{N}\sigma_\eta^4 + \frac{\sigma_\eta^4}{N^2}\left(\frac{\sin(\pi\alpha NT_s)}{\sin(\pi\alpha T_s)}\right)^2 & , \alpha \neq 0 \end{cases}\end{aligned}$$

and for hypothesis $\mathcal{H}_1$, is given by

$$\begin{aligned}\mathbf{E}(C_1) &= \sum_{k=0}^{M-1} [|h|^4 E(|R_x^{\alpha}(kT_s)|^2) + E(|R_\eta^{\alpha}(kT_s)|^2)]\\ &\approx \frac{M|h|^4}{16N^2}\left(\frac{\sin(\pi\epsilon NT_s)}{\sin(\pi\epsilon T_s)}\right)^2\\ &+\frac{M+1}{N}\sigma_\eta^4 + \frac{\sigma_\eta^4}{N^2}\left(\frac{\sin(\pi\alpha NT_s)}{\sin(\pi\alpha T_s)}\right)^2.\end{aligned} \tag{3.7}$$

The ratio of the test statistics at the two hypothesis can be approximated as

$$\gamma_1 \approx 1 + \frac{M|h|^4}{16(M+1)\sigma_\eta^4 N}\left(\frac{\sin(\pi\epsilon N T_s)}{\sin(\pi\epsilon T_s)}\right)^2 \tag{3.8}$$

If $\epsilon = 0$, there is no cyclic frequency mismatch, accordingly

$$\gamma_1 \approx 1 + \frac{M|h|^4 N}{16(M+1)\sigma_\eta^4}, \qquad \epsilon = 0. \tag{3.9}$$

Similarly for the DM signal the ratio of the test statistics at the two hypothesis can be approximated as shown in [17]

$$\begin{aligned}\gamma_1 \approx 1 &+ \frac{I(d+c)\delta_2^2 + L_s d(\delta_4 - 2\delta_2^2)}{(I-2)\varpi_2^2 + \varpi_4} \\ &+ \frac{L_s^2 d\delta_2^2}{N((I-2))\varpi_2^2 + \varpi_4}\left(\frac{\sin(\pi\epsilon N T_b/L_s)}{\sin(\pi\epsilon T_b)}\right)^2, \epsilon \neq 0.\end{aligned} \tag{3.10}$$

where $E(|\eta(n)|^2)=\varpi_2$, $E(|\eta(n)|^4)=\varpi_4$, $E(|a(n)|^2)=\delta_2$, $E(|a(n)|^4)=\delta_4$ and $L_s = T_b/T_s$ with T_s as the sampling period of the DM signal and d, c are some constants related to the pulse shaping filter $p(t)$.

If $\epsilon = 0$, there is no cyclic frequency mismatch and (3.10) reduces to

$$\begin{aligned}\gamma_1 \approx 1 &+ \frac{I(d+c)\delta_2^2 + L_s d(\delta_4 - 2\delta_2^2)}{(I-2)\varpi_2^2 + \varpi_4} \\ &+ \frac{N d\delta_2^2}{(I-2)\varpi_2^2 + \varpi_4}, \epsilon = 0.\end{aligned} \tag{3.11}$$

Clearly from (3.8), (3.10) γ_1 is a decreasing function of N and

$$\lim_{N\to\infty} \gamma_1 = 1,$$

which implies that the performance degrades as the number of samples increase (sensing time increases) [17],[18]. This is a serious limitation of Cyclostationary Detection.

3.3 Block based Cyclostationary Detection

In this section we propose block based CSD which overcome performance degradation of traditional CSD caused due to sensitivity for cyclic frequency mismatch. Analysis of (3.8),(3.10) clearly shows that the performance of CSD will be very good (corresponding to when γ_1 is maximum) for certain extent of values of N. Based on this analysis, we divide the number of samples into b blocks each of length $\hat{N}$ ($b = \left\lfloor \frac{N}{\hat{N}} \right\rfloor$) such that for which γ_1 is maximum. Performing CSD for each block and then comparing it with the test statistic gives b logical decisions. By combining these logical decisions from the b blocks (say by using the majority rule) we can improve performance of CSD. Figure 3.1 gives the block diagram of the proposed block CSD and is described as follows:

1. Block the N input samples into b blocks with $\hat{N}$ samples in each blocks. (The block size $\hat{N}$ will be optimized subsequently.)

