

PRACTICAL IMPLEMENTATION OF SENSING RECEIVER IN COGNITIVE RADIOS

by

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B.S., University of Southern California, 1988

M.S., University of Southern California, 1991

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AN ABSTRACT OF A DISSERTATION

submitted in partial fulfillment of the requirements for the degree

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Department of Electrical and Computer Engineering  
College of Engineering

KANSAS STATE UNIVERSITY  
Manhattan, Kansas

2009

## **Abstract**

Due to the underutilization of scarce radio spectrum resources, there is a major shift from an exclusive use model to a spectrum sharing model. This model has been successful in unlicensed bands where wireless standards such as Bluetooth and WLAN IEEE802.11 have flourished. In order to extend this model to licensed bands, a smart radio is required. The cognitive radio is a smart radio that leverages its knowledge of the local environment in order to operate with acceptable interference to primary users in the band. The brain of the cognitive radio is encapsulated in the sensing receiver. Using the sensing receiver, the cognitive radio observes the activity in a band and adapts its communication link parameters so that the interference is minimal. The sensing receiver attributes include wideband operation and agility. The receiver must be able to make fast decisions accurately of the spectrum activities. To date, a practical sensing receiver has been a major stumbling block in the realization of cognitive radio.

In our research, we demonstrate that a practical sensing receiver is in fact feasible. Our research objectives are: (1) to develop a sensing receiver based on current technology, and (2) to augment our solution with an associated algorithm that makes fast and accurate sensing possible. Our proposed solution is a Dedicated Sensing Receiver (DSR) that employs an adaptive algorithm to enable the cognitive radio to effectively operate in a competitive environment. The DSR algorithm divides the frequency band into coarse sensing bins which are further divided into fine sensing bins. We derive a mathematical model that is optimized for minimizing sensing time. The optimization algorithm generates the optimum number of coarse and fine sensing bins based on environment conditions and the radio implementation. Our results illustrate the DSR's ability to reduce open spectrum recognition time by over 10X while operating over a broad range of frequencies. This drastic improvement opens the door for the cognitive radio technology to be used in commercial applications in the near future and more importantly, it allows for the cognitive radio to succeed in time sensitive applications such as voice or video. We have also developed an end to end simulation platform to model the algorithm and the dedicated sensing receiver performance.

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Major Professor  
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## Dedication

Glory be to God, the Almighty for, by His grace and His grace alone, that this work has been completed.

*God grant me the serenity;  
To accept the things I cannot change;  
Courage to change the things I can;  
And wisdom to know the difference.*

To my beautiful and loving wife, Hanan: You are the best thing that ever happened to me. Without your courage, support and love, I would not have achieved success. You are a bright ray of light in my life. I love you.

To my witty and genius son, Ali: You are smarter and brighter than I ever was. You have the potential to be great! Son, please remember that nothing in this world can take the place of persistence. Talent will not; nothing is more common than unsuccessful people with talent. Genius will not; unrewarded genius is almost a proverb. Education will not; the world is full of educated dumb people. Persistence and determination alone are what's most important.

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## Glossary

|        |   |
|--------|---|
| 2G     | Second Generation   |
| 3G     | Third Generation  |
| A/D    | Analog to Digital Conversion  |
| ADC    | Analog to Digital Conversion  |
| ARCEP  | Autorité de Régulation des Communications Électroniques et des Postes |
| ASIC   | Application Specific Integrated Circuit                               |
| BB     | Baseband  |
| BER    | Bit Error Rate  |
| BOM    | Bill of Material  |
| BPF    | Band Pass Filter  |
| BPSK   | Binary Phase Shift Keying   |
| BT     | Bluetooth   |
| CDMA   | Code Division Multiple Access   |
| CMOS   | Complementary metal–oxide–semiconductor                               |
| CNR    | Carrier to Noise Ration   |
| Codecs | Coder/Decoders  |
| CPE    | Consumer Premise Equipments   |
| CR     | Cognitive Radio   |
| dB     | Deci Bell   |
| DSA    | Dynamic Spectrum Access   |
| DSP    | Digital Signal Processing   |
| DSR    | Dedicated Sensing Receiver  |
| FCC    | Federal Communications Commission                                     |
| FDMA   | Frequency Division Multiple Access                                    |
| FER    | Frame Error Rate  |
| FFT    | Fast Fourier Transform  |
| G      | Gain  |



|       |  |
|-------|--|
| GHz   | Giga Hertz ( $10^9$ Hz)                          |
| GSM   | Global System for Mobile communications          |
| Hz    | Hertz  |
| IC    | Integrated Circuit                               |
| IEEE  | Institute of Electrical and Electronic Engineers |
| IF    | Intermediate Frequency                           |
| ISM   | Industrial, Scientific and Medical               |
| KHz   | Kilo Hertz ( $10^3$ Hz)                          |
| Km    | Kilometer  |
| LNA   | Low Noise Amplifier                              |
| LO    | Local Oscillator                                 |
| LPF   | Low Pass Filter                                  |
| LSI   | large-scale integrated                           |
| LUT   | Look Up Table                                    |
| MAC   | Middle Access                                    |
| MHz   | Mega Hertz ( $10^6$ Hz)                          |
| MIMO  | Multiple Input Multiple Output                   |
| ms    | MilliSeconds                                     |
| mW    | MilliWatts                                       |
| NF    | Noise Figure                                     |
| NPRM  | Notice of Proposed Rule Making                   |
| Ofcom | Office of Telecommunications                     |
| OFDM  | Orthogonal Frequency Division Multiplexing       |
| PDA   | Personal Digital Assistant                       |
| PHY   | Physical   |
| PLL   | Phase Lock Loop                                  |
| PSD   | Power Spectral Density                           |
| PU    | Primary User                                     |
| QAM   | Quadrature Amplitude Modulation                  |
| QoS   | Quality of Service                               |
| QPSK  | Quadrature Phase Shift Keying                    |

|       |  |
|-------|--|
| RF    | Radio Frequency                                |
| RFIC  | Radio Frequency Integrated Circuit             |
| RFID  | Radio Frequency IDentification                 |
| SDMA  | Space Division Multiple Access                 |
| SDR   | Software Defined Radio                         |
| SER   | Symbol Error Rate                              |
| SNR   | Signal to Noise Ratio                          |
| SU    | Secondary User <sup>1</sup>                    |
| TDM   | Time Division Multiplexing                     |
| TDMA  | Time Division Multiple Access                  |
| TV    | Television                                     |
| U-NII | Unlicensed National Information Infrastructure |
| UWB   | Ultra Wide Band                                |
| VCO   | Voltage Controlled Oscillator                  |
| W     | Watts  |
| WCDMA | Wideband Code Division Multiple Access         |
| WLAN  | Wireless Local Area Network                    |
| WPAN  | Wireless Personal Area Network                 |
| WRAN  | Wireless Regional Area Networks                |

---

<sup>1</sup> SU in this document refer to an unlicensed user operating in a licensed band

# 1 Introduction

Under the current spectrum management policy in the United States, the Federal Communications Commission (FCC) licenses the rights of a spectrum to a single user in a geographical location for wireless operation. The radio spectrum from 9KHz to 300GHz has already been allocated to various users/services. As a scarce resource, spectrum is regulated such that licensed users have the exclusive rights to operate in a given frequency band. With wireless becoming so ubiquitous in our everyday lives, the sheer volume of wireless users is overwhelming the existing infrastructure and is taxing the spectral management policies in the United States and around the globe [3]. Ad-hoc networks and infrastructure networks have proven the ability to coexist in unlicensed bands. For example, in the Industrial Scientific and Medical (ISM) band operating at 2.4GHz, Bluetooth Wireless Personal Area Network (WPAN) coexist with the IEEE802.11 Wireless Local Area Network (WLAN), cordless phones, RFID cards and microwave ovens. With such success, the wireless community is wondering about the relevance of an exclusive use policy. Two possible scenarios for spectrum allocation are:

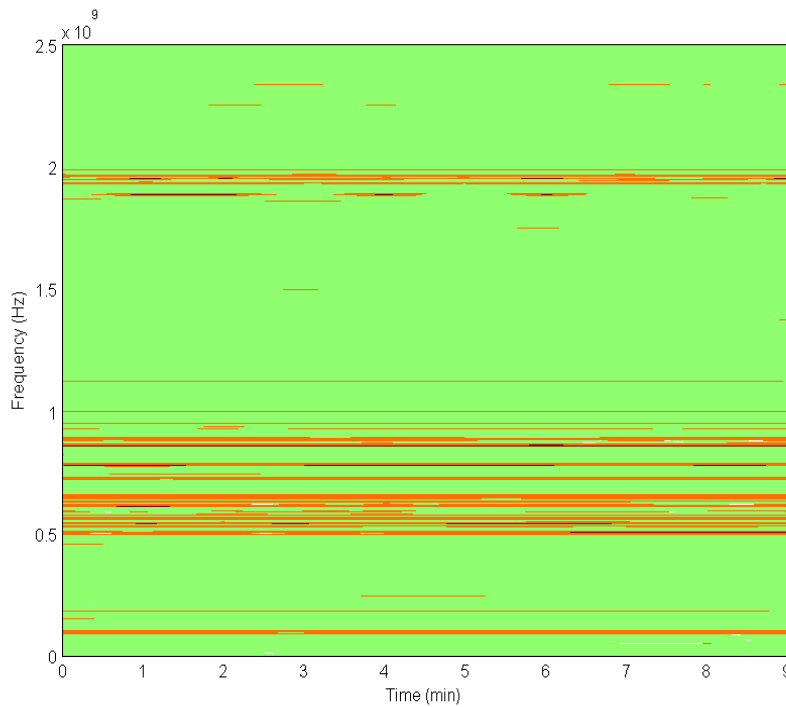
1. Owned spectrum, in which the owner has exclusive use rights of the spectrum
2. Shared spectrum, in which users agree upon an etiquette for cooperation or competition for spectrum resources.

The owned spectrum model is a legacy spectrum allocation policy that leads to significant underuse of the spectrum.

## 1.1 Spectrum Allocation Inefficiencies

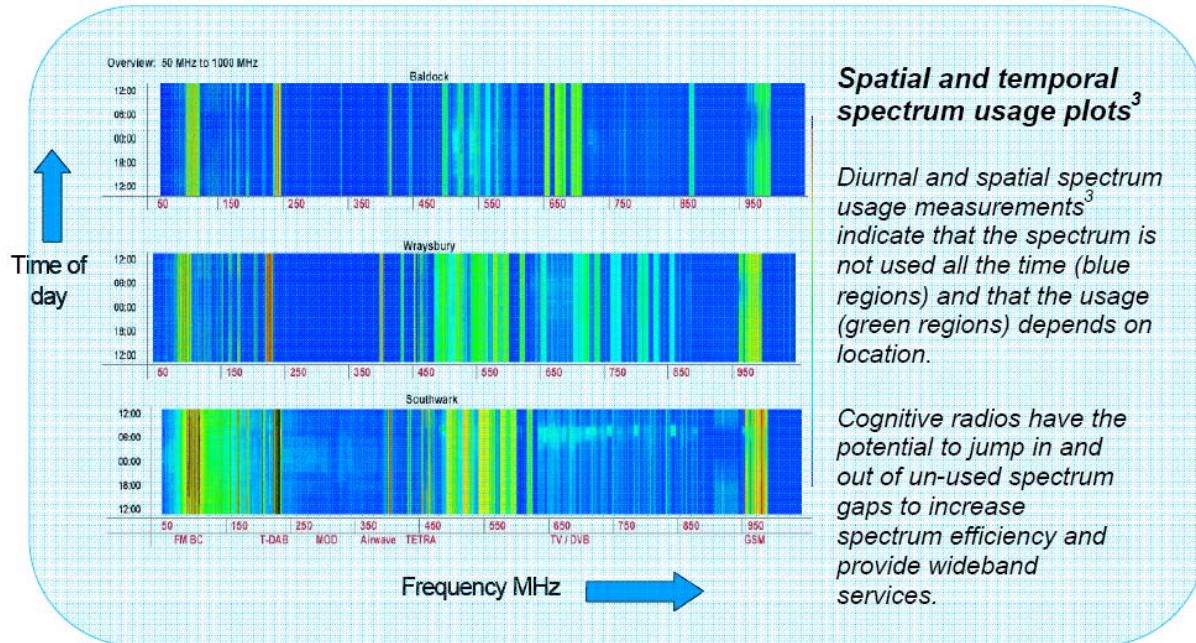
The advent of wireless communication has made static spectrum allocation a daunting challenge for the regulatory bodies around the world. In the US, the Federal Communications Commission (FCC) is chartered with regulating communications and with oversight responsibility for new technology. The FCC allocated some frequencies as unlicensed bands where users can operate

without a license. For example, in 1985, the FCC opened up the Industrial, Scientific and Medical (ISM) bands for wireless LANs and mobile communications (Part 15.247). This policy started a revolution towards spectrum sharing. Today, the utilization of the unlicensed bands is getting sufficiently saturated which is forcing wireless standards to migrate towards higher frequency bands. Over the past decade, we have seen a consistent migration of the shared spectrum from the 900MHz and 2.4GHz ISM bands towards the Unlicensed National Information Infrastructure (U-NII) 5GHz frequency band, and lately wireless standards workgroups have started working in the newly opened spectrum in the 60GHz millimeter wave frequency band [1]. However, lower frequencies (<5GHz) are generally more desirable because of better propagation property, wall penetration, and lower radio complexity. In spite of this apparent scarcity of spectrum, spectrum measurements [2,18] in various cities illustrate that both the licensed and unlicensed bands are not used efficiently and there exists a lot of “white spaces” in the spectrum. The measurements taken by the Berkeley Wireless Research Center show the allocated spectrum is vastly underutilized [2]. Measured results of radio activity in downtown Berkley, CA are shown in Figure 1-1 below where green signifies no spectrum activities.



**Figure 1-1 Radio Activity in Downtown Berkley, CA [2]**

Similar results were also measured by the Office of Telecommunications (Ofcom) in the United Kingdom. Ofcom commissioned a study in 2007 [18] and the spectral activity was recorded in three different British cities. Results up to 1GHz were recorded.



**Figure 1-2 Spectrum Usage in 3 English Cities [18]**

In the lower bands, the occupancy may be even lower. In the 30 – 300MHz spectrum, measurements show that the utilization rate is only 5.2% [26]. In fact, the FCC Spectrum Task Force has reported that utilization of spectrum below 3GHz can be as low as 15% [3]. Yet, the primary users (users with exclusive rights to a spectrum band) in the owned or licensed spectrum are reluctant to share the spectrum with secondary (SU) or opportunistic users. Spectrum sharing promises to increase spectrum efficiency and it has been proven to be successful and commercially practical. Hence, the FCC has been considering reallocating some of the white spaces for secondary wireless use since 2004 [4].

In 2002, DARPA launched the first phase of the neXt Generation (XG) communications research program that focuses on opportunistic radios and dynamic spectrum sharing [5]. In 2007, the XG program reported a 10X increase in spectrum utilization and cited that high-speed, wideband

sensing remains a technical challenge [6]. Similarly, University of Kansas in collaboration with Rutgers University and Carnegie Mellon University developed CogNet [125] which is an experimental protocol stack for cognitive radio. CogNet research [126] is providing innovations that “promise to form a foundation for cognitive adaptation of network layers.” However, the main challenge is mostly due to today’s radio design limitations of RF receivers. In addition, there is a need for a revamped and dynamic allocation policy that addresses such inefficiencies.

## **1.2 Increasing Spectral Efficiency and Utilization**

Spectral efficiency has become the primary concern of wireless engineers in the design of next generation wireless systems. Considering the cellular phone market alone, the industry is predicting shipments of more than 1.4 billion new devices in 2009 [110]. This mind boggling number does not take into account the existing number of phones already in use or other wireless standards such as WLAN and BT devices. With the number of users in the billions, cell phone providers are continuously looking for new ways to decrease capital investment in basestation (i.e. decrease the number of basestations deployed) while increasing the number of users and maintaining customer satisfaction. The only way to achieve such a feat is by increasing spectral efficiency and spectrum reuse. Over the past three decades, wireless communication has focused on multiple access schemes such as OFDM and CDMA. Improvements in radio architecture and especially in RFIC technology have yielded incremental performance improvement. However, with all advances we have witnessed over the past few years, the underutilization of the spectrum still persists.

One of the main reasons for the inefficiencies is the fact that the FCC licenses or assigns spectrum to a single owner be it; a television broadcast channel, a cell phone provider or a microwave backhaul operator. As discussed earlier, the spectrum under 3GHz is less than 15% utilized [3]. The other 85% of the time the spectrum remains unused by the PU and with the appropriate technology, the unused spectrum could be allocated to secondary users. Opponents of the spectrum sharing concept fear that undisciplined secondary users may cause interference in addition to increasing in-band noise which may degrade overall performance. Proponents of spectrum sharing

point to the overwhelming success in the unlicensed bands. Regardless on which side of the debate users fall, one thing is abundantly clear; spectrum sharing if applied properly may improve spectrum utilization. One way to share the spectrum is to allow secondary users to operate with minimal or no interference to primary users.

In order to enable secondary users to coexist with other users in a band, the radio must be opportunistic. Opportunistic users must quickly recognize and exploit available frequencies while they must be willing to quickly interrupt the link and look for other available channels to complete transmission. This concept is called Dynamic Spectrum Access (DSA). DSA requires the radio to have the following features:

- 1) Cognitive: the radio must sense the environment and recognize spatial, temporal or spectral voids. It must quickly identify opportunities to use the unused channel.
- 2) Programmable: the radio must be programmable such that it is able to change parameters such as modulation, power level, operating frequency and transmission bandwidth.
- 3) Agility: The radio must quickly scan and / or jump to available channels.
- 4) Broadband: the radio must have the ability to scan a large number of frequencies in order to increase the probability of correctly identifying available channels as to avoid long interruption in transmission.
- 5) Low cost: the radio must be within the cost envelope of today's technology.
- 6) Low power consumption: the radio must have the ability to save power similar to today's wireless device.

Cognitive radios originated as a possible solution to dynamic access in a DSA system. Cognitive radio adapts to the environment as a function of space and time by sensing the spectrum and making short term decisions on appropriate transmission parameters.

### 1.3 Cognitive Radio

With greater demand for high data rates and mobility, cognitive radios represent one of the most significant paradigm shifts in wireless engineering in the past three decades. Cognitive radios are smart radios operating as secondary users that exploit Dynamic Spectrum Access (DSA) techniques. The cognitive radio concept was originally introduced in the software defined radio community but since evolved into its own field of study [22]. Cognitive radios are capable of adapting to user needs as well as to environment conditions. The radio senses the local environment, adapts itself to the environment conditions and adjusts its transmission parameters to meet its objectives and to minimize interference with other occupants of the spectrum. Several advances in approaches to spectrum management, radio policies, radio design, and network design are inching cognitive radios towards commercial reality.

One of the key elements to Cognitive Radio (CR) is the ability to learn and monitor the environment. At the core of cognitive radios is a sensing receiver which assesses channel conditions, determines presence of users, and exchanges information with other users if required. The addition of spectrum sensing function and built-in intelligence is what transforms a classical radio to a “smart” radio with a capability to make decisions. The sensing function requires the radio to step through a set of frequencies and perform signal detection at each channel. Even weak signals must be detected because the CR might be experiencing a fade or a hidden terminal problem [47]. The traditional signal processing approach is to treat this problem as a hypothesis testing problem for which the test statistics are averaged over time. As the average time increases, the resolution improves. Therefore in general, sensing time is challenged by the radio ability to tune to a specific frequency and by the signal processing overhead required by power detection. The challenge becomes even more daunting when one considers time sensitive applications such as video or voice transmission. Time sensitive applications require minimal end to end delays, and more importantly, can not afford long interruptions in transmission.



### *1.3.1 Time Sensitive Applications*

A large fraction of wireless devices are used for voice or video transmission. Unlike data transmission, long delays for channel assignment or constant interruption would hurt the quality of the transmission. Acceptable time delay requirements for supporting voice applications are subjective. Time delay tests are calculated based on industry standard [10,11,12,13,14,15]. It is difficult to define a universal minimum delay number or an end to end latency. However, it is certain that the latency within a system must be minimized for real time applications such as voice, interactive video, mission critical data, and network management data.

We choose to focus on delay requirements and packet loss for voice as an illustrative real time application. In voice transmission, delays cause two main noticeable issues: echo and speech overlap. Total delays in the system can be a combination of factors: 1) voice Codecs delays, 2) processing delays, and 3) network delays. Furthermore, packets carrying voice information could be lost or dropped during transmission. Packet losses create a phenomenon known as clips which further degrades voice quality. Voice degradation has been measured through subjective testing [10]. There exists numerous publications and standards providing guidance and metrics for acceptable voice performance into time delays. The following requirements have been identified based on these tests:

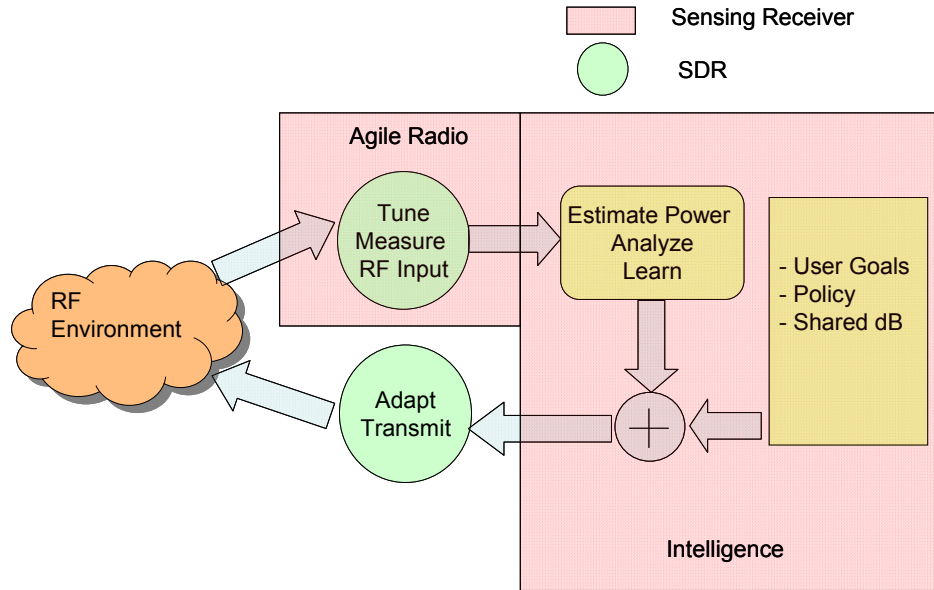
- Total delay from mouth to ear should be less than 150ms [10]
- Karam et al. [11] show that inherent delays are due to queuing delays, formation time and propagation delay vary according to encoding schemes anywhere from 5 ms to 85ms.
- Packet Loss Concealment (PLC) is a technique used to mask clips and is dependent on the codec. Popular codecs such as G.711 [12], G.729A [13] and G.723.1 [14] are capable of replaying the last received sample (interpolation) at different attenuation and hence concealing the packet loss.
- PLC is capable of concealing up to 20ms of missing voice samples [15].

Based on acceptable voice delays and popular codec standards described above, we propose a rule of thumb for total time delay between packet transmissions to be less than 20 ms. In order to better understand the impact of the 20ms limit on a communication link, the radio system design must define an allocated time budget for the different components contributing to latencies in the receiver.

### ***1.3.2 Cognitive Radio Architecture***

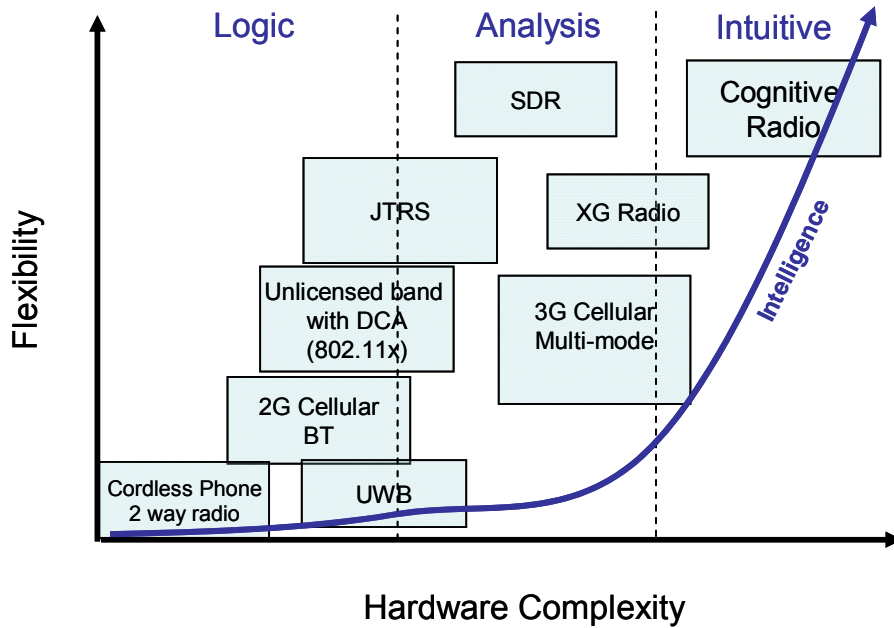
Radio receivers remain limited by available technology. Radio designers continue to push the technology envelope. In the recent decades, multiple access schemes such as TDMA, CDMA and OFDM became popular and contributed to easing some of the congestions. However, radio architecture alone does not address the underutilized spectrum. The “white spaces” in the spectrum are substantial and may be temporarily reassigned to secondary users. A proactive solution is required in order for secondary users to operate on a temporarily available channel in the background with minimal interference with the primary users. Hence, the secondary user must carry the burden of continuously monitoring the channel and must be able to interrupt communications to avoid colliding with the primary user. Due to the burden of continuously monitoring and adapting to a changing environment, such radios are dubbed “smart radios.”

The cognitive radio is a smart radio composed of two parts: 1) a sensing receiver and 2) a software programmable radio. A cognitive radio block diagram is shown in Figure 1-3.



**Figure 1-3 Cognitive Radio Block Diagram**

The sensing receiver scans the environment for available channels suitable for secondary operation. The sensing receiver tunes to a desired frequency, estimates power level, analyzes and learns the environment activities. It then applies user goals, available regulatory policies and any shared information from other users or a central command to set the parameters of transmission. The CR adopts the transmission parameters and begins transmission. The sensing receiver provides the intelligence necessary for the CR to operate as a secondary user. Because of the added intelligence, the radio complexity increases. A comparison of the different standards versus hardware complexity is shown in Figure 1-4.



**Figure 1-4 Hardware Complexity versus Flexibility**

Cognitive radios are concerned with timely detection of available channels. The radio must be agile and wideband in order to take advantage of available transmission opportunities.

### *1.3.3 Standards and Policies*

Cognitive radio is driving significant effort in standard and regulatory bodies. With operating frequency in sub 1GHz bands and low utilization rates, the TV broadcast bands are prime for cognitive radio operation. In response to the FCC spectral task force efforts in 2002 and the Notice of Proposed Rule Making (NPRM) dated May 2004 [3,111], the IEEE initiated the IEEE802.22 working group in order to standardize a worldwide unlicensed operation in the TV broadcast bands [4] standard. The most prominent target application of IEEE802.22 WRANs is wireless broadband data access to rural and remote areas, with performance comparable to DSL and cable modem. The standard addresses the needs of both the PHY and the MAC layers [55]. The system is setup as a point to multipoint air interface, with the data traffic flowing through a

basestation. In 2005, the IEEE established the IEEE1900 project to develop standards targeted at DSA and CR. The IEEE 1900 task force later evolved into IEEE Standards Coordinating Committee (SCC) 41 Standards Committee on Next Generation Radio and Spectrum Management [27].

As policy makers move towards creating standards to adopt dynamic spectrum sharing, scientists and engineers are scrambling to design and develop a practical cognitive radio architecture.

In summary, the complexity of the radio renders the realization of cognitive radio almost impossible. One possible implementation solution is a brute force approach where the focus is on the radio design while other approaches focus on an overall system solution based on collaboration and cooperation techniques within the cognitive radio network. In the next section, we highlight different techniques used to address the challenges involved in CR implementation.

## **1.4 Prior Efforts**

In order for Cognitive Radio (CR) to become a practical technology, it must overcome the implementation challenges. The CR must perform the measurements swiftly, make a decision and tune to available bands over wide frequency bands. The broader the scanned frequency spectrum, the higher the probability of finding a viable channel suitable for CR operation. However, broadband radio design has been a challenge for many decades. The sensing problem can be addressed at two levels:

- 1- Cognitive radio network efforts
- 2- Wideband agile radio architecture efforts

Each level attempts to either improve the sensing performance or to reduce the complexity of the design.

### ***1.4.1 Cognitive Radio Network Efforts***

There are two types of network for secondary users: competitive or cooperative network.

Competitive or non-cooperative radios [112] work to optimize their own performance with disregard to other users in the band. The goal is the competitive maximization of the information throughput sustained by each link active over the network. Such non-cooperative techniques have been heavily researched for network traffic applications. Selfish non-cooperative players are interested in connecting to a source by making a routing decision that minimizes cost [112,113,114,115]. However, competitive networks do address the issue of minimizing the sensing complexity.

On the other hand, cooperative sensing has the ability to reduce the workload and the complexity of the sensing receiver. The cooperative network concept is based on information sharing among users. The information about the environment conditions is shared or broadcast to other users [44]. In addition, users may agree to divvy up the sensing workload among them. Cooperative networks have shown a lot of promise and have been heavily researched in both wired and wireless communications [116,117,118]. Similarly, cooperative networks or cooperative sensing have been popular in cognitive radio [44,70,98]. Hence, a significant effort is spent on addressing the sensing challenge through cooperative network or cooperative sensing. Several research topics in a distributed [54] and centralized network [97] have yielded improvements in important sensing parameters such as detection probability, low SNR detection, false alarm minimization, collision avoidance, and scheduling.

There are several proposed approaches to cognitive radio operation:

- **Gap Filling Approach (White Space Filling):** This is the most common and most obvious approach. The CR transmission occurs during spatial, temporal or spectral voids. This interference avoidance is predicated on the ability of the CR to accurately and quickly sense available gaps in transmission [38].

- Simultaneous Controlled Transmission: In this case, the CR must adjust transmit power level such that the PU receiver is able to operate with acceptable interference or noise temperature from other occupants [39].
- Cognitive Transmission: This approach assumes that the CR has prior knowledge of the PU link [40]. Assume that the CR is operating in a GSM band. The CR listens to the basestation command and control information and decides when and where to transmit in the band. If a CR has the ability to demodulate and decode the information contained in the TS0 slot in a GSM frame [21], it would possess the ability to predict where and when GSM users would transmit. Hence, the CR is able to use this knowledge to avoid interference. Another example is that of CDMA based PU. In this case, the CR would have the ability to use one of the available orthogonal spreading codes so that the interference is minimized [40].
- Opportunistic Interference Cancellation: This approach adds additional intelligence to the cooperation in a cognitive network. It assumes cooperation between the PU and the CR transmitter [41]. It assumes that the CR knows the transmission message and is able to provide orthogonal message in essence cancelling the effect at the PU receiver. The advantage of this technique is that the CR is able to transmit at full power with real effect on the PU receiver
- Network with Beacon: This approach is an extension of interference avoidance described earlier. The PU transmits a beacon before each transmission. The beacon signal warns cognitive users of an upcoming PU transmission [42]. This early warning minimizes the probability of interference.
- Network with Primary Exclusive Regions: This approach is more suitable in a broadcast environment where one PU transmitter is communicating with multiple receivers. The network imposes exclusive regions near the PU transmitter that forbid CR operation [43].
- Single or Multi- Hop Network: This is a cooperative model where the CRs agree to communicate with a nearby receiver with self imposed transmit power. With such restrictions on the transmit signal, the operating distance between the transmitter and receiver is limited [45]. In multi-hop networks, there is a stronger cooperation among users. As an example, the network allows nearby receivers in a coalition to strip off stronger transmissions.

The research around sensing is also very extensive

- **Blind Sensing Algorithms:** the technique is based on oversampling the received signal or by employing multiple receive antennas. The algorithm does not require knowledge of the channel or of the noise power (i.e. blind). When the primary signal is present, the signal statistics computed will differ much more in value from each other, than when the primary signal is not present [[129],[130].
  
- **Cooperative Sensing:** Defines 2 protocols:
  - *Non Cooperative (NC) Protocol:* All the users detect the primary user independently. However the first user to detect the presence of the primary user informs the other users through the central controller (distributed sensing)
  - *Totally Cooperative (TC) Protocol:* Two users operating in the same carrier, if placed sufficiently near to each other, cooperate to find the presence of the primary user. The first user to detect the presence of the primary user informs the others through the central controller.
  - Agility is measured as the probability of detection of non cooperative divided by prob. of detection of cooperative protocol. The paper estimates that maximum gain in using the technique is 11% [131]
  
- **PU LO Leakage Detection:** Technique is based on the possibility of detecting primary receivers by exploiting the local oscillator (LO) leakage power emitted by the RF front end of primary receivers [132].
  
- **Radio Identification Based Sensing:** A complete knowledge about the spectrum characteristics can be obtained by identifying the transmission technologies used by primary users.
  - Several features are extracted from the received signal and they are used for selecting the most probable primary user technology by employing various classification methods. Features obtained by energy detector based methods are used for classification. Channel bandwidth and its shape are used in reference



features. Channel bandwidth is found to be the most discriminating parameter among others [133]

- Cyclostationary Feature Detection: To improve spectrum sensing sensitivity, cyclostationary feature detection computes the autocorrelation of received signal before the spectral correlation detection. The technique is based on the fact that modulated signals are in general coupled with sine wave carriers, pulse trains or cyclic prefixes which result in built-in periodicity. The periodicity help extract information about the received signal such as modulation, pulse shape and bandwidth [134].

However, none of these approaches addresses the requirements for time sensitive applications. As a matter of fact, several of these techniques actually lengthen the time required to search for appropriate CR channels.

**Key:** ● Best Solution  
 ● Good  
 ○ Adequate  
 ● Inadequate  
 ● Un-workable

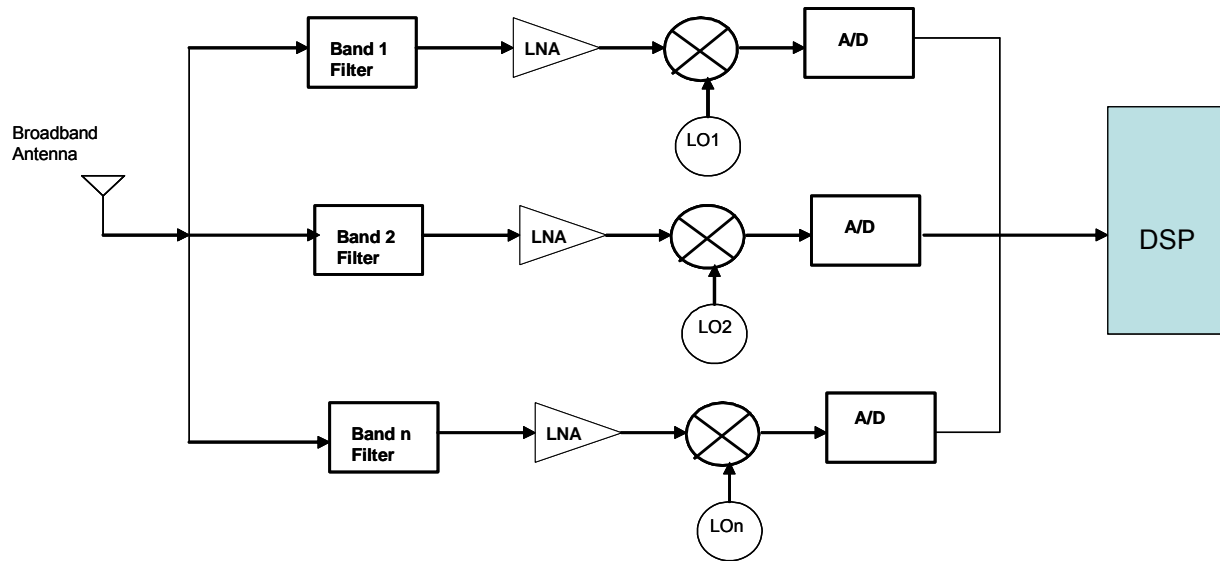
|   | Detection Time   | Detection Ability   | Complexity | Dependencies  | Overall Performance   |
|---|--|---|------------|---|---|
| Base Sensing Receiver                     | ● Too slow   | ● OK in narrowband apps   | ○          | ○   | ● Solution workable in low bandwidth solutions                                |
| Blind Sensing                             | ○ Fear of false positive   | ● Because of "comparative sensing" might miss low SNR solutions | ○          | ○   | ● Fear of missing avail. channels or false positives.                         |
| Cooperative sensing - Distributed         | ● Each user must still scan and detect the band                        | ● Sharing helps improve detection                               | ○          | ● Requires the cooperation of others in the network | ● Too slow and needs input of others  |
| Cooperative sensing - Centralized         | ● Time may be accelerated with help from BS                            | ● Sharing helps improve detection                               | ○          | ● Requires the cooperation of others in the network | ○ Improved time, but requires infrastructure and may be limited in freq. ops. |
| Cooperative sensing – Totally cooperative | ○ Time is gated by 2 or more CR sensing the same channel               | ● Sharing helps improve detection                               | ○          | ● Requires the cooperation of others in the network | ● Improved time, but requires infrastructure and may be limited in freq. ops. |
| PU LO Leakage Detection                   | ○ Limited to the PU bands  | ● Solution very limited to a known band                         | ○          | ● Need prior knowledge of PU.                       | ● Very limited solution   |
| Radio Identification Based Sensing        | ○ Limited to the PU bands  | ● Solution very limited to a known band                         | ●          | ● Need prior knowledge of PU.                       | ● Very limited solution   |
| Cyclostationary                           | ● Detection time slows down considerably                               | ●   | ○          | ○   | ● Better detection ability but much worse time                                |
| Network with Beacon                       | ○ Leverages beacon to detect signal, but limited to beacon freq, bands | ○ Solution very limited to a known band                         | ●          | ● Requires cooperation from Beacon                  | ● Very limited application – can help avoid interference                      |

**Table 1: Prior Work Summary**

In Table 1, “detection time” is the time required to scan the entire bandwidth, “detection ability” is the ability to correctly predict the presence or absence of a signal, “complexity” refers to the implementation complexity, “dependency” is the need for the sensing receiver to depend on another user, a base station or a master controller to perform sensing and finally the “overall performance” is summarized in the last column. It is clear that none of the previous work actually address the timely sensing requirement of CR. Besides the focus on sensing receivers and CR networks, the radio architecture needs to be addressed.

#### ***1.4.2 Wideband Agile Radio Architecture Efforts***

A wideband and agile radio architecture has eluded radio designers for decades. Multi-gigahertz designs with minimal delays have been demonstrated in military applications [119,120,122]. The design objectives of such projects are not suitable for commercial applications. In consumer electronics, receiver channelization has been commonly used to divide a broad band into narrow band channels [59,60,62,64,67,68]. An implementation is shown in Figure 1-5. In this illustrative example, the broadband spectrum is divided into “n” separate narrow band receivers. Each receiver is capable of addressing a specific frequency which reduces the stress on the components of the receiver chain.



**Figure 1-5 Channelization of a Broadband Receiver**

However, channelization adds cost, size, power consumption, and complexity. Worse yet, the receiver chains may interfere with one another and the performance degrades.

All proposed approaches add cost, complexity and/or power consumption. A neglected aspect of realizing the potential of cognitive radio is the consumer acceptance of the technology. In order for a technology to be widely accepted, it must provide economic benefits or performance improvements over existing solutions. In other words, in order for the market to accept CR as a viable solution, the radio must provide quantifiable benefit on incremental benefits it provides. These barriers of entry have killed several technologies that were once very promising.

### ***1.4.3 Summary***

While all the proposed approaches have merit, none of these techniques have focused on minimizing sensing time. As a matter of fact, several of the proposed approaches may actually increase the sensing time by adding overhead to the network. For example, in a centralized CR network, a spectrum coordinator assigns channels for short term use. The coordinator must deal with two conflicting objectives 1) maximizing spectrum use and 2) minimizing interference. Similar to other sensing solutions, this technique improves sensing but does not address the delay

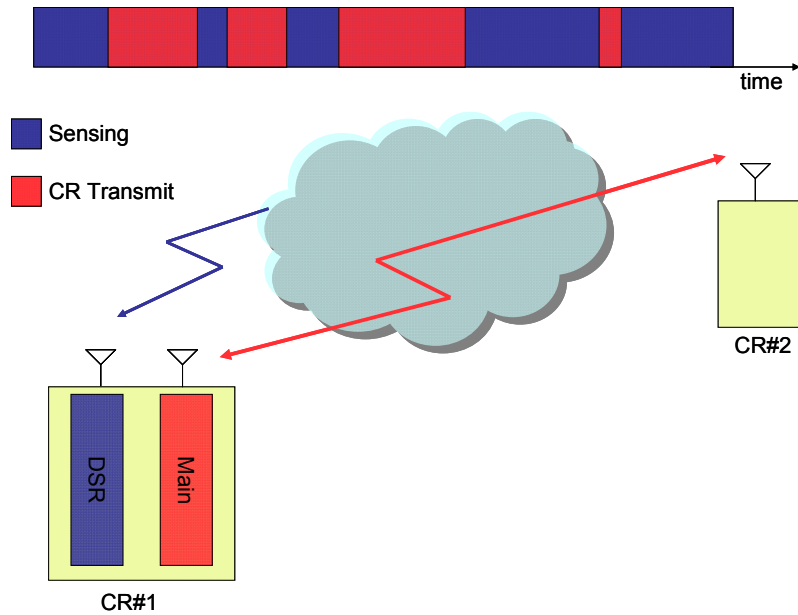
or overhead associated with cooperation. Moreover, the delays and overhead required for centralized or cooperative networks lengthen sensing time and may render the CR highly ineffective. In order to address the sensing challenge, an innovative architecture and algorithm is needed.

## **1.5 Proposed Approach**

Due to the highly dynamic spectrum activity, the sensing receiver must make fast decisions, minimize false alarms or false positives, and must be able to operate over broad frequency band. The focus of our research is a practical implementation of a sensing receiver for cognitive radio use. Our solution is divided into a radio implementation and an associated algorithm.

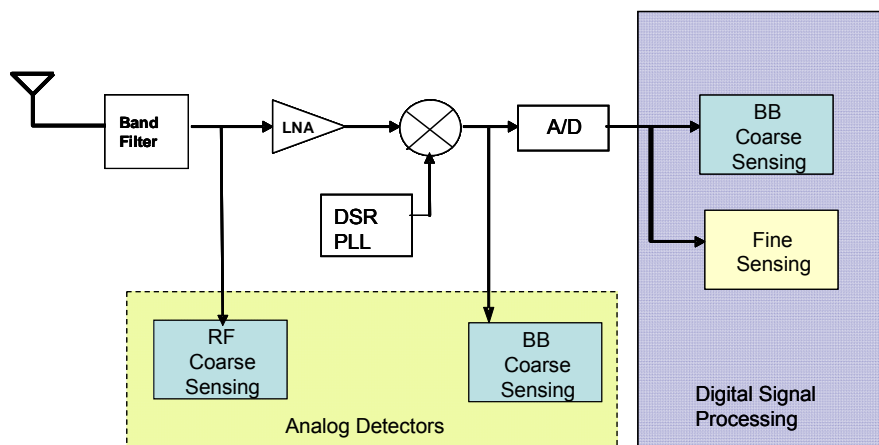
### ***1.5.1 Radio Implementation***

In time sensitive applications, the goal is to reduce the overall sensing time and to minimize in-communication delays. We propose to dedicate a receiver for sensing purposes only. As the main radio is communicating with another radio, the sensing receiver is tasked with continuously searching for available channels suitable for CR operation. Without a dedicated receiver, the main receiver must perform sensing as well as sustain a communication link which adds delays. The sensing receiver performs searches for time, frequency and/or code availability.



**Figure 1-6 Cognitive Radio with Dedicated Sensing**

In order to accelerate spectrum sensing, analog detectors are used. Analog detectors are faster relative to spectral estimation in the digital domain [29,31,68,92,122,123]. In our proposed DSR, sensing is performed in the analog domain prior to power detection in the digital domain.



**Figure 1-7 Dedicated Sensing Receiver Block Diagram**

Analog detector provides the DSR local detection capability and may be extended to add local decision making. As an example, the analog detector may be used as a blanking receiver [122]

where a transmission is immediately terminated as soon as the analog detector senses the presence of a PU in the band. The DSR decision mechanism is controlled by an adaptive algorithm.

### 1.5.2 DSR Algorithm

The algorithm controlling the radio is an integral piece of the DSR. The algorithm divides the desired spectrum into coarse and fine bins which are scanned for users. The power detection feature of the algorithm is time sensitive and varies as the sensing objectives vary.

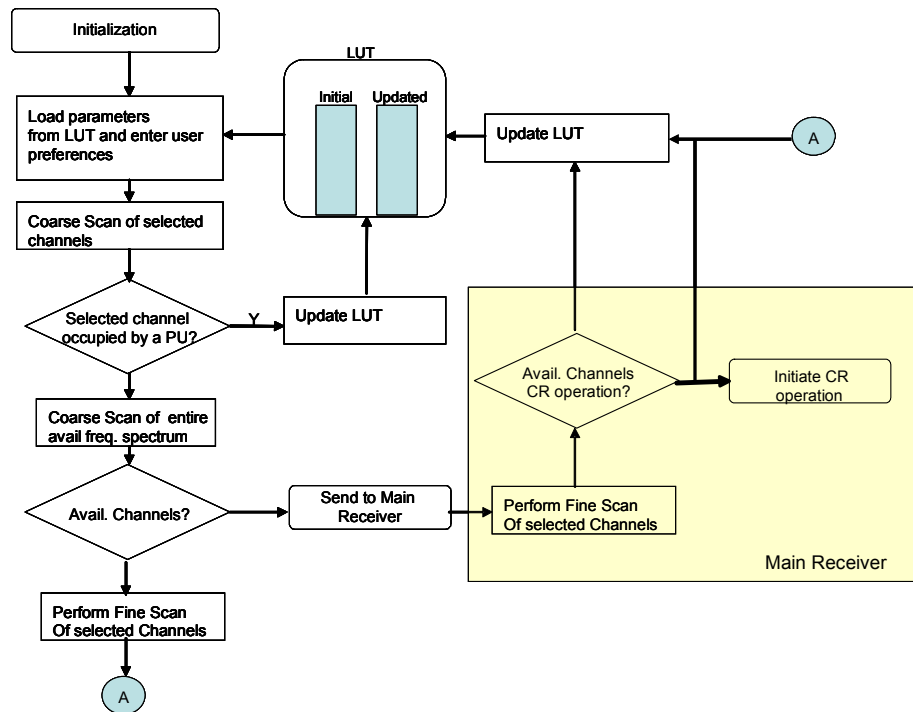


Figure 1-8 DSR Algorithm

The algorithm modifies the sensing parameters as the environment changes states. The DSR first performs a coarse sensing to determine occupancy. Each coarse bin is then divided into fine bins for fine sensing. The fine sensing is meant to provide better resolution but also requires additional processing time. The number of scanned frequencies directly affects the sensing time. If the DSR must sense 1000 fine channels, but the coarse sensing ruled out 50% of the channels, then fine sensing is only required for 500 bins. Prior to transmission, the DSR uses all of the

available resources; it uses the main receiver to perform parallel fine sensing, hence, reducing the overall time.