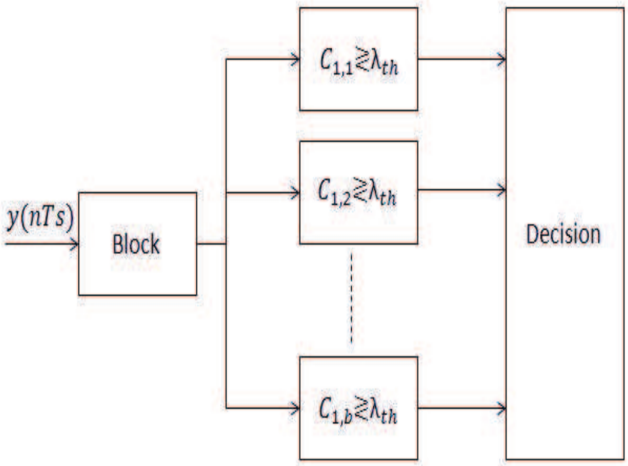

Figure 3.1: Block Diagram for Block CSD

2. Compute CAC for each block; which is easily adapted from (3.5) as

$$R^{\alpha}_{j,y}(kT_s) = \frac{1}{\hat{N}} \sum_{n=0}^{\hat{N}-1} y((n+k)T_s)y^*(nT_s)e^{-j2\pi\alpha nT_s} \tag{3.12}$$

   where lag $k = 0, 1, 2, ..., M-1$ with $M << \hat{N}$ and $j = 1, 2, ..., b$

3. Obtain test statistics for each block, given by

$$C_{1,j} = \sum_{k=0}^{M-1} |R^{\alpha}_{j,y}(kT_s)|^2 \tag{3.13}$$

4. Compare test statistic with threshold and decide $\mathcal{H}_0$ or $\mathcal{H}_1$ for each block.

5. Combine the decisions from the b blocks (for eg. use majority logic to make final decision).

In what follows we assume majority logic decision fusion as it is sufficient to bring out the effectiveness of the blocking strategy. However the results can be easily extended and analyzed for other fusion rules [23, 24, 25].

Remark 1. *Note that we have assumed that number of samples is an integral multiple of b. Note that this is not a limiting restriction and the algorithm can be generalized by defining b as number of blocks with $\hat{N}$ samples. The incomplete block is not used in the decision making process.*

3.4 Robustness of Block Based Cyclostationary Detection

In this section we show that the performance of Block Based Cyclostationary Detection presented in previous section is robust to cyclic frequency mismatch. But first, we derive the optimal block length $\hat{N}_O$ that optimizes the performance of block CSD. Note that this result on optimal block length is independent of the fusion rule.

Theorem 1. For a Single Carrier signal given by, $x(t) = \cos(2\pi f_o t)$, the optimal block length N_O belongs to the set

$$\left\{ \left\lfloor \frac{1.165}{\pi \epsilon T_s} \right\rfloor, \left\lceil \frac{1.165}{\pi \epsilon T_s} \right\rceil \right\} \tag{3.14}$$

where $\lfloor a \rfloor$, $\lceil a \rceil$ denote the floor and ceil of a.

Proof. As in subsection 3.2.2 we consider SC signal $x(t) = \cos(2\pi f_o t)$ as the primary signal. For this signal, at cyclic frequency $\pm 2f_0$ CAC is non-zero that is $\alpha = 2f_0$. Let this SC be sampled with a sampling period of T_s which results in N number of samples. We can now express the expected value of the test static for j^{th} block when received signal is only noise (hypothesis $\mathcal{H}_0$) as

$$\mathbf{E}(C_{1,j}) = \sum_{k=0}^{M-1} E(|R^{\alpha}_{j,\eta}(kT_s)|^2)$$

$$= \begin{cases} \frac{M+1}{\hat{N}}\sigma_\eta^4 + \sigma_\eta^4 & ,\alpha = 0 \\ \frac{M+1}{\hat{N}}\sigma_\eta^4 + \frac{\sigma_\eta^4}{\hat{N}^2}\left(\frac{\sin(\pi\alpha\hat{N}T_s)}{\sin(\pi\alpha T_s)}\right)^2 & ,\alpha \neq 0 \end{cases}$$

Similarly, the expected value of the test static for the j^{th} block when received signal is signal plus noise (hypothesis $\mathcal{H}_1$), is given by

$$\mathbf{E}(C_{1,j}) = \sum_{k=0}^{M-1} [|h|^4 E(|R^{\alpha}_{j,x}(kT_s)|^2) + E(|R^{\alpha}_{j,\eta}(kT_s)|^2)]$$

$$\approx \frac{M|h|^4}{16\hat{N}^2}\left(\frac{\sin(\pi\epsilon\hat{N}T_s)}{\sin(\pi\epsilon T_s)}\right)^2$$

$$+\frac{M+1}{\hat{N}}\sigma_\eta^4 + \frac{\sigma_\eta^4}{\hat{N}^2}\left(\frac{\sin(\pi\alpha\hat{N}T_s)}{\sin(\pi\alpha T_s)}\right)^2 \tag{3.15}$$