By applying architectural modifications and the DSR algorithm, we reveal an improvement of over 10X in initial identification of available channels and assuming that the DSR identifies more than two viable channels, the in-call delays are less than 3ms [122]. This number is significantly lower than the 20ms maximum delay provided by the rule of thumb for voice communication in section 1.3.1 above.

### ***1.5.3 DSR Algorithm Optimization***

One of the challenges of the algorithm is to identify an appropriate frequency bandwidth for the coarse and fine bins. The number of bins is directly proportional to the bin size (bin bandwidth) and it considerably affects the mean sensing time. As an example, assume that the DSR is tasked to perform channel sensing on a 1GHz band. The band is first divided into 10 coarse channels and each coarse channel is further divided into 10 fine channels (10 coarse and 100 fine sensing steps for a total of 110 sensing steps). Hence, each fine channel is 10MHz wide. If the sensing receiver detects activity within a fine bin, then the whole bin is discarded. The issue is that disregarding 10MHz of spectrum is a considerable waste if you consider that voice transmission only requires a few KHz of bandwidth (over 500 voice channels at 20KHz / channel). On the other hand, if fine channels are divided into 1MHz bands, then the DSR needs 1010 sensing steps which may take an unreasonably long time to scan. To further improve our DSR results, we optimized the solution by allowing the algorithm to find an optimal bin size for coarse and fine sensing [127][128]. The bin size changes as the environment conditions vary. The optimization process improves the DSR and the implementation cost is minimal. We derive a mathematical expression for mean sensing time of the DSR that takes account of the environment parameters and the radio implementation. The mathematical model helps determine the optimal number of coarse bins and fine bins to use in order to minimize the overall mean sensing time.

### ***1.5.4 DSR Simulation Platform***

Modeling and simulation are essential for performance evaluation of wireless systems. MATLAB and Simulink are popular tools and widely used in the industry and academia [106,107,108,109]. We developed an end to end DSR simulation platform in MATLAB and Simulink. The platform is divided into 3 parts: 1) environment simulation model, 2) the radio receiver, and 3) the DSP sensing algorithm. The platform demonstrates the adaptive application of the DSR algorithm. The simulation correctly identifies holes in the spectrum and ranks the available channels in a look up table that are appropriate for immediate radio operation. The fast response and continuous update further establish the DSR as a viable and practical solution to practical sensing for cognitive radios.

### ***1.5.5 Summary***

We demonstrate through our research [122,123,124,[127][128] that our implementation of a Dedicated Sensing Receiver (DSR) has the ability to meet the strict requirements of cognitive radio sensing in today's technology. The DSR realization requires execution at both the physical and MAC layers. Most importantly, we demonstrate that the radio is capable of supporting time sensitive applications.

## **1.6 Our Contribution to the Cognitive Radio Field**

Our key contributions to the field are summarized below:

- For the first time, we develop a practical implementation of a sensing receiver [122]. The DSR architecture investigates the use of sensing in both analog and digital domains in order to deal with time sensitive applications (see Chapter 4, Section 2).



- We derive a closed form equation that predicts the sensing time of the DSR [123]. The mathematical model takes into account 1) the radio implementation such as the PLL lock times, DSP running frequency and number of channels, and 2) the channel conditions such as SNR, available channels and scan bandwidth (See Chapter 4, Section 2).
- For the first time, we propose an adaptive algorithm that balances performance versus detection time [124]. It is critical that the algorithm minimizes false detection of available channels. At the same time, the CR should not waste an opportunity to capture an available channel. The algorithm continuously updates a look up table (LUT) with available frequencies. The algorithm is notable for the continuous update and ranking of channels in the LUT. The DSR quickly identifies available channels and begins CR operation, while the algorithm continues to test for channel availability by increasing the detection performance in the known and new channels (See Chapter 4, Section 2.1).
- We introduce the concept of adding a priori information to aid initial channel acquisition time [122]. The a priori information includes geographical location, user preference, running statistics collected over time where the radio learns what channels to use and others to avoid. The a priori algorithm reduces the set of frequencies to be searched and speeds up sensing. This technique is especially powerful during broadband scans (See Chapter 4, Section 2.2.)
- We have made a significant contribution in cognitive radio implementation at the MAC [122]. We provide MAC layer techniques that help the cognitive radio minimize interruptions and improve quality of service for time sensitive application. We propose the use of unallocated bands, use of multiple non-contiguous bands concurrently and handshaking / look ahead techniques that lower the number of interruption of service. We show that such techniques improve the performance (See Chapter 4, Section 1).
- We determine the incremental cost of implementing a DSR [123]. We design the DSR and the main receiver from which we generate a bill of material (BOM). The BOM is evaluated versus catalog prices of components. We determine that the DSR does not add

any significant cost to the cognitive radio in spite of the addition a second RF chain (See Chapter 4, Section 2.4)

- We augment the DSR algorithm with detection and false alarm probability which takes into account the statistical aspects of the digital communication [124]. We supplement the mathematical model for total sensing time with detection and false alarm probabilities in addition to a detection time penalty. We derive a closed form expression for mean detection time of the DSR (See Chapter 5, Section 1).
- A significant contribution to the field is the optimization of the DSR algorithm [127],[128]. The choice of coarse and fine sensing bandwidths has a considerable effect on the sensing time. Sensing receivers struggle with two mutually exclusive goals: 1) large sensing bandwidths which provide faster sensing time, and 2) the opportunity cost which increases as sensing bandwidth increases. The opportunity cost is defined as the CR disregarding a large portion of the spectrum without testing for available channels. We develop an adaptive algorithm that defines the optimal coarse and fine sensing bandwidths based on the channel conditions and the radio implementation. The sensing bandwidths continuously change as the channel conditions change. This is another important and significant contribution of this work since the concept of dynamic sensing bandwidths greatly improves the cognitive radio performance (See Chapter 5, Section 3).
- We showcase an end to end MATLAB and Simulink simulation platform of the DSR. Simulations are important to system designers during the design and tuning of the radio performance. Simulation is the one vehicle to demonstrate algorithm performance and how the LUT is populated. We successfully demonstrate how the sensing receiver unearths available channels and populates the LUT (See Chapter 6).

The above contributions significantly advance commercial deployment of cognitive radios.

## 1.7 Organization of the Dissertation

This dissertation is divided into seven chapters; there are six chapters in addition to this introduction.

- Chapter 2, we introduce the concept and motivation for the cognitive radio technology. We provide a summary of regulatory bodies and IEEE standards, and the evolution of software defined radio to cognitive radios. We then provide an extensive review of cognitive radio network approaches. We then examine the cognitive radio architecture by introducing the difference between classical radio architecture versus a smart radio. We then focus on the power spectral detection that is usually done in the digital domain.
- In chapter 3, we highlight the complexity of spectrum sensing. We review the requirements driving the radio architecture and the signal processing needs. We address the consumer acceptance challenges. We survey prior work related to solving the sensing problem and the shortcomings of prior research.
- In chapter 4, we introduce the Dedicated Sensing Receiver (DSR). We define the goals of the algorithm and our approach to resolving the sensing receiver. We present an architectural solution as well as the DSR algorithm. We derive a mathematical model of the DSR and we present our results and improvements directly attributed to our implementation.
- In chapter 5, we incorporate probability of detection and false alarm to the sensing time derivation. We derive a statistical equation for the mean detection time. We work to optimize the mean detection time based on optimal choice of coarse and fine bins. We then present the results of the optimized algorithm and we perform a sensitivity analysis on the mean detection time.
- In chapter 6, we showcase our DSR simulation platform. We breakdown the simulation into three parts: the environment, the receiver and the signal processing. We deep dive

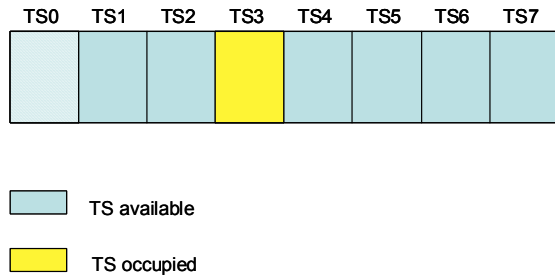
into each block and provide results from each block. We then show the overall platform and the LUT performance.

- Chapter 7 summarizes the overall dissertation. We provide a list of potential future research work.

## 2 The Cognitive Radio

In this chapter, we introduce the cognitive radio. We present the motivation behind the technology and we familiarize the reader with radio design practices. We focus on the challenge of spectrum sensing approaches and power estimation techniques.

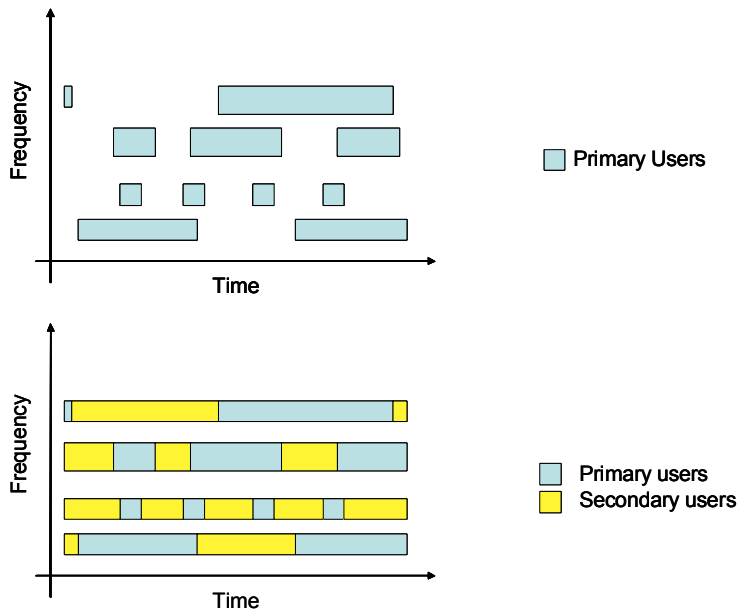
Most of today's radios are oblivious of their environment and they are usually regulated by a central management system such as a basestation controller (cellular) or a master user (BT). The centralized management makes decisions on the frequency of operation, the modulation scheme, frame structure, payloads, power levels, and even allowable power savings of the radio. The end user is usually given rigid guidelines to follow with minimal input on the decision. With such inflexible guidelines, the regulatory bodies usually assign a band of operation and they provide operation guidelines. These guidelines are usually captured in ratified standards. With high spectrum underutilization, the wireless community has been looking at new ways to increase spectral efficiencies. Multiple access schemes have been used to multiplex several users onto a single communication resource [19]. Multiple access schemes such as Frequency Division Multiple Access (FDMA), Time Division Multiple Access (TD), Code Division Multiple Access (CDMA), Space Division Multiple Access (SDMA) and Polarization Division (PD) have been heavily used in recent years. For example since the early 1900's, FDM has been used in telephony which enabled transmission of several telephone signals simultaneously [20]. Similarly, TDMA and FDMA have been used in 2G cellular such as GSM [21]. However, multiple access schemes do not fully resolve the underutilization issue. Consider the GSM cellular standard in Europe, the mobile user operates in the 890-915MHz (transmit band). The band is subdivided using FDMA into bands spaced at 200KHz for a total of 124 carrier frequencies. Each frequency is divided in 8 time slots for TDMA operation. At each frequency, a maximum of 7 users can operate simultaneously (time slot 0 or TS0 is usually used to carry command and control information). Hence, GSM allows for a maximum of 868 simultaneous voice users. Now assume that there is only one user at a given frequency.



**Figure 2-1 GSM Frame**

In the example above, only one user is present at time slot TS3. Since TS0 is used for control, there are six available time slots that are capable of carrying data that are sitting idle. Regardless of the number of users, the frequency channel is tied up. Now imagine that a secondary user (non-GSM) is able to make use of the available time slot in Figure 2-1 for data or voice communication. The secondary user must first recognize that there are available time slots and then must adapt its communication parameters such that it does not interfere with an on-going GSM transmission in TS3. Therefore, the secondary user must adopt its modulation bandwidth to match that of GSM, its power level to be within the GSM limits, and its time slots not to interfere with other time slots (i.e. TS0 and TS3). That same secondary user might have a similar opportunity to operate in a TV band, in which it must adopt the TV band requirement as not to interfere with its operations. Hence, the secondary user is a “smart user.” It is not bound by a single frequency, modulation scheme, power level or multiple access schemes. It has to be cognizant of its environment and must adapt to the environment such that its operation does not negatively impact the primary user of the band. Such smart user is commonly named Cognitive Radio (CR).

Cognitive Radios (CR) are smart radios operating as secondary users that employ Dynamic Spectrum Access (DSA) techniques. CRs are capable of adapting to user needs as well as to environment conditions. CRs are relegated to searching and finding available frequency bands or time slots in which it is able to communicate with other devices. The spectrum is fairly crowded and coexistence with other users in a chaotic and unpredictable environment is highly competitive. In order to survive, a CR senses, monitors and tracks the activities of other occupants in the band while adapting to the channel conditions. A CR is able to opportunistically operate in a band without adversely affecting primary users.



**Figure 2-2 Cognitive radios operate as secondary users and fill the unutilized spectrum**

By leveraging white spaces in Figure 2-2 Cognitive radios operate as secondary users and fill the unutilized spectrum. The new found efficiency allows for additional users at no additional cost.

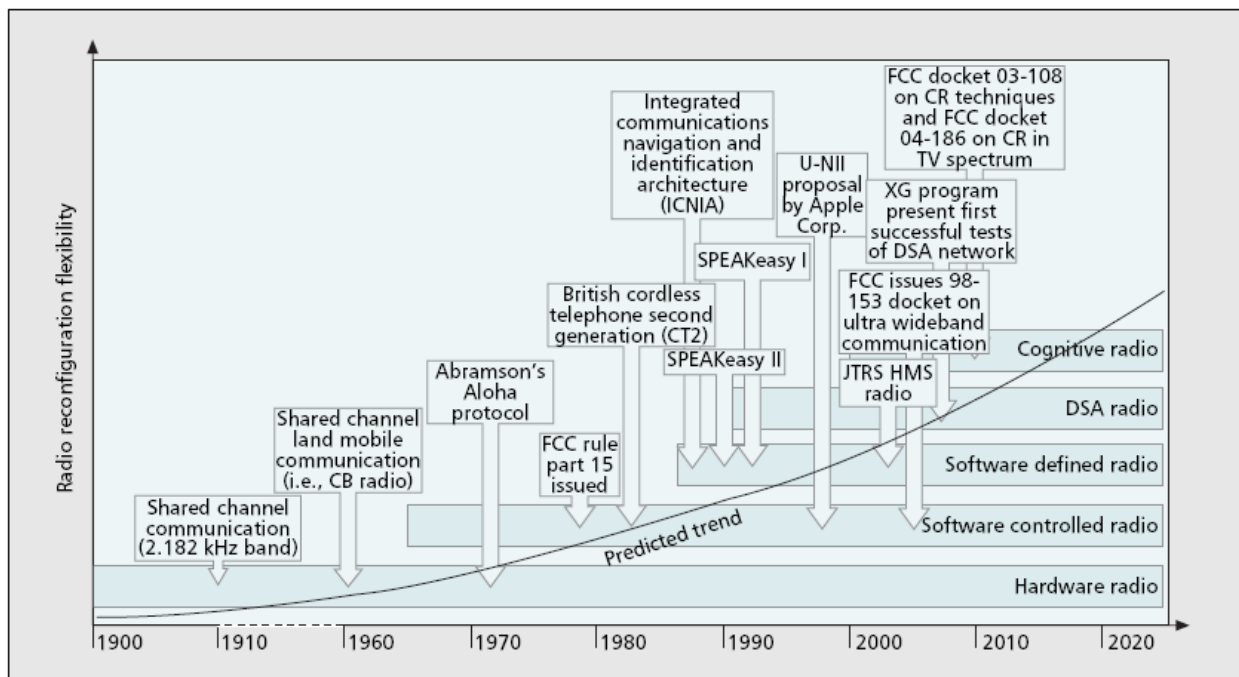
## 2.1 History and Background

Cognitive Radio was originally suggested by Mitola and Maguire [22] as an extension of the Software Defined Radio (SDR) work. Mitola and Maguire suggested that CRs extend the software radio with radio-domain and model-based reasoning. Spectrum sharing in its purest form is a superset of the SDR.

### 2.1.1 Regulatory Bodies

The first realization of the SDR was funded by DARPA in 1990. The SPEAKeasy project was later extended to a second phase as the SPEAKeasy II project in 1996. SPEAKeasy II later evolved into the Joint Tactical Radio System (JTRS). DARPA later launched the nXet Generation program (XG) that focused on “the enabling technologies and system concepts to

dynamically redistribute allocated spectrum.” After the successful conclusion of the XG project [5], the FCC realized that CR was the answer to stimulate growth of open spectrum [3]. The IEEE Communications Society and the IEEE Electromagnetic Compatibility Society setup the IEEE 1900 task force in order to address the deployment of next generation wireless. The IEEE 1900 task force later evolved into IEEE Standards Coordinating Committee (SCC) 41 Standards Committee on Next Generation Radio and Spectrum Management [27].



**Figure 2-3 Evolution of Cognitive Radio**

In addition to the regulatory body work in the US, several countries have initiated similar efforts in releasing analog broadcast TV bands in the digital switch over. In Europe, the French telecom regulator ARCEP (Autorité de Régulation des Communications Électroniques et des Postes) commissioned a study in 2008 to recommend an optimum solution to releasing available spectrum to the market [16]. Similarly, a study was performed in the United Kingdom by Ofcom (Office of Communications) in 2008 [17]. The report concluded that “... we should allow cognitive access as long as we were satisfied that it would not cause harmful interference to licensed uses...This could potentially bring substantial benefits to citizens and consumers in the



form of new devices and services.” As the world looks to release additional premium spectrum for smart radio operation, the IEEE Wireless Regional Area Network (WRAN) established a IEEE802.22 standard with the intent to standardize operation in white spaces within the television spectrum.

### ***2.1.2 IEEE802.22 Standard***

In response to the FCC proposed rulemaking for unlicensed operation in the TV broadcast bands [3,4], the IEEE initiated the IEEE802.22 working group in order to standardize a worldwide operation in the TV band. The most prominent target application of IEEE802.22 WRANs is wireless broadband access to rural and remote areas, with performance comparable to DSL and cable modem. The standard addresses the needs of both the PHY and the MAC layers [55]. The system is setup as a point to multipoint air interface, with the data traffic flowing through a basestation. The PHY layer is based on operation in 6MHz, 7MHz or 8MHz operation to leverage the existing international standards. The 802.22 system specifies spectral efficiencies in the range of 0.5 bit/(sec/Hz) up to 5 bit/(sec/Hz) which translates to 3Mbps to 30Mbps throughput in a single 6MHz channel. This throughput is shared among users and the Quality of Service (QoS) may easily be controlled by the basestation. Just like the offered services in DSL and cable modem, the basestation may guarantee a minimum data rate based on the purchased level of service. The 802.22 PHY has to provide high flexibility in terms of modulation and coding. The proposed modulation is OFDM but the PHY adapts to the environment to sustain the minimum level of service. In addition to the modulation and coding, the end user must have the ability to provide Transmission Power Control (TPC) in order to minimize interference with other users.

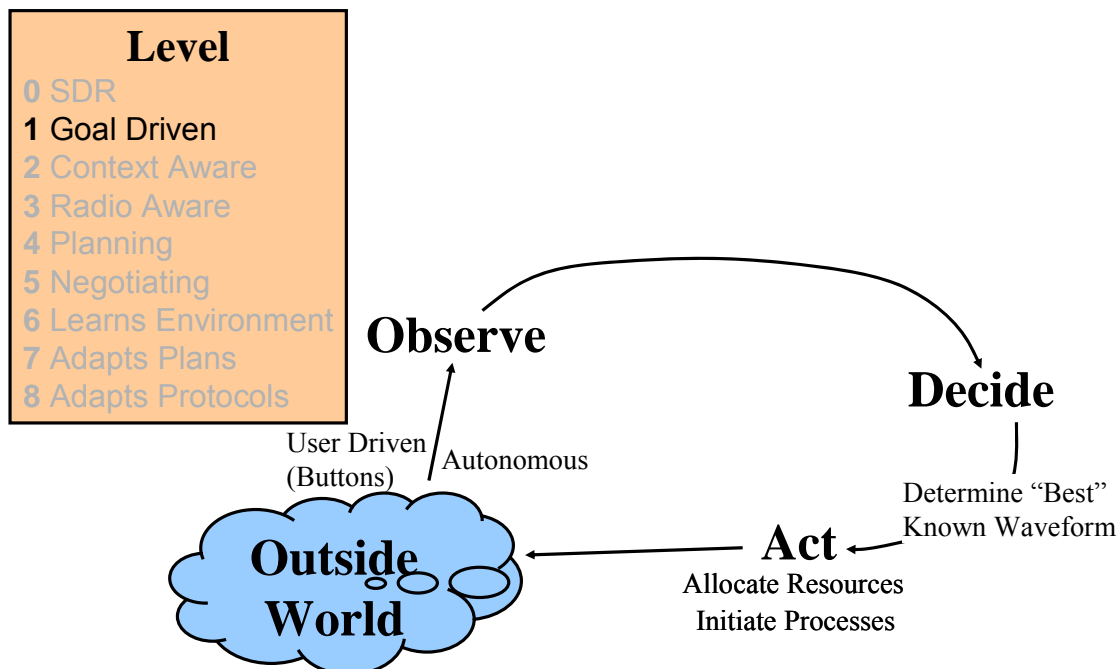
Since the MAC is based on cognitive radio technology, it adapts to the dynamic environment and the channel conditions. At initialization, the Consumer Premise Equipments (CPE) must scan in order to find a suitable basestation. Since the available bands might not be contiguous, the CPE must perform a bonding technique in which the CPE groups several bands together to improve performance. However, since the network is centralized via a master controller, the sensing and decision strain on the CPE is greatly reduced. The basestation is responsible for spectrum

management function in the system. It has the capability of providing guidance to the end user under a cooperative sensing model. The spectral management component improves the coexistence and reduces interference.

The IEEE802.22 standard was launched in 2004 and is well on its ways to becoming the first commercial application of the cognitive radio.

### 2.1.3 The Mitola Radio

Joseph Mitola III has defined 9 layers of cognitive radio [53]. At its lowest form at Level 0, the cognitive radio is a software defined radio (SDR). The SDR is a programmable platform that is programmed by the user. At Level 1, it is the simplest form of the cognitive radio; it is a goal driven CR where the radio observes, decides on the best known waveform, and acts by allocating resources and initiating the communication link. Level 1 is shown in Figure 2-4.



Adapted From Mitola, "Cognitive Radio for Flexible Mobile Multimedia Communications", IEEE Mobile Multimedia Conference, 1999, pp 3-10.

**Figure 2-4 Level 1 – Goal Driven Cognitive Radio [54]**

As the CR progresses, it adds more intelligence and more functionality. The 9 levels are shown in Figure 2-5.

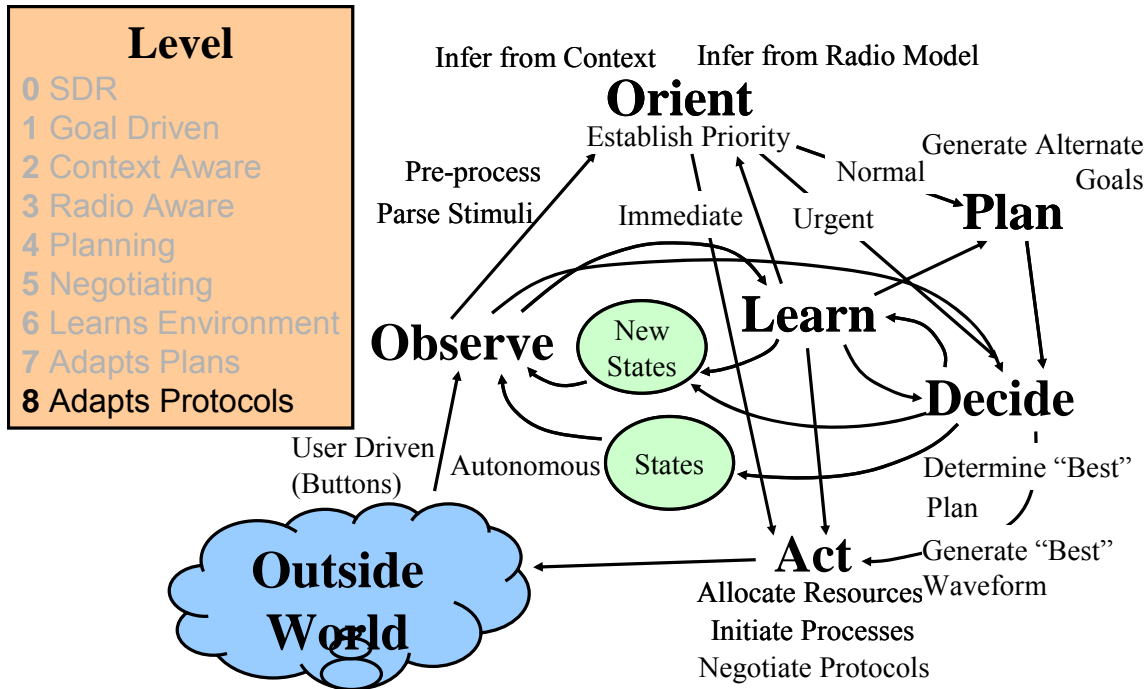
| Level | Capability            | Comments   |
|-------|-----------------------|--|
| 0     | Pre-programmed        | A software radio   |
| 1     | Goal Driven           | Chooses Waveform According to Goal. Requires Environment Awareness.                      |
| 2     | Context Awareness     | Knowledge of What the User is Trying to Do   |
| 3     | Radio Aware           | Knowledge of Radio and Network Components, Environment Models                            |
| 4     | Capable of Planning   | Analyze Situation (Level 2& 3) to Determine Goals (QoS, power), Follows Prescribed Plans |
| 5     | Conducts Negotiations | Settle on a Plan with Another Radio  |
| 6     | Learns Environment    | Autonomously Determines Structure of Environment   |
| 7     | Adapts Plans          | Generates New Goals  |
| 8     | Adapts Protocols      | Proposes and Negotiates New Protocols  |

Neil, J; “**Analysis and Design of Cognitive Radio Networks and Distributed Radio Resource Management Algorithms**” Dissertation submitted September 06, 2006.

**Figure 2-5 Cognitive Radio Level Definitions [53,54]**

As the radio evolves from Level 1 towards Level 8, its awareness of the environment increases and its adaptation becomes more sophisticated. The CR analyzes the environment, observes the users, establishes a priority, recognizes the interaction, learns of the desired goal of each interested user, proposes alternate goals then acts on allocating resources. At the level 8, the CR is at the highest intelligence level as it proposes new plans, new protocols to other CR in the immediate area.

A Level 8 adaption of the Mitola radio is shown in Figure 2-6.



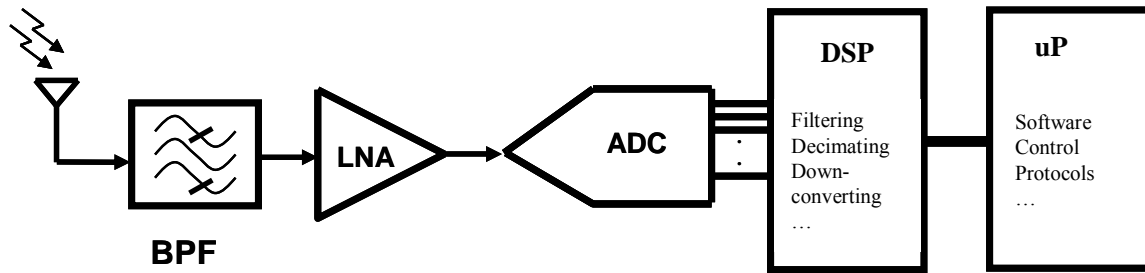
**Figure 2-6 Level 8 Cognitive Radio [54]**

At the root of cognitive radio is still a programmable radio which is commonly called a Software Defined Radio (SDR).

### ***2.1.4 Evolution of the Software Defined Radio***

As the number of wireless standards continues to increase and as consumers continue to demand additional wireless protocols in devices, an elegant and low cost solution for multi-mode radio is needed. With recent advances in large-scale integrated (LSI) circuits [68], microprocessors and Analog to Digital converters (A/D converters), one attractive solution for a multi-mode receiver is a Software Defined Radio (SDR).

An ideal SDR is shown in Figure 2-7.



**Figure 2-7 Software Defined Radio Architecture**

The basic underlying concept of Software Defined Radio (SDR) has been around for many years. As defined in the literature [28,29,30], SDR migrates the hardwired receiver to a flexible software programmable platform. A DSP may be used for tuning, signal selectivity, filtering, and down conversion. As a result, a microprocessor (or ASIC) is able to accommodate more than one wireless standard (providing flexibility). Beyond the flexibility and economic benefits, Software Defined Radio offers a wide range of far reaching benefits:

- Flexibility: SDR executes different software (in a microprocessor or a DSP code) for different protocols limiting the need to hardwire a receiver for each standard.
- Fewer components: because the received signal is sampled at RF, there are fewer external components.
- Lower cost: less components means lower material cost.
- Reduced complexity and improved reliability: the reliability of the product is improved because the number of external components are reduced.
- The complexity of the hardware is much lower on software defined radios than the classical approach.

Software radios have started the migration from research to practical implementation and have found application in cellular base stations, satellite ground stations, and other such applications in which power consumption and size are not major concerns. As of today, SDR remains impractical in battery powered handheld devices. In order for SDR to become practical, several challenges need to be overcome:

- A/D Converter limitations: The resolution of an A/D converter is a key criteria. In 3G, receivers are requiring an A/D with a spurious free dynamic range of greater than 105dB, which translates to a 17.5bits of resolution [31]. (Practically speaking, 20-bit resolution is required to account for non-linearity and implementation margin.) A 20-bit resolution in SDR handhelds leads to two problems: (1) the voltage resolution is too low. For example, a 20-bits A/D converter operating at 3V (typical cellular voltage 2.7 – 3.3V). The voltage per bit is  $3V / 2^{20} = 2.86\mu\text{V/bit}$ . This resolution is unusable as voltage, temperature and DC offset cause variations in the range of tens of milliVolts. (2) Additionally, sampling frequencies of the A/D converter must be high enough (for instance, 48 MSPS based on 8X oversampling of CDMA [32]). However, sampling rate and number of resolution bits work against one another, with high resolution A/D converters (>20 bits) running at low sampling frequencies (<10 MSPS), and high sampling frequency (> 1GSPS) employing low resolution (<10 bits). Attempts to build A/D converters with higher sampling rates and higher resolution lead to unpractical high power consumption. Assuming a sample and hold technique, an equation for power consumption of an A/D was derived by Kenington [32], and is characterized by

$$P_i = \frac{kT}{t_s} 10^{\left(\frac{6n+1.76}{10}\right)} \quad (1)$$

where  $P_i$  is the power consumed in Watts,  $k$  is Boltzman's constant ( $=1.38 \times 10^{-23}$  J/K),  $T$  is the device temperature in Kelvin,  $t_s$  is the sampling interval and  $n$  is the number of resolution bits.

From (1), the power consumption increases an order of magnitude due to implementation in software radio. Assume classical sampling theory, a 20-bit A/D converter sampling a 1GHz signal needs to run at 2GHz, consumes around 475W of power. Hence, high resolution, high sampling rate A/D converters required in battery operated software radios are not yet practical.

- Software management: An engineering challenge is the management of the different protocols without duplication and added complexity.
- Power management: Wireless standards have built-in standby or sleep modes where most of the electronic circuitry in the receiver is shut off. Minimum sections of the radio are awakened at preset time intervals to listen for a page from a base-station or sniff for local wireless devices. It is unknown how these operations may be implemented in SDR.
- Clock generation: Most of the power in a radio is consumed by clock and frequency generation circuits. Additionally, the sampling clock of the ADC should be jitter free, noise free, and spur free. In SDR, every wireless standard requires a different set of clocks such as frame reference, chip rate, and radio channel spacing. Additionally, the distribution of clocks throughout the radio must be carefully planned.
- Computational efficiency: As more and more functions are moved to the software, computational efficiencies in both the DSP and the microprocessor become critical to power savings. One solution uses Java as a development environment based on an object oriented language with proper modification for real-time operations [33]. This solution allows standardization of a common hardware platform. A Java kernel downloads and executes different operating protocols. The layered structure of the software would use (or re-use) modules, for example, coding modules can be reused between two standards. Although Java reduces the download and run time of a protocol, it is still orders of magnitude higher than needed.
- RF filtering: Enough filtering must be present in the front end (RF) to insure successful operation of the A/D converter. Narrow Filtering must be present at the receiver front end to filter out adjacent channels and blockers. One solution sweeps the filter characteristics (frequency) by sweeping the dielectric constant of the substrate [34]. Another solution uses MEMS (micro-electro mechanical structures) to switch filter elements in and out [33]. Both solutions are still in the research stages and are impractical for high volume deployment.

To date, in both industry and academia, most of the above challenges remain unsolved.

With today's technology challenges, the concept of a fully programmable SDR is not practical. However, a partially programmable radio is feasible. SDR in conjunction with DSA has evolved into cognitive radio.

## **2.2 Cognitive Radio Networks**

Cognitive radios make use of spectrum sensing capabilities to assess spectrum activity. Once availability is established with certainty, the CR chooses the transmission parameters and begins transmission. The use of Cognitive Radios (CR) is a way of enabling opportunistic sharing of the spectrum commonly called Dynamic Spectrum Access (DSA) among licensed primary users (PUs) and unlicensed secondary users (SUs). The objectives of CR are 1) to protect the primary user of the spectrum, and 2) allow for secondary users to benefit from otherwise unused frequency bands. The goal to protect primary user is upheld for several reasons:

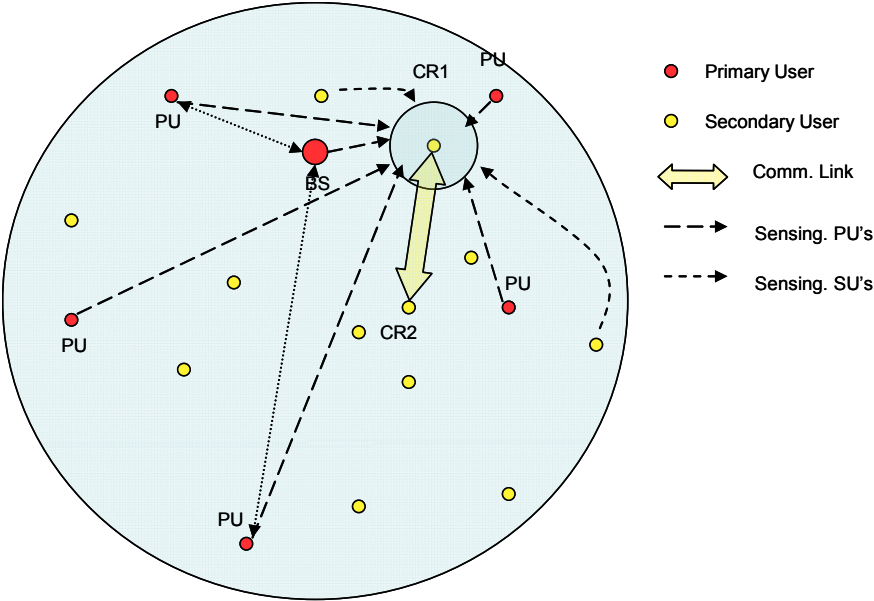
- 1- licensee paid for the band
- 2- PU is providing social and economical services that require priority such as safety or broadcast TV
- 3- Legacy product might not be able to change. Moving a channel or service out of a frequency band is costly and tedious.

The keys to effective CR operation are 1) wideband and fast frequency hopping capability, 2) reliable spectrum sensing and reliable prediction of the presence of PUs and 3) ability to use spectrum with minimal interference to other users. In general, a CR may find an available or underutilized spectrum and begin operations. However, the CR must cease operation or lower power levels such that it does not interfere with the PU or in most cases, it interrupts data transmission, and hops to a new channel.

In an example in Figure 2-8, a CR network assumes that all radios operate on the same frequency channel; CR1 is initiating a communication link with CR2. CR1 performs a channel sensing in which activities of PUs and SUs are recorded. CR1 must adapt to the PU environment while it uses spectrum sharing etiquette to avoid interference with other SUs. Spectrum sharing etiquette

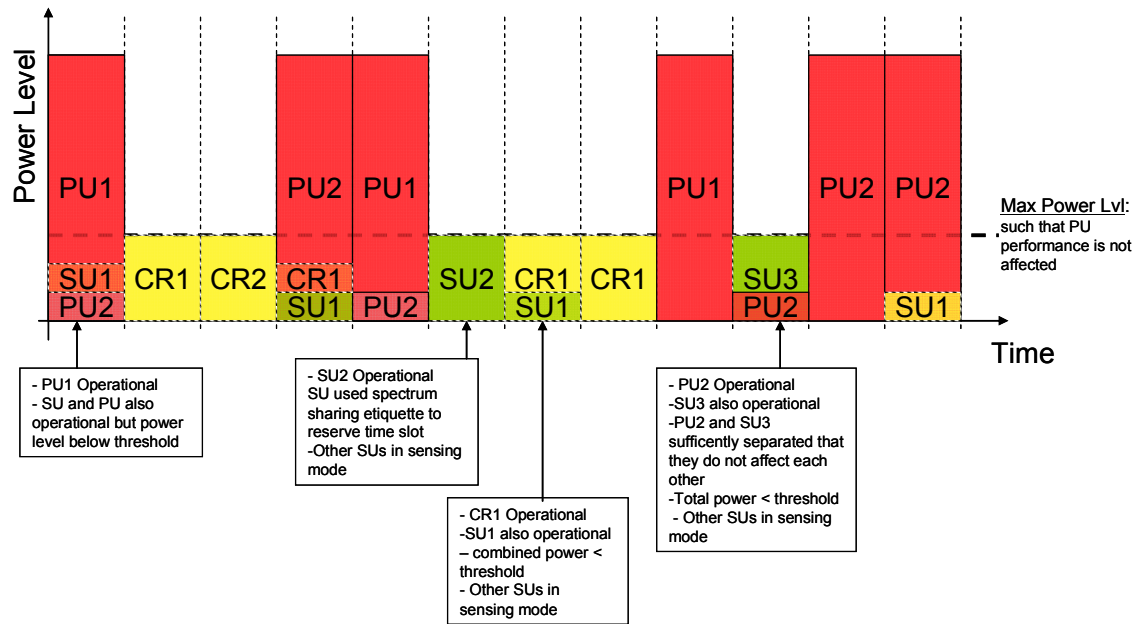


[35] improves cooperation and reduces interferences among SUs. Once CR1 deems it safe to transmit, it communicates the control parameters to CR2. Control parameters include transmission parameters such as modulation type, modulation parameters, power level and time slots.



**Figure 2-8 CR Network**

Since the radios are operating at the same frequency, a multiple access scheme is needed to exploit available resources. In this case, assume that time division multiplexing is used.



**Figure 2-9 CR Operation over Time**

Figure 2-9 shows a timing example of how a CR and other SUs and PUs may coexist in a cognitive network. In this example, all devices are operating on the same frequency; hence all SU devices must continuously be in spectrum sensing mode. In the first time slot, we have three devices operational at the same time; we show PU1 with the highest power level and dominating the airwaves, but PU2 and SU1 are also operational. This approach is called controlled transmission or interference temperature approach [39]. One possible reason is that SU1 and PU2 are sufficiently far away from PU1 such that their combined power does not affect PU1. PU1 reception may operate in the presence of other users up to the point where other users adversely affect performance. Users in the band are seen as noise at the P1 receiver. Therefore, as long as the PU1 receiver does not degrade below a set threshold they can still operate in the background. While the devices are operational, other SUs must continue to sense and collect information about the environment and other tenants in the system.

Another option shown in Figure 2-9 is interference avoidance. This example is shown in slot 2, where CR1 is operating in time slots where no other occupants are visible.

There are two approaches to a cognitive network design: cooperative or competitive

### ***2.2.1.1 Competitive Network***

Just as the name suggests, in a competitive network, every CR fends for itself. Each CR must perform spectrum sensing, and must compete with other users for a frequency or time slot.

### ***2.2.1.2 Cooperative Network***

In a cooperative network, users share information among each other to allow for improved overall performance [44]. Cooperation can increase performance of the overall network. However, diverting resources for cooperation may degrade individual performance and may add additional delay and overhead. Hence, one of the challenges of cooperative networks is to provide incentives for all users to participate. If users are not willing to join the coalition, they are treated as noise or jammers within the network. Cooperation may occur at all levels: between PUs and SUs or between SUs and other SUs. Users may also cooperate on power level where each user agrees to transmit up to a self imposed  $P_{\max}$  [45].

The question remains how the SUs operate in an orderly manner without harming other users when a spectrum resource is made available. The solution is part of the cognitive radio network design.

## ***2.2.2 Cognitive Radio Networks Approaches***

Cognitive radios hold a promise to increase spectral utilization with acceptable degradation if any, in the licensed and unlicensed bands. The license holder buys the exclusive rights to the spectral band. However, more often than not, the spectral utilization is in the low teens. Cognitive radio network provides guidance in choosing the transmission parameters such that the interference with other users is minimal. The cognitive radio uses its ability to adapt to the environment and chooses its parameters to meet its needs. For instance, assume that the CR receiver is far from the CR transmitter. The CR pair must agree on 1) power level, 2) modulation scheme, and 3) transmission schedule. As the physical distance between the transmitter and the receiver increases, the transmission link can be maintained via 1) additional transmission power, 2) increase in effective signal power due to additional gain at antenna such as directional

antenna, multiple antenna or beam steering techniques, 3) choice of robust modulation scheme better suited for low signal to noise ration (SNR) such as OFDM or BPSK, 4) use of signal processing techniques such error correction, and 5) collaboration with other CRs or a controller within the network. The receiver must also contend with classical wireless issues such as fading, interference, multipath effect, and shadowing [37].

With the central goal of increasing channel capacity, the CR network uses several techniques to increase spectral capacity [36].

#### ***2.2.2.1 Gap Filling Approach (White Space Filling)***

This is the most common and most obvious approach. The CR transmission occurs during the spatial, temporal or spectral voids. This interference avoidance is predicated on the ability of the CR to accurately and quickly sense available gaps in transmission [38]. This approach was explained in the example earlier in the section on Cognitive Radio .

#### ***2.2.2.2 Simultaneous Controlled Transmission***

In Figure 2-9, we demonstrated an example where PUs, SUs and CRs may operate simultaneously. In this case, the CR must adjust its transmit power level such that the PU receiver is able to operate with acceptable interference or acceptable noise temperature from other occupants [39].

#### ***2.2.2.3 Opportunistic Interference Cancellation***

This approach assumes that the CR has prior knowledge of the PU link [40]. Assume that the CR is operating in a GSM band. The CR listens to the basestation command and control information and decides when and where to transmit in the band. If a CR has the ability to demodulate and decode the information contained in the TS0 slot in a GSM frame [21], it would possess the ability to predict where and when a GSM users would transmit. Hence, the CR is able to use this knowledge to avoid interference. Another example is if the PU is a CDMA user. In this case, the CR would have the ability to use one of the available orthogonal spreading codes so that the interference is minimized [40].

#### ***2.2.2.4 Cognitive Transmission***

This approach adds additional intelligence to the cooperation in a cognitive network. Cognitive transmission assumes cooperation between the PU and the CR transmitter [41]. It assumes that the CR knows the transmission message and is able to provide orthogonal messages in essence cancelling the effect at the PU receiver. The advantage of this technique is that the CR is able to transmit at full power with real effect on the PU receiver.

#### ***2.2.2.5 Network with Beacon***

This approach is an extension of the interference avoidance described earlier. The PU transmits a beacon before each transmission. The beacon signal warns cognitive users of an upcoming PU transmission [42]. This early warning minimizes the probability of interference.

#### ***2.2.2.6 Network with Primary Exclusive Regions***

This approach is more suitable in a broadcast environment where one PU transmitter is communicating with multiple receivers. The network imposes exclusive regions near the PU transmitter that forbid CR transmission [43].

#### ***2.2.2.7 Single or Multi- Hop Network***

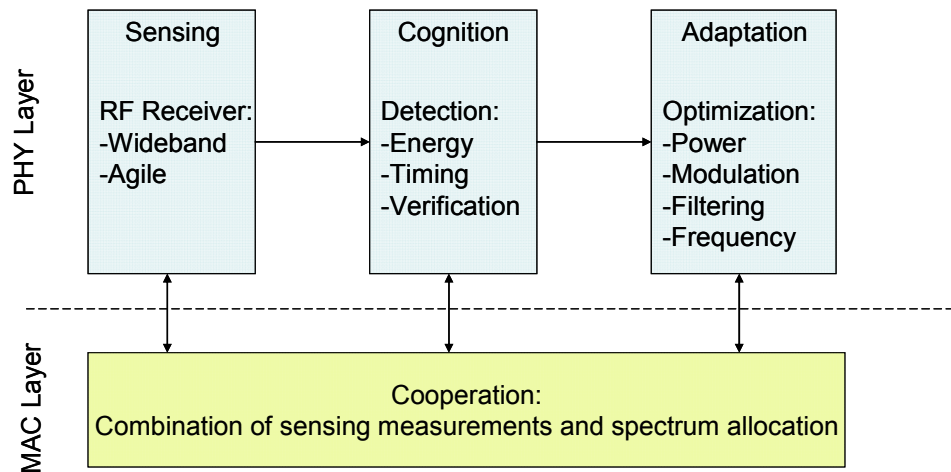
This is a cooperative model where the CRs agree to communicate with a nearby receiver with self imposed transmit power. With such limit on the transmit signal, the distance between the transmitter and receiver is limited [45]. In multi-hop networks, there is a stronger cooperation among users. As an example, the network allows nearby receivers in a coalition to strip off stronger transmissions.

It is hard to rank the above network approaches. Each network has advantages and disadvantages based on different environment parameters such as the intended use, geographical location, legacy products, and regulatory restrictions. The right answer for a CR network in downtown Tokyo might be different than that of downtown Manhattan, KS. Regardless of the network type, the CR device must perform a very basic function: the ability to learn and adapt to its environment.

## 2.3 Cognitive Radio Architecture

One of the key elements to Cognitive Radio is its ability to learn and adapt to the environment. Non-smart radios are usually constrained by a master controller. In cellular networks, the frequencies are hardwired during the frequency planning stages and the basestations provide rights of transmissions. In BT, a device becomes a master and directs traffic to avoid interference. In theory, the CR is frequency agnostic and is free to choose the link parameters as long as the device operates below a given threshold. Based on its sensing algorithm, the CR has the flexibility to choose [46] link parameters that affect bandwidth use, spectrum efficiency, bit error rate, probability of dropped call, throughput, goodput (as opposed to throughput), power consumption, and system delay / computational complexity. For example, a CR may choose the frequency, pulse shape, symbol rate, modulation, frame length, power level, power guard bands and timing of the transmission.

The sensing function is distributed across the physical and the MAC layers. Garbic et al. [47] proposed a cross layer functionality shown in Figure 2-10.



**Figure 2-10 Cross Layer Functionalities of Spectrum Sensing**

### 2.3.1 Sensing: RF Receiver Design

In past decades, several standards have entered the wireless arena such as WCDMA, GSM, CDMA, Bluetooth, IEEE802.11, IEEE802.16 etc. Several of these standards must coexist in a competitive environment. Furthermore, Radio Frequency (RF) design engineers have been hard pressed to include multi-band, multi-mode radios in a small form factor battery operated device. The standards operate at different frequencies, bandwidths, modulation and performance requirements. The RF receiver performs frequency conversion by using mixers, phase locked loops (PLL), amplifiers, and filters.

#### 2.3.1.1 Classical Radio Architectures

The transmitter function of a radio is to modulate a signal. Modulation is a process in which one or more parameters of the carrier frequency are varied according to the baseband information. The frequency band is then shifted to a suitable region within the spectrum where the electrical signals are converted to an Electro Magnetic (EM) wave by the antenna and then transmitted in space. Conversely, a receiver captures a signal off the airwaves and converts it down to either an intermediate frequency (IF) or to baseband (Frequency 0Hz). The receiver filters out unwanted signals (blockers) that are within or around the band of operation of the receiver.

The two widely used techniques in RF receivers are heterodyne and homodyne receivers which is commonly called a direct conversion receiver [7]. A classical super heterodyne receiver is shown in Figure 2-11 and the direct conversion receiver is shown in Figure 2-12.

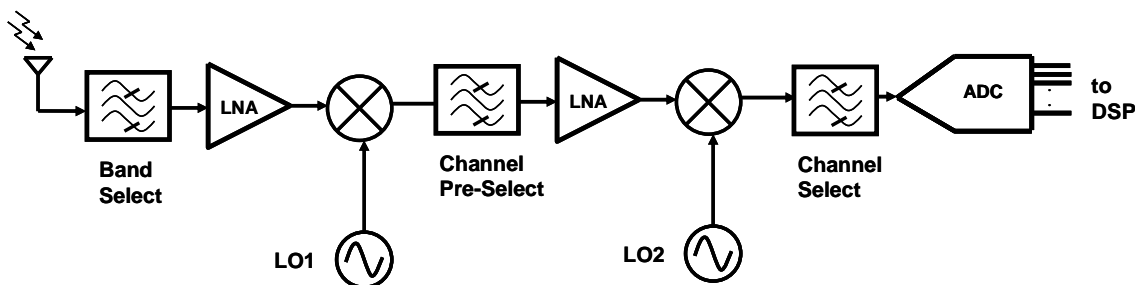
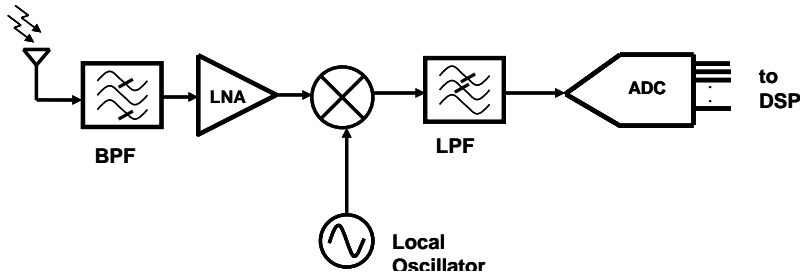


Figure 2-11 Super Heterodyne Receiver



**Figure 2-12 Direct Conversion Receiver**

Referring to Figure 2-11, the antenna receives a broad spectrum, which is immediately filtered by the front-end filter before entering the first stage of amplification. The Low Noise Amplifier (LNA) amplifies the received signal (the desired as well as the undesired signals). Furthermore, the LNA adds noise in-band. The first mixer converts the signal to the first intermediate frequency (IF). At the IF, the signal is amplified and filtered further. A second and final stage of mixing is then used to convert the IF to a baseband signal (0 Hz) where further amplification and/or filtering may take place prior to sampling the signal. The digitized received signal is then sent to the modem for processing.

Referring to Figure 2-12, the direct conversion receiver operates in a similar manner as the super heterodyne receiver in Figure 2-11 but the downconversion to baseband is done in a single stage. The local oscillator (LO) is set to the same frequency as the desired receiver signal. Since  $F_{LO} = F_{RF}$ , then the mixer output is given by  $DC_{out}$ :

$$F_{LO} = F_{RF} = F_{mix} = \frac{\omega_{mix}}{2\pi} \quad (2)$$

$$DC_{out}(t) = \frac{1}{2} A_{LO} \cdot A_{RF} \cdot [\sin\{(\omega_{mix} - \omega_{mix})t + \phi\} + \sin\{(\omega_{mix} + \omega_{mix})t + \phi\}] \quad (3)$$

where  $A_{LO}$  is the amplitude the LO signal,  $A_{RF}$  is the amplitude the received signal and  $\phi$  is the phase difference between the LO and the received signal. When  $\phi=0$ , then equation (3) simplifies into equation (4):



$$DC_{out}(t) = \frac{1}{2} A_{LO} \cdot A_{RF} \quad (4)$$

Hence, when  $F_{LO} = F_{RF}$  there are two components at the output of the mixer: one component is frequency independent (freq. = 0, i.e. baseband) and the other component is operating at  $2 \times LO$  (2 times the local oscillator frequency). After a low pass filter is applied at the output of the mixer, the  $2 \times LO$  component is filtered out and only the baseband component remains. The advantages of direct conversion are cost and low complexity. The disadvantages of direct conversion receivers are LO Self-Mixing, DC offsets, LO leakage and higher linearity requirements.

### **2.3.1.2 Receiver Figures of Merit**

The receiver figure of merit defines the performance of a receiver. Among those measures, the sensitivity and the selectivity are the primary figures of merit.

#### **2.3.1.2.1 Receiver Sensitivity**

The sensitivity describes the smallest signal level that a receiver can acquire and demodulate within a desired Signal to Noise Ratio (SNR) or Carrier to Noise Ratio (CNR). The SNR or CNR is usually dictated by the modulation type and the receiver Bit Error Requirement (BER). The received signal strength is directly proportional to the distance of the receiver from the transmitter source. The signal strength decreases by a factor inversely proportional to the distance from the transmitter squared ( $1/r^2$ ) in theory, where  $r$  represents the distance from the transmission source. The estimated free-space path loss is given by (5).

$$L \approx 20 \cdot \log(f) + 20 \cdot \log(d) + 32.44 \text{ (dB)} \quad (5)$$

where  $L$  is defined as the path loss in dB,  $f$  is the frequency of operation in MHz and  $d$  is the distance from the source in Km.

The free space loss in (5) is dependent on both the distance and the frequency of operation. For example, if we have a cellular phone operating at 2GHz and is located 10Km away from the basestation, the free space loss is 118.46dB, other losses such as rain, imperfection in antenna

apertures, multipath, shadowing of the transmitter from the receiver will add to the losses. Thus, it is critical to improve the sensitivity of the radio in order not to adversely affect the performance of the wireless device. In cellular phones for example, the sensitivity determines how far a user moves away from the basestation before the call is drop.

Besides path losses described above, the receiver sensitivity is also limited by the noise generated within the receiver. There are several types of noise generators: thermal, shot, flicker, impulse solar, and galactic radiation. The thermal and shot noises are the most prominent.

The thermal noise usually determines the noise floor of the receiver. The thermal noise is generated by the Brownian motion of the electrons in a conductor that would cause instantaneous voltage fluctuations. Because of the high velocity of the electrons and the large number of charge carrier present, the noise extends to a very high frequency. The spectral density of the noise is approximated to be:

$$S_n(f) \approx \frac{kT}{2} \text{ W/Hz} \quad (6)$$

where k is Boltzman's constant ( $1.38 \times 10^{-23}$  J/K) and T is the absolute temperature in Kelvin (T=290K at room temperature).

The spectral density in equation (6) does not contain any frequency dependence. Hence, the noise is uniform across a very large bandwidth and hence its name "white noise." The thermal noise power present in a given bandwidth B (Hz) is given by (7).

$$P_n = kTB \text{ W} \quad (7)$$

At room temperature, the temperature is T=290K,  $P_n = (1.38 \times 10^{-23})(290)B$  in Watts or  $P_n \text{ (dB)} = -174 + 10 \log B$  (in dBm).

Another type of noise present in the system is called "shot noise." Active electronic components such as amplifiers, oscillators and active filters generate noise. The noise is due to irregularities in the current flows across the active devices. Unlike thermal noise, shot noise is not dependent on temperature but it is proportional to the square root of the current.

As stated earlier, a receiver adds noise above the thermal noise floor. The noise figure of the receiver determines its sensitivity to demodulate signals. The noise figure is defined as the signal to noise ratio (SNR) at the input of the receiver divided by the SNR at the output as shown in (8).

$$F = \frac{SNR_{in}}{SNR_{out}} \quad (8)$$

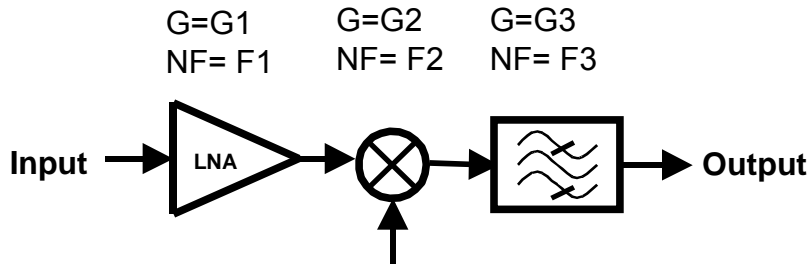
Combining equations (7) and (8), the noise factor is show in (9).

$$F = \frac{P_n - kT_s B_n}{kT_0 B_n} + 1 = 1 + \frac{T_s}{T_0} \quad (9)$$

where  $T_0 = 290\text{K}$  and  $T_s$  is defined as the effective noise temperature of a circuit or a system. The noise factor is generally provided in dB and is called the noise figure.

$$NF = 10 \cdot \log(F) \text{ in dB} \quad (10)$$

The noise figure defines how much additional noise the receiver would add to the received noise density at the antenna which is the thermal noise floor at  $-174 \text{ dBm/Hz}$ . A typical receiver is composed of cascaded blocks such as filters, amplifiers, and mixers. Each block has an associated gain and noise figure. Luckily, the noise figures in a cascaded chain do not add linearly. The noise contribution of the follow-on stages is attenuated by the gain (cascaded) of the previous stages. Assume we have a cascaded system as shown in Figure 2-13.



**Figure 2-13 Cascaded System**

where  $G_1$  represents the gain of the first stage,  $NF_1$  represents the noise factor of stage 1. The cascaded gain ( $G_{sys}$ ) and the cascaded noise figure ( $F_{sys}$ ) of the equivalent network are shown respectively in (5) and (6).

$$G_{sys} = G_1 \cdot G_2 \cdot G_3 \quad (11)$$

$$F_{sys} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} \quad (12)$$

The main contributor to the cascaded noise figure is the noise factor of the first stage. The contribution of the third stage is diminished by the gain of the first and the second stage. Once the noise figure of a cascaded system has been calculated, the sensitivity of the receiver can also be derived. As an example, assume that we have a receiver with a noise figure of 10dB. The signal has a 10MHz bandwidth, and the modulation is QPSK with a required minimum probability of bit error to be less than  $10^{-5}$ . Assuming static conditions, the sensitivity of the receiver is calculated in Figure 2-14 below.

|   |                    |
|---|--------------------|
| Noise Floor                                 | = -174 dBm / Hz    |
| Noise Figure of Receiver                    | = 10 dB            |
| Signal BW ( $10 \cdot \log(10\text{MHz})$ ) | = 70 dBHz          |
| SNR ( $P_e < 10^{-5}$ , QPSK)               | = 10dB (no fading) |

$$\text{Sensitivity} = -174 + 70 + 10 + 10 = -84 \text{ dBm}$$

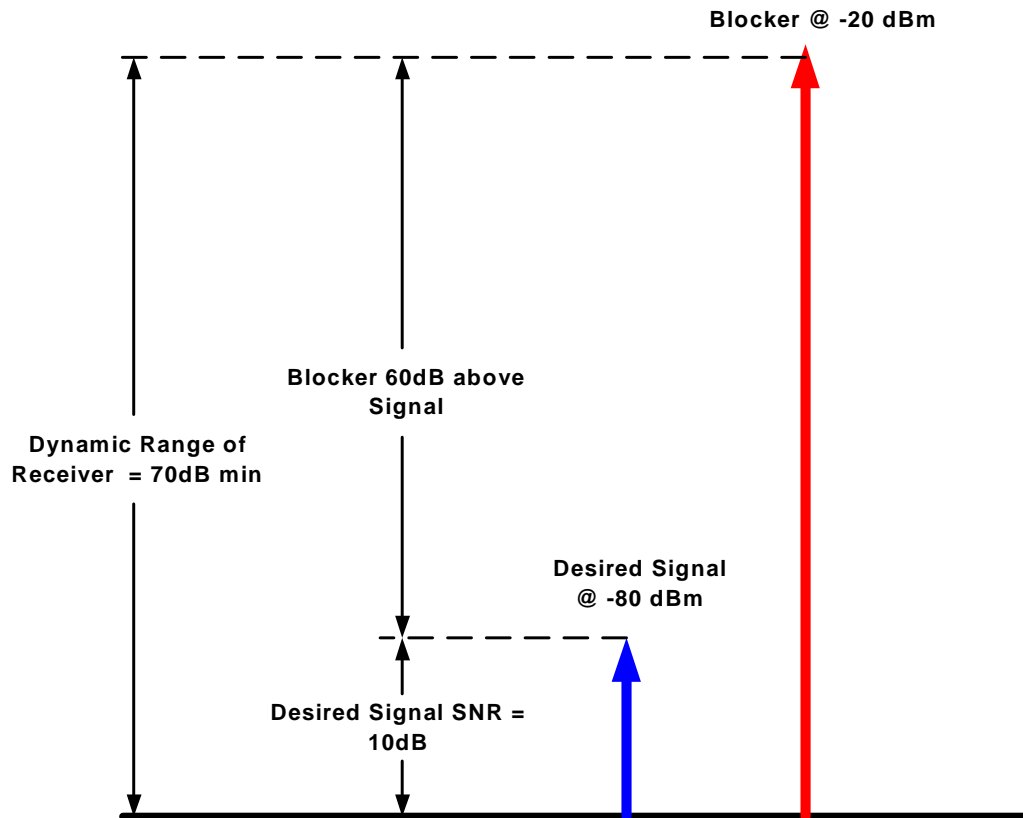
**Figure 2-14 Receiver Sensitivity (QPSK Modulation)**

Thus, the lowest signal the receiver can demodulate with a  $P_e < 10^{-5}$  bit error is  $-84\text{dBm}$ .

Since the major noise contributor to a system is dictated by the first stage, it is advisable to use a Low Noise Amplifier (LNA) with the lowest possible noise figure and as much gain as possible to reduce the noise contribution of the upcoming stages. However, more gain in the front stage increases distortion in the amplifier that is commonly measured by the linearity requirements of the latter stages of the receiver.

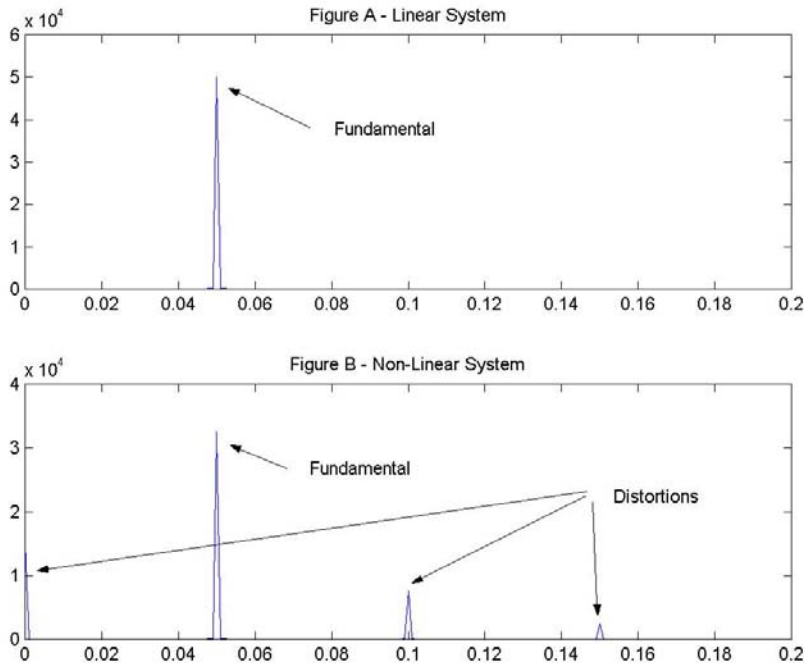
#### ***2.3.1.2.2 Receiver Selectivity and Linearity***

As described earlier, the airwaves are becoming increasingly crowded. Undesired signals are received at the antenna and unless filtered, they interfere with the desired signal and degrade the performance of the receiver. The selectivity of a receiver determines the rejection level of the undesired signals or the robustness of the receiver under these conditions. The blockers limit the dynamic range of the receiver, which determines the range of amplitudes that the radio can accept and assimilate. For example, if a radio is required to receive signals at levels ranging between  $-100\text{ dBm}$  and  $-20\text{ dBm}$  then the dynamic range of the radio must be at least  $80\text{dB}$  ( $-20 - (-100) = +80\text{ dB}$ ) as shown in Figure 2-15. The adverse effects of the blockers depend on the radio architecture.



**Figure 2-15 Effect of Blockers on the Receiver Dynamic Range**

The linearity of the receiver is an important factor in a radio design. Receiver non-linearity causes distortion of the received signal which degrades the signal quality and sensitivity of the receiver. As an example, a simulation was run in MATLAB to show the effect of non-linearity in a system. In Figure 2-16, Figure A shows the output of the system when the receiver is operating in the linear region while Figure B shows the effect of the non-linearity in the system.



**Figure 2-16 Effect of Non-Linearity in a System**

Note that the output power of the fundamental signal in the non-linear system in Figure B is lower than that in Figure A. This reduction is due to the fact that the total power is distributed between the fundamental and the harmonics. The power levels within the fundamental signal and the harmonics vary depending on the level of non-linearity; as the system becomes more non-linear, power is transferred from the fundamental (desirable) to the harmonics (undesirable).

In order to lower distortion, the device must operate in the linear region of the device. In other words, the higher the signal at the input of the system, the higher the probability that the device will operate in a non-linear mode.

### **2.3.1.2.3 The Noise Figure versus Linearity Compromise**

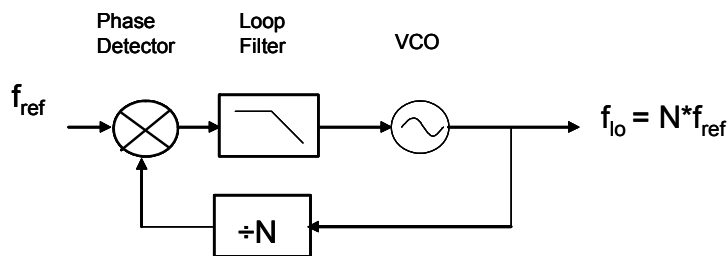
As shown above, in order to lower the equivalent noise figure of the system (i.e. increase sensitivity), the gain of the first stage should be as high as possible because the noise figure of the subsequent stages are reduced by the gain of the stage. On the other hand, if the amplification of

the first stage is high, then a large signal saturates the second stage which may cause non-linearities. Obviously, the second stage would rather see as small of a signal as possible from the previous stage in order to maintain the linearity of the system. In other words, if we cascade blocks in a system, noise figure requires that the gain is shifted to the earlier stage, while linearity requires that the gain is shifted to the later stages. RF system engineers continuously compromise system noise figure for system linearity. This compromise depends on the implementation complexity.

### 2.3.1.3 Frequency Generation and Phased Locked Loops

Frequency generation in a receiver is usually accomplished via a Phased Locked Loop (PLL) locked to a master reference clock. A PLL is essential to modern receivers as it generates a frequency used as a local oscillator (LO). The LO is then mixed with the received signal (RF) to generate an intermediate frequency (IF) or a baseband version of the RF signal (Figure 2-11 and Figure 2-12).

The critical PLL parameters that affect the receiver performance (besides center frequency and power consumption) are switching time, phase noise, and spurs (also called reference sideband). While phase noise and spurs are directly proportional to the loop bandwidth, switching time is inversely proportional [8]. A typical PLL is shown in Figure 2-17.



**Figure 2-17 Phase Locked Loop Block Diagram**

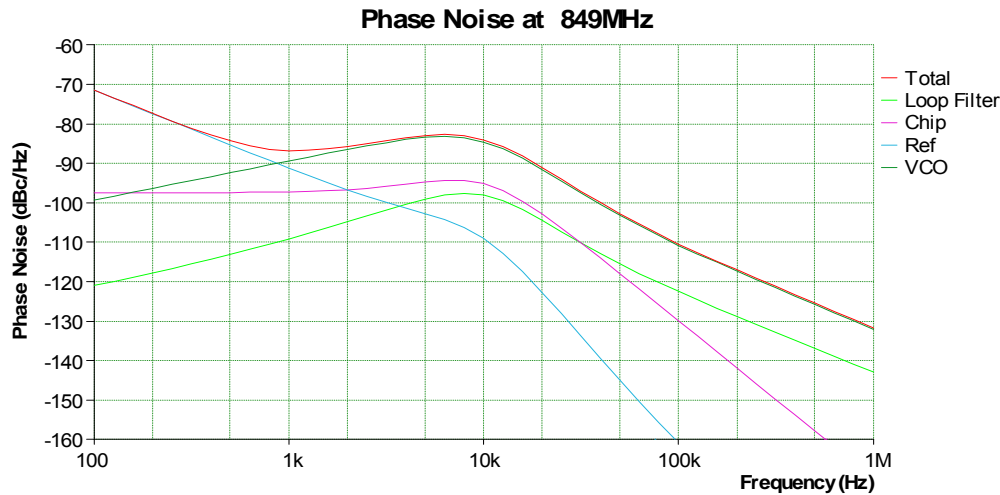
The Voltage Controlled Oscillator (VCO) is locked to a multiplied reference frequency  $N \cdot f_{ref}$  where  $N$  is the multiplication factor. The closed loop tracks any voltage variation within the loop bandwidth and the phase noise is therefore equivalent to  $20 \cdot \log(N)$  of the reference oscillator ( $f_{ref}$ ) phase noise. Outside of the loop bandwidth, the free running VCO phase noise is dominant. A PLL



design example is shown in Figure 2-18. In this example, an Analog Devices PLL chip part number ADF4360-3 was used in the simulation of a PLL operating at 850MHz. The total phase noise is dominated by the reference inside the loop bandwidth and tracks the VCO phase noise outside the loop bandwidth. Hence, in order to reduce the phase noise (integral under the “Total” curve in Figure 2-18), the loop bandwidth must be reduced. However, the loop bandwidth is inversely proportional to the lock time. As the loop band decreases, the lock time increases [9]. The normalized lock time of a PLL is given by  $\Delta\tau_i$  in (13) below.

$$\Delta\tau_i \propto \frac{1}{\psi}, \quad \psi = \frac{\omega_n}{(K_{vco} \cdot K_d)} \quad (13)$$

where  $\omega_n$  is the loop natural frequency,  $K_{vco}$  is the VCO gain and  $K_d$  is the phase detector gain. From (13), the lock time is reduced as the loop bandwidth is increased which works against the PLL phase noise.



**Figure 2-18 Phase Noise Contributor in a PLL using Analog Devices part number ADF4360-3 with a built-in VCO.**

The PLL phase noise is critical to the receiver as it proportionally degrades the performance of the received signal. Hence, as the phase noise increases, the sensitivity of the receiver decreases.

### 2.3.2 Detection

The goal of the detection block as shown in Figure 2-10 is spectrum sensing. The objective is to make a decision quickly and with accuracy on whether a signal is present at a given frequency. After the received spectrum has been demodulated and digitized, the power detection is performed in the digital domain via a Digital Signal Processing or DSP. In practical implementation, spectrum sensing consumes power and introduces delays. Sensing the spectrum with sufficient dynamic range remains a challenge. Luckily, there exists several spectrum estimation techniques heavily used in communication today and they are well documented in the literature [48,49].

For a wide sense stationary random process, the power spectral density (PSD) is the Fast Fourier Transform (FFT) of the autocorrelation function:

$$S_{xx}(f) = \int_{-\infty}^{+\infty} r_{xx}(\tau) e^{-j2\pi f\tau} d\tau \quad (14)$$

where,

$$r_{xx}(\tau) = E[x(t)x^*(t-\tau)] \quad (15)$$

Two of the classical techniques using FFT for PSD estimation are 1) Periodogram Spectral Estimator or PSE, and 2) Blackman-Tukey Spectral Estimator or BTSE. Marple [48] outlined the advantages of the two techniques:

- Computationally efficient
- PSD estimate directly proportional to the power for the sinusoid processes

The disadvantages are:

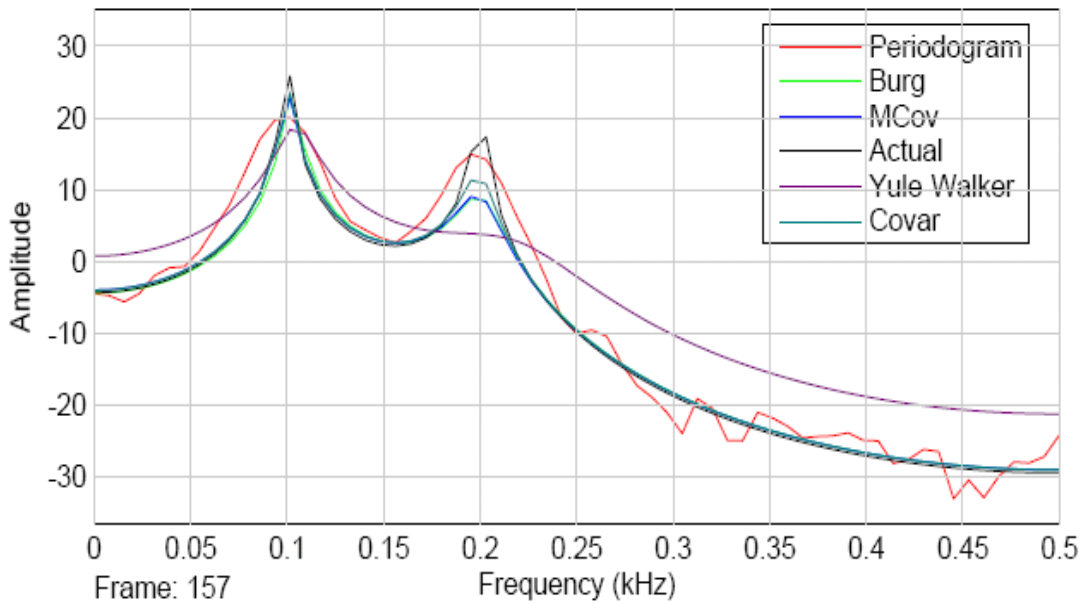
- Frequency resolution is limited by the availability of data
- Introduction of distortion due to sidelobe leakage
- Need for ensemble averaging (periodogram)

In addition to the PSE and the BTSE, there are other parametric techniques that can be used to estimate power such as Burg, Yule-Walker, Multitaper Method, and Modified Covariance. The performance of each technique has been compared and summarized in Table 2.

| <b>Technique</b>  | <b>Resolution</b>                             | <b>Complexity</b> | <b>Comments</b>                    |
|-------------------|---|-------------------|------------------------------------|
| Periodogram       | Good  | Low               | Need for averaging                 |
| Blackman-Tukey    | Good  | Low               | High Sidelobe Leakage              |
| Burg              | High  | Medium            | Produces a stable model            |
| Covariance        | Good  | Low               | May produce unstable models        |
| Yule Walker       | Low for short records<br>Good for long record | Medium            | Always produces a stable model     |
| Minimum Variance  | Good  | High              | Not a good candidate for CR        |
| Multitaper Method | High  | High              | High Complexity solution           |
| Filter Bank       | High  | Low to Medium     | Mid to high realization complexity |

**Table 2 Comparison of Spectrum Estimation Techniques**

A MATLAB simulation was performed in order to compare different techniques. For the purposes of the simulation, the estimation order is set to 4 and the FFT length is 128. The comparison is shown in Figure 2-19.



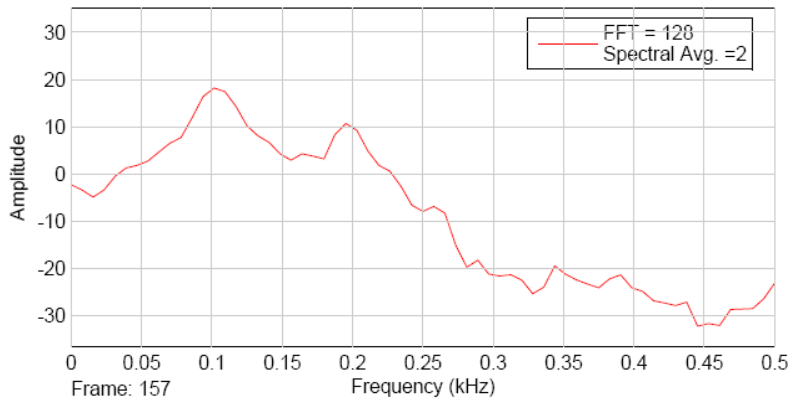
**Figure 2-19 Spectrum Estimation Techniques**

From Figure 2-19, the minimum variance technique provides the best resolution. However, minimum variance has the highest complexity and computational delays. In a sensing receiver, it is critical to make an assessment of the environment quickly as conditions change. It is important not to miss an opportunity or worst yet, interfere with another user because the information is stale.

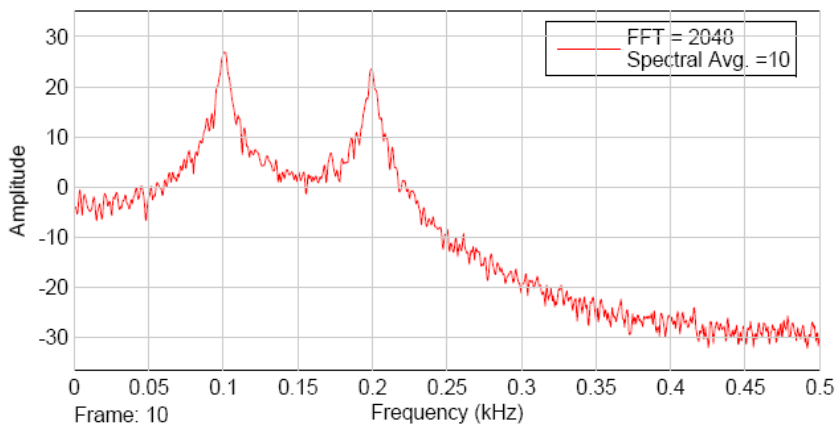
For the purpose of the CR, a PSD solution is suitable. It provides fair to good resolution while the complexity is low.

### **2.3.2.1 PSD and FFT Computation Complexity**

Like any FFT calculation, the higher the number of points, the better the resolution. Comparing Figure 2-20 and Figure 2-21, it is very clear that the resolution in Figure 2-21 is superior.



**Figure 2-20**  $2^N$ -point FFT with  $N=7$  and Avg. = 2



**Figure 2-21**  $2^N$ -point FFT with  $N=11$  and Avg. = 10

However, as the number of points increases, the processing time also increases. In practical implementations, FFTs have widely used the split-radix FFT algorithm [50, 51]. The split-radix algorithm is commonly used in DSP in order to reduce implementation of operations to be performed by the DSP. Duhamel [50] derived a reduced number of additions and multiplications per FFT operation. The fundamental idea behind the split-radix algorithm is the parallelization of the computation. The work is split into different work flows. Different decompositions can be used for different parts of the algorithm. In fact, the different parts are computed independently and use the best possible computational approach. The net of the split-radix algorithm is that the number of real additions and multiplications needed for a  $2^N$  points FFT (with  $N > 1$ ) is given by C.

$$C = 4 \cdot N \cdot \log_2 N - 6 \cdot N + 8 \quad (16)$$

Assuming we use a DSP that performs one addition and one multiplication per cycle, the total time to perform a discrete Fourier Transform (DFT) is given by:

$$T_{DFT} = \frac{1}{F_{DSP}} (4N \log_2 N - 6N + 8) \quad (17)$$

where,  $F_{DSP}$  is the clock frequency of the DSP.

From (17), it is clear that as the number of FFT points increases, the time to perform a DFT also increases. However, the DFT calculation time is inversely proportional to the DSP clock frequency.

### ***2.3.3 Adaptation***

Once the presence or absence of the user in band is determined, the CR adapts to the channel conditions. The CR must perform the following tasks:

- 1- Compare spectral density with threshold
- 2- Determine the CR network environment. If cooperative, it shares the information with appropriate users or controllers
- 3- Compare the bandwidth of the white space
- 4- Compare the time slot of the white space
- 5- Determine channel conditions
- 6- Determine the receiver distance
- 7- Alert the receiver. In some cases, a pilot tone is used to sync up to the receiver
- 8- Determine the appropriate modulation and coding schemes for the link
- 9- Determine the required filtering and pulse shaping
- 10- Determine the power level
- 11- Initiate communication

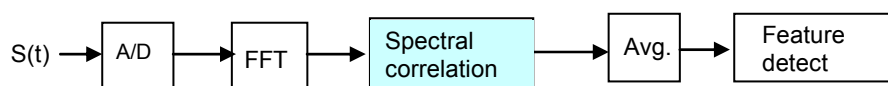
The CR must operate swiftly to complete its adaptation to the channel conditions and to the white space. Obviously, there are several scenarios depending on the network configuration and channel requirements.

### 2.3.3.1 Cyclostationary Feature Sensing

In some cases, power spectral density is not enough. Additional information about the received signal may be required (recall the discussion around the opportunistic interference cancellation approach discussed earlier). In such case, the CR must have the capability to sense and classify channel availability and link quality with lowest number of false positives.

The burden is on the CR not to interfere with primary users of the spectrum. Energy detection is an effective method of determining if energy is present within the measurement bandwidth. The measured energy does not differentiate between a modulated signal or high noise level. Additional signal processing is required to better discriminate between a modulated signal and noise. One approach is to better assess and classify the energy in-band and to take advantage of the cyclostationary nature of modulated signals. The detectors are called *Cyclostationary Feature Detectors*.

Cyclostationary feature detectors take advantage of the fact that modulated signals are generally riding on a carrier with built-in periodicity. The resulting modulated signal is classified as cyclostationary and hence its random processes characteristics also display periodicity. The cyclostationary detector performs a spectral correlation within the measurement bandwidth [52]. It consists of an augmentation of classical energy detection with a spectral correlation process. The block diagram of a cyclostationary feature detector is shown in Figure 2-22.



**Figure 2-22 Cyclostationary Feature Detector**

The spectral correlation provides a different signature for each modulation such as signal modulation type and information bandwidth. The advantage of the feature detector is that it

provides more information about the received signal. However, it adds delay and additional processing to the sensing operation.

## 2.4 Summary

Opportunistic use of underutilized and oversubscribed frequency spectrum makes Cognitive Radio an ideal wireless technology. CRs are smart radios that adapt to the environment and are able to operate with acceptable interference with other users. At the physical layer, the CR has three distinct functions:

- 1- Spectrum Sensing
- 2- Detection
- 3- Adaptation

At the core of the cognitive radio is a sensing receiver that is responsible in assessing channel conditions, presence of users, and sharing of the information with other users if required. The addition of spectrum sensing function and the built-in intelligence differentiates a radio from a “smart” radio. After the spectrum is downconverted and digitized, the detection of the signal occurs in baseband. The most common form of power spectral detection is achieved via Fast Fourier Transform (FFT). FFTs have been heavily used over the past half century and are well understood. However, FFT processing time is proportional to the number of points, and hence the CR must strike a delicate balance between resolution and processing time.

By making use of the sensing information, the CR adapts to channel conditions. Hence, a software programmable radio is essential to the CR operation. The CR must choose the link parameters such as operation frequency, modulation type, spectral shaping, bit timing, power level. At the highest level of cognition, the CR would adapt and propose a protocol level.

The CR technology, although exciting, has yet to be practical. One key detractor of practically implementing a CR is the challenge in designing a practical sensing receiver. The sensing receiver ability to scan and sense available channel and maintain the ability to minimize



interference is still unpractical. The delays and overhead costs associated with the sensing receiver are rendering the CR technology unusable in its basic form. Additional enhancements are required.

### 3 Spectrum Sensing Challenges

In this chapter we review the challenges of the cognitive radio. More specifically, we focus at the heart of the challenge that is the spectrum sensing function. We present methods proposed to address spectrum sensing by the research and the engineering communities. We then highlight the limitations and shortfalls of the proposed solutions.

The difference between a classical receiver and a cognitive radio is the ability of a radio to adapt to its local environment. In order to adapt, the radio must recognize the state of the channel. Cognition is achieved via a sensing receiver. The sensing receiver scans the spectrum and identifies opportunities for transmission. However, the spectrum is highly dynamic and hence the decision on availability must be made almost instantaneously. In order to optimize its chances to find an available channel, the sensing receiver must scan several frequencies. The broader the scanned spectrum, the higher the probability of finding a viable channel suitable for CR operation.

The sensing receiver is challenged at three different levels:

- 1- radio design challenges
- 2- spectral detection and digital signal processing challenges
- 3- “consumer acceptance” challenges

The radio implementation decisions have a great effect on the sensing receiver. First, a broadband receiver architecture adds noise and spurs within the chain which degrades the sensitivity of the receiver. Second, the PLL introduces delays in the system. The delays hurt the sensing since the environment is fast changing, and by the time the PLL tunes to a channel and onto the next channel, the environment might have changed thus rendering the results invalid. Radio engineers have been grappling for decades with the challenge of designing an agile and broadband radio.

The digital signal processing also introduces delays. The challenge remains in identifying a practical signal processing technique such that delays are minimized and acceptable. This challenge becomes more difficult if we consider real time operation of the CR.

One of the most neglected aspects of cognitive radio implementation is consumer acceptance. In order for the technology to be widely accepted, it must provide economic benefits or performance improvements over existing technology. In other words, in order for the market to accept CR as a viable solution, the radio or the service must cost less than today's technology. In addition, the new technology must provide benefits such as reduced power consumption or applications that users desire. These barriers of entry have killed several technologies that were once very promising.

The research community identifies several solutions to the sensing challenges. As we review the different techniques, we determine that the solutions are usually optimized for a specific application and might not necessarily address the broader market. We identify the benefits and the disadvantages of each solution.

### **3.1 Sensing Receiver challenges**

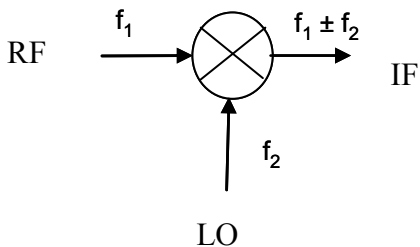
Building a broadband and agile sensing receiver has proven to be a difficult task. The task becomes even more intimidating when one considers that the receiver must be highly reliable in order to avoid interference. The sensing challenges include RF receiver design, spectral sensing in the DSP, and practical implementation restrictions that reduce consumer acceptance of the product.

#### ***3.1.1 RF Receiver Design Challenges***

In section 2.3.1.1, we highlighted two common radio architectures: the super heterodyne receiver and the direct conversion receiver. Both architectures are implemented by mixers in which the RF frequency is translated to an intermediate frequency (IF) or baseband before sampling and baseband processing.

### 3.1.1.1 Mixers and Spurious Output

The mixer is a non-linear device that performs a multiplication function in the analog domain. A mixer is used in the receiver to downconvert (i.e. translate the frequency to lower frequencies or to baseband – 0Hz ). In a receiver mode, the radio frequency (RF) signal received from the antenna is “mixed” or multiplied by a frequency generated by the radio called the local oscillator (LO) and the output signal is called the intermediate frequency (IF) [56]. An ideal mixer is shown in Figure 3-1.



**Figure 3-1 Ideal Mixer Model**

Assume that the RF frequency is  $f_1$  and the LO frequency is  $f_2$ . In an ideal mixer, the output of the mixer is:

$$IF = f_1 \pm f_2 \quad (18)$$

Recall that the mixer performs a multiplication of two sinusoidal signals. Assume that the RF and the LO are represented in (19).

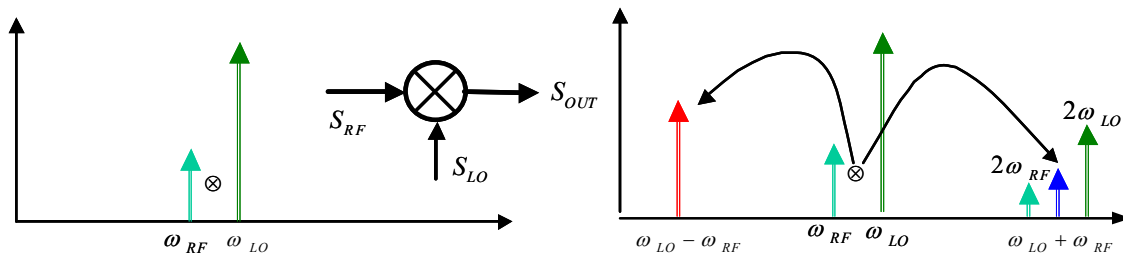
$$RF = A_{RF} \cdot \cos(\omega_{RF}t) \text{ and } LO = A_{LO} \cdot \cos(\omega_{LO}t) \quad (19)$$

The output of an ideal mixer is shown in (20).

$$IF = [A_{RF} \cdot \cos(\omega_{RF}t)] [A_{LO} \cdot \cos(\omega_{LO}t)] = \frac{A_{RF} \cdot A_{LO}}{2} [\cos(\omega_{RF} - \omega_{LO})t + \cos(\omega_{RF} + \omega_{LO})t] \quad (20)$$

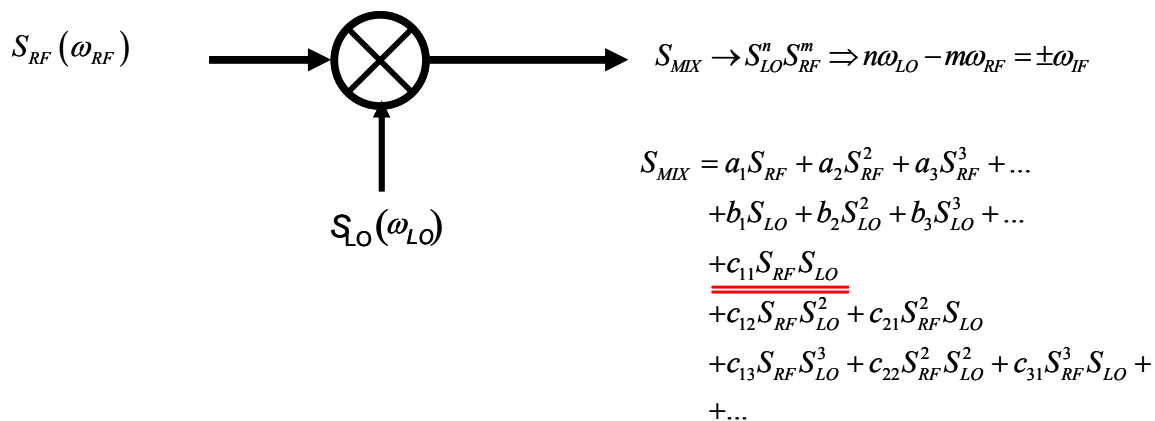
However, mixers do not only multiply the fundamental frequencies of RF and LO, but also all of the harmonics components.

Graphically in the frequency domain, a mixer performs a frequency translation as shown in Figure 3-1.



**Figure 3-2 Mixer Function**

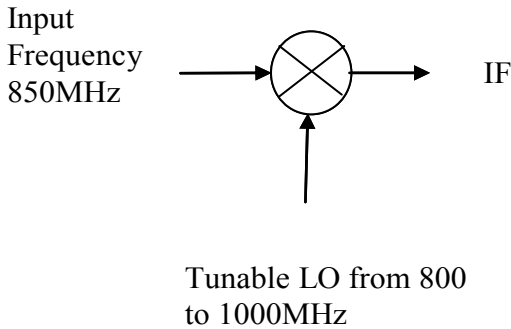
Mathematically, a mixer performs a multiplication of two sinusoidal signals.