The ratio of test statistic for the two hypothesis (approximated) for the j^{th} block is given by

$$\gamma_{1j} \approx 1 + \frac{M|h|^4}{16(M+1)\sigma_\eta^4\hat{N}}\left(\frac{\sin(\pi\epsilon\hat{N}T_s)}{\sin(\pi\epsilon T_s)}\right)^2 \tag{3.16}$$

If $\epsilon = 0$, there is no cyclic frequency mismatch and

$$\gamma_{1j} \approx 1 + \frac{M|h|^4\hat{N}}{16(M+1)\sigma_\eta^4}. \tag{3.17}$$

To find optimal $\hat{N}$, we maximize the ratio of test statistic given in (3.16) by differentiating it wrt $\hat{N}$ and equating it to zero to obtain

$$\frac{2\sin(\pi\epsilon\hat{N}T_s)\cos(\pi\epsilon\hat{N}T_s)(\pi\epsilon\hat{N}T_s) - \sin^2(\pi\epsilon\hat{N}T_s)}{\hat{N}^2} = 0 \tag{3.18}$$

Equation (3.18) simplifies as

$$\sin(\pi\epsilon\hat{N}T_s) = 0 \text{ or } \frac{\tan(\pi\epsilon\hat{N}T_s)}{(\pi\epsilon\hat{N}T_s)} = 2. \tag{3.19}$$

The maximum is obtained from the solution of $\frac{\tan(\pi\epsilon\hat{N}T_s)}{(\pi\epsilon\hat{N}T_s)} = 2$ and accordingly

$$\hat{N}_O \approx \frac{1.165}{\pi\epsilon T_s} \tag{3.20}$$

and (3.14) is immediate. □

For Digital Modulated signal we similarly have

Theorem 2. *For a Digitally Modulated signal,* $x(t) = \sum_{n=-\infty}^{+\infty} a(n)p(t-nT_b)$, *the optimal block length* N_O *belongs to the set*

$$\left\{ \left\lfloor \frac{1.165}{\pi\epsilon T_s} \right\rfloor, \left\lceil \frac{1.165}{\pi\epsilon T_s} \right\rceil \right\} \tag{3.21}$$

where $\lfloor a \rfloor$, $\lceil a \rceil$ *denote the floor and ceil of* a.

Here we just outline the proof for *Theorem* 2. Note that for the Digitally Modulated signal the cyclostationary feature occur at cyclic frequency k/T_b for any integer k.Let this DM signal be sampled with T_s sampling period resulting in N number of samples. Then ratio of test static(approximated) for the j-th block is given by

$$\begin{aligned} \gamma_{1j} \approx 1 &+ \frac{I(d+c)\delta_2^2 + L_s d(\delta_4 - 2\delta_2^2)}{(I-2)\varpi_2^2 + \varpi_4} \\ &+ \frac{L_s^2 d\delta_2^2}{\hat{N}((I-2))\varpi_2^2 + \varpi_4} \left(\frac{\sin(\pi\epsilon\hat{N}T_b/L_s)}{\sin(\pi\epsilon T_b)} \right)^2, \epsilon \neq 0. \end{aligned} \tag{3.22}$$

If $\epsilon = 0$, that is, there is no cyclic frequency mismatch, and the ratio is

$$\begin{aligned} \gamma_{1j} \approx 1 &+ \frac{I(d+c)\delta_2^2 + L_s d(\delta_4 - 2\delta_2^2)}{(I-2)\varpi_2^2 + \varpi_4} \\ &+ \frac{\hat{N} d\delta_2^2}{(I-2)\varpi_2^2 + \varpi_4}, \epsilon = 0 \end{aligned} \tag{3.23}$$

To find optimal $\hat{N}$, we maximize the ratio of test statistic given in (3.22) by differentiating it wrt $\hat{N}$ and equating it to zero to obtain (3.21).