**Figure 3-3 Mathematical Model of a Mixer**

where  $n$  and  $m$  are integer. From Figure 3-3, it is obvious there are several spurious (undesired) outputs at the IF besides the desired signal. These spurs reduce the effectiveness of the receiver.

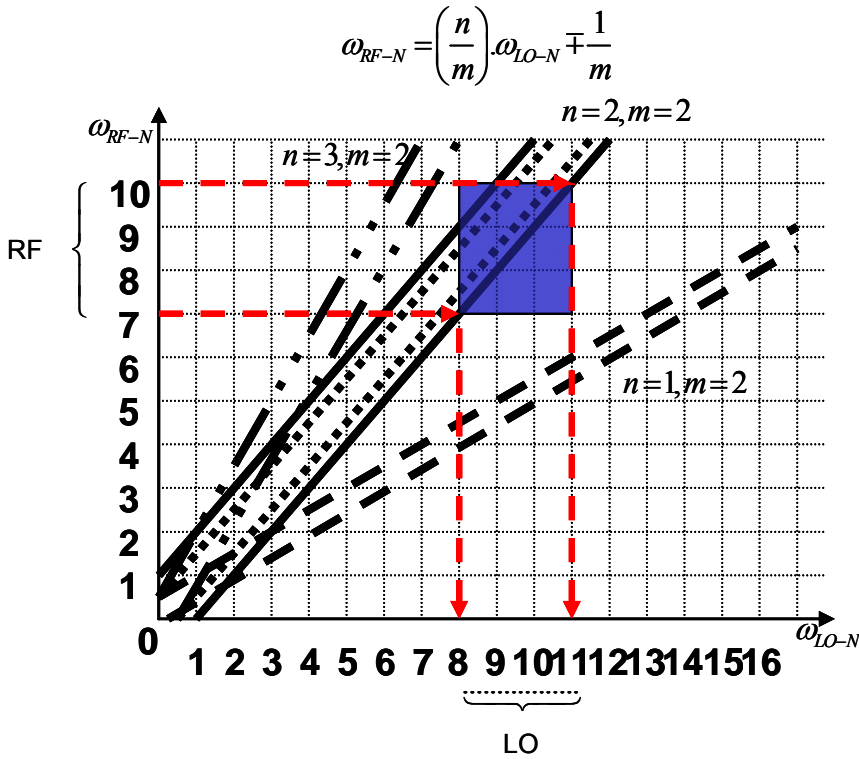
Let's use the example in Figure 3-4 to better describe the number of spurious outputs at the IF port.



**Figure 3-4 - Mixing Example**

In this example, the RF received frequency is 850MHz. However, the front end of the mixer sees the entire band and is usually limited by the use of a front end bandpass filter. For the purpose of this example, let's assume that the frequency at the input varies from 800MHz to 1000MHz.

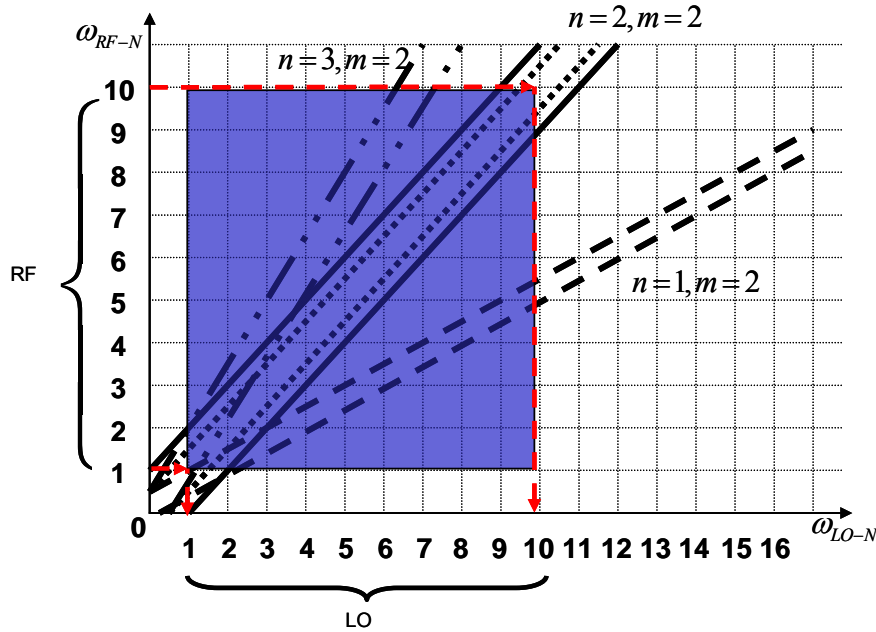
In order to display the IF output, we first normalize the LO and the RF (divide by 100) and plot the mixer output as shown in Figure 3-5 [57,58 ].



**Figure 3-5 Narrowband Operation – Mixer Output Spur Chart (up to 3<sup>rd</sup> harmonic)**

Each line in Figure 3-5 represents a term of the mixer output. The blue box indicated the desired frequency. However, it shows that the  $m=2$  and the  $n=2$  term is present at IF. This spurious term harms the received signal and decreases the received sensitivity. The degradation is usually a function of the signal and the spur levels. Although in this example, we have only considered the 3<sup>rd</sup> harmonic for each signal, there are spurs beyond the 3<sup>rd</sup> harmonics that fall in the IF band that do interfere with the desired signal. For the purposes of this example, assume that all frequencies outside the blue box in Figure 3-5 can be filtered out.

Now assume that we want to design a broadband receiver. The receiver covers the 100MHz to 1000MHz. As the RF signal increases, the LO signal must also track the receiver. Hence, the LO must also cover 900MHz of frequency bandwidth. Keeping the same assumption as the previous example, we only consider up to the 3<sup>rd</sup> harmonics. The resulting spurious response is shown in Figure 3-6.



**Figure 3-6 Broadband Operation – Mixer Output Spur Chart (up to 3<sup>rd</sup> harmonic)**

As expected, the blue blocks representing the IF output has now expanded. As the mixer expanded to cover more frequency band, additional spurs are now in band and can not be filtered out. This example demonstrates the challenge of broadband receivers design.

RF designers choose the LO and IF frequencies such that there are no high order spurs in-band. This exercise is called frequency planning and is a key element of the design.

### 3.1.1.2 PLL Design Limitations

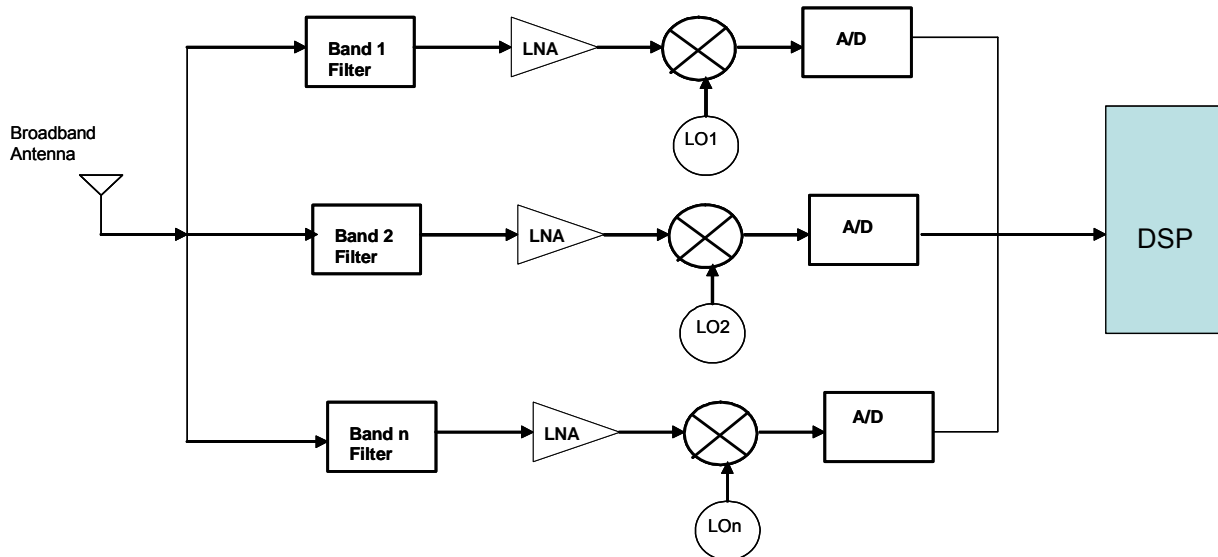
As described in section 2.3.1.3, the PLL is a key element of the receiver design. In Figure 2-17, the phase detector determines the phase difference between the reference frequency and the feedback clock. The phase detector generates a control signal that is integrated by the loop filter before being fed as a control voltage into the VCO. The control voltage increases or decreases the frequency of the oscillator in proportion to magnitude of the error.

As described in the section 3.1.1.1, the receiver requires a local oscillator that is usually generated by a PLL. In case of a broadband sensing receiver, a broadband PLL is required. The challenges of a broadband are the phase noise, lock time and spurious sideband.



The demand for broadband operation of the VCO requires a higher VCO tuning sensitivity named  $K_v$ . As  $K_v$  increases, the phase noise degrades and unwanted frequency deviation due to voltage ripple in the system or noise on the control line, manifests itself as spurious components or increased phase noise.

As an example, let's assume that the PLL is operating in a battery operated device. The available voltage reference is 2.7V. Since the total available voltage is 2.7V, as a rule of thumb [58], the VCO control voltage should be designed between 1V and 2V in order to account for circuit and temperature variations. Hence, if a VCO is designed to operate in the low cellular frequency band, it requires that the VCO tunes 25MHz over 1V control voltage (or  $K_v = 25\text{MHz/V}$ ). Hence, a 10mV offset on the control line will shift the VCO frequency by 250Khz. Now consider that in the same system, the frequency band of interest is 10GHz. The VCO tuning sensitivity  $K_v$  is now 10GHz/V. Using the 10mV voltage offset, the VCO now moves by 100MHz. Hence, the feasibility of a low phase noise over broadband operation is simply not feasible nor practical. In recent past, several techniques were developed to address this problem. A common solution to wideband receiver is channelization of the receiver [67]. Effectively, the wideband receiver is transformed into a cluster of narrow band receiver shown in Figure 3-7.



**Figure 3-7 Channelization of a Broadband Receiver**

Broadband receiver channelization solves the performance problem but it increases cost and power consumption of the overall solution. If the entire receiver is not channelized, other techniques are focused on the PLL and more specifically on the VCO. Medi et al.[59] demonstrated a low power fully integrated frequency synthesizer that is capable of operating from 4GHz to 7GHz. The design was implemented in 0.25 $\mu$ m CMOS via a bank of PLL's each operating in a 1GHz frequency band. Again, the additional switching adds to delays and cost. Rokita [60] experimentally proved that replacing a broadband VCO with a bank of narrowband oscillators improves performance. Although we see considerable improvement, a brute force approach to sweeping the broad frequency bandwidth remains a major challenge.

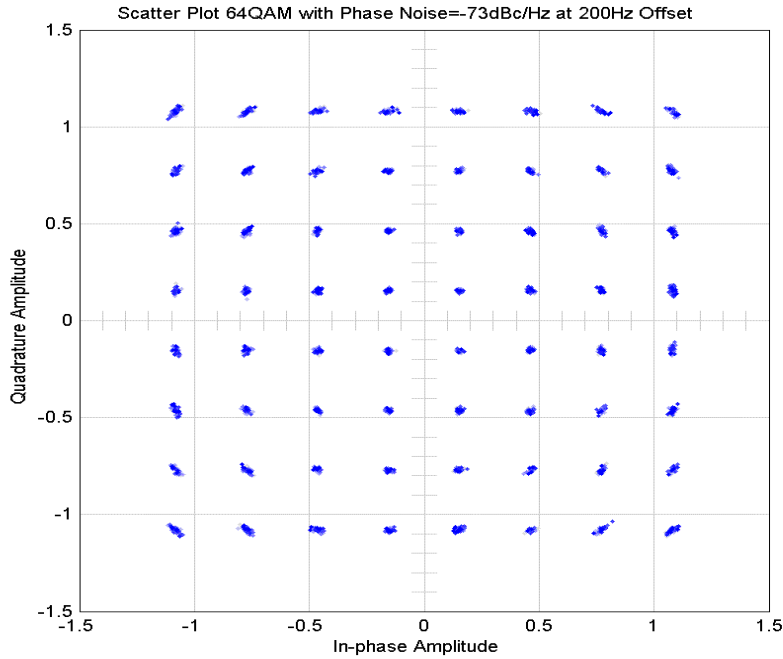
### ***3.1.1.3 Noise Degradation***

Another challenge for radio designers is noise. As described earlier in section 2.3.1.1, the noise either received or internally generated hurt the demodulation of process. Noise is generating in all electronic circuitry.

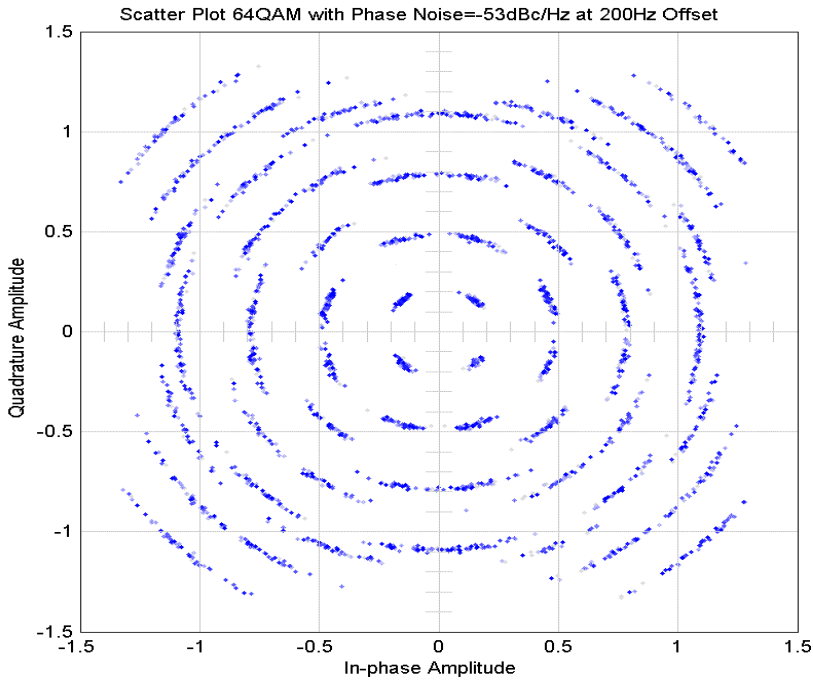
In order to show the effects of noise on received signals, consider the effect of PLL phase noise on the demodulation of a 64QAM signal. Phase noise generated by the PLL is mixed with the received signal. Although the signal remains constant, the noise is added onto the signal which degrades the received Signal to Noise Ratio (SNR).

As an example, consider a case where the center frequency is set to 850MHz as shown in Figure 3-4, and in second case, the center frequency is moved to ten times the frequency of case 1, i.e. the new center frequency is set to 8500MHz. A received signal is demodulated and the symbol error rate (SER) is calculated and the demodulated is displayed in Figure 3-8 and Figure 3-9.

From Figure 3-8, the phase noise at 850MHz is -73dBc / Hz at 200Hz offset. As described earlier, as the frequency increases by a factor of 10, the phase noise degrades by a factor of 20.log (multiplication factor). In case 2, we have changed the frequency by a factor of 10. Therefore, the theoretical (and best case) degradation is 20.log(10)=20dB. The PLL phase noise at 8500MHz is -53dBc/Hz at 200Hz offset. In Figure 3-9, we show the effect of 20dB additional phase noise on a 64QAM scatter plot of case 1 (Phase noise = -73dBc/Hz at 850MHz) and case 2 (Phase noise = -53dBc/Hz at 8500MHz).



**Figure 3-8 Scatter Plot of a 64QAM modulated Signal with Phase Noise =-73dBc/Hz at 200Hz Offset (RF=850MHz)**



**Figure 3-9 Scatter Plot of a 64QAM modulated Signal with Phase Noise =-53dBc/Hz at 200Hz Offset (RF=8500MHz)**

In case 1 shown in Figure 3-8, the symbol error rate (SER) is simulated to be 0%. In case 2 shown in Figure 3-9, the SER is calculated to be 15.05%. From the example described above, it is clear that phase noise and phase jitter significantly impair the sensitivity of a receiver and must be kept in check. In order to minimize this impact, the PLL must be designed for lower phase noise performance which translates in additional cost (better reference frequency oscillator). In addition to the phase noise effects, other RF impairments such as IQ imbalance have similar effects on the received signal.

### ***3.1.2 Spectral Detection Challenges***

Once the signal is demodulated and digitized, digital signal processing is performed on the received signal. The goal of the digital signal processing remains to quickly and accurately determine the presence or absence of a signal.

The spectral detection has been sufficiently discussed in section 2.3.2. The question remains to find ways to strike balance between the FFT resolution and the processing time. As the resolution or spectral averaging increases, the processing increases.

Per equation (17), one way to decrease the processing time is to increase the DSP operating frequency as the processing time is inversely proportional to  $F_{\text{DSP}}$ . However, faster DSP are generally expensive and consumes more power which adversely affect customer acceptance of the technology.

### ***3.1.3 Consumer Acceptance Challenges***

Several technologies with very promising market future never materialized and failed to gain market acceptance. For example, IBM's OS/2 operating system was far technically superior to Microsoft Windows product [61]. However, OS/2 never gained the market acceptance and the success of Microsoft Windows for two key reasons: 1) business model and 2) user interface. Windows is far more user friendly than OS/2, and while IBM charged for a well designed and

feature rich product, Microsoft chose to offer Windows free of charge in OEM machines or for a minimal fee. This business model proved successful as the install base for Windows grew fast and the momentum allowed Microsoft to sustain windows and fund the development of future generations of Windows. Based on technical performance alone, OS/2 was by far the better performer. Another example in wireless communication is the Ultra WideBand (UWB) standard or the IEEE 802.15.4a [65,66]. The technology was originally conceived [63,62] as very low power, low cost and high data rate. The transmission as originally conceived is to be an impulse function generated directly from baseband or by connecting an antenna directly to baseband. The narrower the pulse in the time domain, the broader the spectrum becomes in the frequency domain [64]. However, due to regulatory restrictions, the frequency band was restricted from 3GHz to 10GHz. With this new restriction, an upconversion was needed and the original advantages of low cost, low power and low complexity were erased. UWB has not been successful as defined in [66].

In order for a technology to gain acceptance in the commercial market, there are three drivers:

1. Power consumptions – long standby and talk time are expected. Cellular phones have set standard.
2. Low cost – the technology must be comparable to today's cell phones cost for voice communications and comparable to WLAN CPE prices for data applications.
3. Performance – the new technology must show incremental improvement over the current technology.
4. Miscellaneous advantages – such as user interfaces, branding, business models which we will not address in this section.

In order for CR to be successful, it must at least provide comparable cost, power consumption and performance to today's technology. A CR technology that degrades any of the above requirements drastically reduces its chances to succeed in the market.

## 3.2 Sensing Options

With numerous broadband radio design challenges and the need for fast and accurate sensing, the CR community began to think “outside the box.” The brute force approach and linear scaling of the technology fails to deliver the required performance for a sensing receiver. The CR community pushed towards system and architectural improvements instead of circuit design incremental improvements.

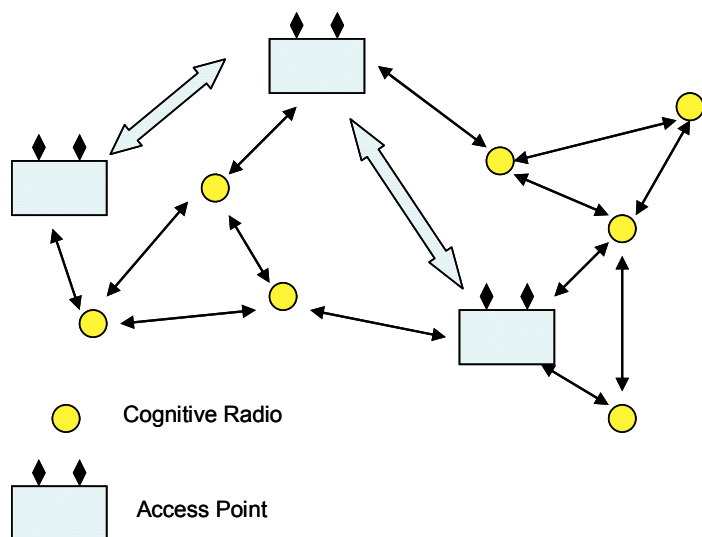
### 3.2.1 *Spectrum Etiquette and “no talk” Zones*

The simplest form of cognitive radio comes in the form of spectrum etiquette where the CR uses a “listen before talk” strategy. This strategy is commonly used in application such as spectrum leasing. The concept of the spectrum etiquette is not new. In 2007, the FCC has proposed etiquette for the ISM band [76]. The NPRM seeks to establish spectrum etiquette for unlicensed transmitters that operate under Sections 15.247 and 15.249 of the rules in the unlicensed bands at 915MHz, 2.4GHz, and 5.8GHz. Therefore, the NPRM had a direct effect on popular standards such as IEEE 802.11, 802.15.4, 802.16, and 802.18. However, [77] showed that spectrum etiquette improves spectrum network performance at the expense of an increased overhead required for information exchange.

### 3.2.2 *Cooperative Spectrum Sensing*

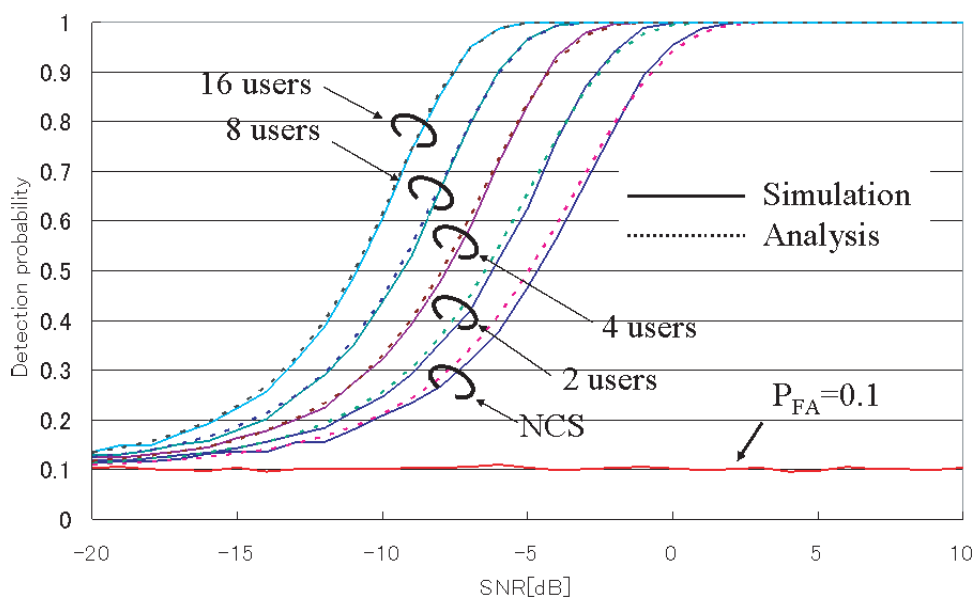
Cooperative spectrum sensing is a solution that relies on the variability of signals strength in various locations within a CR network [69]. The concept assumes a large number of users and sensing information is exchanged between neighbors. Cooperative spectrum sensing is a solution that has the potential to accelerate sensing while reducing the workload on the individual sensing.

The solution assumes the presence of access points [71], a gateway and nodes denoted as end users [70]. This model employs an overlay of several techniques such MIMO, cognitive radio, mesh network, and cross layer communication to improve performance. An implementation is shown in Figure 3-10.



**Figure 3-10 Cooperative Sensing in Cognitive Radio**

The main concern of the cooperative sensing is about increasing the probability of detection of a PU while collaborating with other PU. The results from [70] are shown in Figure 3-11. The simulation assumes BPSK modulation operating at 2.4 GHz.



**Figure 3-11 Detection Probability versus SNR in a CR Cooperative Sensing Network [70]**

where  $P_{FA}$  is the probability of false alarm and NCS is a non cooperative network. The improvement of probability detection is shown in Figure 3-11. However, there are some key questions about application and realization of the cooperative sensing:

1. How can cognitive radios cooperate?
2. How much can be gained from cooperation
3. What is the overhead associated with cooperation?
4. How can cooperative sensing improve the sensing time?

These questions above remain unanswered. Although improving detection probability is important, detection time and accuracy remain a challenge. Also, the addition of access points adds to the complexity, deployment, maintenance, and cost of the CR solution.

### ***3.2.3 Centralized CR Network and Spectrum Leasing***

The FCC published a Notice of Proposed Rulemaking (NPRM) [75] in which the FCC presented four scenarios on effective spectrum lease to cognitive radios:

1. The most common understanding of cognitive radios operation is the use of white spaces. This common understanding of CR use assumes does not rely on authorization from the PU as long as the interference is held in check. This is called the *Commons Model* where the PU is oblivious to the presence of secondary users in the spectrum.
2. The PU is aware of the presence of secondary users in the band and chooses to lease its spectrum under three different models. This is called the *Property-rights* or *Spectrum leasing Model*.
  - a. A frequency band is licensed to a single user. The user may use CR technology to increase spectrum efficiency. Unlicensed CRs are not authorized to use the channel.
  - b. The licensees sublicense the band to a CR user in return for collaboration. The cognitive radio devices could possibly "negotiate" with a licensee's system in real



time and use spectrum only if an agreement is reached between devices and/or their owners.

- c. The cognitive radio technologies can facilitate automated frequency coordination among co-primary spectrum users. This scenario can be extended to spectrum sensing cooperation as well.

In a centralized CR network, a spectrum coordinator assigns short term spectrum leases to end users [73,74]. Unlike cellular spectrum licenses, where licenses are multi-year in large contiguous spectrum chunks, the CR spectrum leases are short term leases to cooperating secondary users. The spectrum bandwidth and center frequency vary from lease to lease. The broker must deal with two conflicting objectives: 1) maximizing spectrum use, and 2) minimizing interference. Under this model, the PU can dictate the model parameters. In this model, the PU sets a censored or a “no talk” zone such that the CR is not allowed to transmit [72].

Similar to other sensing solutions, this technique improves sensing but does not address the delay or overhead associated with cooperation. Worse yet, in the case of the spectrum leasing model, the primary user solely makes the decision if and when the CR is allowed to operate. For example, assume that the PU is a cellular network in peak time. Chances are the PU denies service to SUs during the peak utilization time. This delay renders the CR highly ineffective.

### ***3.2.4 Control Channel***

Under this model, the CR may predetermine an agreed upon channel to facilitate collaboration and cooperation. The control channel can either be implemented as a dedicated frequency channel or as an underlay UWB channel [71]. Wideband RF frontend tuners/filters can be shared between the UWB control channel and normal cognitive radio reception/transmission. Control channels facilitate communication and control among users and help avoid interference or clashes.

The control channel again does not address the sensing time. This option has many drawbacks including having a master control decide when a CR is allowed to initiate transmission. This solution does not address sensing time nor does it solve the real time operation of a CR.

### **3.3 Summary**

From RF impairments to complex spectrum power measurement techniques, the challenges to perform wideband sensing are many. In order for the CR to become a practical technology, it must overcome the sensing challenges. The CR must perform the measurements swiftly, make a decision, and tune to available bands over wide frequency bands or return to a previous channel. Several research topics in a distributed or centralized network have yielded improvements in important parameters such as detection probability, low SNR detection, false alarm minimization, collision avoidance, and scheduling. However, none of these techniques have focused on minimizing the sensing time. As a matter of fact, several of the ideas described earlier have actually increased the sensing time by adding overhead to the network.

In the next section, we show RF impairment effects highly reduced with the addition of a Dedicated Sensing Receiver (DSR). The DSR's main function is power detection and spectrum estimation. It possesses the ability to detect the presence of other users in the band. It does not however need to perform demodulation of the signal. In order to meet the timing restrictions, novel techniques must be integrated into the sensing receiver. The addition of the DSR provides additional benefits to the overall radio. For example, the presence of two independent receive chains in the system provides channel diversity, as well as the ability to share the work load between the two receivers.

With the stated challenges and delay restrictions, the architecture of the DSR and a sensing algorithm is presented in the next chapter. The performance improvements are also made evident by the results.

## 4 Dedicated Sensing Receiver

In this chapter, we present a practical solution to the sensing receiver limitations. In chapter 3, we have shown that most of today's solutions add complexity and cost, but do not however solve the delay issue that plagues cognitive radio channel sensing. We present our solution, our results and additional performance gains if our solution is used in conjunction with other techniques listed in chapters 2 and 3.

Cognitive Radio has shown a lot of promise. If implemented properly, it promises to increase utilization of the spectrum without affecting legacy or licensed users, also known as Primary Users (PU). A CR leverages built-in intelligence and local decision making capability to search for available or underutilized spectrum. With this knowledge of the environment, the CR determines the best communication parameters (modulation, frequency, power level, etc.) that would allow operation as a Secondary User (SU) without negatively impacting others in the band. After operation begins, the CR needs to continuously sense and monitor the spectrum. The key to spurring the wide adoption of cognitive radio in the market is a practical realization of the sensing receiver in the cognitive system. Cognitive Radios (CR) require an accurate assessment of the activities in a desired frequency spectrum in order to determine the availability of idle channels suitable for opportunistic secondary use. Technological advances in recent years have addressed some of the challenges of broadband and frequency agility. However, real challenges still persist such as the physical implementation of broadband frequency synthesizers and high sensitivity receivers with sufficient dynamic range [2,5,6,22]. In order to address these shortcomings, the CR community must emphasize innovative architectures and algorithms.

Prior research has focused on novel techniques such as centralized network sensing, cooperative sensing, and the use of statistical methods and game theory to improve sensing performance [72,73,74,77,78]. However, these techniques are not well suited for use of CR in time sensitive operations; they do not address the reduction of the sensing time. By utilizing cooperative sensing techniques, the CR is dependent on outside inputs to make decisions on its operation.

With a centralized network sensing, additional costs, overhead and delays are introduced by the traffic controller. In the cooperative model, the CR performs energy detection and uses time division slots to communicate with other users. As the number of users increase, the delay becomes intolerably long.

Improvement at both the network level and the radio level are required. In this section, we present different techniques to improve the performance. However, any network proposals require regulatory approval and ratified standards. Hence, our research objective is focused on the sensing receiver. The improvements at the radio level are self contained and in our research, we were able to show significant improvements to the sensing time with the use of a dedicated sensing receiver (DSR). The DSR is solely focused on channel sensing and runs in parallel with a main receiver. The key to the DSR is an efficient and adaptive algorithm that performs spectrum detection and continuously improves the quality of the collected data. We present three levels of sensing: an initial fast mode, a coarse, and a fine sensing mode. The initial sensing is done in the analog domain at the RF or IF frequencies prior to additional processing in the digital domain. We demonstrate that the use of a dedicated sensing receiver (DSR) is necessary and required for fast and reliable sensing in broadband operation. In addition, the overall time delay is also greatly reduced which paves the way for voice and video operation in cognitive radio. We show that the DSR architecture provides up to a five-fold reduction in total mean time detection.

In this section, we present an architecture in which the majority of the burden for accurate sensing and quick decision making on channel conditions and spectrum availability is off-loaded to the Dedicated Sensing Receiver (DSR). The sensing receiver is equipped with power detectors at RF, IF, and baseband for coarse and fine sensing. In addition to coarse and fine sensing, we introduce a fast sensing mode that leverages available a priori knowledge of the CR environment or the prior state of the CR. The advantages of the fast mode become apparent during the setup of the CR at initialization. Simulation results show a drastic improvement in initial sensing over a broad spectrum. The overall time delay is also greatly reduced which opens the way for voice operation in cognitive radio.

In addition to the radio architecture, an algorithm is presented. The algorithm parameters vary as the channel conditions vary. More importantly, the algorithm utilizes an escalation process in which the number of FFT points or averaging increases as it progresses through the algorithm. The algorithm assumes self reliance and does not require input from the network or other CR in the system. By adapting its sensing time, the DSR is able to make quick decisions and it uses the analog detectors as the first defense mechanism against interference.

This section is divided in three parts:

- 1) CR network improvements
- 2) Sensing receiving improvements
- 3) Results

As part of the radio architecture, we will also present the analog detection.

## **4.1 Proposed CR System Improvements**

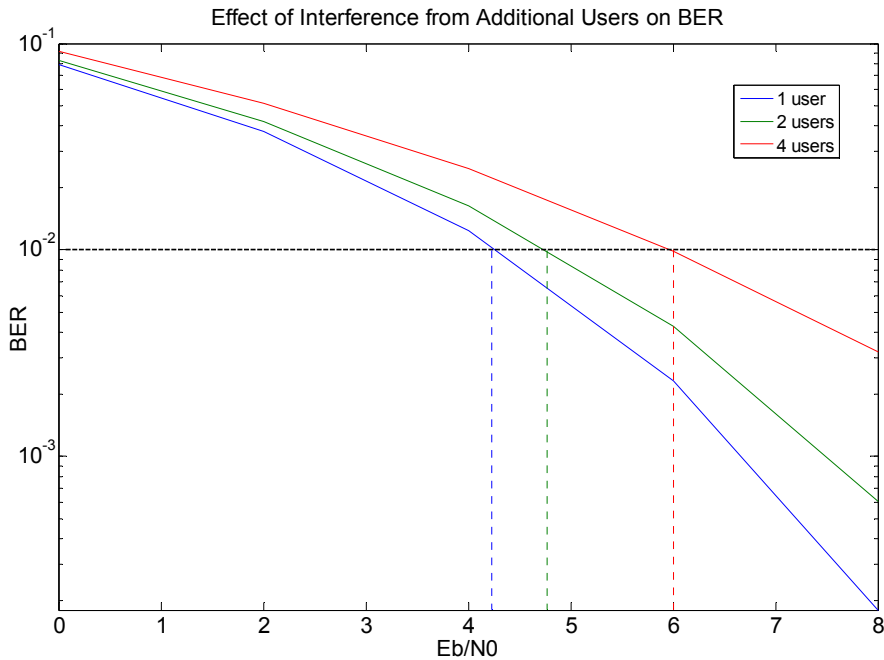
As we established in chapter 3, the sensing receiver is challenged on two fronts:

- 1) Agile frequency tuning over a broad spectrum
- 2) Delays introduced by the need for accurate detection

In section 2.3.1.1, we introduce the background on modern receiver. The frequency generation is a key aspect of the receiver. While sensing, the receiver must tune and settle on each frequency before the power detection is initiated. In this section, we introduce different concepts at the physical (PHY) layer and Middle Access Control (MAC) layer that improve the performance of the CR.

### 4.1.1 Figure of Merit

In order to distinguish between the advantages of different solutions, a measurable figure of merit is required. Common figures of merit in digital communication are Bit Error Rate (BER), Symbol Error Rate (SER), Frame Error Rate (FER) or Packet Error Rate (PER) [64]. The parameters assume that a received signal is demodulated and the error rate is calculated. Although the conditions of operation vis a vis a PU may be fairly well defined, the CR may still interfere with other wireless systems. As an example of performance degradation due to addition of users, a CDMA link with one, two or four users is simulated. Modulation is BPSK and we are interested in a  $BER=10^{-2}$ . With one user, there is no interference, the  $E_bN_0$  required for this BER is 4.1dB. When two users are present, the required  $E_bN_0$  is now 4.8dB or a degradation of 0.7dB. As the number of users increase to four, the  $E_bN_0$  degrades even further to 6dB. That is a total degradation of 1.9dB due to presence of other users in the cell. The BER curves are plotted in Figure 4-1.



**Figure 4-1 Effect of added interference from users on BER**

One way to reduce the interference is by implementation spectrum etiquette. The FCC [4] and several research papers [22,80,81,82] have proposed an operation etiquette for Cognitive Radios.

However, interference occurs even in the case where CRs have agreed to play by the rules. As discussed in 3.2.1, sharing etiquette enables spectrum coordination among multiple wireless devices.

Users utilizing a well defined protocol help disciplined operation and more importantly, a predictable performance measured by the error probability of the system. The packet error probability ( $p_e$ ) of a CR operating as a secondary user (SU) in a frequency band is highly dependent on noise ( $p_n$ ), interference from the presence of a primary user ( $p_{i,PU}$ ), and interference from other secondary users of the band ( $p_{i,SU}$ ). The packet error probability  $p_e$  is therefore defined as the sum of the probabilities.

$$p_e = p_n + p_{i,PU} + p_{i,SU} \quad (21)$$

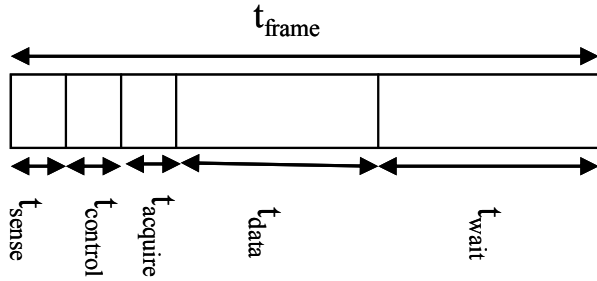
However, in the case of disciplined operation and a well defined etiquette, it is assumed that other SUs are multiplexed on the same channel and hence packet error probability due to interference by another SU is negligible ( $p_{i,SU} \approx 0$ ). Also, for the same reason,  $p_n$  will also be negligible ( $p_n \approx 0$ ). Hence, equation (21) is reduced to:

$$p_e \approx p_{i,PU} = p \quad (22)$$

The packet error probability  $p_e$  along with traditional BER, SER and FER are used as figure of merit of the system performance of the cognitive radio.

#### ***4.1.2 Communication Link and Sensing Time***

Assume that the frame in Figure 4-2 is used in a cognitive radio.



**Figure 4-2 Cognitive Radio Frame**

where  $t_{frame}$  is the total frame time,  $t_{sense}$  is the time allocated to spectrum sensing,  $t_{control}$  is the time allocated exchange information with other users,  $t_{acquire}$  is the time allocated to the receiver to acquire a new channel when interference is present,  $t_{data}$  is the transmitted data and  $t_{wait}$  is part of the spectrum sharing etiquette in which the user “waits” and “listens” to other users [83].

Willkom et al. [83] proposed a reliable link maintenance protocol model for CRs where the frame structure is composed of link maintenance (sense, control, acquire) and the transmitted data as shown in Figure 4-2.

In our definition, the sensing time is the sum of  $t_{sense}$ ,  $t_{control}$  and  $t_{acquire}$ . Our main research objective remains to reduce the total sensing time such that the CR becomes practical while sustaining a communication link.

One aspect of sustaining a communication link is to insure that the link is robust. One way to insure robustness is to include redundancy [64,83]. Redundancy is the process of the transmitting the information or part of the information ( $t_{data}$ ) at least twice just in case the first transmission of the data is corrupted, then subsequent transmissions are recovered.

If  $N$  is the total number of packets and  $X$  is the number of redundant packets in the frame, the message error probability  $P_{err}$  can be determined as [83],



$$P_{err} = \sum_{i=1}^N \binom{N+X}{X+i} p^{X+i} (1-p)^{N-i} \quad (23)$$

where  $p$  is defined as the packet error probability.

From equation (23), the probability of error decreases as the number of redundant  $X$  increases. Furthermore, a figure of merit termed Goodput or  $G_{bit}$  is introduced in order to characterize the transmission performance. Goodput is defined as

$$G_{bit} = \frac{(1-P_{err})Nb_{sc}t_{data}}{t_{sense} + t_{control} + P_m t_{acquire} + t_{data}} \quad (24)$$

where  $b_{sc}$  is the bit rate used by the channel and  $P_m$  is defined as the probability that at least one of the channels can no longer be used.

As defined earlier, Probability error and Goodput is used as a figure of merit of the performance of the channel.

### ***4.1.3 Proposed Approaches for Improving Performance***

Innovation is required to meet the intent of a practical CR. Innovation does not necessarily mean invention but rather novel ways of using existing techniques in addition to invention if any. The proposed system combines PHY and MAC layer techniques to improve performance. There are some very well documented approaches that improve performance such as adaptive waveform [84], adaptive and reactive modulation and power control [85]. However, additional innovative ideas are required to make Cognitive Radios a reality especially in a voice or video transmission application.

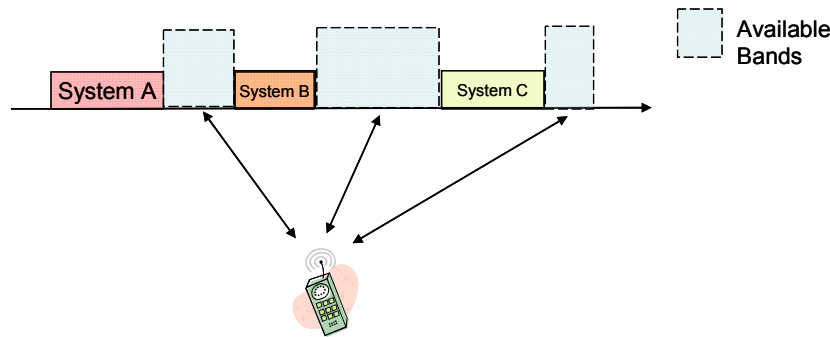
The proposed ideas target improving performance of the system.

- (1) Use of multiple unallocated bands
- (2) Use of multiple bands concurrently to increase data transmission and redundancy
- (3) Handshaking and look-ahead techniques.

For simplicity, we assume that each technique operates in a stand alone. Obviously, several of these techniques may be combined for improved performance which may be subject of our future work.

#### ***4.1.3.1 Use of multiple unallocated bands***

One proposed concept that helps cognitive radio applications is a de-fragmentation of the spectrum. The concept requires frequency reassignment which is usually politically difficult to achieve and practically impossible. The easier option is the “selective reuse of frequency” where a CR uses the unoccupied chunks of unallocated spectrum [86] as per Figure 4-3.



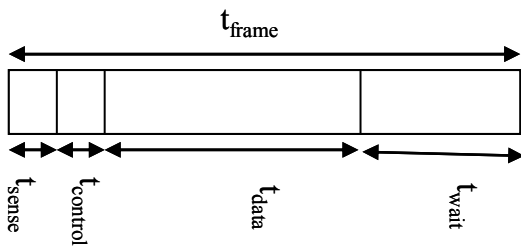
**Figure 4-3 Selective Frequency Reuse**

In the unallocated spectrum, although there are no primary or licensed users, there might be some regulatory restrictions. Some of the unallocated bands may act as guard bands for a particular system or more importantly, a CR carrier may disturb the operation of the nearby system. For example, CR operation may affect the receiver by: 1) reducing receiver sensitivity,

2) adding noise, 3) adding interferers to the receiver front end, or 4) affecting the transmitter such as creating intermodulation products with the carrier.

Assuming that operation in the unallocated or scarcely populated bands is permissible, the CR needs only to take into account the etiquette for operation as other CRs might also be competing for the same spectrum. By leveraging the etiquette,  $t_{acquire}$  is significantly reduced and becomes negligible.

The frame structure shown in Figure 4-2 is then reduced to:



**Figure 4-4 Reduced Frame Structure**

Hence, from Figure 4-4, the total frame time is shown in (25).

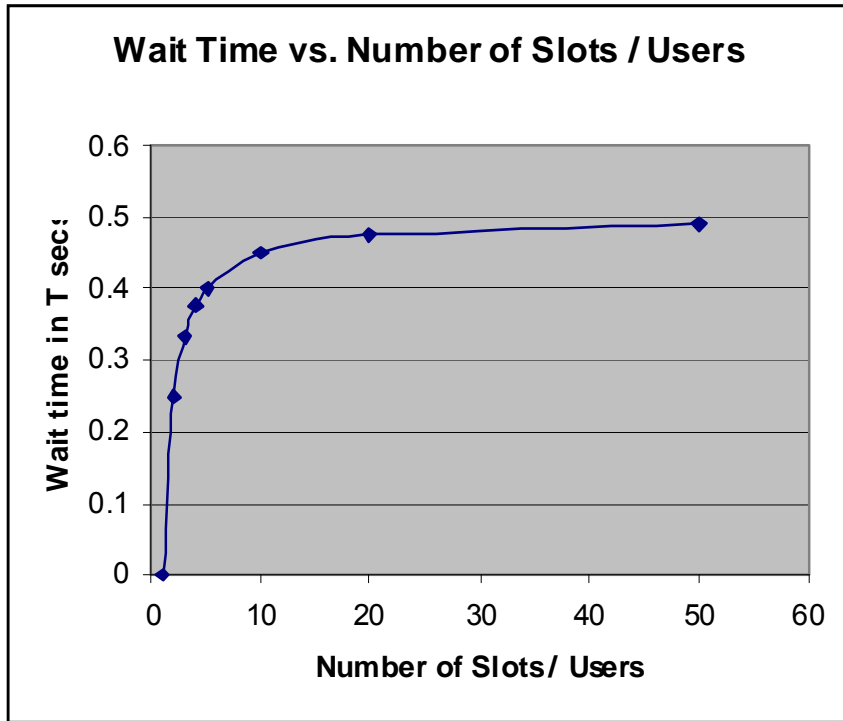
$$t_{frame} = t_{sense} + t_{control} + t_{data} + t_{wait} \quad (25)$$

Sklar [87] defines the message delay as  $D = t_{wait} + t_{frame}$  where  $t_{wait}$  is determined by a TDM protocol that is dependent on the number of users in the frequency band.  $t_{wait}$  is the average wait time of a packet before transmission. The goal is to reduce the wait time before retransmit.

In a TDM protocol,  $t_{wait}$  is given by:

$$t_{wait} = \frac{T}{2} \left[ 1 - \frac{1}{M} \right] \quad (26)$$

where  $M$  is the number of orthogonal time slots and where the bits are transmitted in  $T/M$  seconds. For a dedicated channel, i.e.  $M=1$  then the wait time equals 0 sec. However, as the channel is shared among users, the wait time increases. As an example, the data rate is normalized and the wait time is plotted in Figure 4-5.



**Figure 4-5 Normalized Wait time versus Number of Users**

Our proposed approach is to use the unallocated spectrum with an agreed upon etiquette and a centralized control center to assign and direct communication among secondary users. Each radio senses and reports back to central command. The central command keeps track of the different carriers and any newcomers. In addition, depending on the required or subscribed quality of service, the command might assign a higher number of time slots in order to lower delays or increase the number of redundant slots.

The main advantages of using fixed and known frequency bands in concurrence with a central control system are:

1- A well disciplined interaction and better coexistence among the many users of the frequency slot where the controllers assign time slots as required. In some cases, the controller might reassign a CR to a different frequency. This concept is similar to a cellular network with a network initiated handoff [21].

2- A synchronized and predictable traffic flow which reduces the interference among the users. In addition, the reduction in complexity also reduces the need for extensive processing and reduction in cost and power consumption.

3- A reduction or elimination of the link maintenance.

Because of the above improvements, equation (24) approximates to:

$$G_{\text{bit}} \sim (1-P_{\text{err}}) N b_{\text{sc}} \quad (27)$$

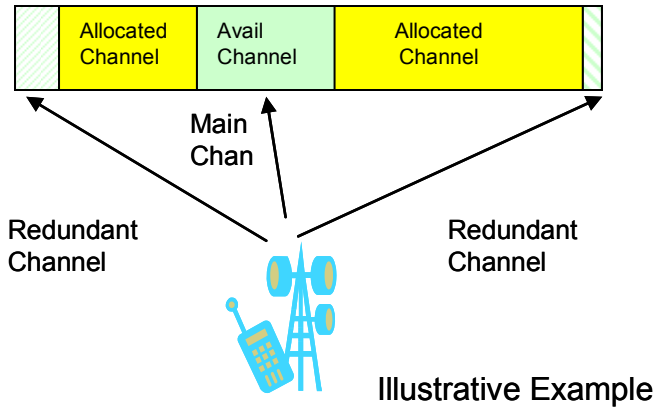
The probability of goodbits in equation (27) is higher than the probability in equation (24).

Hence, by adding spectrum etiquette, an overall improvement in performance is achieved.

#### ***4.1.3.2 Use of Multiple Frequency Bands Concurrently***

The proposed approach suggests that CR transmits and receives concurrently on two or more different frequencies. Depending on traffic, the CR may transmit and receive the desired data on the less congested channels, while using the more congested frequency channels to transmit and receive redundant data. The advantages of such technique are: 1) built-in redundancy, 2) frequency and spatial diversity, and 3) asynchronous operation where the time slot is dynamically assigned. However, this technique challenges the PHY and MAC implementation. The technique uses a single channel as a main communication channel and scavenges other channels for redundancy. The redundant channel may be using channels that are too narrow for normal CR operation and a frame may be divided among several frequencies. The frame reconstruction is only needed when the main channel is being jammed.

An illustrative example is shown in



**Figure 4-6 Concurrent Use of Non-Contiguous Frequency Bands**

In using one or more redundant channels different from the main transmission as shown in Figure 4-6, the CR is able to 1) keep the data channel clear of any redundancies and retransmits, and 2) use the redundant channels to reduce the probability of error on BER is degraded. An interesting benefit of this technique is that the “best” channel is dedicated to sending data with no redundancy. The redundant channels in turn send a subset of the data on available channels. Since only a fraction of the info bits are needed to be redundantly transmitted, the data may be transmitted via a more robust channel, i.e. lower modulation order or higher power. In addition, depending on the number of channels available for redundant transmission and the proximity of the channels, a single radio may be used to sequentially transmit the redundant data on several channels.

With limited redundancy, equation (23) reduces to:

$$P_{err} = \sum_{i=1}^N \binom{N}{i} p^i (1-p)^{N-i} \quad (28)$$

$$P_{err} = \sum_{i=0}^N \binom{N}{i} p^i (1-p)^{N-i} - (1-p)^N \quad (29)$$

$$P_{err} = (p+1-p)^N - (1-p)^N \quad (30)$$

$$P_{err} = 1 - (1 - p)^N \quad (31)$$

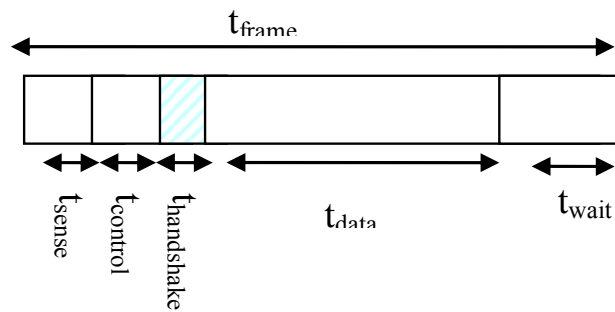
Equation(31) shows that the error probability decreases exponentially as 1) the interferers are well understood and disciplined, and 2) the overall throughput of the system is not decreased by redundant packets as was the case in 4.1.3.1. Since there are no primary users defined in the unallocated channels, the redundant channel is used to correct errors generated by the channel conditions.

The challenge of this technique is 1) the ability to concurrently transmit at several frequencies which requires several transmit chains and 2) MAC layer complexity. The complexity of the MAC is mostly due to dynamically modifying the frame structure based on the environment.

#### ***4.1.3.3 Handshaking and Look Ahead Techniques***

One of the keys to CR operation is handshaking between the transmitter and the receiver [4]. Since the operation of a CR is opportunistic, there are inherent delays due to searching, control and handshaking before transmission. The delay may be reduced if innovative handshaking and look ahead techniques are used.

Our proposed approach is to include a look-ahead handshake. The transmitter and the receiver use the packet structure as defined in Figure 4-4 but with the addition of look ahead data.



**Figure 4-7 Modified Frame to Include Handshaking Information**

Both the transmitter and the receiver scan and sense channels. The channels are prioritized. The transmitter and the receiver exchange their information in  $t_{handshake}$  and would agree ahead of time on the next channel frequency. There are several look ahead algorithms that may be implemented.

The advantages of the proposed protocol are:

- 1- Shared work load and collaboration between the transmitter and the receiver. For example, the frequency band may be divided in two sub bands. Each CR senses half of the band and exchanges the results with the paired CR. Hence approximately, each CR performs half of the work. In addition, this concept maybe shared among several CRs. Assume that there is a three way handshake as defined in [4]. The three CRs may agree to pool their efforts and subdivide the frequency band in three sub-bands and hence reduce the sensing workload.
- 2- Faster and more efficient decision making and handshake between the transmitter and the receiver translates into better system performance and less delay.
- 3- The concept may be extended to a central controller concept. Assume that there is a central basestation that is in-charge of managing traffic. The basestation will in turn use the look-ahead control data to create a complete understanding of the cell it is covering.

The proposed technique reduces the amount of sensing work the CRs have to do in order to coexist in a competitive space.



#### 4.1.3.4 CR Network Improvement Results

In this section, we present simulation results to illustrate the impact of the proposed solutions in sections 4.1.3.1 and 4.1.3.2.

The first scenario uses the following setup: the CR employs QPSK modulation with a spectral roll off of 0.9, an AWGN channel with an SNR= 30dB, 100 frames, 100 symbols/frame,  $t_{\text{sym}}=0.01\text{s}$ , 4x oversampling and introducing a maximum variable delay of four samples.

For the purposes of this study, we compare the performance with and without a PU present (unallocated channels). In the case of the PU present, we assume that the PU appears every Y number of frames and will wipe out X frame of data. For completeness, we record the Symbol Error Rate (SER) and the delay before a retransmit of the corrupted data. (Note: we do not drop the bad frame from the error calculations).

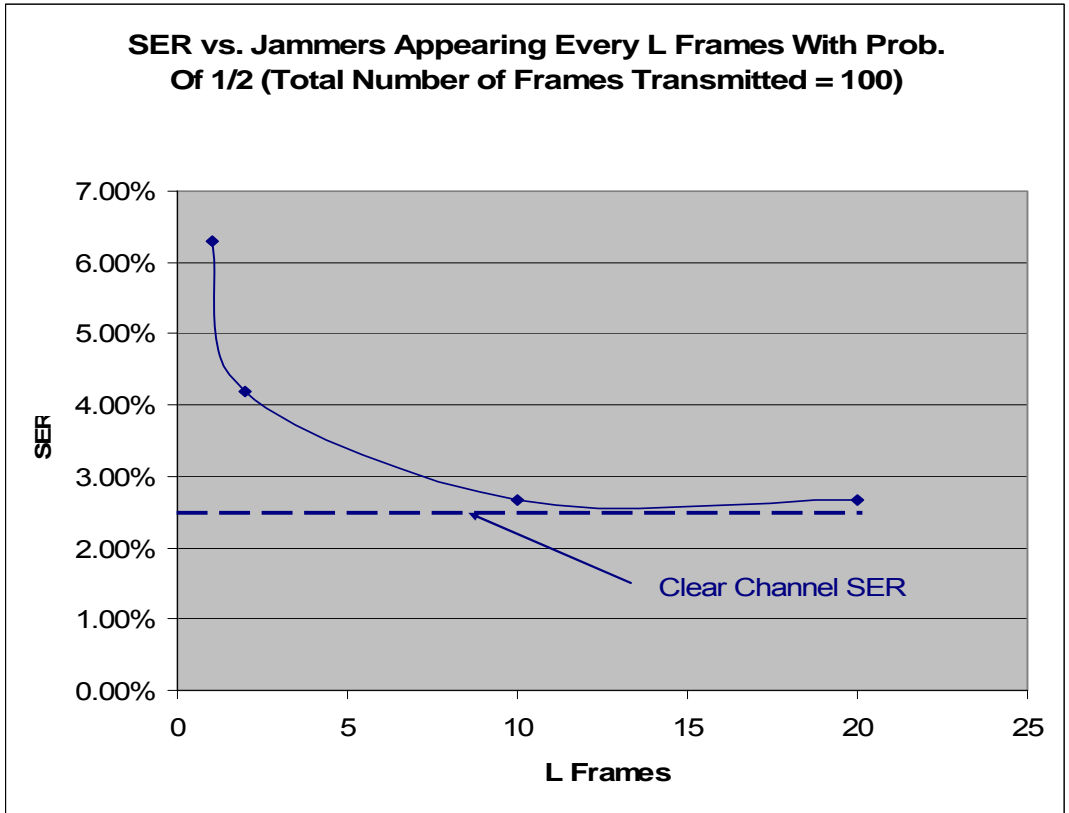
| Approach      | PU appears every Y frames | PU lasts for X frames | Retransmit request (over 100 frames) | Min Delay (sec) | SER    |
|---------------|---------------------------|-----------------------|--------------------------------------|-----------------|--------|
| Traditional   | 5                         | 1                     | 20                                   | 20              | 17.14% |
|               | 10                        | 1                     | 10                                   | 10              | 9.85%  |
|               | 10                        | 5                     | 50                                   | 50              | 38.46% |
|               | 20                        | 5                     | 25                                   | 25              | 20.97% |
|               | 20                        | 1                     | 5                                    | 5               | 6.21%  |
|               | 50                        | 20                    | 40                                   | 40              | 31.37% |
| Clear Channel | 0                         | 0                     | 0                                    | 0               | 2.68%  |

**Table 3 System Symbol Error Rate (SER) and Min Delay in the presence of a PU**

As a baseline, we simulate the performance of the CR in a clear channel. In the model, the SER performance is measured to be 2.68%. In Table 3, a CR operates opportunistically across multiple bands. We measure the effect of a jammer on the CR receiver. When a PU appears, the data transmitted by the CR is corrupted for the entire frame. When the channel is clear, the CR receiver requests a retransmit. However, the CR can not make the request until the channel is clear. For the purposes of this simulation, we assume that the CR receiver makes the request as soon as the jammer stops transmitting.

In Table 3, we vary two parameters: the frequency at which the PU appears and the length of time the PU occupies the channel. It is obvious that the performance of the receiver degrades as PUs appear more frequently and as they last longer. For example, if the PU appears every five frames and lasts for one frame or said differently, the PU occupies the channel 20% of the time, the SER degrades from 2.68% to 17.14%. This receiver becomes almost unusable. As the PU occupies the spectrum less time, we expect the SER to improve. In the second case, a PU appears every 10 frames and lasts for 1 frame (or 10% of the time), the SER degrades from 2.68% to 9.85%.

Now, assume that we are using unallocated bands as discussed in 4.1.3.1, under this model there are no PUs but other CRs may occupy the channel. Without a PU present, the jamming/interference is due to other “opportunistic” (or cognitive) radios. We assume that another CR can appear once every L frames with probability  $\frac{1}{2}$ . That is, if  $L=1$ , every frame has a  $\frac{1}{2}$  probability of being interfered by a competing CR. Once the competing CR appears, we assume that this radio interferes for one frame. In Figure 4-8, we show the SER performance of the proposed operating in unallocated bands as a function of L. For reference, the SER for a “clear channel” is also recorded. The results using technique 1 over 100 frames are displayed below. These results may also represent the performance of a CR as the number of CRs increase.



**Figure 4-8 SER vs. Other SUs Jamming the CR Every L Frames**

Using results Figure 4-8 and Table 3, it is clear that operating a CR in unallocated bands where the CR is competing with other opportunistic radios but there are no dominant PUs, the performance does not only reduce BER but also reduces the incremental delay due to retransmit requests from the receiver. As a drawback, this approach adds complexity to the radio. Disjointed frequency bands might require additional radio chains which might increase cost and power consumption.

Using the same setup as described above, the benefits of multiple frequency bands concurrently are demonstrated next. In order to model the multiple data streams, a deinterlacer splits the transmit data into odd and even bits. Then the odd bits were transmitted on channel 1 and the even bits were transmitted on channel 2. The data was transmitted on two separate AWGN channel and the output of the demodulators were combined using an interlacer. The SNR level of the channels were varied and recorded. Since we assume two uncorrelated channels, the signal

levels received from each channel are independent. We record the effect on the CR receiver BER when operating on two channels concurrently.

|      | Channel 1 | Channel 2 | BER   |
|------|-----------|-----------|-------|
| AWGN | 5 dB      | 5 dB      | 7.12% |
|      | 5 dB      | 6 dB      | 5.76% |
|      | 5 dB      | 7 dB      | 4.75% |

**Table 4 CR BER Performance vs. Use of a Two Channel System in an AWGN Channel**

The improvements are even more impressive if one considers a Raleigh fading channel. The same simulation was run with a single channel setup versus a two channel system. The single channel had the power delay profile characteristic of [0 -3dB -7dB -10dB] at [0 1s 3s 10s]. The single channel versus two channel improvements and the channel characteristics are shown in Table 5. In addition, a 4 channel model (where the transmit signal is split and transmitted on four separate channels) was created. In this case, BER improvements were recorded over 1000 frames.

|                | Channel 1<br>Delay Profile         | Channel 2 (3 and 4)<br>Delay Profile  | BER<br>Improvement |
|----------------|------------------------------------|---|--------------------|
| Single Channel | [0 1 3 10] sec<br>[0 -3 -7 -10] dB | NA  | Reference          |
| Dual Channel   | [0 1 3 10] sec<br>[0 -3 -7 -10] dB | [0 1 3 10] sec<br>[0 -3 -7 -10] dB  | 0%                 |
|                | [0 1 3 10] sec<br>[0 -3 -7 -10] dB | [0 3 10] sec<br>[0 -3 -10] dB   | 7.96%              |
|                | [0 1 3 10] sec<br>[0 -3 -7 -10] dB | [0 1 10] sec<br>[0 -3 -10] dB   | 6.71%              |
|                | [0 1 3 10] sec<br>[0 -3 -7 -10] dB | [0 3] sec<br>[0 -3] dB  | 9.91%              |
| Quad Channel   | [0 1 3 10] sec<br>[0 -3 -7 -10] dB | [0 3] sec<br>[0 -3] dB – Ch2<br><br>[0 3 10] sec<br>[0 -3 -7]dB Ch3<br><br>[0 5] sec<br>[0 -7] dB – Ch4 | 15.48%             |

**Table 5 BER improvement of Multiple vs. Single Raleigh Channels**

The delta BER improvements have increased with additional channels added. This is due to the channel diversity and built-in interleaving of the data (over the channels). Hence, the probability that the channel be wiped out by a primary user or other users is significantly lowered. In addition, the impairments and characteristics of each RF channel are different. Hence, as one of the channels degrades and loses packets, the other channels remain strong and uncorrelated. One interesting aspect of using such a technique is sending the redundant data on the lower bandwidth channels while keeping the “best” channel to send the main data.

In summary, the CR performance improves as different cross layer techniques are implemented. However, most of these techniques assume that the CR already knows that the channels are clear for operation. We still require that we recognize the availability of the channel.

In the next section, we concentrate on the sensing receiver and on minimizing the sensing delay.

## **4.2 Proposed Sensing Receiver Architecture**

The improvements in 4.1.3 assume that the channel or groups of channels are available. Once the operation of the CR starts, the performance may degrade due to interference. Now assume that the CR finds a way to always have a backup channel. This backup channel is used as soon as the CR channel is interfered with such that the system delay is not noticeable. Such technique is commonly used in mobile wireless network. For example in GSM [21], while in a call the device is required to scan continuously for channels and stores a list of neighboring cells (or frequencies). The concept is that as the GSM mobile unit moves through the cell, it is ready to perform a handoff to neighboring cells as the channel deteriorates and a minimum threshold is triggered. Parametric information stored in the mobile device contains channel number, Received Signal Strength Indicator (RSSI) information, channel BER and power levels. Similarly, a CR may take advantage of such techniques. However, the CR does not usually have a basestation feeding it the information as commonly done in cellular communication.

The proposed approach is to define an agile broadband radio receiver dedicated to scan and record channel conditions in a look up table (LUT). More importantly, an algorithm is needed for learning the channel condition and prioritizing the different channels for use by the channel transmitting and receiving the actual data. In order to avoid interference or saturation of its front end, the sensing receiver operates when the transmitter is not actively transmitting. A high level diagram is shown in Figure 4-9.

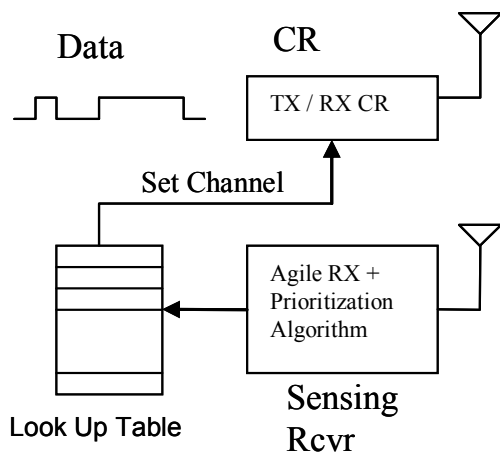


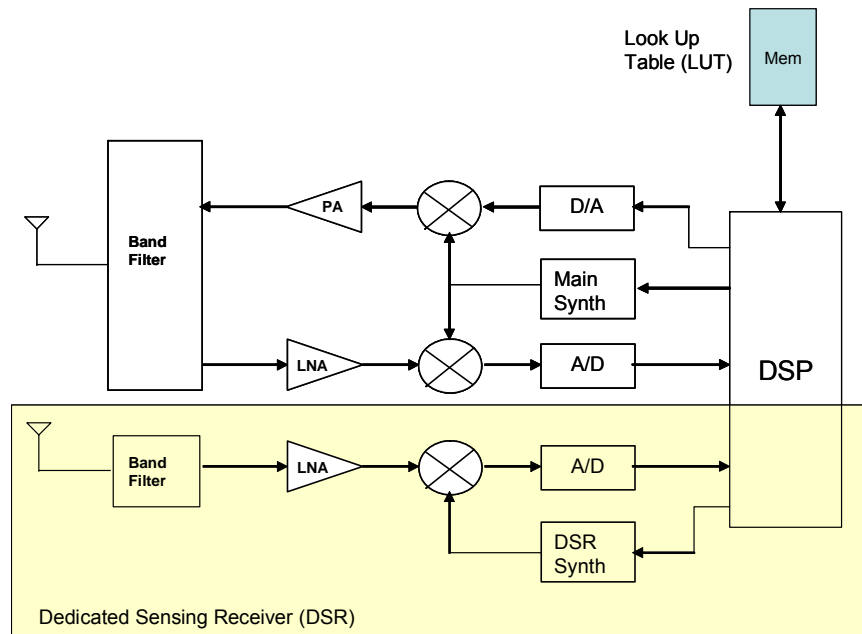
Figure 4-9 Proposed addition of a sensing receiver and LUT

The advantages of such techniques are:

- 1- Better decision making: with a dedicated receiver, the radio continuously senses and learns the channel without added complexity or urgency to send and receive data packets. The sensing receiver scans desired frequency bands and records the frequencies where a CR is allowed to operate. Assuming the CR can coexist with a PU, the receiver learns and may generate a prediction algorithm of the probability of PU appearance or interference with other CRs.
- 2- Faster scan: the sensing receiver prioritizes and saves the list of “best” channels in the look up table. In addition, the look up table may save the “preferred channels” where the CR has historically operated. Hence, upon power up of the radio, the main receiver searches the “preferred channels first” prior to doing a broadband sweep.
- 3- Cost and complexity – A broadband agile receiver is problematic and difficult to implement [88]. In addition, during large frequency swings, the synthesizer settling time is longer while it is critical to maintain low phase noise as the phase noise of the synthesizer directly affects BER. The Voltage Controlled Oscillator (VCO) used in the synthesizer has a limited frequency tuning range and may require the use of different VCO’s for different bands while keeping the phase noise in check. With a dedicated sensing receiver, the complexity of the PLL is not passed onto the main receiver since settling time may be

longer (i.e. the VCO settling time is not gating the packet transmission) and the phase noise may be worse since the sensing receiver does not require demodulation of the information received. Hence, because of lower requirements on the receiver, the cost, power and performance might improve.

The overall CR implementation is shown in Figure 4-10.



**Figure 4-10 Cognitive Radio and DSR High Level Diagram**

We believe that the Dedicated Sensing Receiver (DSR) is the catalyst for implementing a practical cognitive radio.

### ***4.2.1 The DSR Architecture***

The DSR is a radio that continuously strives to find available spectrum and prioritizes channels based on the environment and needs of the user. As part of the scanning process, the radio tunes to different channels in the bandwidth of interest. As discussed earlier, in order for CR to be effective, the radio must be agile and must be able to operate in broadband.

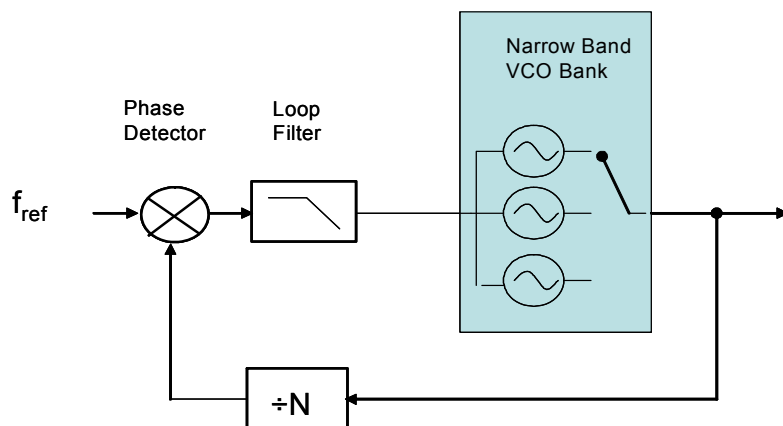


### 4.2.1.1 Practical PLL

The key radio element that allows frequency tuning is a frequency synthesizer. The frequency synthesis is done via a phase locked loop (PLL) as discussed in section 2.3.1.3. In this section, we propose a sample PLL design that is based on commercially available components. The goal of this exercise is to choose a PLL lock time that is practical for our analysis.

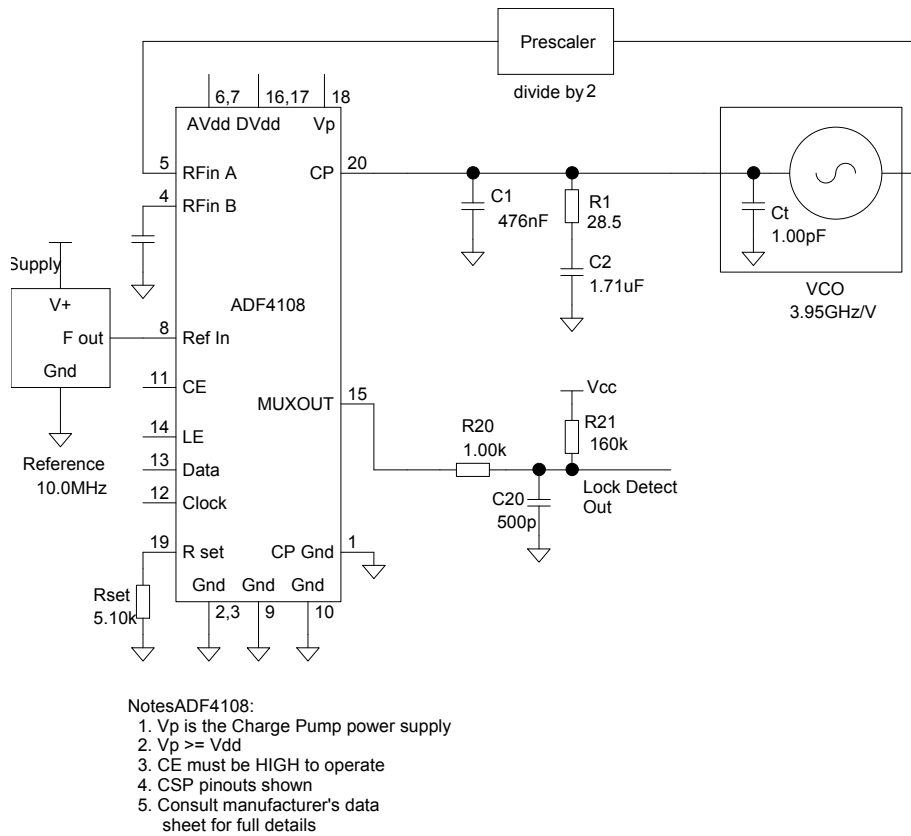
Based on available PLL chips and VCO vendors, we have chosen to design a PLL based on the Analog Devices family [89]. As an example, we have chosen to design a PLL capable of operating up to 10GHz. Unfortunately, broadband and high frequency VCOs are not readily available. The design of the PLL proved to be difficult in today's available technology. A search of typical VCO vendors such as UMC, Z-Com, Crystek and Fox [90] from catalogs and the web did not yield a suitable VCO. A 5GHz VCO with a multiplier is used. The VCO gain is  $K_v=4\text{GHz/V}$ . In addition, commercial PLL requires an external pre-scalar to reach 10GHz. A divide by 2 pre-scalar is used in this simulation. The noise is assumed to be ideal and hence no phase noise is added to the simulation.

Since a monolithic broadband 10GHz VCO is not readily available, the commonly used practice is to use a bank of narrowband VCOs as shown in Figure 4-11.



**Figure 4-11 DSR PLL Implementation**

The schematic of the PLL is shown in Figure 4-12.



**Figure 4-12 PLL Schematic**

There are many different types of loop filters. In Figure 4-12, the loop filter is a common second-order integrator. Analog Devices recommends that the loop filter bandwidth should be 1/10 of the desired frequency step (channel spacing). As discussed in chapter 3, increasing the loop bandwidth reduces the lock time, but the filter bandwidth should not be more than 1/5 of the frequency step in order to avoid significantly increasing the risk of instability.

A loop filter's bandwidth can be increased by increasing either the frequency or phase detector charge-pump current. The variation of loop bandwidth with Kv presents a major design challenge in wideband PLL designs, where the Kv can vary by more than 300%. Increasing or decreasing the programmable charge-pump current is the easiest way to compensate for changes in the loop bandwidth caused by the variation in Kv.

As we discussed earlier and demonstrated in Figure 3-8 and Figure 3-9, the phase noise is a key component. In this simulation, we derived the phase noise generated by the PLL as shown in Table 6 .

| <b>Frequency Domain Analysis of PLL</b>  |              |            |            |             |               |
|--|--------------|------------|------------|-------------|---------------|
| Analysis at PLL output frequency of 5GHz |              |            |            |             |               |
| <b>Phase Noise Table</b>                 |              |            |            |             |               |
| <b>Freq</b>                              | <b>Total</b> | <b>VCO</b> | <b>Ref</b> | <b>Chip</b> | <b>Filter</b> |
| 100                                      | -81.98       | -167.2     | --         | -82.01      | -103.8        |
| 1.00k                                    | -79.59       | -147.2     | --         | -81.65      | -83.83        |
| 10.0k                                    | -73.75       | -136.1     | --         | -82.42      | -74.38        |
| 100k                                     | -112.5       | -156.9     | --         | -121.6      | -113.1        |
| 1.00M                                    | -151.8       | -160.0     | --         | -161.6      | -153.1        |

**Table 6 Phase Noise Contribution at 5GHz**

However, the phase noise contribution is more common used in digital communications as phase jitter [8,91]. Next we derive the phase jitter based on the phase noise in Table 6.

A sine wave in time domain is represented by:

$$v(t) = A \cdot \cos(\omega_0 t + \phi(t)) \quad (32)$$

Where,

$\Delta\phi(t)$  is the phase deviation and

$$\omega_0 = 2\pi \cdot f_0 = \frac{2\pi}{T_0}, \quad f_0 \text{ center frequency and } T_0 \text{ is the period} \quad (33)$$

$$(32) \text{ can be rewritten as } v(t) = A \cdot \cos\left(\frac{2\pi}{T_0} \left(t + \frac{\Delta\phi(t)T_0}{2\pi}\right)\right) \quad (34)$$

Since the sine wave is periodic, then:

$$t = T_0 + \Delta T = T_0 \left( 1 - \frac{\Delta\phi(t)}{2\pi} \right) \quad (35)$$

The jitter is defined as:

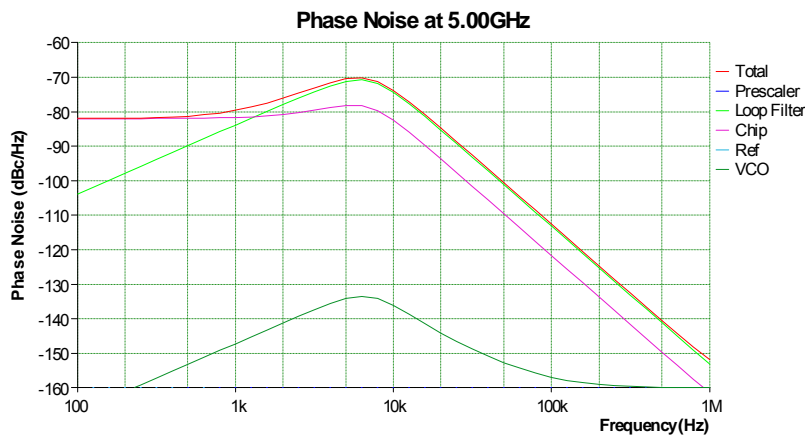
$$J = \frac{\Delta T}{T_0} = \frac{\Delta\phi(t)}{2\pi} \quad (36)$$

$S_\phi$  is defined the spectral density of phase fluctuations and is defined as the phase fluctuation in a 1Hz BW.

(36) can be rewritten as:

$$J = \frac{\sqrt{2 \cdot \int_{f_1}^{f_2} L_\phi df}}{2\pi} \quad (37)$$

where  $L_\phi$  is defined as the phase noise contribution. The phase noise at 5GHz is shown in Figure 4-13.



**Figure 4-13 Phase noise contribution at 5GHz**

Hence, the phase jitter is usually defined by the frequency at which the phase noise is measured. In our case, we calculated the phase jitter based on (37) with  $f_1=100\text{Hz}$  and  $f_2=100\text{KHz}$  offset from carrier.

### Phase jitter using brick wall filter

from 10.0kHz to 100kHz

**Phase Jitter** 1.03 degrees rms (38)

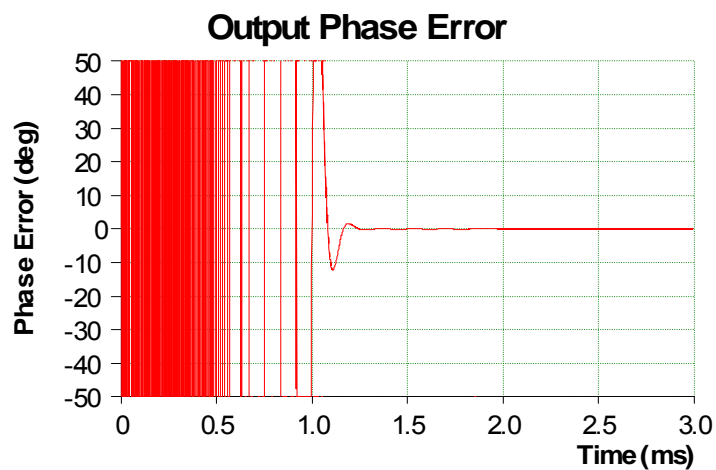
We then measure the transient response of the loop for a 4GHz step.

### Transient Analysis of PLL

Frequency change from 1GHz to 5GHz

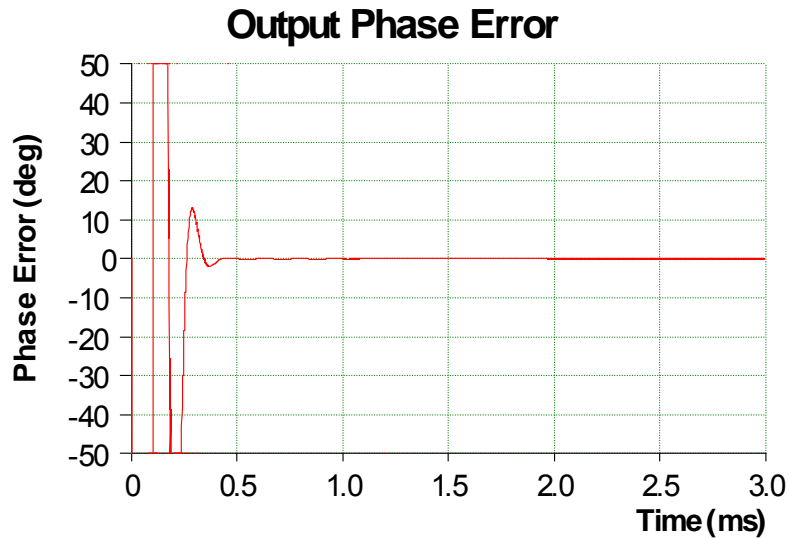
**Simulation run for 3.00ms** (39)

The locktime is shown in Figure 4-14. The locktime for a 4GHz jump is 1.10ms to within 1kHz of centre frequency.



**Figure 4-14 PLL Locktime based on a 4GHz jump (from 1GHz to 5GHz)**

The locktime for a 1Mhz jump is shown in Figure 4-15.



**Figure 4-15 PLL Locktime for a 1Mhz Jump (from 4.999GHz to 5.00GHz)**

For a single channel jump (from 4.999GHz to 5.0GHz), the lock time is 0.6ms shown above.

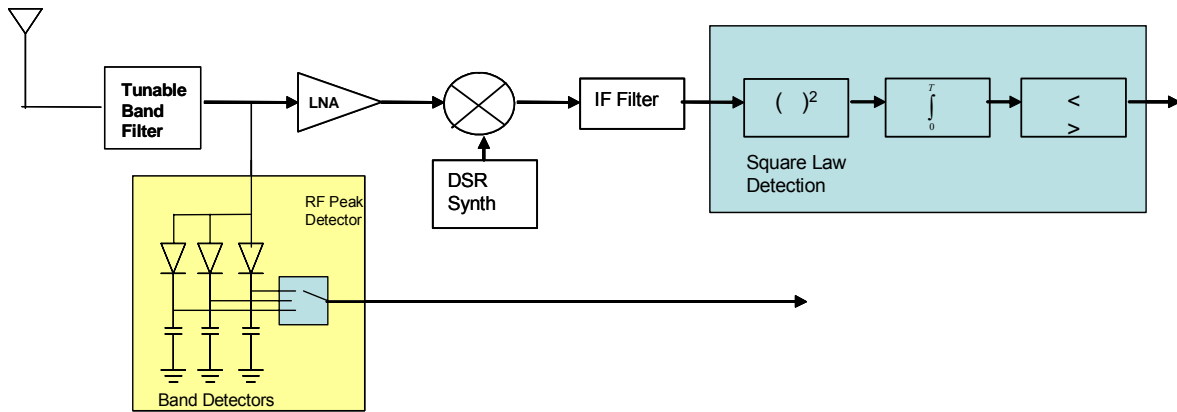
For the purposes of our analysis, we would use the locktime and phase jitter from this example. These numbers are practical and feasible in today's technology.

To put these numbers in perspective, assume we want to scan a 1GHz band in 1MHz steps. The delay required to sweep the entire band is therefore 6s ( $1000 \times 0.6$ ). The 6s does not take into account time required for detection or FFT calculations. This simple calculation alone proves the challenge that faces CR sensing receiver. The question then becomes, does every channel need to be scanned? This question is answered in the subsequent sections.

#### **4.2.1.2 Analog Domain Sensing**

Signal detection may be achieved either in the analog or digital domain. In general, the detection time is lengthened as the signal is processed and converted through the receiver chain. In order to optimize its sensing performance, the DSR distributes the workload throughout the receiver. In other words, the DSR attempts to detect power at RF, IF (if available) and at baseband in the analog domain prior to yielding to energy detection in the digital domain as discussed in section 2.3.2. The advantages of analog domain detection are 1) fast response 2) relatively low cost and

3) relatively low power consumption. However, analog domain detection does not work well at low SNR. It can not differentiate among received signals and the frequency detection accuracy is mediocre at best. For the purposes of the sensing receiver and the DSR, the analog domain detection can be used to do a cursory search for the presence of energy in a band. The proposed implementation is shown in Figure 4-16.



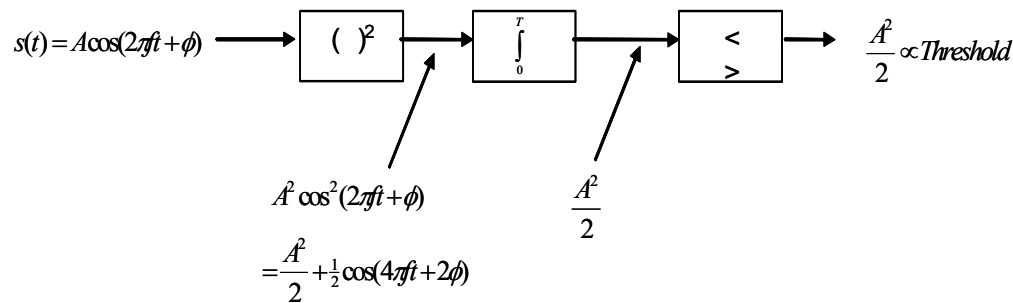
**Figure 4-16 Analog Domain Energy Detection**

The implementation of analog domain detectors consists of a bank of diodes at RF. Different diodes with different threshold levels are used for different frequency and amplitude levels. RF peak detectors effectively follow the current-voltage characteristic curve or IV curve of a Schottky diode. They are cheap, effective and easily integrated on an IC. Peak detectors are usually calibrated because of non linear characteristics of the diode (one advantage of using RF detector in CRs is that calibration may be minimal or not required since the radio is mostly interested in relative energy level). The measured voltage output of a peak detector is given by [92]:

$$V = K(\sqrt{P})^\alpha \quad (40)$$

where V is detector voltage output, K is a constant characterizing the diode, P is the detected power and  $\alpha$  is a variable dependent on the biasing condition of the diode.

The non-linearity of the detector is reflected in  $\alpha$  which is a variable that depends on biasing conditions of the diode. As the signal power increases, the diode is more linearly biased and hence  $\alpha \leq 1$ . However, at low power level, the diode operates in the square law region of the diode or  $\alpha = 2$ . Hence, peak detectors are not very practical at low power level because of the voltage required to turn on the diode. Hence, a detector more tuned to operating in the square law biasing region of the diode is needed. Simplistically, a square law detector can be seen as multiplying a signal by itself and then using the integrator to remove the high order term, hence keeping the DC term to be compared against a threshold.



**Figure 4-17 Square Law Detector**

where  $s(t)$  is the received signal,  $f$  is the frequency and  $\phi$  is a phase shift.

By allowing power detection decisions at different levels in the chain, it provides the following advantages:

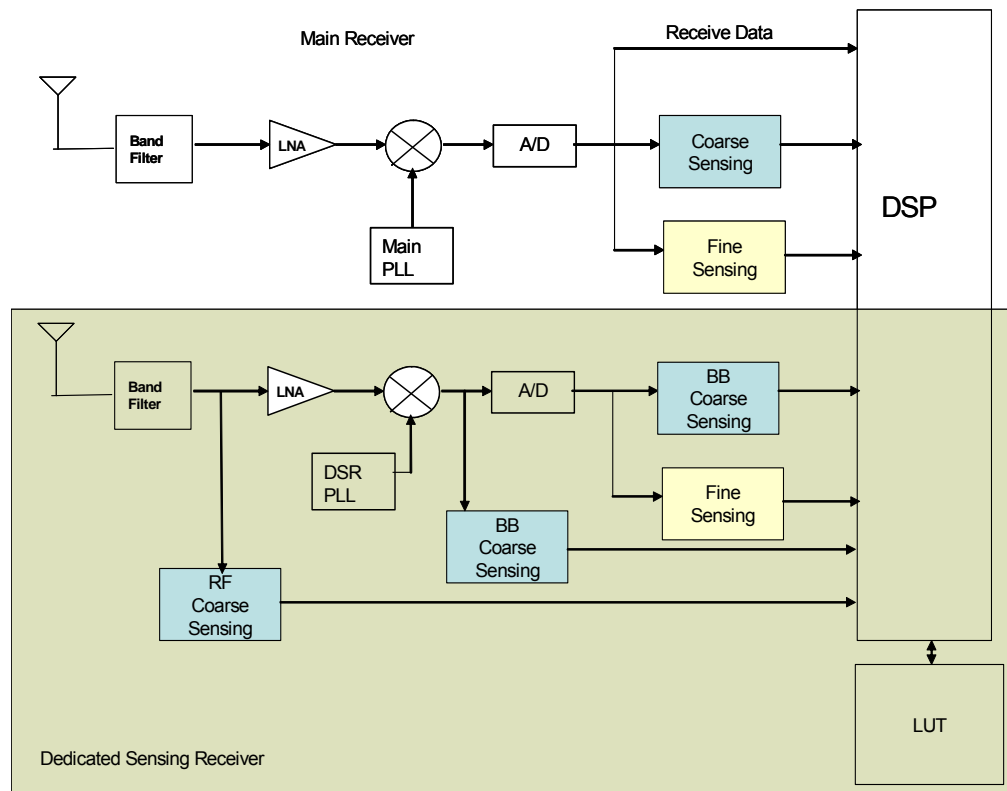
- 1- distribute sensing where the workload is distributed between the analog and digital domains
- 2- faster detection as the decisions are localized
- 3- broadband detection since the diodes are fairly broad and would detect power across a relatively wideband.
- 4- protection against interference as the analog domain detector may be used to quickly detect a potential PU or SU presence and may request interruption



Although not very accurate, power detection in the analog domain is still very effective for the purposes of CR since the absolute power level is not as critical as a fast decision on the presence of users in the band.

#### 4.2.1.3 DSR Model

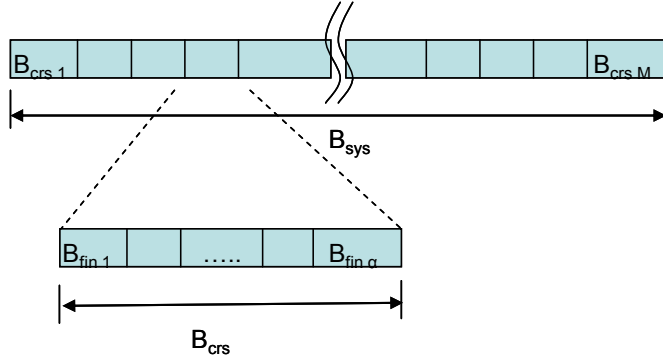
Based on the addition of analog and local sensing, the CR shown in Figure 4-10 is now expanded to the architecture shown in Figure 4-18.



**Figure 4-18 Dedicated Sensing Receiver**

The sensing function is mostly provided by the dedicated sensing receiver. However, the main receiver is also capable of assisting with frequency sensing. As long as the CR is not scheduled to transmit or is not presently in communication, the main receiver provides support.

In order to scan the spectrum, the overall bandwidth is divided into frequency bins. The DSR performs a multi resolution sensing i.e. coarse and fine bins. Assume that the overall bandwidth to be scanned for users is  $B_{sys}$ .  $B_{sys}$  is divided into coarse bins and the coarse bins are further divided into fine bins as shown in



**Figure 4-19 Coarse and Fine Frequency Bins**

Multi-resolution sensing has proven to be beneficial in broadband sensing. Neihart et al. [94] derived an estimated improvement on processing time for a multi-resolution sensing technique for multiple antenna system in CR operations. However, the work is limited and it does not address different sensing modes and it does not optimize the frequency search algorithm.

In our implementation, the overall system bandwidth  $B_{sys}$  is divided into frequency bins.  $B_{sys}$  is divided into coarse sensing bins  $B_{crs}$  and fine sensing bins  $B_{fin}$ , where  $B_{crs}$  is a multiple integer of the  $B_{fin}$ .

$$B_{crs} = \alpha B_{fin} \quad , \text{ where } \alpha = 1, 2, 3, 4, \dots \quad (41)$$

The resolution of the estimation is proportional to  $N$ . Hence, the resolution increases as  $N$  increases. For fine sensing:

$$B_{fin} = N.F_{res} \quad (42)$$

where  $F_{res}$  is the resolution of the sensing.