Remark 2. *Note that the value of the test static* γ_{1j} *is same for each block. Therefore the probability of detection and false alarm will be same for each block. Therefore to demonstrate the robustness of the BCSD scheme we need to show that the probability of detection and false alarm, increases and decreases, respectively, as the number of blocks, b increases.*

Let the probability of detection for j-th block be $P_{dj} = P_d$ and probability of false alarm for the

j-th block be $P_{fj} = P_f$. Using majority logic, the overall probability of detection, P_D is given by

$$P_D = \sum_{j=\lceil b/2 \rceil}^{b} \begin{pmatrix} b \\ j \end{pmatrix} P_d (1 - P_d)^{b-j}. \tag{3.24}$$

where $\begin{pmatrix} b \\ j \end{pmatrix}$ is the binomial coefficient. Similarly the overall probability of false alarm, P_F is given by

$$P_F = \sum_{j=\lceil b/2 \rceil}^{b} \begin{pmatrix} b \\ j \end{pmatrix} P_f (1 - P_f)^{b-j}. \tag{3.25}$$

When P_f is small, P_F can be approximated as

$$P_F \approx \begin{pmatrix} b \\ \lceil b/2 \rceil \end{pmatrix} P_f^{\lceil b/2 \rceil} \approx \begin{pmatrix} b \\ b/2 \end{pmatrix} P_f^{b/2}. \tag{3.26}$$

Using the approximation [26]

$$\begin{pmatrix} b \\ b/2 \end{pmatrix} \approx \frac{4^{b/2}}{\sqrt{\pi b/2}} \tag{3.27}$$

we have

$$P_F \approx \frac{(4P_f)^{b/2}}{\sqrt{\pi b/2}} \tag{3.28}$$

which clearly shows that the probability of false alarm decreases as the number of blocks increase. Similarly,

$$P_D \approx 1 - \frac{(4P_m)^{b/2}}{\sqrt{\pi b/2}} \tag{3.29}$$

where P_m is the probability of miss. As b increases $\frac{(4P_m)^{b/2}}{\sqrt{\pi b/2}}$ decreases and it follows that the probability of detection improves with the number of blocks.

3.5 Simulation Results

We consider a two cases for simulation *Case one:* single carrier signal of frequency 1.5 MHz is chosen as primary signal. Sampling frequency is chosen as 10 MHz. Block based CSD and CSD uses CAC at cyclic frequency α =3 MHz (without mismatch) and α = 3*(1+2*10^{-5}) MHz with number of lags M=1024 and sample size varies from 10240 to 600 x 1024 and samples are blocked into $\hat{N}_O = 61440$ by using (3.14).

The probability of detection is found at -20dB and -25dB SNRs with a probability of false alarm 0.1. Figure 3.2 gives the probability of detection for SC signal as a function of the sample size when the mismatch error 2x10^{-5} at -20 dB SNR. Observe that the probability of detection of Block CSD (denoted by BCSD) with mismatch is same as that of CSD without mismatch error. However the performance of CSD with mismatch degrades significantly as is corroborated by [17, 18] and follows a squared sinc pattern as is expected by referring to (3.8).

Figure 3.3 gives the probability of detection for SC signal as a function of the sample size when the mismatch error 2x10^{-5} at -25 dB SNR. Observe that the probability of detection of Block CSD (denoted by BCSD) is much better than that of CSD with mismatch. However it is inferior to that of CSD without mismatch. Also note that the probability of detection of BCSD increases in steps. This is due to the fact that the increase in sample size effects the probability of detection when the next block is filled.

Case two: Digitally modulated BPSK signal is chosen as primary signal with symbol duration $T_b = 0.4\mu s$ and pulse shaping filter $p(t) = 1$ for $t\epsilon[0, T_b)$. Sampling frequency is chosen as 10 MHz. Block based CSD and CSD uses CAC at cyclic frequency α =2.5 MHz (without mismatch) and α = 2.5*(1+2*10^{-5}) MHz with number of lags M=256 and sample size varies from 10240 to 400 x 1024 and samples are blocked into $\hat{N}_O = 71680$ by using (3.14).

Figure 3.4 gives probability of detection for DM signal as a function of the sample size when the mismatch error 2x10^{-5} at -15 dB SNR with a probability of false alarm 0.1. It can be observed that probability of detection of BCSD is better than CSD with mismatch as observed for SC signal.