In practical implementations, FFTs have widely used the split-radix FFT algorithm [51]. The number of real additions and multiplications needed for a  $2^N$  points FFT (with  $N > 1$ ) is given by:

$$\text{number of real operations} = 4N \log_2 N - 6N + 8 \quad (43)$$

The total time to perform a discrete Fourier Transform (DFT) is given by:

$$T_{DFT} = \frac{1}{F_{DSP}} (4N \log_2 N - 6N + 8) \quad (44)$$

where  $F_{DSP}$  is the DSP operating frequency. For simplicity, assume that the DSP is capable of performing one addition and one multiplication per clock cycle, the total sensing time for coarse and fine sensing of the total bandwidth is given by:

and fine sensing of the total bandwidth is given by:

$$T_{crs} = \frac{B_{SYS}}{B_{crs}} T_{DFT} \quad (45)$$

Combining (41), (44) and (45) and assuming that in coarse mode,  $M$  receivers are sharing the sensing load; we can write  $T_{crs}$  as

$$T_{crs} = \frac{B_{SYS}}{\alpha M N_{crs} F_{res} F_{DSP}} [4N_{crs} \cdot \log_2(N_{crs}) - 6N_{crs} + 8] \quad (46)$$

In fine mode, the total fine processing time is given by:

$$T_{fin} = \alpha T_{DFT} \quad (47)$$

Combining (41), (44) and (47),  $T_{fin}$  is defined as:

$$T_{fin} = \frac{\alpha}{F_{DSP}} \left[ 4N_{fin} \cdot \log_2(N_{fin}) - 6N_{fin} + 8 \right] \quad (48)$$

The total time  $T_{DFT\_SYS}$  needed to perform the FFT operation across coarse and fine sensing modes is given by (49):

$$T_{DFT\_SYS} = \frac{B_{SYS}}{\alpha M N_{crs} F_{res} F_{DSP}} \left[ 4N_{crs} \cdot \log_2(N_{crs}) - 6N_{crs} + 8 \right] + \frac{\alpha B_{SYS}}{B_{crs} F_{DSP}} \left[ 4N_{fin} \cdot \log_2(N_{fin}) - 6N_{fin} + 8 \right] \quad (49)$$

In order to complete the overall system sensing time, we need to include the radio tuning time which is mostly dominated by the PLL lock times. Let's define three different PLL locks times:  $T_{init}$  which is the initial lock time,  $T_{PLL\_crs}$  which is PLL lock time for a coarse step and  $T_{PLL\_fin}$  which is PLL lock time for a fine step. Hence, the total PLL sweep time  $T_{PLL\_SYS}$  during the sensing operation is give by:

$$T_{PLL\_SYS} = T_{init} + \alpha \beta T_{PLL\_fin} + \beta T_{PLL\_crs} \quad (50)$$

Where,

$$\beta = \frac{B_{sys}}{B_{crs}} \quad (51)$$

From (49) and (50), the total sensing time  $T_{sys}$  is given by:

$$\begin{aligned}
T_{SYS} &= \frac{B_{SYS}}{\alpha M N_{crs} F_{res} F_{DSP}} [4N_{crs} \cdot \log_2(N_{crs}) - 6N_{crs} + 8] + \\
&\dots \frac{\alpha \cdot \beta}{M \cdot F_{DSP}} [4N_{fin} \cdot \log_2(N_{fin}) - 6N_{fin} + 8] + T_{init} + \frac{\alpha \beta}{M} T_{PLL\_fin} + \frac{\beta}{M} T_{PLL\_crs}
\end{aligned} \tag{52}$$

$$\begin{aligned}
T_{SYS} &= \frac{B_{SYS}}{\alpha M N_{crs} F_{res} F_{DSP}} [4N_{crs} \cdot \log_2(N_{crs}) - 6N_{crs} + 8] + \\
&\dots \frac{\alpha \cdot \beta}{M \cdot F_{DSP}} [4N_{fin} \cdot \log_2(N_{fin}) - 6N_{fin} + 8] + T_{init} + \frac{\alpha \beta}{M} T_{PLL\_fin} + \frac{\beta}{M} T_{PLL\_crs}
\end{aligned} \tag{53}$$

As discussed earlier, the lock time of a PLL is dependent on several variables such as loop bandwidth, loop order, filter implementation (active or passive), step size, etc. Hence in section 4.2.1.1, we simulated a PLL lock time that is dependent on its implementation. In addition, we expand equation (52) with a term  $\rho$  whereas  $\rho$  is defined as the percentage of coarse bins that are identified as available after coarse sensing. In other words, if the entire band was available then  $\rho = 1$  which means that every coarse bin must be submitted for fine sensing. Conversely,  $\rho = 0$  means that the coarse sweep identified that all bins are occupied and hence no need for fine sensing. In this case, continue coarse sensing until  $\rho > 0$ . Equation (52) becomes:

$$\begin{aligned}
T_{SYS} &= \frac{B_{SYS}}{\alpha M N_{crs} F_{res} F_{DSP}} [4N_{crs} \cdot \log_2(N_{crs}) - 6N_{crs} + 8] + \\
&\dots \frac{\alpha \cdot \beta \cdot \rho}{F_{DSP} \cdot M} [4N_{fin} \cdot \log_2(N_{fin}) - 6N_{fin} + 8] + T_{init} + \frac{\alpha \cdot \beta \cdot \rho}{M} T_{PLL\_fin} + \frac{\beta}{M} T_{PLL\_crs}
\end{aligned} \tag{54}$$

We use the formula in (53) above in the following section to determine the scan time of the system.

As a next step, we define the accompanying algorithm to be used with the DSR.

## ***4.2.2 The DSR Algorithm***

The DSR is charged with detecting the presence of users on specific channels with a goal to reduce sensing time and to avoid false positives. The DSR makes use of several architectural implementations and algorithmic techniques to reduce the detection time. As discussed in section 4.2.1, analog and local detection techniques were added to accelerate identifying a suitable CR channel while protecting against interference.

In this section, we define the algorithm used and number of FFT points needed at each level of the detection. More importantly, we introduce a concept of initializing the sensing function with available input such as geographical location, time of operation, or a priori information about the channels or users. The goal is to leverage as much information as possible in order to reduce sensing time. We focus on the total solution and operational sensing times.

### ***4.2.2.1 A Priori Algorithm***

One aspect of the algorithm is to leverage a priori information about the use model of the CR and channel conditions. In order to assist the initial scan time at power on, initial conditions are pre-loaded into the LUT. Initial conditions direct the CR to avoid known frequencies where CR operation is either forbidden to operate or unfavorable. Initial setup is dependent on geographic location, operation times and known channel conditions. For example, if the user is operating in the US, then the CR may apply the 802.22 guidelines for CR operations in the TV bands. Additionally, the CR may be directed to avoid the cellular or military bands. Alternatively, this information is provided to the user from collaborating PU, where the PU provides authorized and unauthorized time windows for CR operation in a licensed channel.

Intuitively, it is obvious that an initial setup reduces the scanning time and may improve the quality of decision making by avoiding false positives. The CR continuously refreshes the information in the LUT with information about the channel and the user. The best time to scan and update the LUT is during off-peak hours (such as the middle of night).

The a priori algorithm may be improved by allowing the sensing receiver to demodulate certain control channels. For example, assume that the DSR is scanning a cellular band. The DSR may listen to control channels of the cellular basestation and may “know” what channels may or may not be occupied by the cellular PU. This idea may be extended to other standards such as access points or TV stations. The disadvantage of such technique is that the DSR is now required to perform full demodulation of the received signal which adds delay, complexity, and cost versus simple energy detection. The advantages depend on the particular use of the CR.

#### ***4.2.2.2 Coarse Sensing***

Coarse scanning relies on the initial data and a priori information about the channel or user preferences that are saved in the LUT described in section 4.2.2.1. The sensing receiver detects the presence of other users in the band and classifies them as either primary or secondary users. A simple matched filter is used in the case of coherent detection of a known signal. However, in the case where the transmitted signal is not known, a non-coherent detector is needed. The challenge of non-coherently detecting an unknown signal is amplified when the receiver is operating at low Signal to Noise Ratio (SNR). The presence (or absence) of a weak signal is especially difficult when the signal does not contain a deterministic component [93] which is frequently the case of CR operation. The CR must be able to reliably detect the presence (or absence) of weak signals with unknown modulation, frequency, phase, and amplitude in order to avoid interference with other users of the band. In our realization, we consider the use of spectrum matching detector (SMD) which is suited for detecting signals with low or negative SNR. The SMD technique suffers from two shortfalls: (1) it requires knowledge of the power spectral density (PSD) of the signal which is cross correlated with the received signal, and (2) it is computationally intensive. An approximate version with slight degradation, called Approximate SMD (ASMD), is based on the periodogram and can be implemented efficiently by the discrete Fourier transform [95]. In our proposed DSR realization, we recommend the use of coherent detectors in bands where the PU is well known and the use of ASMD when the PU is unknown. The a priori initialization provides information to tenants in the channel so the coarse sensing algorithm makes a decision about coherent or ASMD detection. For example, if the

DSR is scanning in the cellular band, the CR is looking for a PU that complies with published standards. The scanning performance and results are discussed in the upcoming sections.

#### ***4.2.2.3 Fine Spectrum Sensing***

After completion of the coarse scan, the DSR prioritizes a list of candidate frequencies for use by the main receiver. The main receiver performs fine sensing on the smaller set of prioritized channels. The DSR continues on its wideband coarse scan of the remaining spectrum. Because of the need for fast scanning, the DSR uses non coherent energy detection to complete its coarse scanning before it reverts to fine sensing.

#### ***4.2.2.4 Coarse versus Fine Spectrum Sensing***

After the completion of coarse sensing, the DSR prioritizes the available frequency bands. Both the main receiver and the DSR engage in fine spectrum sensing. The FFT is performed on a captured segment of the frequency spectrum. The FFT resolution increases with the FFT size which also increases the observation time. FFTs are computationally intensive, and increased resolution requires longer observation time, higher power consumption and higher cost (more gates if implemented in hardware). Lim et al [96] compared the implementation of an FFT fully in TMS320C6416 DSP from Texas Instruments and the benefits of adding a hardware implementation co-processor implemented in an FPGA. Typical performance for a 1,024-point 16-bit complex FFT running on a 720 MHz C64x family DSP is about 6,526 cycles or about a 9.06  $\mu$ s transform time and 1.2W (for internal logic only). In contrast, an FFT implementation that utilizes an FFT accelerator implemented on Altera Stratix FPGA requires less than 4.6  $\mu$ s transform time while only 884mW. There are several combinations for implementing a sensing receiver that heavily depends on its parameters and its uses. The DSP running frequency and the number of FFT points are contributors to the overall system delay, radio cost and power consumption.

In order to minimize sensing time and power consumption, we use an adaptive FFT with the number of FFT points varying according to 1) operating conditions and 2) the urgency to complete the scan. The algorithm uses one of 4 FFT resolutions – 256, 512, 1024 and 2048



points respectively. The resolution is triggered by information stored in the LUT. In our implementation, we used the following algorithm:

| Number of FFT points | Condition 1   | Condition 2                                     | Comments   |
|----------------------|---|---|--|
| 256                  | Initial setup identified this band as not available | Analog detector is positive                     | This condition is used to 1) confirm the presence of signal in a band , 2) define the bandwidth, and 3) monitor and record the activity of a user  |
| 512                  | Analog detector is positive                         | Coarse sensing is detecting a narrowband signal | This condition is used to 1) confirm the presence of a signal in the band that has been identified in coarse scans, 2) define the bandwidth of the signal, and 3) monitor and record the activity of a specific user |
| 1024                 | Analog detector is negative                         | Coarse sensing is negative                      | No users were identified, use this scan to confirm that no signal is present   |
| 2048                 | Coarse sensing is negative                          | 1024 point FFT is negative                      | This is the last check to confirm that the band is available. After the 1024 point FFT scan is negative, this mode is used to avoid false positives.   |

**Table 7 DSR FFT Resolution Conditions**

After the fine tuning is completed and a candidate frequency band is identified, the sensing receiver continuously monitors and records activities in the band of interest.

### 4.2.2.5 Overall Algorithm

With sensing modes and techniques described above, an adaptive algorithm is used to quickly and efficiently determine the availability of a spectrum. A sensing algorithm that benefits from the distributed sensing is described in Figure 4-20.

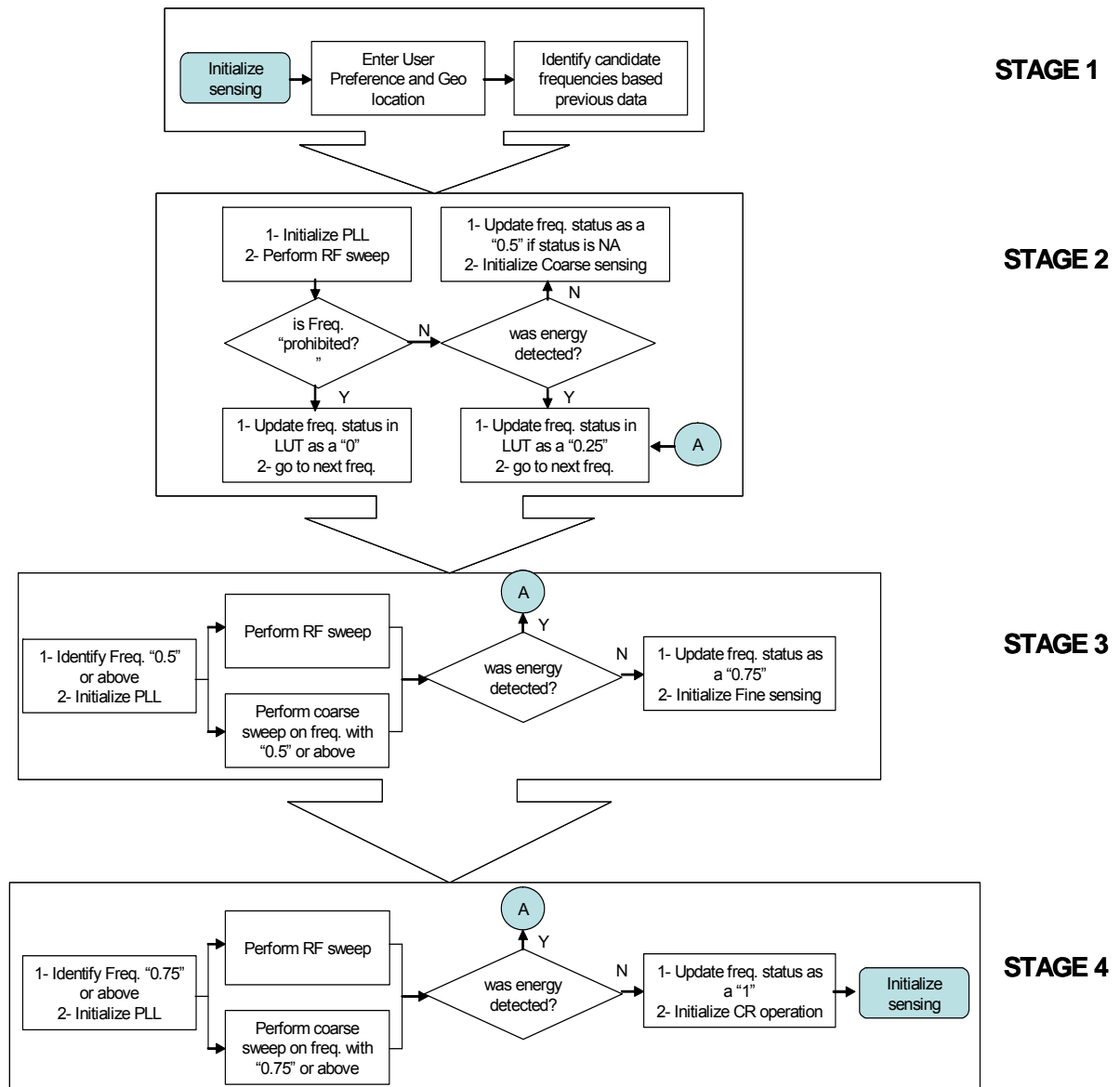


Figure 4-20 DSR Algorithm

At the core of the algorithm is the ability to continuously update a select number of candidate frequencies for CR operation and learn over time the best bands for CR operation. The algorithm adds a merit factor to each frequency and the ability to continuously monitor the presence of an interferer. Continuous monitoring is done via RF detectors in the analog domain because of its fast response time. As described in previous sections, the disadvantage of RF detectors may be its undesired ability to detect broadband signals which may falsely record energy outside the desired band of interest. This issue can be mitigated by adding filtering. However, the addition of filtering is counter productive, unpractical, and costly. In our implementation, we focus on recording relative energy or simply changes in energy in the band. Relative energy for each band is recorded and saved along with other parameters in the LUT.

As each frequency band is scanned, a figure of merit is attached to it and continuously updated. The DSR assigns 0, 0.25, 0.5, 0.75 or a 1 weight. A “0” associated to a frequency means that the CR is prohibited of operating in the band, while a “1” defines a frequency band at which the CR is cleared to operate. As shown in Figure 4-20, the figure of merit increases as the DSR moves from stage 1 through stage 4. At each stage, the resolution of the frequency detection increases as the sensing bandwidth decreases. The ability of the algorithm to quickly discard large frequency band as non-candidates via the initial setup or RF scanning is important since the CR does not waste time and power on known non-candidates. Obviously, this algorithm can be easily augmented with additional levels of sensing. As an example, if a cyclostationary feature sensing is added to the sensing receiver prior to initiation of CR operation, a “0.85” figure of merit may be added to the sensing weights.

The algorithm above takes full advantage of the combined benefits of sensing at the different stages of the receiver. In addition, it is able to localize some of the decision making close to the sensing point. The advantages of this technique are:

- 1- The proposed approach takes advantage of the availability of two antennas and two receivers (Main and DSR) in essence providing the benefit of diversity during sensing.
- 2- Better decision making: with a dedicated channel, the sensing receiver continuously senses and learns without the added complexity or urgency to send and receive data packets. The sensing receiver scans the desired frequency bands and records the frequency

bands where a CR is allowed to operate. Assuming the CR may coexist with a PU, the receiver learns and generates a prediction algorithm of the probability of PU appearance or interference with other CRs.

- 3- Faster scan: the sensing receiver prioritizes and saves the list of “best” channels in the look up table. In addition, the look up table may save the “preferred channels” where the CR has been historically operating. Hence, upon power up of the radio at stage 1, the main receiver searches the “preferred channels first” prior to doing a broadband sweep.
- 4- Built-in diversity leverages the availability of two receiver (main radio and DSR) chains and two antennas. Diversity is an effective tool to combat fading and multipath in the wireless channel. Although the DSR is dedicated to sensing at all times, the main receiver assists the DSR in 1) coarse sensing after the initial scan, and 2) fine sensing once an initial list of candidate frequency bands are identified for CR operation or during “quite” time.

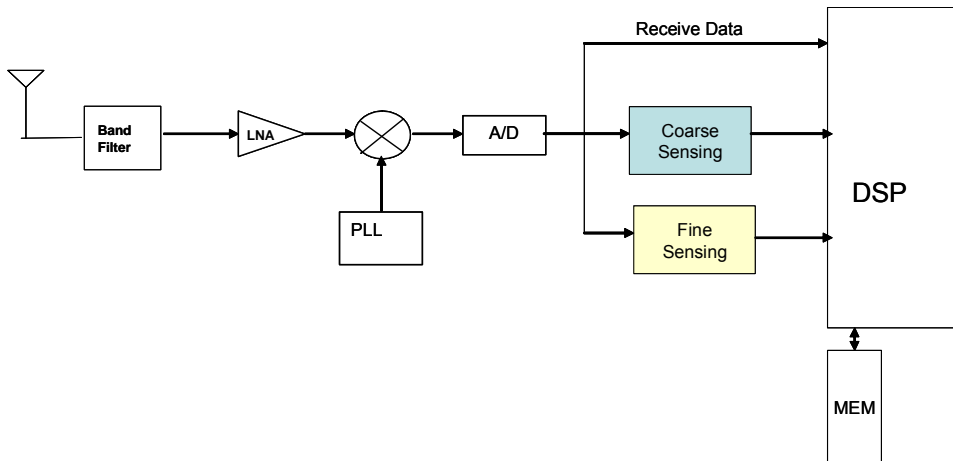
In order to take full advantage of the DSR and its algorithm, the radio architecture and especially the phase locked loop must be able to quickly hop and settle onto a desired frequency. Without an agile PLL, the system scan time would be gated by the radio hardware. The overall PLL design is critical to the performance, cost, and complexity of the CR specifically across wideband operation.

### ***4.2.3 The DSR Results***

In this section, we present the results of a base CR with and without DSR. The results are presented in the next two sections.

#### ***4.2.3.1 Reference Architecture***

For reference, a typical CR receiver without a dedicated sensing receiver chain is shown in Figure 4-21.



**Figure 4-21 Typical Cognitive Radio Receiver**

The receiver in Figure 4-21 is used as a reference for sensing delay and cost improvements. The simulation parameters are shown in

|  |              |
|--|--------------|
| Delay goal   | <20ms        |
| System frequency of operation  | 0 - 10GHz    |
| Bcrs = 100 MHz   | 100MHz       |
| Bfin = 10 MHz  | 10MHz        |
| DSP Freq (TI TMS320C55x Family) –<br>F <sub>DSP</sub>                    | 100MHz       |
| Sensing Resolution – F <sub>res</sub>                                    | 10KHz        |
| VCO control voltage  | 1V           |
| PLL Phase Jitter max   | 1.0 deg. rms |
| Percentage yield of candidate channels<br>after Coarse Sensing - $\rho$  | 40%          |
| Percentage of coarse bins known as<br>“bad” channels                     | 30%          |
| Number of fine channel scan between<br>frames during normal CR operation | 100          |

**Table 8 Simulation Parameters**

Meeting all the design constraints is difficult and in some cases impossible with commercially available components. For example, VCOs from 0 – 10GHz are not commercially available. In addition, it is difficult to keep lock of the PLL within the phase jitter requirements stated above. To maintain a PLL jitter of 1.0 deg. rms, the loop bandwidth is reduced and hence the lock time increases. The initial lock time is simulated to be 1.1ms. (to within 1Khz of desired frequency). For a single coarse channel jump of 100MHz step, the lock time –  $T_{PLL\_crs}$  is 0.6ms per channel and for a single fine step of 10MHz step,  $T_{PLL\_fin}$  is 0.35ms. See Appendix B for additional data.

For the CR receiver in Figure 4-21, we define the number of coarse bins as:

$$\beta = \frac{B_{sys}}{B_{crs}} = \frac{10Ghz}{100Mhz} = 100 \quad (55)$$

and,

$$\alpha = \frac{B_{crs}}{B_{fin}} = \frac{100MHz}{10Mhz} = 10 \quad (56)$$

Hence, the total time to scan the entire bandwidth in coarse mode is composed of the initial lock time added to the scan time for each band multiplied by the number of bands in (54). The total scan time in coarse mode is given by:

$$\text{Total } T_{PLL\_crs} = \beta \cdot T_{PLL\_crs} + T_{init} = 100 * 0.6ms + 1.1ms = 61.1ms \quad (57)$$

Similarly, the scan time for fine mode is calculated.

$$\text{Total } T_{PLL\_fin} = \beta \cdot \alpha \cdot T_{PLL\_fin} + T_{init} = 100 * 10 * 0.35ms + 1.1ms = 351.1ms \quad (58)$$

The total scan time in fine mode contribution is directly proportional to the number coarse bins candidates that were generated in the coarse scan mode (i.e.  $\rho$ ). In this example,  $\rho = 0.4$ .

The additional processing time due to the FFT processing is also calculated. For simplicity, use  $N_{\text{crs}} = 128$  points and for fine mode  $N_{\text{fin}} = 1024$  points.

| Mode   | FFT points – N | Number of operations | DSP Frequency (MHz) | Time (ms) |
|--------|----------------|----------------------|---------------------|-----------|
| Coarse | 128            | 2824                 | 100                 | 0.02824   |
| Fine   | 1024           | 34824                | 100                 | 0.34824   |

**Table 9 FFT Delay Contribution**

Hence, the total scan time in the above mode depends on the number of bands available through the coarse mode.

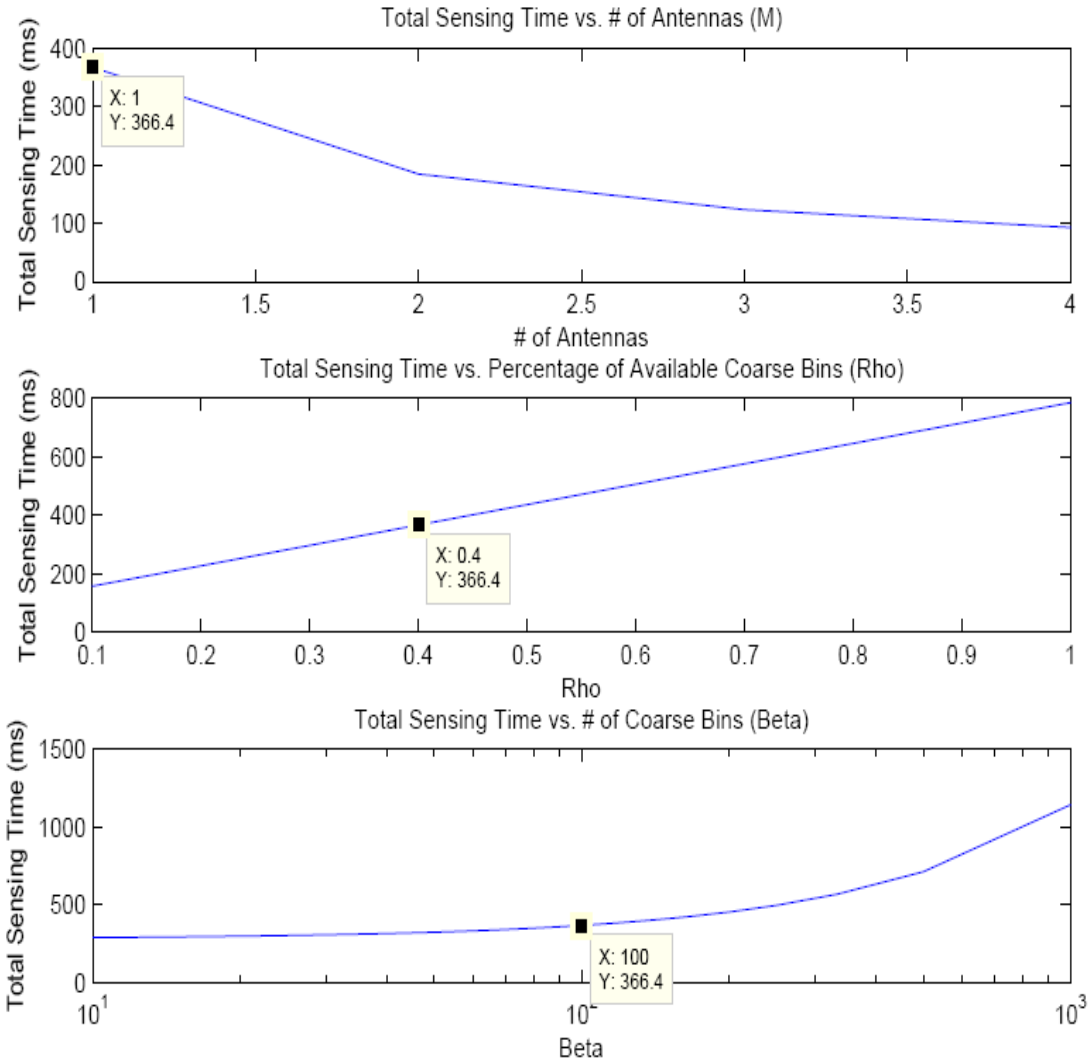
The total scan time is equivalent to sum of the total coarse scan, the total fine scan, total coarse FFT ( $100 \times 0.028$ ), and total fine scan time ( $10 \times 0.34 \times 100$ ). The results are shown in

$$T_{\text{sys}} = T_{\text{PLL}_{\text{crs}}} + 0.2824\text{ms} \times 100 + T_{\text{PLL}_{\text{fin}}} + 10 \times 100 \times 0.34\text{ms} = 755.2\text{ms} \quad (59)$$

From equation (59), 755.2ms represents the total time required prior to initiating CR operations which is assuming that the base CR is able to reduce the number of fine scans after the coarse scan by 60% per Table 8 ( $\rho = 40\%$ ). Since the bands are not contiguous, we assume there is a 20% overhead on the lock time.

$$T_{\text{sys}} = 1.2 \times (T_{\text{PLL}_{\text{crs}}} + 0.2824\text{ms} \times 100 + (T_{\text{PLL}_{\text{fin}}} + 10 \times 100 \times 0.34\text{ms}) \times 0.4) = 366\text{ms} \quad (60)$$

One important aspect to better understand these results is to get a feel for the sensitivity of these results versus various parameters.



**Figure 4-22 T<sub>sys</sub> Sensitivity**

Figure 4-22 provides the sensitivity of  $T_{sys}$  versus the number of antennas ( $M$ ), the number of frequency candidate after coarse sensing ( $\rho$ ), and the number of coarse bins ( $\beta$ ). The (x,y) points highlighted on each graph represent the total sensing time required for the conditions set in the above examples. Hence,  $T_{sys}$  in the base configuration is 366ms. Prior to initiating CR operation, the radio must go through a final fine sweep to eliminate any false positives. The sensing phase (equivalent to stage 4 in Figure 4-20) is assumed to hold a total of 100 channels which are assumed not to be contiguous.



Now assume that the CR must revalidate the channel availability during normal CR operation. Based on the assumption in Table 8, the CR must cease operation and fine sense a total of 100 channels which are assumed not to be contiguous. The channels are chosen based on the update LUT ranking of available channels. Hence, the delay during CR operation in fine mode is the sum of the initial PLL setup time,  $\alpha * T_{PLL\_fin}$ , the computation time for a 1024points FFT and initial PLL setup time (PS: the two “initial PLL setup times” are needed because the PLL must unlock and relock at the beginning of the scan and then again at the end). During this operation, the CR is not authorized to transmit or receive data.

$$\text{Total } T_{PLL\_fin} = 1.1\text{ms} + 100*(0.6+0.35) \text{ ms} + 100*(0.34824) + 1.1\text{ms} = 132 \text{ ms} \quad (61)$$

Total delay due to channel sensing during CR operation is 132ms which is much higher than the maximum target delay of 20ms.

#### ***4.2.3.2 DSR Architecture***

In order to accelerate initial CR operation, the proposed architecture shown in Figure 4-18 adds additional detectors at RF in addition to a regularly updated Look Up Table (LUT). Besides the added peak and energy detectors in the DSR, a key addition is a regularly updated LUT. As discussed earlier, the LUT is initially loaded with either good channels to search or bad channels to avoid based on user information (i.e. geographical location, time of day, priority....) that is constantly updated based on a self learning algorithm. The self learning algorithm monitors the channel conditions in presence of PUs such as jammers versus time, locations and number of users. In this example, we assumed that the percentage of bad channels to avoid is 20%. Such channels in the 10Ghz spectrum include several military and emergency bands, VHF / UHF TV, upper and lower cellular bands, air to ground, fixed microwave, and ISM bands,

As the CR is enabled, it performs an initial setup and fast sensing on the known good channels as listed in the LUT. For example, the LUT may recommend the CR to avoid the cellular band, the ISM bands, and the known military bands in the 0 – 10Ghz spectrum. Hence, a quick search

which is done via the analog detector, only considers 70 good bins. The total sweep time of the 70 channels is 43.1ms. However, since we have two receivers, each receiver will scan half of the band in parallel. Hence, the total time to coarse scan the entire bandwidth is 21.5ms. From Table 8, considering that 40% (or 28 coarse bins) of the measured channels are considered “good,” then the fine sensing effort is equally divided by the main receiver and the DSR. The total sensing time given by equation (53) is  $T_{sys} = 91.1ms$ .

In this example and simulation parameters, it is highly probable that with the help of a priori information, the CR may begin operation after 91.1ms of sensing compared to 755.2ms under the base model. Thanks to the presence of a dedicated sensing receiver, there are no delays during the normal CR operations because the DSR performs the sensing while the main receiver is operational.

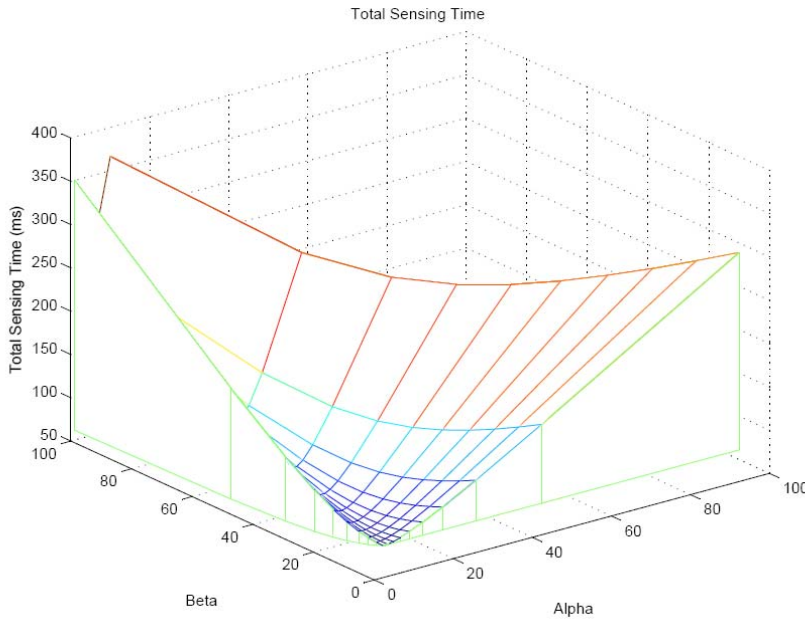
Based on the simulation parameters listed in Table 8, the summary improvement for using the DSR is shown in Table 10.

| <b>Operation</b>                                       | <b>Base Architecture</b> | <b>Proposed Architecture</b> |
|--|--------------------------|------------------------------|
| Total sensing time before initial CR operation         | 755.2ms                  | 91.1ms                       |
| Sensing delays during CR operation (100 fine channels) | 131ms                    | 1.7ms max                    |

**Table 10 Simulated Improvement Base vs. DSR**

Although the initial sensing time shows 88% improvement, the most important result is described as “Sensing Delays during CR operations.” The results show that the base architecture delays are well over the 20ms requirements and hence voice operation in the base architecture is not feasible. In the proposed approach, the DSR continuously scans for available bands and updates the LUT used by the main receiver. In other words, the receiver chooses the best channel as estimated by the DSR. The maximum delay is due to  $T_{PLL}$  tuning to a new frequency band and the initial setup  $T_{init}$ .

These results and improvements are sensitive to the choice of the bin size. As the number increases, there is a better probability of the detecting a user. However, if the number of bins increase, the total sensing time becomes unacceptable. These results can be further improved as  $\beta$  and  $\alpha$  are further optimized. In Figure 4-23, we show the effect on the DSR sensing time versus  $\alpha$  and  $\beta$  are modified (for this example).



**Figure 4-23 Sensing Time vs. Number Coarse and Fine Bins**

In Figure 4-23,  $\alpha$  and  $\beta$  can be further optimized to improve sensing time. However, one must balance detection probability, probability of false alarm versus sensing time. The optimization of these results will be further discussed in the next chapter.

#### **4.2.4 DSR Incremental Cost**

The DSR offers the ability to improve performance of a CR by aggressively addressing the reduction in sensing time. By adding a new receiver chain, one must be cognizant of added complexity and cost to the realization of the circuitry. The impact of adding circuitry for a DSR

logically suggests additional cost. However, the theory does not always hold true for wideband and agile receivers. The overall cost is highly dependent on key specifications of each component. As specifications become harder to achieve and less main stream (not produced in high volume), the economy of scale suggests that cost would be higher.

The cost of the base architecture and the DSR proposed architecture is estimated using catalog part numbers and retail prices as listed in vendor catalogs. In the case that a component is not found in standard catalogs, the cost is estimated based on closest alternatives. In reality, the overall cost of the system depends on volumes, supply channels, and procurement ability of each company. We take a snapshot of available cataloged parts in order to perform a relative comparison. In Table 11, we show a summary of the analysis and the reasoning behind choosing a specific design point.

| Component | Option 1 - no DSR   | Option 2 - With DSR  |   |
|-----------|---|--|---|
|           |   | Main Receiver  | Sensing Receiver  |
| LNA       | the dynamic range must be large enough to handle the sensing range while the the noise figure must be low enough to improve sensitivity   | With a DSR, the main receiver is provided information about the signal amplitude seen by the front end. An attenuator can be used in the front end to attenuate large signals and disabled for small signals- Hence, the DR is smaller and the NF are easier | Since demodulation of the signal is not required, noise figure may be relieved  |
| Mixer     | Because of the high DR requirement, the VCO must be able to handle large signals. A high IP3 mixer is required. If not, the mixer goes into saturation and distorts the received signal | Because of the attenuator in the front end (sometimes shutting off the LNA will work as an attenuator), the IP3 of the mixer is lower  | Because demodulation is not a requirement, the sesing receiver may get away with some non-linearity distortion from the mixer - hence a lower IP3 is acceptable |

**Table 11 Example of Cost Analysis and Reasoning Done on DSR**

Broadband components are not readily available and hence high dynamic range and wideband components tend to be very expensive. Components that are used in consumer products such as cell phones, portable multimedia or computers tend to be produced in high volumes and they are usually lower cost.

In order to reduce complexity, we compare the cost of the key components only. Reference receiver is shown in Table 12 and our proposed architecture is presented in Table 13.

| Component    | Baseline Solution                                       |   |   |  |           |     |                  |
|--------------|---|---|---|--|-----------|-----|------------------|
|              | Required Spec   | Available Spec  | Part Number                                 | Manufacturer                             | Unit Cost | Qty | total Cost       |
| LNA          | Gain = 30dB<br>NF < 3 dB<br>F=0 - 10Ghz<br>RF= 0 -10GHz | Gain = 12.5dB@5GHz<br>NF= 3.2dB<br>@5GHz<br>F=0 - 6Ghz<br>RF=0-3.1GHz | ZX60-6013E+                                 | Minicircuit                              | \$ 49.95  | 2   | \$ 99.90         |
| Mixer        | IP3 = 30dBm   | IP3 = 33dBm   | LAVI-362VH+                                 | Minicircuit                              | \$24.95   | 1   | \$ 24.95         |
| PLL          | 0-10GHz   | 0-8GHz  | ADF4108                                     | Analog Devices                           | \$3.30    | 1   | \$ 3.30          |
| VCO          |   |   | * was not able to find a broadband solution | Price estimated on commercial VCO prices | \$200     | 1   | \$ 200.00        |
| <b>Total</b> |   |   |   |  |           |     | <b>\$ 328.15</b> |

**Table 12 Cost of Reference Sensing Receiver**

|                     | Proposed Solution                     |  |             |                |           |     |                  |
|---------------------|---------------------------------------|--|-------------|----------------|-----------|-----|------------------|
|                     | Required Spec                         | Available Spec                                       | Part Number | Manufacturer   | Unit Cost | Qty | total Cost       |
| Main Receiver       |                                       |  |             |                |           |     |                  |
| LNA                 | Gain = 30dB<br>NF~ 5dB<br>F=0 - 10Ghz | Gain = 24dB@5GHz<br>NF= 5.6dB<br>@5GHz<br>F=0 - 6Ghz | GVA-84+     | Minicircuit    | \$ 1.82   | 2   | \$ 3.64          |
| Attenuator          | 0 - 10GHz                             | 0 - 7GHz   | PAT-10+     | Minicircuit    | \$ 2.95   | 1   | \$ 2.95          |
| Mixer               | RF= 0 -10GHz<br>IP3 = 10dBm           | RF=0-4.2GHz<br>IP3 = 12dBm                           | SIM-43+     | Minicircuit    | \$ 7.49   | 1   | \$ 7.49          |
| PLL                 | 0-10GHz                               | 0-8GHz   | ADF4108     | Analog Devices | \$3.30    | 1   | \$ 3.30          |
| VCO                 | 0-1GHz<br>1-2GHz<br>2-4GHz            | 0-1GHz<br>1-2GHz<br>2-4GHz                           | Several     | Minicircuit    | \$ 20.00  | 5   | \$ 100.00        |
| <b>Main Total</b>   |                                       |  |             |                |           |     | <b>\$ 117.38</b> |
| DSR                 |                                       |  |             |                |           |     |                  |
| LNA                 | Gain = 30dB<br>NF~ 5dB<br>F=0 - 10Ghz | Gain = 24dB@5GHz<br>NF= 5.6dB<br>@5GHz<br>F=0 - 6Ghz | GVA-84+     | Minicircuit    | \$ 1.82   | 2   | \$ 3.64          |
| Mixer               | RF= 0 -10GHz<br>IP3 = 10dBm           | RF=0-4.2GHz<br>IP3 = 12dBm                           | SIM-43+     | Minicircuit    | \$ 7.49   | 1   | \$ 7.49          |
| PLL                 | 0-10GHz                               | 0-8GHz   | ADF4108     | Analog Devices | \$3.30    | 1   | \$ 3.30          |
| VCO                 | 0-1GHz<br>1-2GHz<br>2-4GHz            | 0-1GHz<br>1-2GHz<br>2-4GHz                           | Several     | Minicircuit    | \$ 20.00  | 5   | \$ 100.00        |
| <b>DSR Total</b>    |                                       |  |             |                |           |     | <b>\$ 114.43</b> |
| <b>System Total</b> |                                       |  |             |                |           |     | <b>\$ 231.81</b> |

**Table 13 Cost of Dedicated Sensing Receiver**

In Table 13, we show two receiver chains: a main receiver and the DSR. The estimated cost of the main receiver using a DSR is reduced by 65%. This is due to the relief in specification on the PLL and dynamic range.

The total cost of the overall solution (main and DSR chains) in this example is still 30% cheaper than a non DSR solution. This is mostly due to the fact that with the addition of the main receiver, the dynamic range is able to adapt to the channel conditions based on information provided by the DSR. The proposed approach reduces the complexity of the main receiver.

### 4.3 Summary

Cognitive radios are challenged for coexistence as secondary users for low latency applications. In this chapter, we addressed the implementation challenges of spectrum sensing for cognitive radio applications. The intelligence of the CR is contingent upon accurate and fast sensing of the spectrum. The research community proposed different techniques to address sensing algorithm, but it has yet to address the key issue of practical implementation of the sensing receiver. This problem becomes significantly more challenging if the CR is used in time sensitive applications. We presented a practical CR receiver architecture that is suited for low latency operation.

In order to improve the sensing time and the practical implementation of the sensing receiver, we have presented a dedicated sensing receiver approach. The DSR offers immediate improvement to the sensing time and a mathematical model is derived. The DSR combines power detection in both analog and digital domains. The addition of a DSR in the proposed architecture leverages coarse and fine sensing modes, and introduces an additional mode that leverages a priori information. The DSR utilizes power estimation at several stages in the receiver chain which sped up spectrum sensing and shortened the decision time. More importantly, the DSR leverages a priori information to improve and speed up the decision making. The power estimation from the DSR are prioritized and loaded into an LUT that the main CR receiver uses to determine the next best band for CR data transmission. We show substantial improvement in initial sensing and more importantly, the DSR made a significant improvement when the CR is in communication.

The band of interest is divided into coarse and fine bands, and the total sensing time is significantly dependent on the number of bins and the number of FFT points. In this chapter, we

assume that the detection probability is perfect. Also, we choose the bin sizes without taking into account the detection probability or the probability of false alarm.

In the next chapter, we focus on introducing detection probability and the probability of false alarm. We derive a new equation to take the probability into account. We then optimizing the number of coarse bins and fine bins such that the total sensing time is minimized. The number of bins is therefore dependent on channel conditions such as the number of users, the implementation of the DSR (i.e. PLL locktime), detection, and false alarm probabilities.

## 5 DSR Optimization

In this chapter, we propose an optimized dedicated broadband sensing receiver architecture for use in cognitive radios supporting delay sensitive applications. We derive an expression for mean acquisition/detection time as a function of a number of parameters including the number of coarse and fine frequency bins employed. We then determine the optimal number of coarse and fine bins that minimizes the overall detection time required to identify idle channels under various system conditions. Using analytical and simulation results, we quantify the dependence of optimal coarse and fine bin selection on system parameters such as 1) size of FFT used in scanning, 2) probability of detection and false alarm of the underlying sensing algorithm, 3) signal to noise ratio of the received signal, and (4) expected number of available channels. We present our derivations and discuss results.

Cognitive Radios (CR) require an accurate assessment of the activities in a desired frequency spectrum in order to determine the availability of idle channels suitable for opportunistic secondary use. Prior research has focused on novel techniques such as centralized network sensing [97], cooperative sensing [98] and the use of statistical methods and game theory [99] to improve sensing time. These techniques are not well suited for practical implementation of CR in time sensitive operations. By utilizing such techniques as presented in [97], [98] and [99], the CR is now dependent on outside inputs to make decisions on its operation. With a centralized network sensing, additional costs and delays are introduced by the traffic controller. In the cooperative model, the CR performs energy detection and uses time division slots to communicate with other users. As the number of users increases, the delay may become intolerably long. As discussed in chapter 4, we propose the use of a dedicated sensing receiver (DSR) that is solely focused on channel sensing and runs in parallel with a main receiver. The key to the DSR is an efficient algorithm that performs spectrum detection and continuously improves the quality of the collected data and decision process. The fast and initial sensing is done in the analog domain at the RF or IF frequencies prior to additional process in the digital domain. We demonstrated that the use of a dedicated sensing receiver (DSR) is necessary and



required for fast and reliable sensing in broadband operation. In addition, the overall time delay is also greatly reduced which opens the way for voice operation in cognitive radio. We were able to show that the DSR architecture provides up to a five-fold reduction in total mean time detection.

In this chapter, we focus on optimizing the broadband sensing receiver architecture for use in cognitive radios supporting delay sensitive applications. In our proposed DSR model, we use a two-stage sensing technique for performing broadband sensing. Here, we divide the desired bandwidth into coarse bins which are then subdivided into fine bins. After the initial setup, the receiver performs a cursory scan of the coarse bins in search of idle channels. Once idle channels are identified, the receiver then proceeds to a more thorough scan of the channels using improved resolution in order to avoid misdetection or a false alarm (especially when the primary users of the channel are operating at low signal to noise ratio (SNR)). Higher resolution scans require more time to complete the operation. The coarse scan while faster is not as accurate and might lead to a high number of misdetections. Hence, a delicate balance between the faster coarse scan and the more accurate fine scan is needed. Therefore, we first derive an expression for total mean detection time as a function of the number of coarse and fine bins as well as other system parameters such as phase locked loop (PLL) lock time, digital signal processing (DSP) frequency of operation, and received signal to noise ratio. We then determine the optimal values of coarse and fine bins that minimize the total mean detection time. Using both analytical and simulation results, we quantify the effect of various system parameters on the optimal choice of coarse and fine bins. For example, we show that the optimal number for coarse bins decreases with increase in SNR and the optimal number of fine bins increases with increased interference in the band.

## 5.1 Mean Detection Time

Recall that in equation (53), we derived a formula for the total system scan time given implementation parameters such as PLL, DSP speed, number of FFT points. However, this equation assumes perfect detection and no false alarm during coarse scanning and does not take into account information about the environment such as SNR of the received sign. In order to

characterize the sensing time accurately, the probability of detection and false alarm rate of coarse scanning must be incorporated into the equation (53).

### 5.1.1 Detection and False Alarm Probability

For the purposes of this work, we use energy detection for determining availability of a channel. The received signal is filtered then passed through a square law detector and integrated over a sensing time. The received signal is then sampled and the radio receives a complex baseband signal  $\{x_n\}$ . The complex baseband signal is given by:

$$x_n = \eta \cdot s_n + w_n \quad (62)$$

where  $s_n$  is the PU signal and  $w_n$  is the complex noise process. The value of  $\eta$  determines the presence or absence of the PU with  $\eta \in \{0,1\}$ . Consider the binary hypothesis testing problem:

$$\begin{aligned} H_0 : x &= w \\ H_1 : x &= s + w \end{aligned} \quad (63)$$

where  $H_0$  represents PU absent and  $H_1$  represent PU present.

The SNR of  $x$  is defined by:

$$\gamma \equiv \frac{\sigma_s^2}{\sigma_w^2} \quad (64)$$

where  $\sigma_s^2$  is the power of the signal and  $\sigma_w^2$  is the variance of the Gaussian noise with mean zero. Let's further define the  $P_n$  as the power of noise and  $P_s$  as the signal power at the sensing receiver as measured in bandwidth  $B$ . Therefore,

$$P_n = kTB \quad (65)$$

where  $k$  is the Boltzman constant ( $1.38 \times 10^{-23}$  J/K) and  $T$  is the system temperature in K. Furthermore, we can define  $P_s$  as:

$$P_s = \sigma_s^2 = \gamma \cdot P_n \quad (66)$$

We assume that the  $s_n$  and  $w_n$  are uncorrelated. We assume that we collect a total of  $N$  samples of  $x_n$ ,  $\mathbf{x}=(x_0, x_1, \dots, x_{N-1})$ . The received signal is compared to a threshold  $D_t$ .

$$\begin{aligned} R(x) < D_t & \text{ decide } H_0 \\ R(x) \geq D_t & \text{ decide } H_1 \end{aligned} \quad (67)$$

Assuming that the detector bandwidth is defined as  $B$ , [100,101] have shown that the observed output as:

$$f(x | H_0) = \frac{1}{\sqrt{2\pi}\sigma_n} \exp\left(-\frac{x^2}{2\sigma_n^2}\right) \quad (68)$$

$$f(x | H_1) = \frac{1}{\sqrt{2\pi}\sigma_n} \exp\left(-\frac{(x - \sqrt{P_s})^2}{2\sigma_n^2}\right) \quad (69)$$

where  $x$  represents the power of observed samples.

The detection probability  $P_d$  is defined as the probability of the sensing receiver to correctly detect an idle channel under  $H_0$ . False alarm probability is defined as the sensing receiver falsely identifying a channel idle (PU not present) under  $H_1$ . Assuming that the detection is done via energy detection,  $H_0$  is chi-square distributed and  $H_1$  is non-central chi-square distributed [102].

### 5.1.2 Mean Detection Time

We define 1)  $P_d$  as the detection probability, 2)  $P_{fa}$  as the false alarm probability, 3)  $D_t$  as the threshold level for the detection rule, 4)  $J$  as implementation penalty metric that models the additional wasted time needed to recover from a false alarm and resume the search process and  $M$  as the number of receivers. In the case of the DSR shown in Figure 4-18,  $M$  would be set as 2, since we have two receiver chains. We further define  $L$  as the actual number of idle coarse channels and  $K$  as the actual number of idle fine channels.  $K$  and  $L$  represent the environment and they are a measure of the channel occupancy. Recall that we defined  $\rho$  as the percentage of coarse bins that are identified as available after coarse sensing. Hence,  $\rho$  can be represented in term of  $L$  as:

$$\rho = \frac{L}{\beta} \quad (70)$$

Assuming a serial search is performed, the mean detection time  $\bar{T}_{det}$  [100] corresponds to:

$$\bar{T}_{det} = \bar{S}_{det} \cdot (T_s + T_i) \quad (71)$$

where,

$$\bar{S}_{det} = \frac{(\beta - L) \cdot J \cdot P_{fa} + \beta}{P_d \cdot (L + 1)} \quad (72)$$

$T_s$  is the switching time and is equivalent to  $T_{PLL\_crs}$  and  $T_{PLL\_fin}$  for coarse and fine sensing, respectively.  $T_i$  is the integration time required for making a decision. From (71) and (72) above, we can write down the mean detection time in coarse mode.

$$\bar{T}_{\text{det}_{crs}} = \frac{(\beta - L)J.P_{fa} + \beta}{P_d \cdot (L + 1)} \left[ T_{PLL\_crs} + \frac{1}{F_{DSP}} (4N_{crs} \cdot \log_2 N_{crs} - 6N_{crs} + 8) + T_{init} \right] \quad (73)$$

Similarly,  $\bar{T}_{\text{det}}$  for the fine mode can be derived and plugged into (53) in order to determine the overall mean time detection of system defined by  $\bar{T}_{\text{sys}}$  in (74) below.

$$\begin{aligned} \bar{T}_{\text{sys}} = & \left( \frac{B_{\text{sys}}}{\alpha M N_{crs} F_{res}} + \frac{(\beta - L)J.P_{fa} + \beta}{P_d (L + 1)} \right) \cdot A_{crs} + \left( \frac{\alpha \beta \rho}{M F_{DSP}} + \frac{(\alpha - K)J.P_{fa} + \alpha}{P_d (K + 1)} \right) \cdot A_{fin} + T_{init} \\ & + \left( \frac{\alpha \beta \rho}{M} + \frac{(\alpha - K)J.P_{fa} + \alpha}{P_d (K + 1)} \right) T_{PLL\_fin} + \left( \frac{\beta}{M} + \frac{(\beta - L)J.P_{fa} + \beta}{P_d (L + 1)} \right) T_{PLL\_crs} \end{aligned} \quad (74)$$

where,

$$A_{crs} = 4.N_{crs} \log_2 N_{crs} - 6N_{crs} + 8 \quad (75)$$

and,

$$A_{fin} = 4.N_{fin} \log_2 N_{fin} - 6N_{fin} + 8 \quad (76)$$

As expected, there are several parameters that affect the overall mean time detection of a two stage sensing system. Equation (74) illustrates that sensing time is influenced by environmental parameters such as  $B_{\text{sys}}$ , SNR, K and L, while other factors affecting  $\bar{T}_{\text{sys}}$  are under user control such as number of FFT points,  $\alpha$  and  $\beta$ .

In the next section, we work to minimize  $\bar{T}_{\text{sys}}$  by appropriately choosing user defined parameters.

## 5.2 DSR Optimization

The main goal of the sensing receiver is to detect available channels quickly and reliably. Most importantly, it is critical to minimize false alarms. In examining equation (74) above, there are

some very obvious ways to minimize  $\bar{T}_{sys}$  such as reducing initialization time ( $T_{init}$ ) and the PLL lock times ( $T_{PLL\_crs}$  and  $T_{PLL\_fin}$ ). However, the PLL lock time has physical implementation limitations as discussed in chapter 3. Therefore, we look for other ways to minimize the sensing time and optimize the system.

### 5.2.1 Finding Minimum

Assume that a real function  $f$  is differentiable. A point  $P_0$  is a local minimum point [104] if there is a neighborhood of  $P_0$  in which

$$f(P) \geq f(P_0) \quad (77)$$

and it is called an absolute or global minimum if the inequality in (77) holds true for all points in the domain. In order for  $f$  to have an extreme value (maximum, minimum, or saddle point) at  $P_0$ , the derivative of  $f$  at  $P_0$  must equal zero.

$$\nabla f(P_0) = 0 \quad (78)$$

A point  $P_0$  where (78) holds true is called a critical point. This is called the first derivative test. In order to verify that whether  $P_0$  is a minimum, a second derivative test is required. Assuming that  $f$  is twice differentiable, then:

$$\nabla^2 f(P_0) > 0 \quad (79)$$

$\nabla f$  and  $\nabla^2 f$  called the gradient and Hessian of  $f$  respectively.

Therefore, in order to find a minimum, we perform the first derivative test of a function to find an extreme point or the critical point, then the second derivative test checks whether the critical point is a minimum.

### 5.2.2 Convex Optimization

In the case of optimization, we search for absolute or global minimum of a real function. However, as the number of variables in a function increases, this problem becomes more difficult. Recall that optimization of the total system delay time requires that we optimize equation (74). The equation contains many variables and the convergence may be difficult to achieve. One way to perform such an optimization is to use a technique called convex optimization.

Optimization has several challenges [103]: 1) the function might have several local optima, 2) the algorithm might not converge or a very poor convergence rate and, 3) initial and stopping criteria of an optimization algorithm may be arbitrary or might wander. However, if the function is convex, then the minimum point is in fact a global minimum.

As discussed earlier, a convex function  $f$  can be characterized by conditions of its gradient and Hessian functions. A function is convex if its Hessian is positive, semi-definite on its domain.

$$\nabla^2 f(P_0) \succ 0 \quad (80)$$

If the Hessian is positive definite at  $x$ , then  $f$  attains a minimum at  $P_0$ .

Hence, in order to go beyond minimizing a function, if the function is convex or its Hessian is positive definite, then our solution is optimized.

Some convex properties [103,106]:

- Non negative sum of convex functions is convex
- $x^\alpha$  is convex for  $\alpha > 0$
- $\log x$  is concave
- $x \cdot \log x$  is convex
- $f(x) = \log \sum e^x$  is convex
- $e^x$  is convex if  $x$  is convex
- The inverse of a concave function is a convex function

- $\sqrt{x}$  is concave,  $\frac{1}{\sqrt{x}}$  is convex
- Multiplication of convex functions yields a convex function
- Linear functions are both concave and convex

The above are some of the properties that we use in our analysis.

### 5.2.3 Sensing Time Optimization

From section 4.2 and Figure 4-23, it is obvious that the total sensing time is highly dependent on our choice of coarse and fine bins. One method to reduce sensing time is to make appropriate choices for coarse and fine bins, i.e., selecting optimal  $\beta$  and  $\alpha$  to minimize the overall mean detection time. In order to minimize  $\bar{T}_{sys}$ , we can employ the standard strategy of equating the partial derivatives of (74) with respect to  $\beta$  and  $\alpha$  to zero. However,  $P_d$  and  $P_{fa}$  complicate this computation since they exhibit a dependence on the sensing or detection bandwidth which is directly proportional to  $\beta$  and  $\alpha$ . Therefore, we first simplify (74) by approximating  $P_d$  and  $P_{fa}$ . Assuming non-coherent square-law detection is used,  $P_d$  and  $P_{fa}$  corresponds to [100],

$$P_d \approx Q\left(\frac{D_t - 2T_{sense} B_{sense} (1 + SNR)}{2\sqrt{T_{sense} B_{sense} \sqrt{1 + 2SNR}}}\right) \quad (81)$$

$$P_{fa} \approx Q\left(\frac{D_t - 2T_{sense} B_{sense}}{2\sqrt{T_{sense} B_{sense}}}\right) \quad (82)$$

Where,

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-\tau^2/2} d\tau \quad (83)$$



As described earlier,  $T_{sense}$  and  $B_{sense}$  constitute the sensing time and the sensing bandwidth that is directly proportional to the bin bandwidth. Equations (81) and (82) can be approximated using a sigmoid function. The authors in [105] use a gradient approximation of the sigmoid function that was used for a fast algorithm for learning large scale preference relations. The relationship between the sigmoid function and complementary error function can be approximated as [105]:

$$\sigma(z) = (1 + e^{-z})^{-1} \approx 1 - \frac{1}{2} \operatorname{erfc}\left(\frac{\sqrt{3}z}{\sqrt{2\pi}}\right) \quad (84)$$

Recall that

$$Q(z) = \frac{1}{2} \operatorname{erfc}\left(\frac{z}{\sqrt{2}}\right) \quad (85)$$

By combining (83) and (84),  $Q(z)$  may be approximated as:

$$Q\left(z \frac{\sqrt{3}}{\pi}\right) \approx 1 - (1 + e^{-z})^{-1} \quad (86)$$

We can use the approximation in (86) to find a simplified expression for  $P_d$  and  $P_{fa}$  in (81) and (82). After substituting these updated  $P_d$  and  $P_{fa}$  expressions in (74), the simplified approximation of  $\bar{T}_{sys}$  is shown in (87).

$$\begin{aligned} \bar{T}_{sys}(\alpha, \beta) = & \left( \frac{A_{crs}}{M \cdot F_{DSP}} + \frac{T_{PLL\_crs}}{M} \right) \cdot \beta + \frac{A_{crs} + T_{PLL\_crs}}{L + 1} \left[ J e^{x(y-1)} (\beta - L) + \beta e^x \right] + \\ & \left( \frac{L \cdot A_{fin}}{M \cdot F_{DSP}} + \frac{L \cdot T_{PLL\_fin}}{M} \right) \cdot \alpha + \frac{A_{fin} + T_{PLL\_fin}}{K + 1} \left[ J e^{v(y-1)} (\alpha - K) + \alpha e^v \right] + T_{init} \end{aligned} \quad (87)$$

where,

$$y = \frac{SNR + 1}{\sqrt{1 + 2SNR}} \quad (88)$$

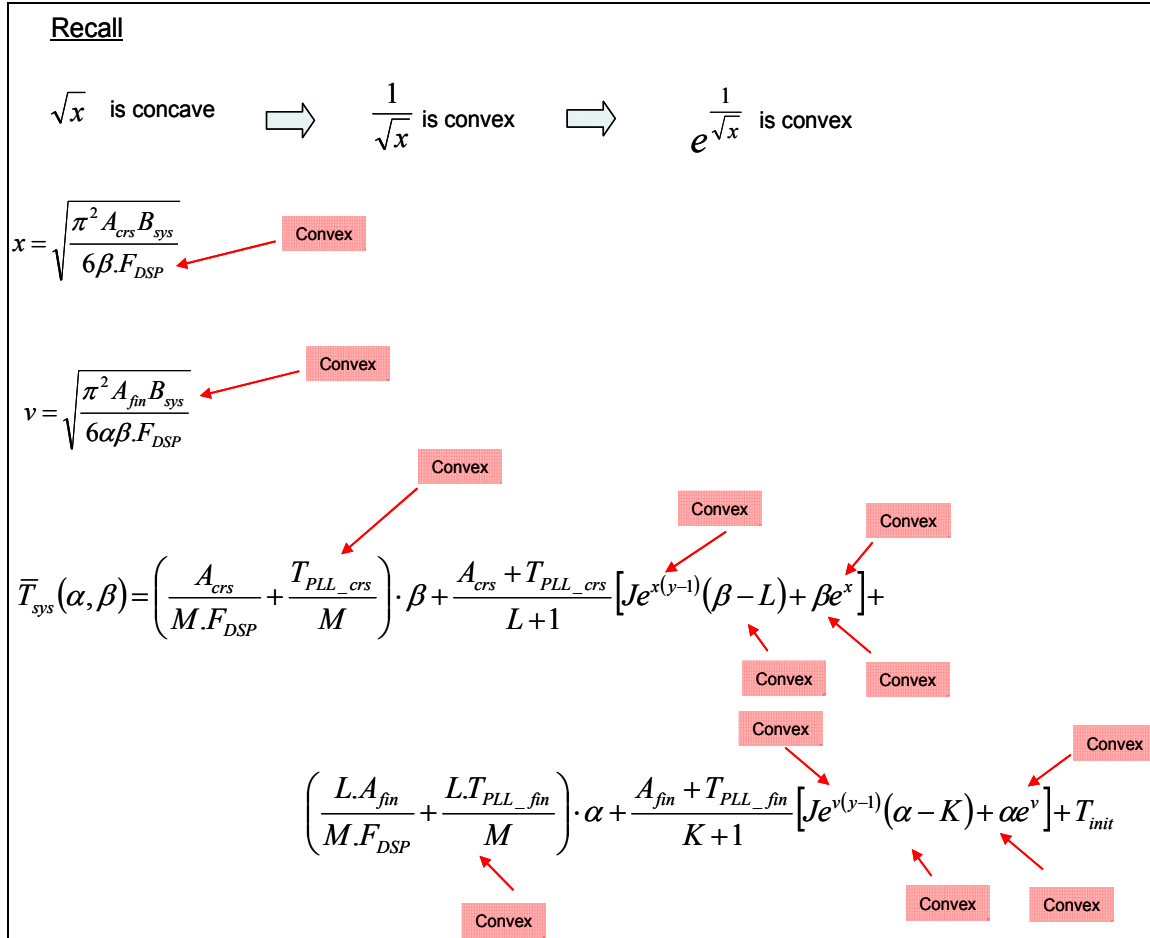
and,

$$x = \sqrt{\frac{\pi^2 A_{cfs} B_{sys}}{6\beta.F_{DSP}}} \quad (89)$$

$$v = \sqrt{\frac{\pi^2 A_{fin} B_{sys}}{6\alpha\beta.F_{DSP}}} \quad (90)$$

The expression in (87) characterizes the mean scan time as a function of a number of system parameters. In order to optimize this function, we need to show that the critical point is a global minimum. As discussed earlier, we can either prove that the Hessian of the (87) is positive definite or we can show that it is a convex function. If the function is convex, then the critical point is also an optimal point.

Specifically, we can show that under certain conditions,  $\bar{T}_{sys}$  in (87) is a convex function with respect to  $\alpha$ ,  $\beta$ . The convexity check is shown in Figure 5-1. The conditions for convexity are  $\alpha > K$  (i.e., the number of idle fine channels is less than the total number of fine channels),  $\beta > L$  (i.e., the number of idle coarse channels is less than the total number of coarse channels), and  $y > 1$  (i.e., the SNR is real). All three conditions for convexity are practical and essential. Since  $\bar{T}_{sys}$  is convex in  $\alpha$  and  $\beta$ , we can determine the optimal choice for the number of coarse and fine bins (that minimize minimum scan time) as the values that force the derivative of (87) to 0.



**Figure 5-1 Convexity Check of Mean Detection Time**

The partial derivate of (87) with respect to  $\beta$  corresponds to:

$$\frac{\partial}{\partial \beta} \bar{T}_{sys} = \frac{A_{crs}}{M \cdot F_{DSP}} + \frac{T_{PLL\_crs}}{M} + \frac{A_{crs} + T_{PLL\_crs}}{L+1} \left[ J e^{x(y-1)} \left( 1 - \frac{1}{2} (y-1)x + \frac{JL}{2\sqrt{\beta}} (y-1)x e^{x(y-1)} + \left( 1 - \frac{1}{2} x \right) e^x \right) \right]$$

(91)

Similarly, we can write down the partial derivative of  $\bar{T}_{sys}$  with respect to  $\alpha$ .

$$\frac{\partial}{\partial \alpha} \bar{T}_{sys} = \frac{L.A_{fin}}{M.F_{DSP}} + \frac{L.T_{PLL\_crs}}{M} + \frac{A_{fin} + T_{PLL\_fin}}{K+1} \left[ J e^{v(y-1)} \left( 1 - \frac{1}{2}(y-1)v + \frac{JK}{2\sqrt{\alpha}}(y-1)v e^{v(y-1)} + \left( 1 - \frac{1}{2}v \right) e^v \right) \right]$$

(92)

As expected, equation (91) depends on both  $\beta$  and  $\alpha$  while equation (91) is only dependent on  $\beta$ . That is, as the sensing receiver initiates a coarse search, the number of coarse bins is not dependent on the fine scan. However, once the coarse scan is completed, the fine scan is dependent on the results of the coarse scan (i.e., dependent on  $\beta$ ) and starts the fine scan according to the priority set in the LUT set after the coarse scan. We can set Equations (91) and (92) to zero and solve for  $\beta$  and  $\alpha$  that minimizes  $\bar{T}_{sys}$ . We employ numerical non-linear solvers in order to find the solution to equations (91) and (92). The results from the optimization and its physical interpretation are presented in the next section.

### 5.3 Optimization Results

In this work, our goal is to find the optimal bin size for coarse and fine sensing under given channel conditions and design implementation of the radio. As the spectrum becomes more and more crowded, the number of idle channels for coarse ( $K$ ) and fine sensing ( $L$ ) decreases and hence, on average, it would take the sensing receiver a longer time to identify an appropriate channel for CR operation (i.e. sensing time increases). Similarly, the physical implementation is mostly defined by the user given restrictions on cost, power, performance, etc. For example, the total time to perform a DFT in (44) is inversely proportional to the speed of operation of the DSP. A brute force approach would be for the designer to choose the fastest DSP available. However, fast DSP come with a premium in cost and power consumption that may or may not necessarily affect the overall system performance. The solution to this problem is fine balance between coarse and fine sensing and the rest of the implementation parameter. However, the

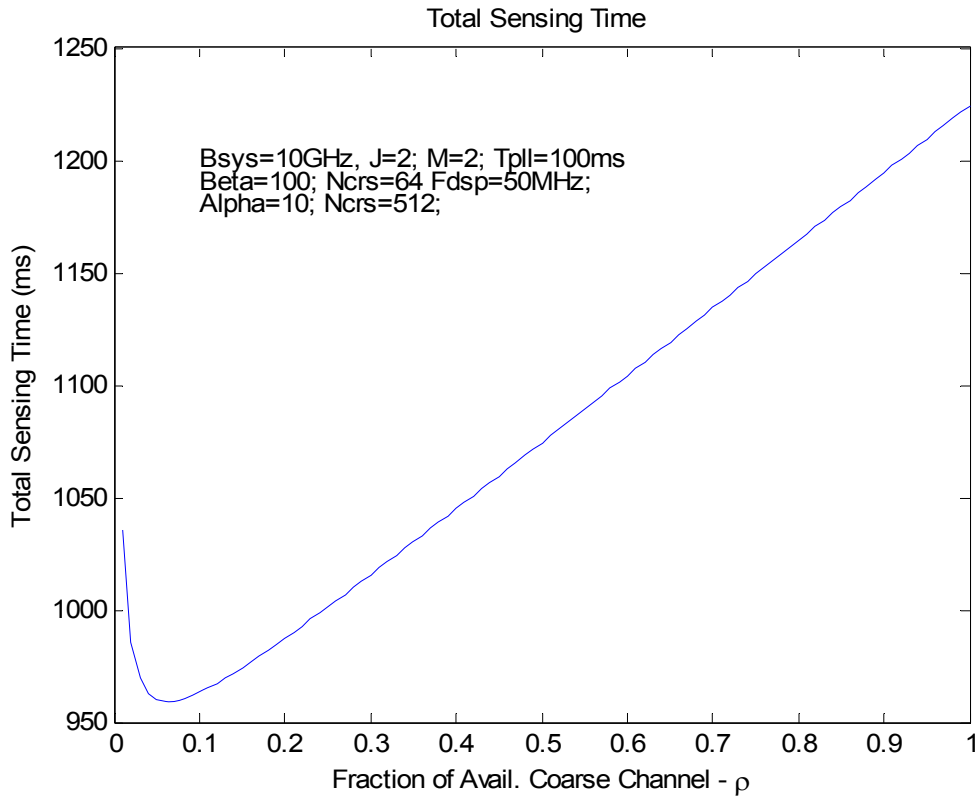
solution is not static and it varies dynamically with the channel parameters such as SNR, L and K.

The simulation results are presented in two parts: in the first part we focus on minimizing  $\bar{T}_{sys}$  in (74). We provide a better insight on the dependence of minimum  $\bar{T}_{sys}$  with respect to input variables such as the number of FFT points, DSP speed, number of available channels (or spectrum crowding), assuming that  $\beta$ ,  $\alpha$ ,  $P_d$  and  $P_{fa}$  are given. In the second part, we concentrate on optimizing  $\beta$  and  $\alpha$  in order to minimize  $\bar{T}_{sys}$  given the channel conditions (e.g. L, K and SNR) and the physical implementation of the radio (such as PLL initialization, PLL lock times, number of FFT points and the DSP frequency  $F_{DSP}$ ).

### 5.3.1 Total Mean Detection Time $T_{sys}$

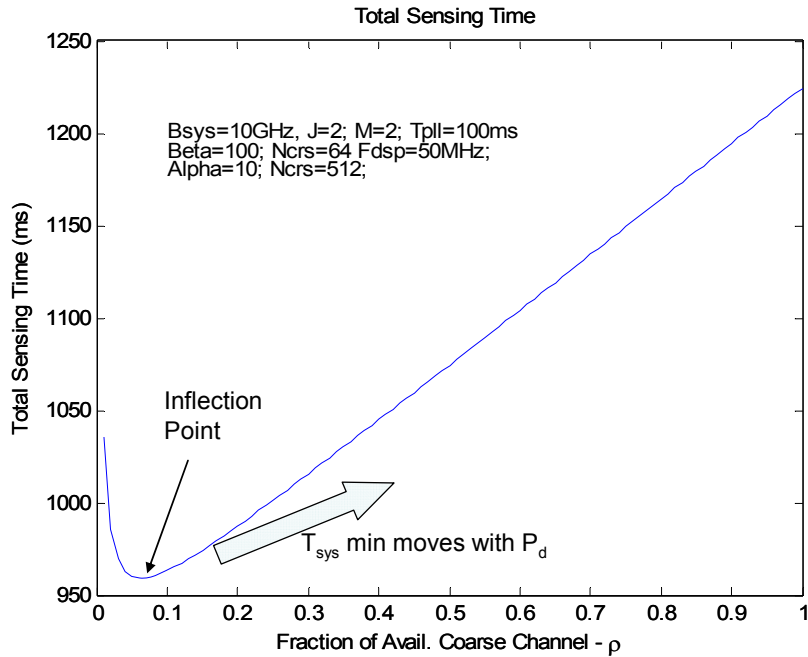
We simulate the total sensing time with respect to channel conditions and our choice of  $\beta$ ,  $\alpha$ . The basic parameters for our environment are  $J=2$ ,  $N_{crs}=64$ ,  $N_{fin}=512$ ,  $M=2$ ,  $F_{DSP}=50MHz$ ,  $B_{sys}=10GHz$  (broadband),  $T_{PLL\_crs} = 10ms$ ,  $T_{PLL\_crs} = 1ms$  and  $T_{inits} = 100ms$ .

In the first simulation, we want to better understand how  $\bar{T}_{sys}$  is affected by the increase in the number of users. As the number of users increases, the occupancy of the spectrum increases and hence the number of idle channel suitable for CR operation decreases. Recall that  $\rho$  constitutes the fraction of available coarse channels that are scanned in fine mode. On one hand, we want  $\rho$  to be as small as possible in order to minimize fine scan. However, if  $\rho$  is too small, then the occupancy of the channel is high and therefore it takes the DSR longer to find an available channel. The simulation of  $\rho$  versus  $\bar{T}_{sys}$  is shown in Figure 5-2.



**Figure 5-2 Total Mean Sensing Time versus Fraction of Available Coarse Channels**

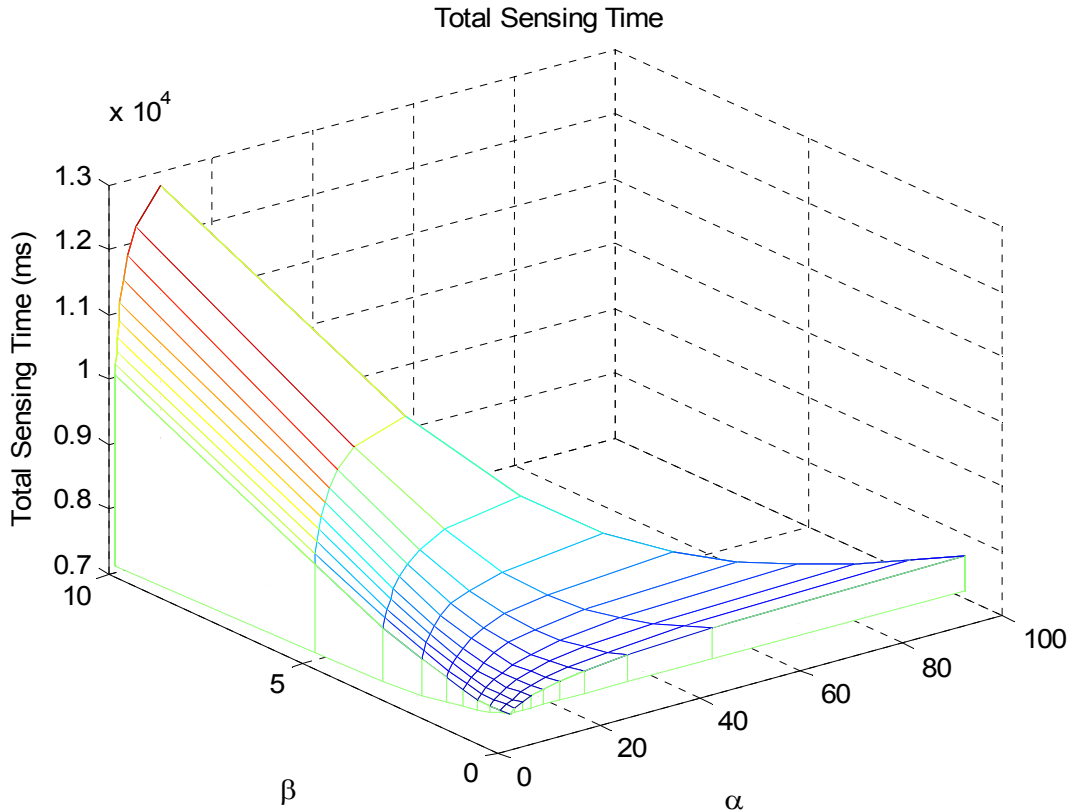
As expected, as  $\rho$  increases the total sensing time increases because more and more coarse channels need to be fine scanned. However, at low  $\rho$  (typically <10%), the sensing time tends to increase due to the fact that at these conditions, it is less probable that the sensing receiver finds an available channel quickly. The dependence in Figure 5-2 indicates an optimal  $\rho$  value for minimizing sensing time. However,  $\rho$  is a system parameter that is outside the control of the designer.



**Figure 5-3 Minimum point moves (to the right) as  $P_d$  increases**

In Figure 5-3, the minimum sensing location is dependent the value of  $P_d$ . As  $P_d$  increases, the  $\bar{T}_{sys}$  minimum location increases while shifting to the right. The main reason for this shift is that as we increase the probability of detection, the false alarm tends to increase and hence delays detection. The shift to the right is due to need for more available channels to make the appropriate decision.

The question remains how would  $\bar{T}_{sys}$  be affected by our choice of  $\beta$  and  $\alpha$ ? Since the coarse sensing time is much lower than the fine sensing time,  $\bar{T}_{sys}$  is reduced if more of the detection is done in coarse mode. On the flip side, the resolution in coarse mode is lower than in fine mode and false alarms or false positive reading of the spectrum would cause the DSR to reset and resume the scanning process. This penalty is captured by parameter  $J$  and the  $F_{res}$  in (74). Using the same variables as defined above, we simulate the total mean sensing time versus  $\beta$  and  $\alpha$  which is shown in Figure 5-4 with  $\rho=.5$



**Figure 5-4 Total Mean Sensing Time versus  $\beta$  and  $\alpha$**

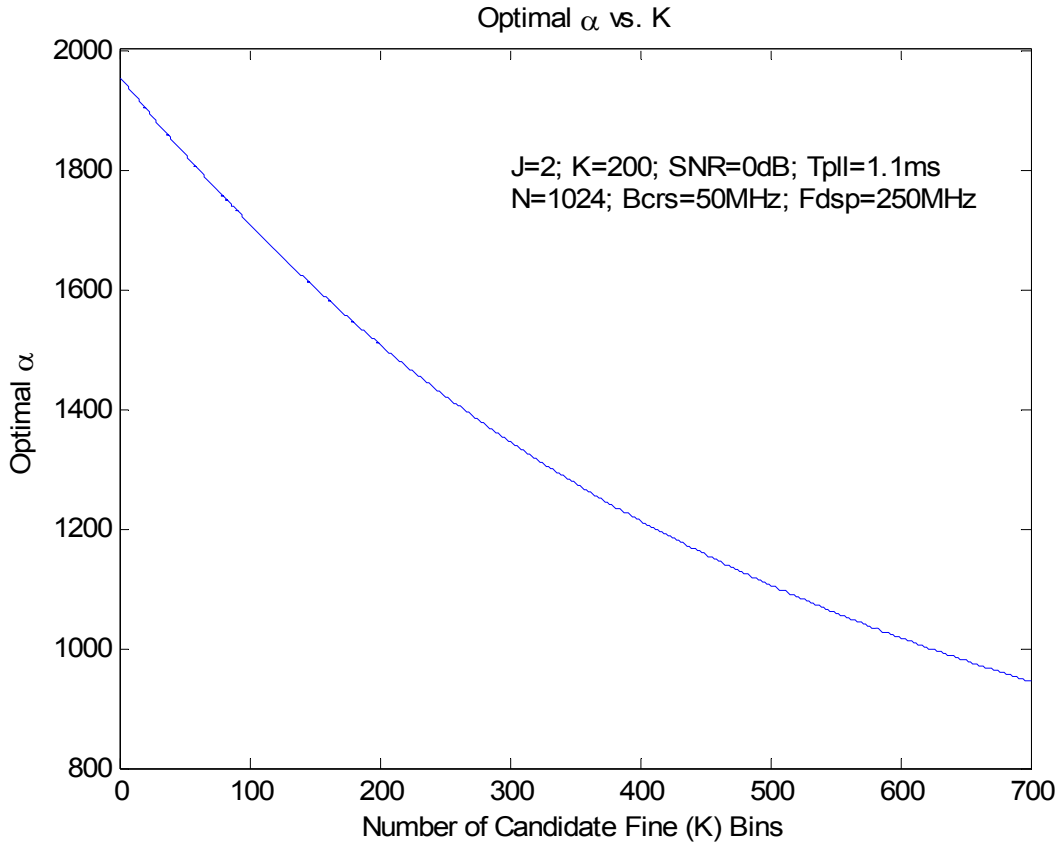
The relationship between  $\beta$ ,  $\alpha$  and  $\bar{T}_{sys}$  is shown in Figure 5-4. The sensing time is typically lower at low  $\beta$  and as expected, increases as  $\alpha$  increases. The resolution and switching time in coarse mode start to have a much greater effect on  $\bar{T}_{sys}$  than the computation of the N-point FFT. Given channel conditions and circuit implementation (on the PLL for example), we expect to find a combination of  $\beta$ ,  $\alpha$  such that the total mean detection time is minimized. One would hope that the combination would give a global minimum and hence provide an optimal solution for the system. In the next sub section, we calculate  $\beta$  and  $\alpha$  such that  $\bar{T}_{sys}$  is optimized.



### ***5.3.2 Optimal $\beta$ and $\alpha$ for Minimum Mean Detection Time***

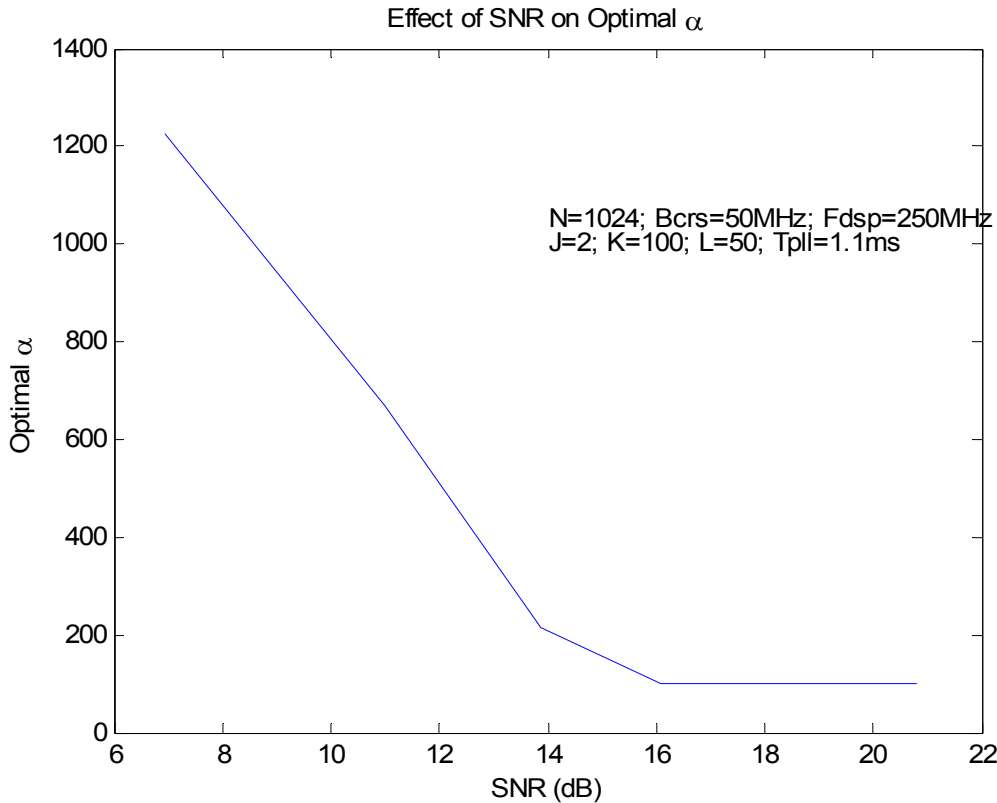
With the detection time highly dependent on the coarse and fine bandwidths, we seek to find an optimal solution. This is a large scale unconstrained optimization with primarily two sets of variables: 1) channel dictated variables such as SNR, and 2) implementation variables such as PLL lock times and the choice of DSP. We want to study the effect of the aforementioned variables on the minimum mean detection time of the system. First, we simulate the effect of the given variables on optimal  $\beta$  and  $\alpha$ . We seek to find the parameters in support of our algorithm such as number of FFT points in coarse and fine mode and bin sizes. Second, we present our results in a summary table format. We use equations (90) and (91) to determine optimal  $\alpha$  with respect to the channel variables (such as  $K$ , SNR). The results are documented in Figure 5-5 and Figure 5-6.

In Figure 5-5, we plot the effect of the number of available fine channels  $K$  (or channel crowding) versus the optimal  $\alpha$ . As the channels become crowded (i.e.  $K$  decreases), the probability of finding an idle channel decreases which requires additional sensing time. This phenomenon can be observed in Figure 5-5, where  $K$  becomes a dominant factor as the number of idle channel decreases. Under the conditions shown in the figure, the effect of  $K$  becomes less dominant when the number of fine available bins reaches  $\sim 700$ . The slope decreases by almost a factor of 5 between  $K=100$  and  $K=700$ .



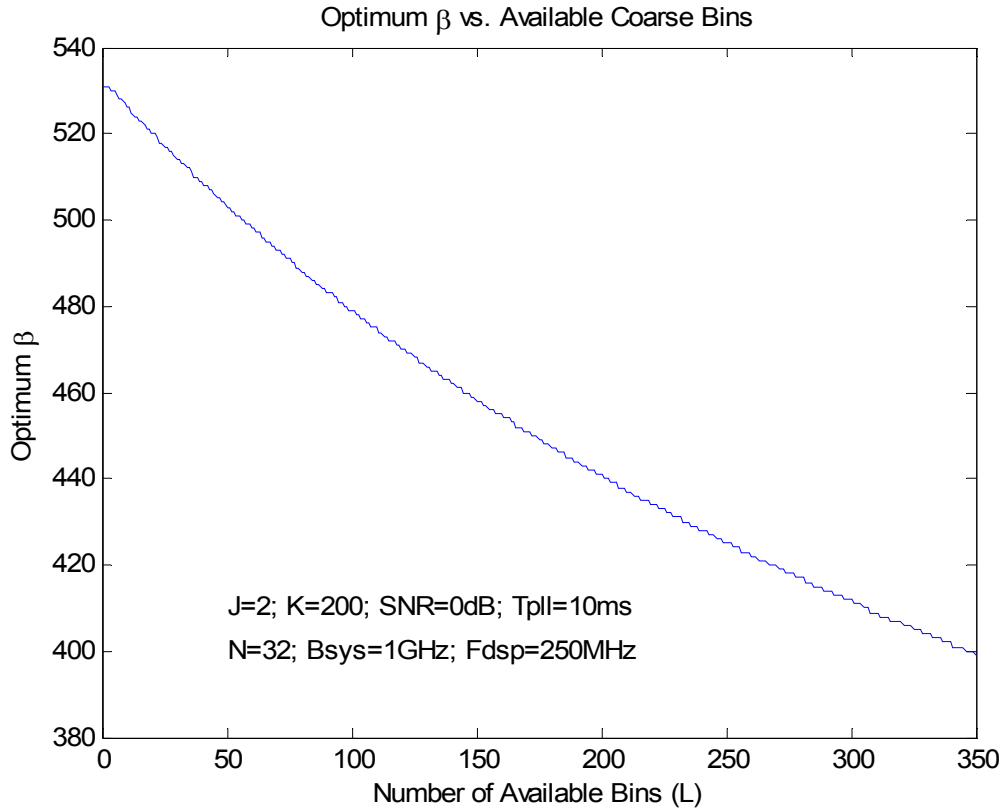
**Figure 5-5 Optimal  $\alpha$  versus Number of Available Fine Bins**

In Figure 5-6, we plot the effect of SNR on choice of  $\alpha$ . We note that as SNR increases, the number of required fine sensing bins decreases until it reaches the limit of our convexity condition  $\alpha = K$ , which basically states that all bins are available and may be used for CR operation. These results support our intuition that in order to minimize the overall scanning time, we need to perform less computation. Since the fine bins require more computation time, we seek to decrease the number of fine bins. That goal becomes more palatable at high SNR value where probability of detection is high and the probability of false alarm is low.



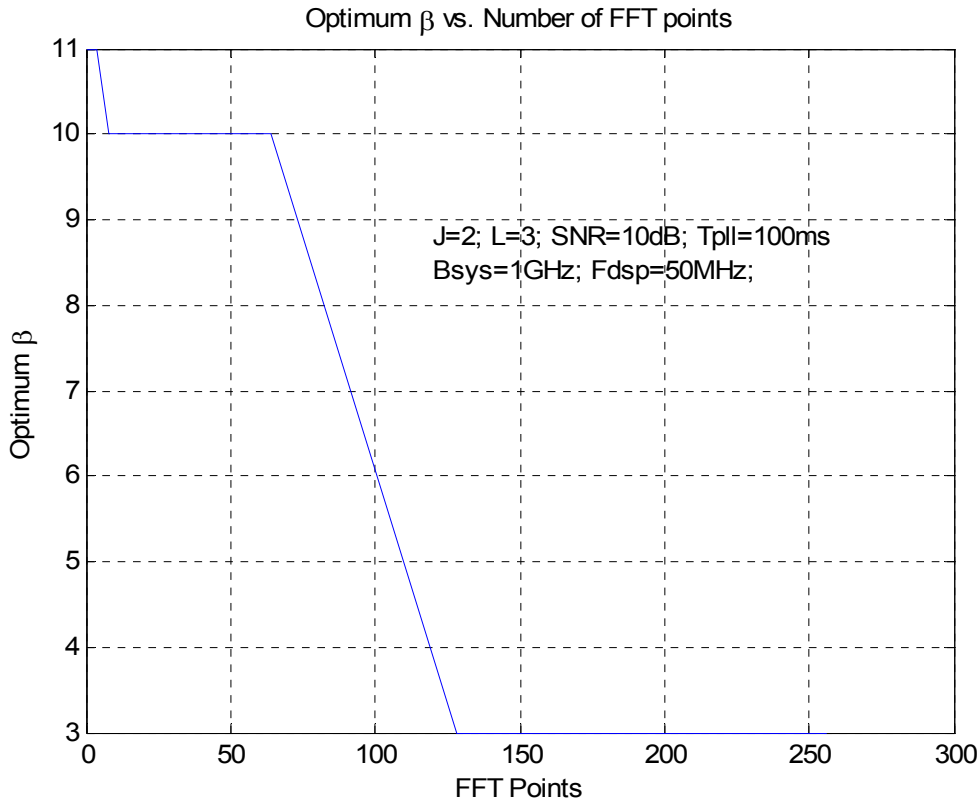
**Figure 5-6 Optimal  $\alpha$  versus SNR of received Signal**

Similarly, we study the effect of the variables on the choice of  $\beta$ . We document our results for  $\beta$ . In Figure 5-7, the number of available coarse bins ( $L$ ) affects  $\beta$  in the same manner as  $K$  affected  $\alpha$ . As the number of available bins increases, we expect a higher probability of detection and a reduction in  $\beta$ , the overall  $\bar{T}_{sys}$  decreases. In this example, we show that when there is only 10% available bins ( $K=50$  of a total of 500), we need a large  $\beta$  which basically states that the bandwidth must be divided into small bands in order to find idle channels. Intuitively, as the SNR increases, we can make decisions on the channel availability easier and hence more time consuming scan (in fine mode) become less necessary.