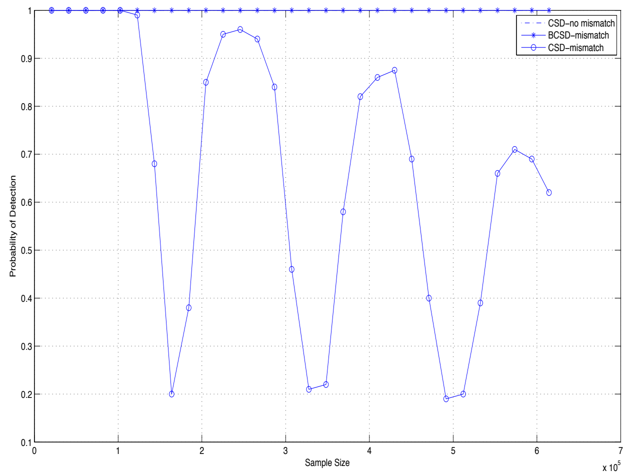

Figure 3.2: Probability of Detection for SC signal with mismatch error 2x10^{-5} at -20 dB SNR.

3.6 Conclusion

In this paper performance of block based CSD with cyclic frequency mismatch is investigated. It is shown that Block based CSD is robust to cyclic frequency mismatch. Simulations results that validate the theoretical analysis is presented. Significantly it is shown that the performance of BCSD improves as the sample size increases, in contrast to that of CSD which degrades with increase in sample size in the presence of cyclic frequency mismatch. Also for low SNR as the sensing time increases probability of detection increases. Computational complexity is same as CSD. Improvement that can be achieved by soft fusion rules will be the focus of a future study.

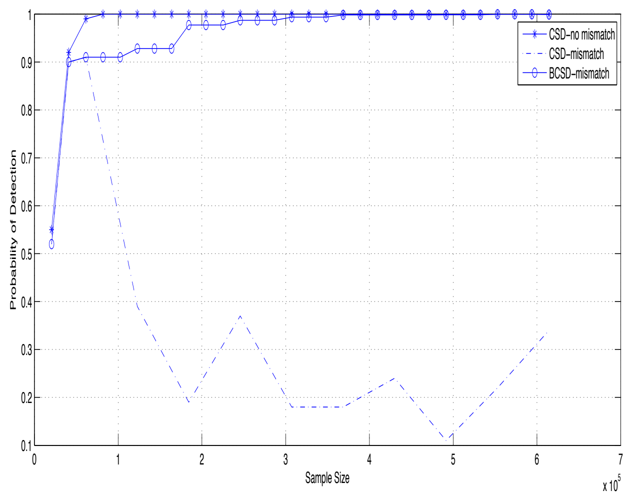

Figure 3.3: Probability of Detection for SC signal with mismatch error 2x10^{-5} at -25 dB SNR.

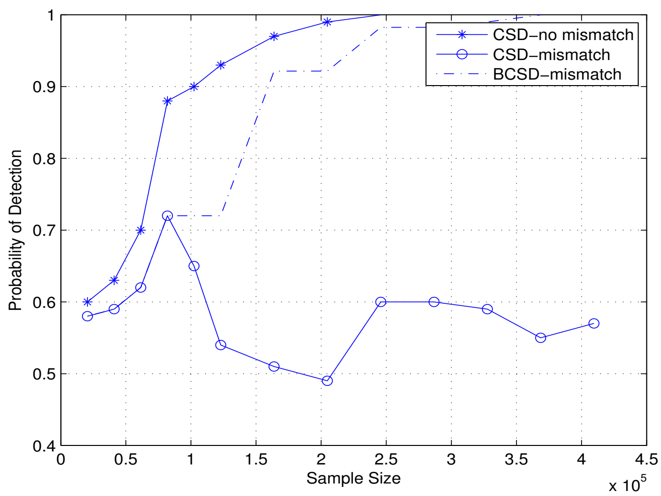

Figure 3.4: Probability of Detection for DM BPSK signal with mismatch error 2x10^{-5} at -15 dB SNR.

Chapter 4

Digital TV White Space in India

4.1 Introduction

Radio spectrum is a premium natural resource that has significant impact on wireless technology and social progress. In India, with 873.36 million mobile subscribers (which is much greater than the original predictions) [27], spectrum congestion and scarcity is a critical issue. The ongoing switchover from analog terrestrial television (ATT) to digital terrestrial television (DTT) broadcasting will provide significant benefits to solve the spectrum congestion problem.

This is because while ATT broadcasting can deliver only a single TV program on each 8 MHz, to avoid interference ATT requires large geographic separation between transmitters that operate on the same frequency. These unused TV channels, designed for protecting adjacent TV transmissions from interference at a given location, are commonly referred to as TV White Space (TVWS) [28]. In contrast, DTT enables multiple TV programs to be carried on a single frequency channel which greatly increases spectrum utilization. Therefore, it is natural to expect that some TV spectrum will be released for new usage after the digital switchover popularly named as Digital Dividend (DD).

However, in India, analog transmission will continue along with digital transmission with no decided cutoff dates for switching off analog transmission. Additionally, India has limited bandwidth available in other internationally harmonised mobile bands such as the 900, 1,800 and 2,100 MHz bands. Therefore DD spectrum is attractive as fixed broadband services are limited in rural areas and DD spectrum might offer unprecedented broadband access capability for rural populations. We calculate the Digital TVWS in South India and plots are provided for the same.

4.2 Frequency Allocation in India

Wireless licenses are allotted, in India, by the Wireless Planning and Coordination (WPC) wing of the Ministry of Communications and Information Technology and an overview is provided by the draft national frequency allocation plan (NFAP) published by the WPC [29]. However, detailed information is not publicly available. Significantly, the draft NFAP 2011 made a very encouraging statement of intent for the utilization of TVWS [29]. Also the following points from NFAP 2011 are of interest

- (Analog) TV broadcast transmitters operate only in the 470-590 MHz band in the UHF band
- Spectrum in the 470-890MHz frequency band is earmarked for Fixed, Mobile and Broadcasting Services
- Digital broadcasting services including (digital TV broadcasting) will operate in the 585-698 MHz
- On completion of digitization,the 470-590 MHz in the UHF band, 54-68 MHz and 174-230 MHz in VHF band; we presume, will be completely free. The important point is the actual date of complete digitization of terrestrial TV.

With regard to terrestrial TV, Doordarshan, a public service broadcaster, has exclusive use of terrestrial TV bands. Doordarshan today has more than 1415 terrestrial transmitters throughout India and transmits on two all-India channels DD National and DD News. The frequency band information of the various transmitters are given in 4.1 .

Table 4.1: Doordharsan TV Transmitter Classification

Band(MHz)	No.of transmitters	Bandwidth (MHz)
VHF Band-I (54-68)	8	7
VHF Band-III (174-230)	1034	8
UHF Band-IV (470-590)	373	8

As a part of the digitization process, Doordarshan has four DTT units at present; one each in Delhi, Mumbai, Kolkata and Chennai, and plans to completely digitize its terrestrial transmission by end of 2017. From the tenders released by Doordarshan [30] , it looks like India will have a total of 19 DTT 6 KW transmitters by 2014-2015. However, no finite time frame has been decided for stopping the analog transmission and it will continue along with the digital transmission till a substantial percentage of the population switches over to a digital mode of reception.

4.3 TV White Space Maps for South India

In this section we present TV white space Maps for South India. These maps are developed using the code provided by [31, 32] with some modifications. Note that we need a more 'precise' definition of white space. Following [33] 'white space' is the intersection of the spatial holes from 'pollution' and 'protection' viewpoints. The classical communication theory viewpoint or the 'pollution' viewpoint takes into consideration the fact that even though a region is in use by a secondary device, the interference at the secondary receiver might be higher than the tolerable interference level. The 'protection' viewpoint stipulates that a secondary radio can only operate in locations where it cannot generate harmful interference to the primary system. Both viewpoints exclude regions around a transmitter and the actual size of the 'white space' is the intersection of the spatial holes from these two viewpoints. As India does not have TV white space regulations, we use the regulations of the FCC (US). Also, as data on microphones is not available they are ignored. We study the Digital TV white spaces in India based on the the Hata Model. Using the TV transmitter information provided by Doordarshan for South India, we quantify TV white space in the UHF bands under the technical specifications given by the FCC for Digital terrestrial TV.

4.4 Digital TV White Space

Since India has a separate band allocated for DTT, we plot

- The white space available at present for the four installed DTT transmitters,
- The white space that will be available when 192 DTT transmitters will be installed (by 2017),
- The white space that will be available when the digitization is completed.

These plots give us an idea of how much white space is available in the DTT band as the digitization proceeds. Note that the proposed specifications for the 19 proposed DTT transmitters is available [30] and we have plotted the 'Digital' White Space for the whole country in the first two plots and; for south India in the last plot. The digital plots are made assuming that the transmitters are FCC compliant for DVB-T2. Accordingly, in the code of [32], the minimum equivalent field strength at receiving place is set to 51 dB mu/m for minimum C/N = 20 dB.