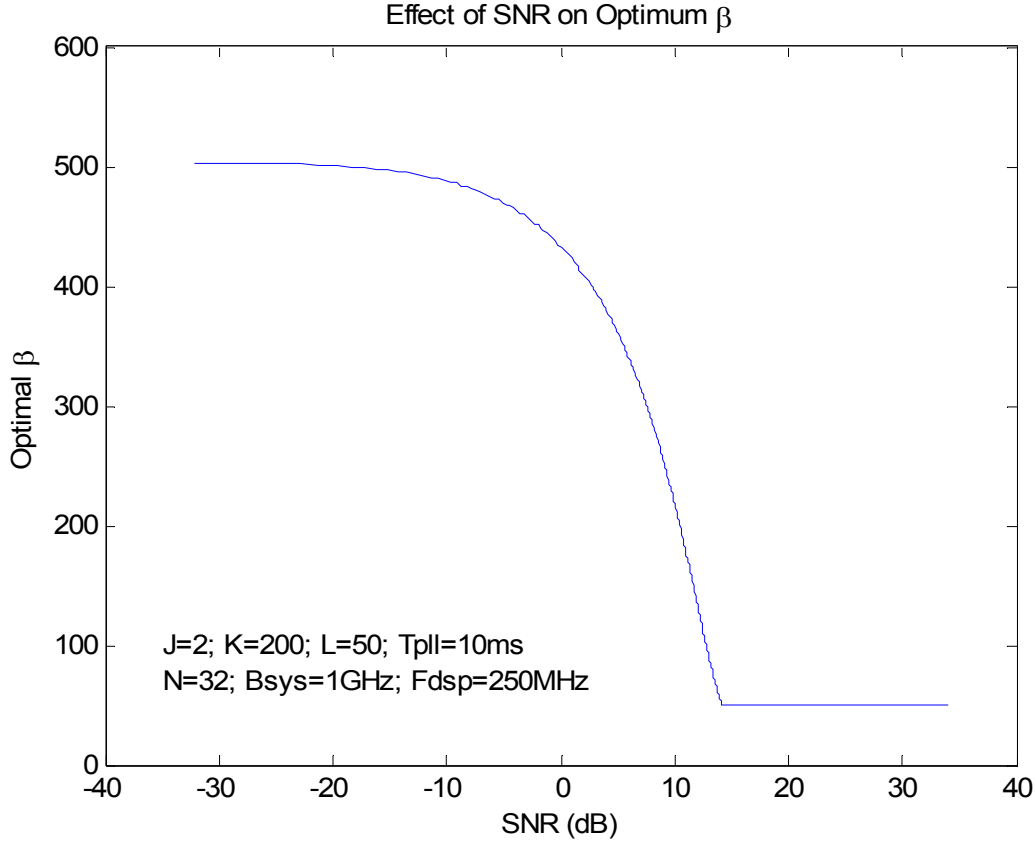
**Figure 5-7 Optimal  $\beta$  versus Available Coarse Channels**

In Figure 5-8, we show the number of coarse N-point FFT calculations versus our choice of  $\beta$ . In this example, we set the fine scan mode FFT points to 512. The number of bins decreases as the number of FFT points increases, until the limit condition for convexity of  $\beta = L$  is reached. Another interpretation of the results is as the number of FFT points increases, it becomes less viable that a 2-stage scanning process is needed. One of the main advantages of going to a 2-stage sensing technique is to reduce the number of calculations by allowing the coarse mode to do a cursory search for available channels. As the number coarse FFT points start to approach that of a fine sensing mode, the advantage and effectiveness of the coarse sensing mode is reduced.



**Figure 5-8 Optimal  $\beta$  versus number of FFT points**

Similar to Figure 5-6, the need for coarse bins decreases as the SNR increases. The results are captured in Figure 5-9. One interesting aspect of the results that was not obvious with  $\alpha$  is the fact that the required number of bins does not vary below a given SNR. This result is the opposite effect of what we discussed earlier in Figure 5-8. As the SNR decreases, more and more bins are needed to a point where the coarse sensing bandwidth is small enough to start infringing on the need for fine sensing. When the SNR is high, the probability of detection increases, and therefore the need for additional coarse search bins is reduced until the limit condition of convexity  $\beta = L$  is reached at which point  $\beta$  can not be reduced further.



**Figure 5-9 Effect of SNR on Choice of Optimal  $\beta$**

In order to better understand the sensitivity of our analysis on  $\bar{T}_{sys}$ , we show selected results below. The parameters used for these simulations are  $T_{PLL\_crs} = 0.5ms$ ,  $T_{PLL\_fin} = 0.1ms$ ,  $M=2$ ,  $J=2$ ,  $F_{DSP}=250MHz$ . In Table 14, we set  $L=6$ ,  $K=22$ ,  $N_{crs}=64$  and  $N_{fin}=2048$ . Please note that by doubling SNR from 15 to 30, the effect on  $\alpha$  is a 32% reduction versus a 7% on  $\beta$ . This discrepancy in variation supports our earlier results: As SNR increases, the need for bins decreases. However, the sensing time is far greater for fine mode sensing than in coarse mode sensing. Hence, the algorithm gives the priority to reducing  $\alpha$  over  $\beta$  which has a greater affect on  $\bar{T}_{sys}$ . Recall that for time sensitive applications, the DSR surveys the desired band of operation, sorts and prioritizes the channels best suited for CR operation. After the channels are identified and stored, the DSR continuously monitors and re-prioritize the channels as needed. In

order to avoid storing “stale” data in the LUT, the overall  $\bar{T}_{sys}$  must be minimized. Our goal is to optimize  $\bar{T}_{sys}$  by minimizing the overall sensing time.

| SNR | $\beta$ | $\alpha$ | $\bar{T}_{sys}$ (s) |
|-----|---------|----------|---------------------|
| 15  | 43      | 998      | 1.57                |
| 30  | 40      | 678      | 1.07                |
| 60  | 39      | 589      | 0.938               |

**Table 14 Mean Detection Time versus SNR**

In Table 15, we set  $L=6$ ,  $K=22$ ,  $N_{crs}=64$  and  $SNR=30$ . We vary the number of FFT points for the fine mode ( $N_{fin}$ ). As expected, by varying the  $N_{fin}$ , there is no impact on  $\beta$  since it is independent of the coarse sensing, but there is a high impact on  $\alpha$  and  $\bar{T}_{sys}$ .

| $N_{fin}$ | $\beta$ | $\alpha$ | $\bar{T}_{sys}$ (s) |
|-----------|---------|----------|---------------------|
| 1024      | 40      | 901      | 0.828               |
| 2048      | 40      | 678      | 1.07                |
| 4096      | 40      | 595      | 1.81                |

**Table 15 Mean Detection Time versus  $N_{fin}$**

As the number of FFT points increase,  $\alpha$  decreases but the overall  $\bar{T}_{sys}$  increase. Another parameter that has a high impact on  $\bar{T}_{sys}$  is the activity in the spectrum. In our results above, we showed that as the number of available channels decreases, we need additional bins (i.e.  $\alpha$  increases) in order to identify idle channels. In Table 16, we set  $L=6$ ,  $N_{fin}=2048$ ,  $N_{crs}=64$  and  $SNR=30$  and we vary the  $K$  variable. Please note that as  $K$  increases,  $\alpha$  decreases and  $\bar{T}_{sys}$  also decreases at a fast rate.

| K  | $\beta$ | $\alpha$ | $\bar{T}_{sys}$ (s) |
|----|---------|----------|---------------------|
| 22 | 40      | 678      | 1.07                |
| 20 | 40      | 539      | 0.867               |
| 10 | 40      | 115      | 0.275               |

**Table 16 Mean Detection Time versus Available Fine Channels (K)**

In this example, the improvement on  $\bar{T}_{sys}$  is drastic. Unfortunately, K is a representation of the channel environment and it is not under user control.

## 5.4 Summary

In this chapter, we quantify the effect of channel variables (SNR, number of idle channels, etc.) and radio implementation parameters (PLL lock time, N-point FFT, etc.) on the total mean detection time. We minimize our detection time by optimizing the coarse and fine bin sizes in our 2-stage sensing algorithm. In order to achieve an equilibrium point, we perform a large scale optimization on the mean detection time with respect to bin sizes. Coarse sensing is faster than fine sensing, however, it is not as accurate. As the number of users in a channel increases, the number of fine bins increases which directly affects the total scan time. Hence, we optimize our sensing time by striking a balance between the fast, lower accuracy coarse detection versus the slower, more accurate fine sensing operation.

We augmented our total sensing time with detection probability and false alarm probability to define a mean detection time for the system. The addition of these probabilities gave us a more realistic model for our sensing receiver. The detection probability is dependent on SNR of the received signal and the channel occupancy. We showed that the mean detection time is sensitive to the bin sizes and the number of bins  $\beta$  and  $\alpha$ .



With the detection time dependent on many variables, we focused on minimizing our mean detection time by optimizing  $\beta$  and  $\alpha$ . The optimization process takes into account all parameters such as number of available channels, detection probability, false alarm probability, SNR, PLL locktimes and the number of channels or antennas.

In the next chapter, we present our simulation platform in MATLAB. We emulate the wireless environment and present the mechanism how the DSR senses and prioritizes the available channels for cognitive radio use.

## 6 DSR Simulation Platform

One important aspect of the research is to develop a simulation platform that emulates the DSR functionality. The improvements of the DSR were demonstrated in chapter 4 and the optimization was presented mathematically in chapter 5. In this chapter, we present the DSR simulation platform that was developed in MATLAB [107]. The platform senses the spectrum and dynamically prioritizes the available channels in a LUT.

The DSR offers immediate improvements to the sensing time and a mathematical model is derived. The DSR uses power estimation techniques at several stages in the receiver which improved spectrum sensing and shortens the decision time. The band of interest is divided into coarse and fine bands, and the total sensing time is significantly dependent on the number of bins and the number of FFT points. The mean detection time is dependent on several factors that may or may not be under user control or implementation. Hence, we focused on optimizing the mean detection time by optimizing the number of coarse and fine bins that we dedicate for a 2-step sensing process. After the bin sizes are chosen, the DSR must sense and prioritize the available channels in a LUT table. The prioritized channels are used by the main receiver for cognitive radio operation. The main challenge is for the DSR to continuously sense the spectrum and appropriately rank the channels. Since the frequency spectrum is very dynamic and users may appear without warning, the burden is on the DSR to make accurate decisions.

We develop a platform in MATLAB and Simulink which is a graphical user interface of MATLAB. The motivation of this effort is to demonstrate the DSR capability to dynamically detect and rank channels based on the power detected. We divide the platform into three parts:

- 1- Spectrum activity model which emulates activity within a spectrum
- 2- The radio receiver which performs the down conversion
- 3- The DSR algorithm which detects power and prioritizes available channels in a look up table

The platform is presented in upcoming sections.

## **6.1 DSR Simulation Platform Description**

MATLAB is one of the most popular simulation environments in the world [108,109]. Its acceptance is mostly due to its user friendly interface and its computing power. It has been widely used in classrooms around the world. Its user interface is based on C and FORTRAN-like programming languages. Its visualization package allows users to display results and data in a simple and easily understandable format. MATLAB was originally developed for the manipulation of system matrix. Simulink is an extension of MATLAB which allows systems to be simulated dynamically. The graphical user interface allows the system to be model at a high level block diagram level.

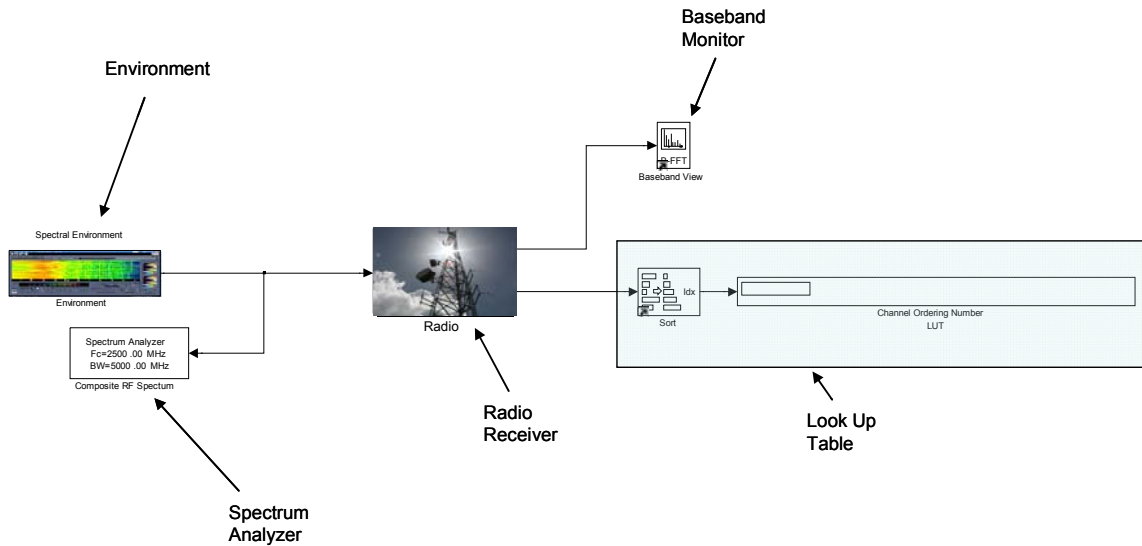
Our motivation to develop this platform is to demonstrate the capabilities of the DSR. We were able to mathematically show that the DSR improves the sensing performance. However, we need to demonstrate the sensing and the prioritization algorithm ability to dynamically choose the right channel.

The simulation is divided into three major blocks:

- 1- The frequency spectrum model
- 2- The radio which receives the signal and converts to digital baseband
- 3- Power detection and LUT prioritization algorithm

The high level diagram of the DSR is show in Figure 6-1.

## DSR Simulation Platform



**Figure 6-1 DSR Simulation Platform in Simulink**

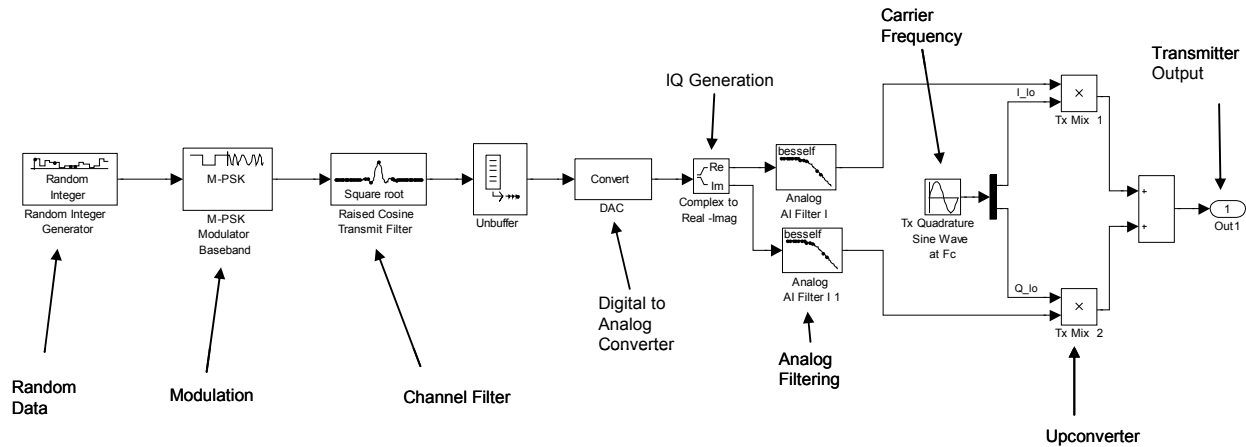
Each block is further described in the upcoming sections.

### 6.2 Spectrum Activity Model

The spectrum activities are dynamic and mostly unpredictable. In general for authorized users, the spectrum is assumed available and ready for transmission. Hence, a licensed user may transmit at any time. In order to represent the environment in the DSR platform, we emulated the activities of 4 transmitters: 900MHz cellular band, 1800MHz cellular band, a WLAN operating in the 2.4GHz ISM band and a UWB user at 4GHz.

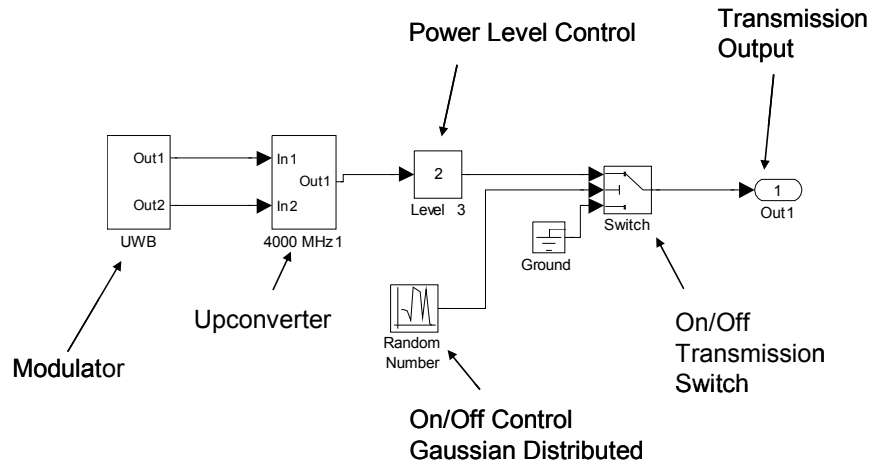
A transmitter block diagram is shown in Figure 6-2.

## Transmitter Block Diagram



**Figure 6-2 Transmitter Block Diagram**

The output of the transmitter for the channel is connected to a manual power level. Power level emulates the signal strength of the PU as seen by the DSR. For example, if the DSR is closer to a cellular base station (900MHz), then the level at 900MHz is set higher than those at other frequencies. The output of the transmitter is then connected to an on / off switch which controls the transmitter output. If the control signal of the switch  $>0.5$  then the transmitter is turned on, otherwise the transmitter is turned off. In order to emulate activities on the channel, the on/off switch control is connected to a Gaussian distributed random signal generator. Depending on the signal type, we can control the probability of the transmitter being on versus off. For example, the spectrum activity at in the cellular band is relatively high (i.e. high traffic). In this case, the mean value of the Gaussian distribution controlling the switch at the cellular transmitter output is set to 0.9 with a variance of 0.1. For a UWB transmitter the traffic is lower, in this case, the mean is set to 0.5 with a variance of 0.25. A transmitter is shown in Figure 6-3.



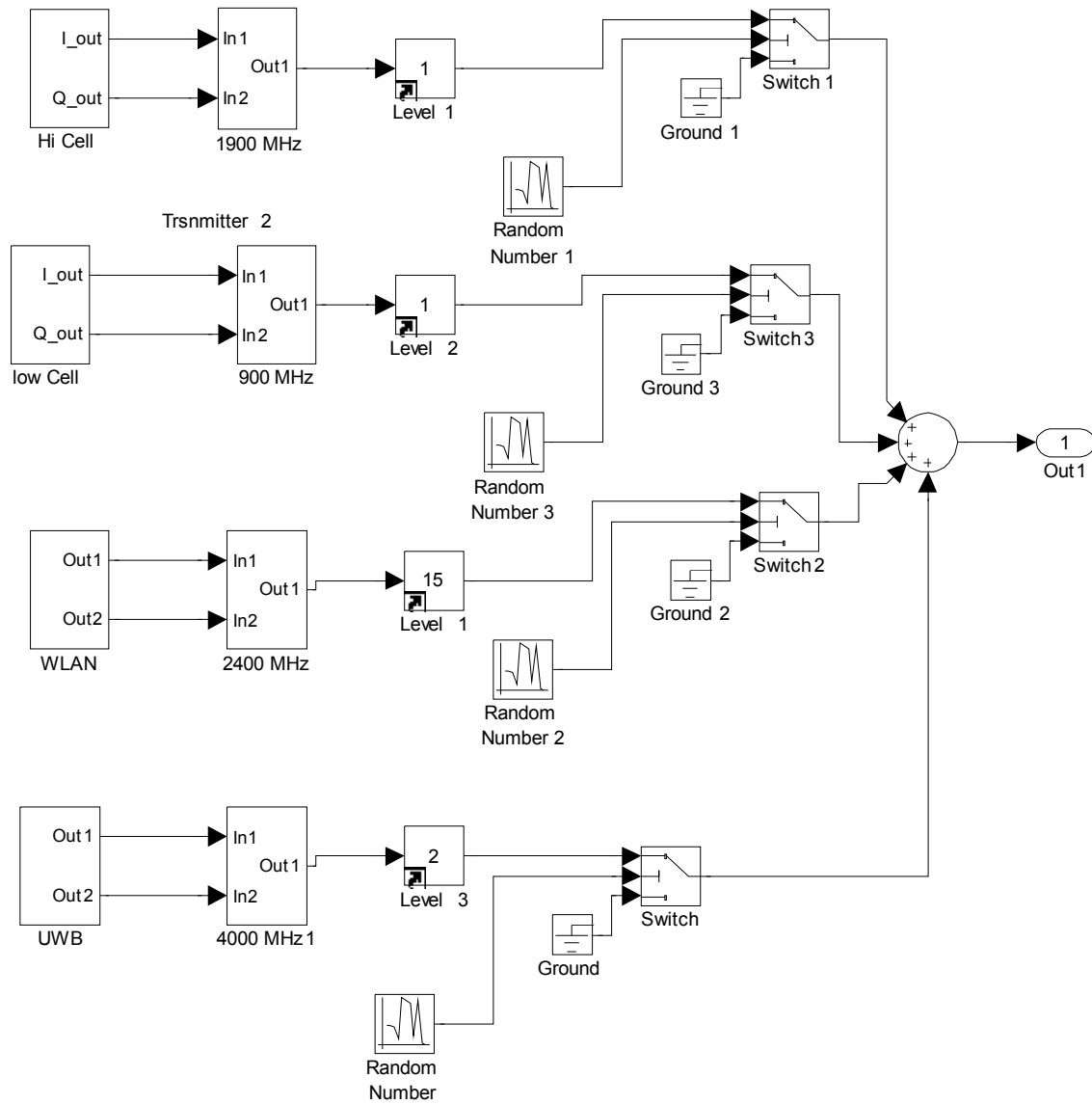
**Figure 6-3 Transmitter Output Control**

The switch control distribution for each standard is shown in Table 17.

| Standard         | Mean | $\sigma$ |
|------------------|------|----------|
| Cellular 900MHz  | 0.9  | 0.1      |
| Cellular 1900MHz | 0.5  | 0.3      |
| WLAN 2.4GHz      | 0.9  | 0.25     |
| UWB 4.0 GHz      | 0.5  | 0.25     |

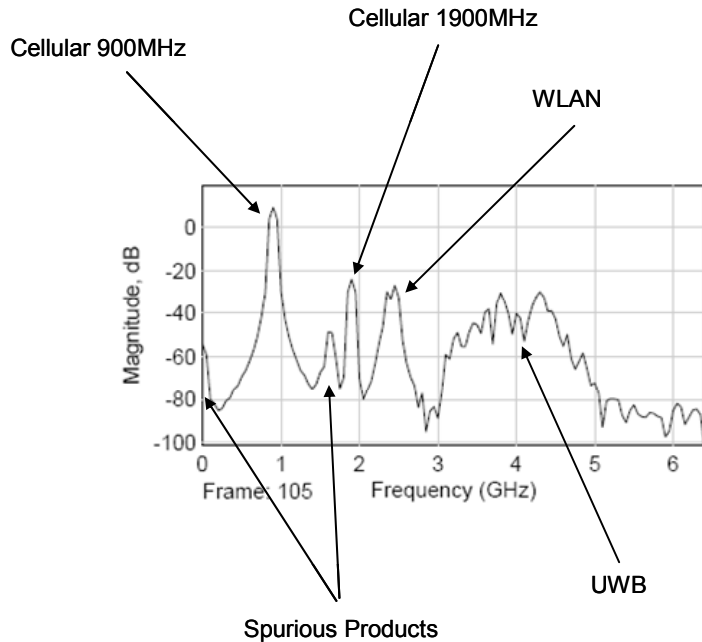
**Table 17 Switch Probability Distribution**

The output of the 4 transmitters are then combined together to simulate a broad spectrum. The overall environment is shown in Figure 6-4.



**Figure 6-4 Environment Platform**

The output of the transmitter is connected to a monitoring spectrum analyzer shown in Figure 6-5.



**Figure 6-5 Simulation Environment Output**

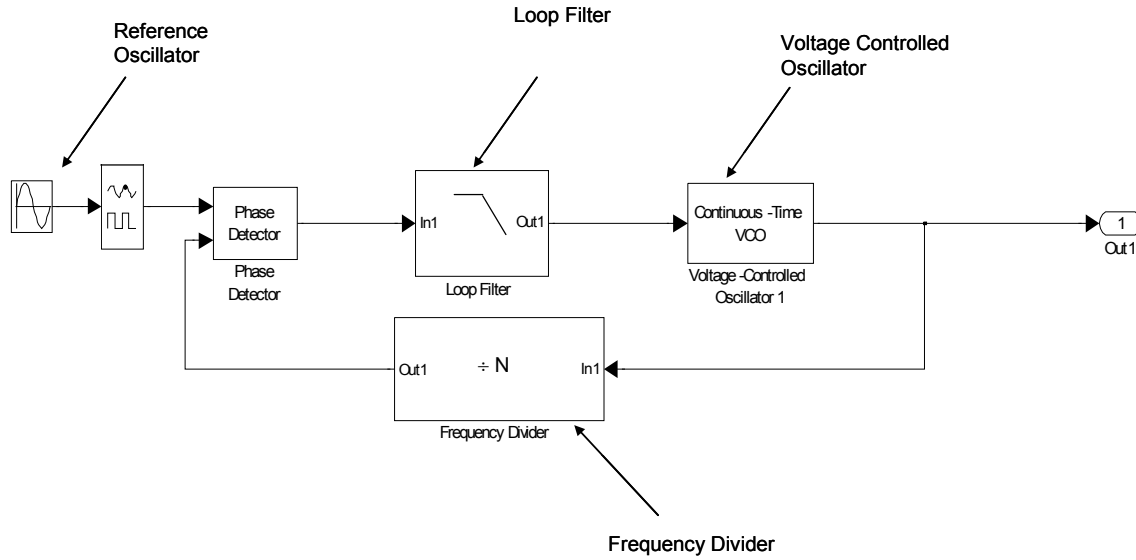
The environment is then fed to the input of the DSR.

### 6.3 Receiver Model

The signal in the DSR is received and demodulated before passing through an analog to digital converter. The receiver downconverts the received signal via mixing the signal with a local oscillator (LO) tuned at the desired frequency. The LO frequency is generally generated by a PLL locked to a reference frequency as shown in Figure 2-17.

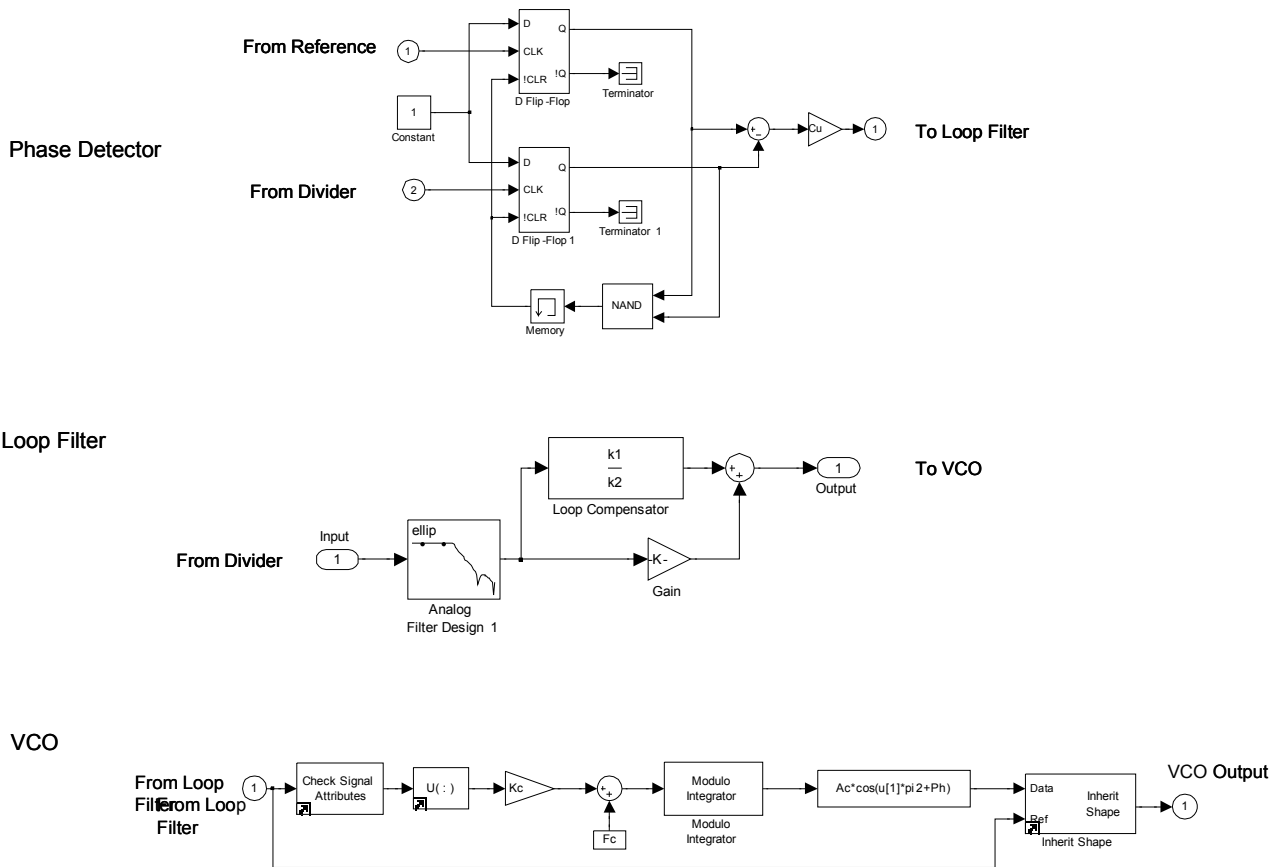
In our platform, a PLL is used to generate the mixing frequency of the LO. The block diagram is shown in Figure 6-6.





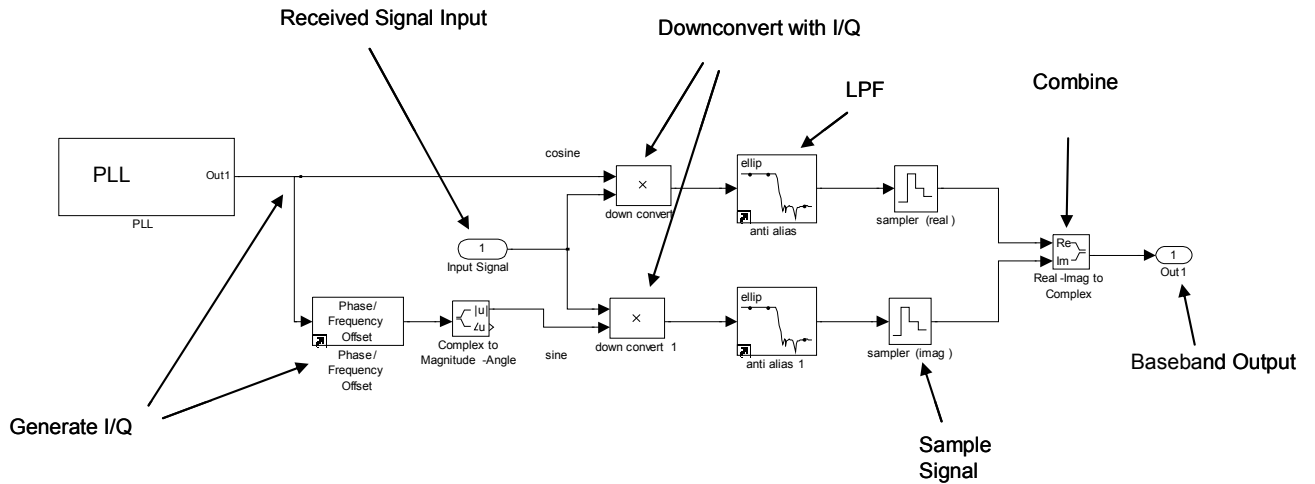
**Figure 6-6 PLL Block Diagram**

The PLL components are shown in Figure 6-7.



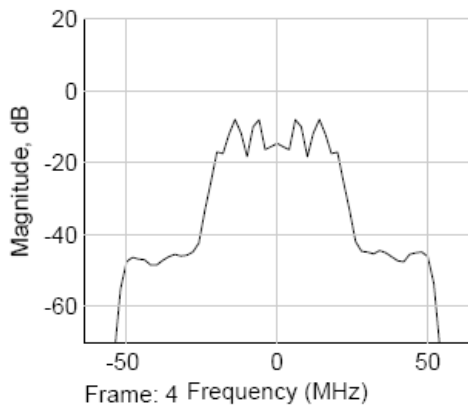
**Figure 6-7 PLL Components - Phase Detector, Loop Filter and VCO**

The output of the PLL is connected to the receiver as shown in Figure 6-8.



**Figure 6-8 Radio Receiver Block Diagram**

At the output of the down converter the signal is shifted to baseband and ready for conversion to baseband via signal sampling. The output of the receiver before sampling is shown in Figure 6-9.

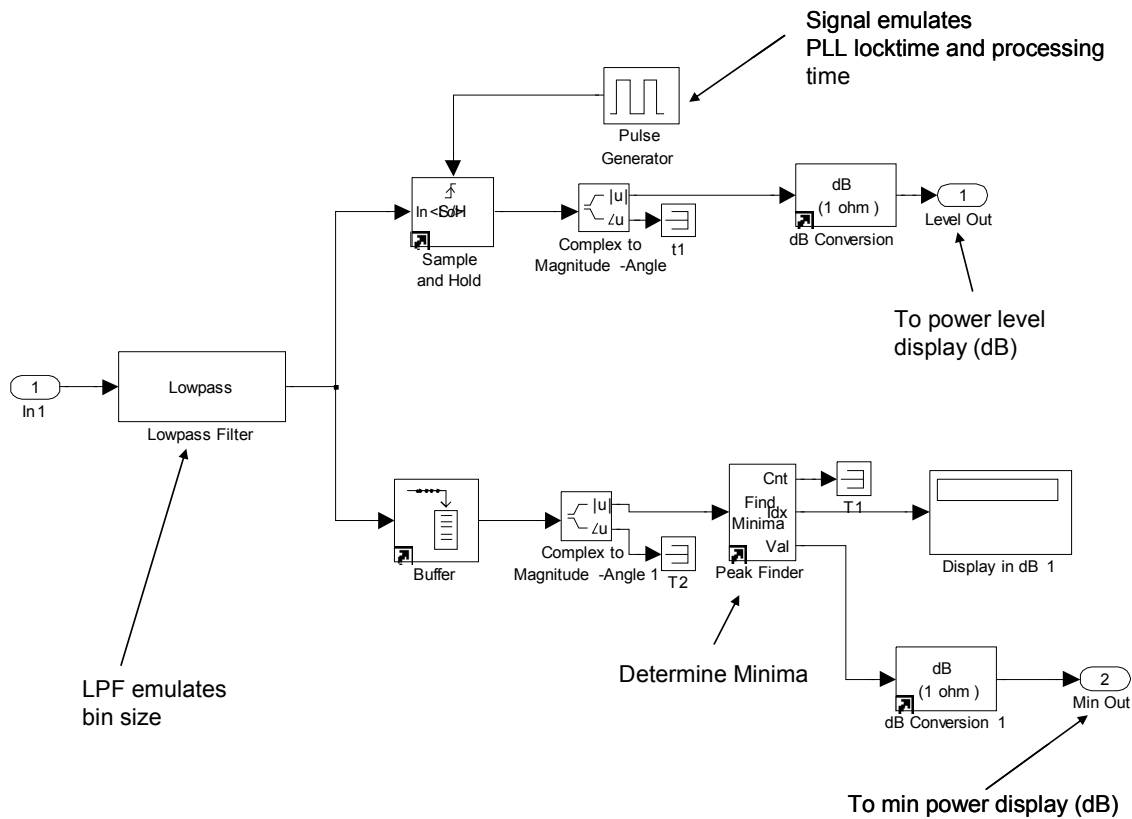


**Figure 6-9 Baseband Output of the Receiver**

Once the received signal is downconverted and digitized, the signal is ready for digital signal processing.

## 6.4 Power Detection and LUT Prioritization

The DSR tunes to a desired band, downconverts to baseband (i.e. direct conversion receiver) and samples the signal. After the conversion of the signal to the digital domain, the DSR is now ready for power detection and algorithm processing. The power detection is shown in figure in Figure 6-10.



**Figure 6-10 DSR Signal Processing**

The lowpass filter is set to represent the bin size. The data is captured and FFT is performed. The data power level is displayed. In order to represent the proper switching time, a trigger is used to capture data via a sample and hold. As the DSR switches to tune to a desired channel, the data processing is not performed. After settling, the data processing is performed.

The other aspect of the signal processing in Figure 6-10 is to find the minimum power and capture the data. Once the data is captured, the DSR sorts the channels with the lowest power at

the top of the LUT. A snapshot of the channel sorting and LUT population is shown in Figure 6-11.



**Figure 6-11 Channel Sorting and LUT Population**

The main CR receiver then uses the information in the LUT to determine which channel to use for its next assignment.

## 6.5 Results Snapshot

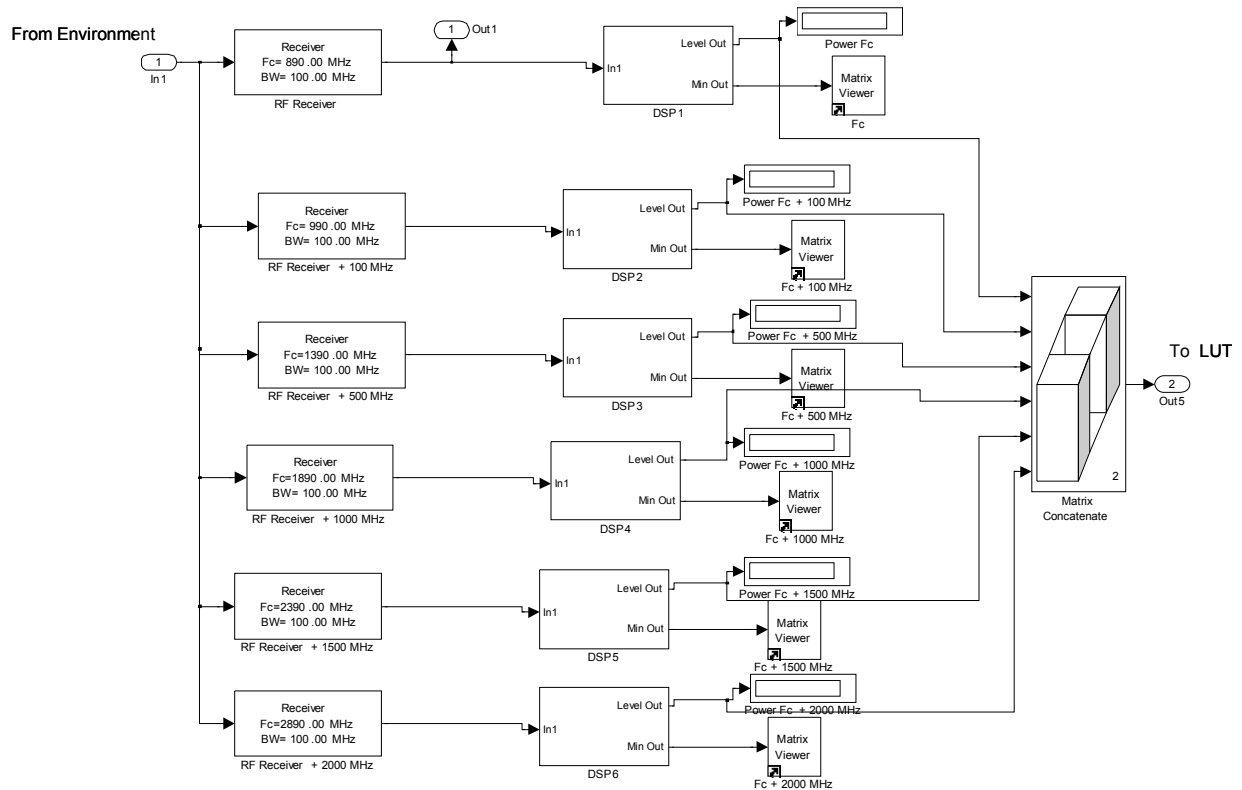
In order to simulate the DSR, we review the power received in 6 channels. This is sufficient to show the efficiency of the DSR. The DSR sets a center frequency noted as  $F_c$  at 900MHz and we measure the power at 6 channels. The channels are shown in Table 18.

| Channel   | Operation (MHz) | Frequency (MHz) |
|-----------|-----------------|-----------------|
| Channel 1 | $F_c$           | 900             |
| Channel 2 | $F_c + 100$     | 1000            |
| Channel 3 | $F_c + 500$     | 1400            |
| Channel 4 | $F_c + 1000$    | 1900            |
| Channel 5 | $F_c + 1500$    | 2400            |
| Channel 4 | $F_c + 2000$    | 2900            |

**Table 18 Channel Numbers and Frequencies**

Recall that our environment has activities at 900MHz, 1900MHz, 2400MHz, and 4000MHz. Without running the simulation, we would expect in general that Channel 2, Channel 3, and Channel 4 to be ranked higher than the other 3 channels. Also, since Channel 2 is operating at 100MHz separation from an active channel, we would expect the noise in Channel 2 to be higher than that in Channel 3, for example, which is 500MHz from an active channel. This condition is not necessarily always true since Channel 1 is not always active and might cease transmission.

The block diagram of the DSR is showing the 6 channels in Figure 6-12.



**Figure 6-12 DSR Channel Processing**

A snapshot of the results is shown in Figure 6-14



**Figure 6-14 Snapshot of Platform Results**

The output of the platform is made to be visual. The power detected is shown on a spectrogram that displays the results from -80dB to 0dB. As the power level increases, the power goes from a cool color (blue) to a hot color (red). For example, the signal at 1900MHz seems to be the highest power on the spectrum analyzer. Consequently the power detection shown for channel 4 is the hottest (more red and orange colors than channel 1 for example). Hence, the user is able to visually correlate the activity in the spectrum with the DSR output with a given implementation delay. In this case, the channels in the LUT are ranked from the lower power detected to the highest with a processing delay. As the channel dynamics change, the power level change and hence the LUT is updated.

As discussed earlier, Channel 3 seems to be better than channel 2. This is due to the proximity and the signal level of Channel 1.

## **6.6 Summary**

The simulation of a system is a very important aspect of system design and implementation. Computer simulations are common and accepted tools for prototyping or modeling a solution. We have developed a simulation platform in MATLAB and Simulink that closely models our DSR system.

The system is divided into three major parts: the environment, the radio and the process. The environment model is designed to be dynamic at different frequencies and different power levels. The radio downconverts the received signal and converts to a digital signal. The signal power is estimated and comparatively ranked versus other channels. The LUT is continuously updated. The simulation visually displays the sensing results.

We were able to show theoretically how the DSR improves the sensing time. One aspect that was challenging to prove was the dynamic operation of the DSR. One of the key contributions of the DSR is the updates of the LUT. In our simulation, we have successfully demonstrated the dynamic operation of the DSR. This concept can easily be extended to other bands, frequencies and more importantly may be augmented with other algorithms and techniques.

## 7 Conclusions and Future Work

This chapter concludes this dissertation with a brief discussion of the work to date and highlights the contributions to the field. It also previews possible research for the future.

Exclusive rights licenses for spectrum use are inefficient and cause a significant underutilization of the spectrum. Due to rapidly increasing demands on wireless resources, dynamic spectrum access is becoming a necessity in order to leverage the underutilized portion of the frequency spectrum. Secondary users may opportunistically use the spectrum when the licensee or primary user of the spectrum is not operational. However, the burden is carried by the secondary user to recognize and to seize the opportunity to utilize an unused spectrum, hence the name Cognitive Radio.

With greater demand for high data rates and mobility, cognitive radios represent one of the most significant paradigm shifts in wireless engineering in the past three decades. A cognitive radio senses the local environment, adapts itself to the environment, and adjusts its transmission parameters to meet its objectives and to minimize interference with other occupants of the spectrum. Several advances in approaches to spectrum management, radio policies, radio design, and network design are inching cognitive radios towards commercial reality. However, the key to cognitive radio realization is a practical sensing receiver. To date, the sensing receiver remains a challenge.

The focus of our research is the practical implementation of a sensing receiver for cognitive radio use. Due to the highly dynamic spectrum, the sensing receiver must make fast decisions, minimize false alarm or false positive, and must be able to operate over broadband.

We demonstrated that our implementation of a Dedicated Sensing Receiver (DSR) has the ability to meet the requirements of cognitive radios. The DSR realization requires execution at the physical and MAC layers. We demonstrate that the radio is feasible in today's technology. The algorithm controlling the radio is an integral piece of the DSR. The algorithm divides the desired



spectrum into coarse and fine bins which are scanned for users. The power detection feature of the algorithm is time sensitive and varies as the sensing goals vary. We developed an adaptive algorithm that modifies the parameters as the environment changes states. We derived a mathematical expression for sensing time of DSR that takes account both environmental parameters as well as models for components of the radio. The mathematical model proves that the sensing time may be minimized such that total mean detection time is acceptable for time sensitive applications such as voice or video application.

The bin sizes considerably affect the mean detection time. To further improve the DSR, we optimized the solution by allowing the algorithm to find optimal bin sizes for coarse and fine sensing. The bin size changes as the environment changes. The optimization process improves the DSR and the cost of implementation is minimal.

Modeling and simulation are essential for the performance evaluation of wireless systems. MATLAB and Simulink are popular tools and widely used in the industry and academia. A DSR simulation platform was developed in Simulink. The platform demonstrated the adaptive application of the DSR algorithm. The simulation correctly identified holes in the spectrum and ranked the available channels in a look up table that is available for immediate radio operation. Fast response and continuous update further establish the DSR as a viable and practical solution to the practical sensing for cognitive radios.

Cognitive radio is still a relatively new research area. While this dissertation lays a strong foundation for the development of a practical cognitive radio, there are significant opportunities for further work in the following areas:

- Expansion of the 2 sensing levels (coarse and fine) to multi-dimensional sensing. Here, the receiver adds new sensing levels. We expect that the additional levels especially in busy area of the spectrum would improve the detection time while minimizing the opportunity cost.

- Adaptive allocation of fine sensing bins within the coarse bins. Here, the number of fine bins may be allowed to be different for each coarse bin. By varying the fine bin size, the sensing receiver may reduce the fine sensing bins where the spectrum is not busy to improve sensing time, while increasing the number of fine bins where spectrum activity is large.
- Expansion of the DSR to take into account some of the network techniques. We expect the DSR overall performance to improve in a cooperative network. Determine the improvements to the DSR as incremental network improvement are added.
- Self-learning and gaming theory adapted to CR operation. We expect the DSR algorithm to benefit from techniques that allow for the algorithm to learn and predict the potential appearance of the primary or secondary users based on the statistical information and data collected on the channel.
- Continuous improvement on radio architecture. As the technology continues to evolve over time, we expect that receiver design and PLL / VCO design will continue to improve.
- Focus on economic impacts and policies. The area of standardization and policy making for cognitive radio is still relatively new. Significant research may be targeted at defining appropriate regulatory policies.
- Improvement in spectral detection techniques. As the DSP technology continues to evolve while the price and power consumption continue to drop, additional spectral detection techniques may be considered.

In summary, spectrum sharing promises to become a viable solution to resolving the apparent scarcity of the frequency spectrum. The DSR is a major step towards finally realizing this goal; the DSR provides drastic improvements to spectrum sensing and could be the catalyst for enabling commercial development of cognitive radios in the very near future.

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