Observe from 4.1 that except regions around the four metros (Delhi, Mumbai, Kolkatta and Chennai) the entire country is a white space for all the 'digital' channels. 4.2 gives the white space that will be available when all the additional 19 DTT transmitters are installed. Note again the entire country is a white space for all the 'digital' channels except for small regions around the 19 DTT transmitters. Also note that 11 channels will be available throughout India in the 'Digital' band. 4.3 gives the white space, for South India, that will be available after completion of digitization. We have assumed that all the (present) 275 locations in South India have been converted to DTT with same antenna height but with a power of 6 KW. Channel allocation was done so as to minimize adjacent channel and co-channel interference and sorting latitude. Channel allocation starts with 21, next location will be allocated channel $max(x1, x2, ..., xn) + 2$ if it is in same range of $x1, x2, x3, ..., xn$ channels. Observe that 6 channels will be available throughout India in the 'Digital' band.

4.5 Conclusion

In this chapter we have presented a study of DTV White spaces for India. We have studied the white spaces available (and that will be available) in the digital UHF bands.

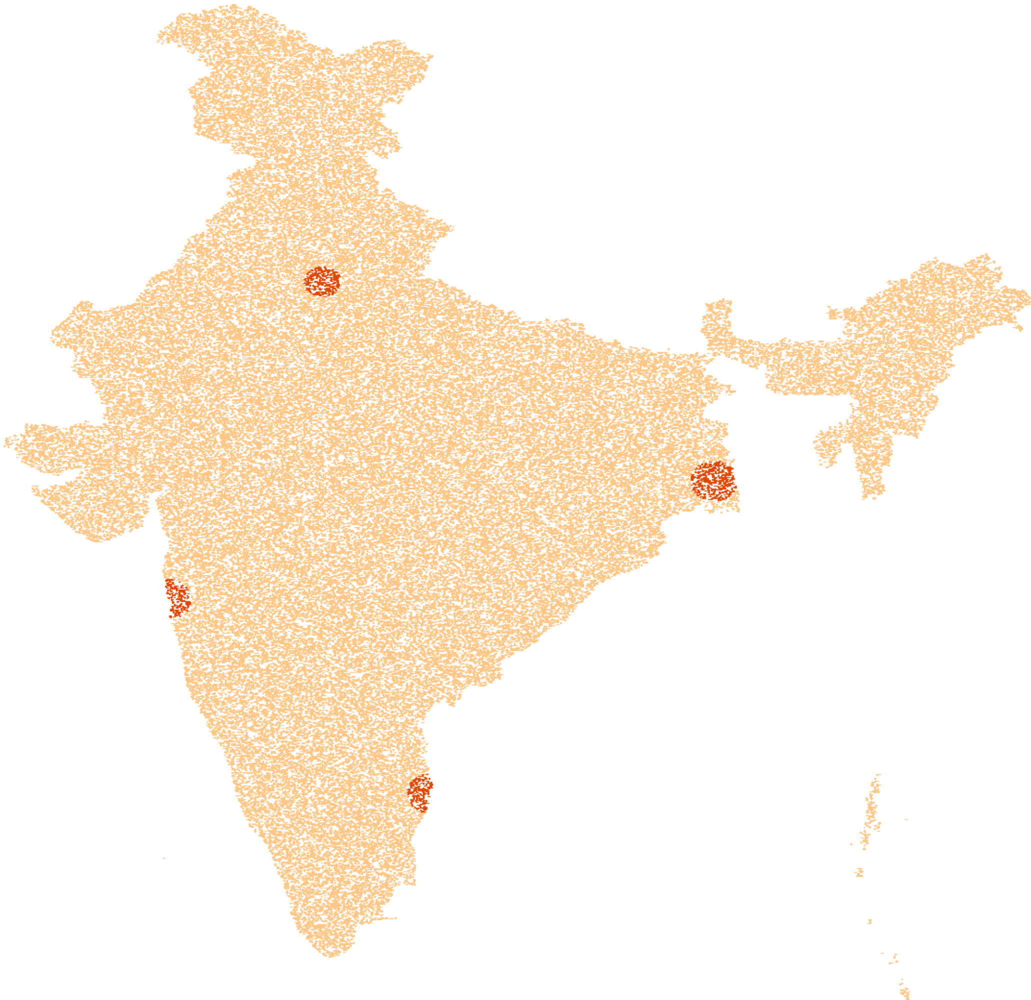

Figure 4.1: Digital white space plot for India for 15 channels of UHF V band with present DTT transmitters

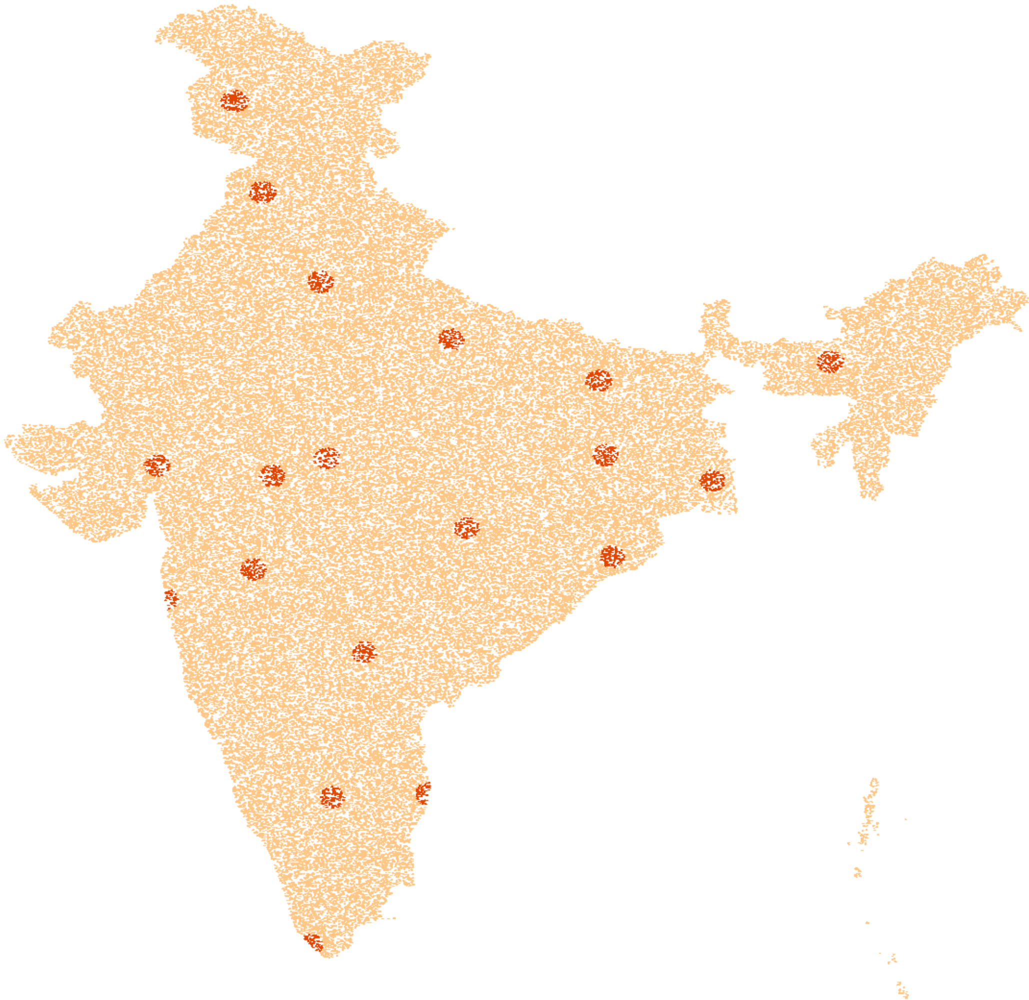

Figure 4.2: Digital white space plot for India for 15 channels of UHF V band with 19 DTT transmitters

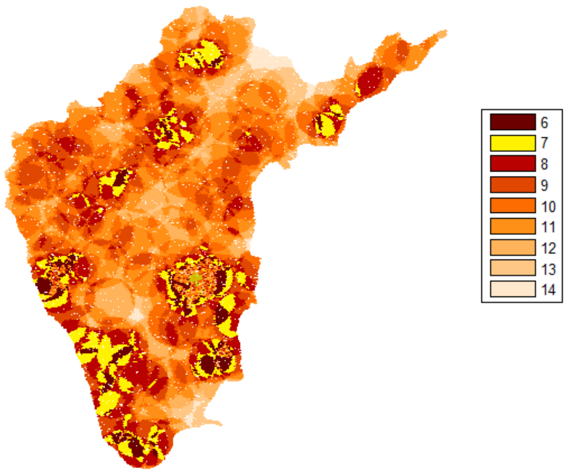

Figure 4.3: Digital white space plot for South India for 15 channels of UHF V band after completion of digitization

Chapter 5

Conclusion

We have demonstrated the feasibility of Cognitive Radio technology using the existing architecture of GSM network using OpenBTS. We also have developed efficient algorithm to overcome pratical problem like frequency offset using block based cyclostationary detection. We have presented oppurtunity for cognitive radio in India by discussing the availability of digital TV white space.

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