The development and practical study of a grey space detector for cognitive radio

by

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A thesis submitted to the
Department of Mathematics & Statistics
in partial satisfaction of the requirements
for the degree Masters of Science

Queen's University
Kingston, Ontario, Canada
August 2010

Acknowledgments

Firstly, I would like to thank my supervisors Dr. David J. Thomson and Dr. Glen Takahara for there guidance and support throughout the past year.

I would also like to thank Wesley Burr for his invaluable assistance in the research and preparation of this thesis.

I am thankful to my family for many years of encouragement, wisdom and love.

Lastly, I would like to give thanks to Erin for all her love and for listening to hours of Statistics without even an introductory understanding of matrices .

Table of Contents

List of Tables

List of Figures

Table of Contents

List of Tables

List of Figures

Chapter 1

Introduction

In 2003 the FCC began evaluating the potential reuse of underutilized spectrum within the TV bands. In 2010, the IEEE 802.22 standard for cognitive radio will be implemented. With this in mind the development of a detection scheme for a cognitive radio system that is both practical and efficient will be beneficial to advancement of the field.

In an effort to avoid creating unacceptable levels of interference, this project will focus on reusing spectrum that has primary signals present. We will need to determine whether a primary signal is fully or partially occupying a frequency at our location.

As more cognitive users begin to reuse frequencies, other areas within the TV bands should be explored. We will design a detector to find TV signals that are present but are below a power limit. This will make additional interference from reuse to the TV receivers acceptable. This detector will be useful in many geographic locations and could be combined with an effective white space detector in a cognitive radio system to optimize spectral use within the TV band.

The detector will use data filtering, Multitaper spectral estimation and the harmonic F-test to determine the presence and strength of digital television signals. We will then test our detector against observed data taken from Queen's University with a residential TV antenna.

We will show the detector is effective at finding TV signals and determining whether spectral bands are fully occupied or not in a realistic urban environment. This study will show the need for further regulation of cognitive radio and provide solutions to possible design issues.

Chapter 2

Background and Literature Review

In recent years greater and greater demands have been placed on the natural resource of the radio spectrum by both transmitters and receivers. Under the current regulations, there is insufficient radio spectrum to meet the allocation demands. Because of this, in 2002 the Federal Communications Commission (FCC) published a report prepared by the *Spectrum Policy Task Force* [8, 9], aimed at improving the management of this resource in the United States. The report found that

> *In many bands, spectrum access is a more significant problem than physical scarcity of spectrum, in large part due to legacy command-and-control regulation that limits the ability of potential spectrum users to obtain such access.*

From this report came increased interest in alternative resource allocation methods for radio spectrum. One of these methods is cognitive radio.

2.1 Cognitive Radio: Fundamentals

The term cognitive radio was coined in 1992 by Joseph Mitola III [15]. Mitola further described how cognitive radio could enhance the flexibility of personal wireless services [16, 17], expanding the ideas of cognitive radio in his doctoral dissertation at the Royal Institute of Technology, Sweden [23]. Following Mitola's work, Simon Haykin published several papers [11, 12, 13] on cognitive radio which became the foundation for much of the current research on the subject. Haykin defines cognitive radio in [11] as

> *... an intelligent wireless communication system that is aware of its surrounding environment (i.e., outside world), and uses the methodology of understanding-by-building to learn from the environment and adapt its internal states to statistical variations in the incoming [radio frequency] stimuli by making corresponding changes in certain operating parameters (e.g., transmit-power, carrier-frequency, and modulation strategy) in real-time, with two primary objectives in mind: 1) highly reliable communications whenever and wherever needed; 2) efficient utilization of the radio spectrum.*

From Haykin's original paper [11], the basic stages for cognitive radio system design were formed. In the sensing stage, the cognitive radio transmitter would sense the radio frequency (RF) environment, attempting to determine the presence of primary signals. Then in the transmission stage, if no such signals were present, the cognitive radio would begin transmitting over this frequency. Periodically, the cognitive radio would need to reassess the availability of the frequency it occupies to ensure it does not interfere with any proprietary signals. If a signal were found, the cognitive radio would stop transmitting and begin to search for a new frequency in which to operate.

Along with the basic procedure for cognitive radio, Haykin presented the idea that when scanning portions of the radio spectrum at a particular time and place, we would find that, 1) some frequency bands in the spectrum are unoccupied; 2) some frequency bands are partially occupied; and 3) the remaining bands are fully occupied by legacy users. These three situations are classified as white (unoccupied frequencies), grey (partially occupied) and black space (heavily occupied). Another definition similar to white space is that of the spectrum hole. A spectrum hole, first defined by Kolodzy [18] in 2002, is a band of frequencies assigned to a legacy user that, at a particular time and geographic location, is not being utilized.

2.2 Spectrum Allocation

Following the increase in scientific interest in cognitive Radio, in 2004 the FCC proposed the use of TV bands for cognitive radio [28]. In the report the FCC stated that under the proposed rules,

> *... unlicensed devices will be able to transmit on channels used for receipt of distant broadcast television signals, therefore increasing the likelihood that there will be interference with a local broadcast signal received from outside the Grade B contour, particularly in rural markets.*

Robert A. O'Connor, an IEEE senior member, defines a Grade B contour as *the outer geographic limits within which the median field strength equals or exceeds the Grade B value* [24]. The specific values for Grade B are found in Table 2.1 below.

Power	Channels	Frequency Range
47 dBu (0.22 mV/m)	2 to 6	54-88 MHz
56 dBu (0.63 mV/m)	7 to 13	174-216 MHz
64 dBu (1.6 mV/m)	14 to 83	470-890 MHz

Table 2.1: Grade B contour field strength values.

Channels	Frequency Range	Average Occupancy
2 to 6	54-88 MHz	12.0%
7 to 13	174-216 MHz	10.1%
14 to 20	470-512 MHz	11.9%
21 to 36	512-608 MHz	24.3%
37 to 51[1]	608-698 MHz	27.1%
52 to 69	698-806 MHz	12.8%

Table 2.2: Spectral Occupancy Measurements Project findings.

Also in 2004, the *Spectral Occupancy Measurements Project*, headed by Mark A. McHenry, published its findings. The goal of the project was to determine the spectrum occupancy (i.e. total *actual* usage) in each frequency band across multiple locations. Their findings are summarized in Table 2.2. The occupancy measurements given in Table 2.2 are the average occupancies for all tested locations. The occupancy observed in urban centres like New York City and Vienna, Virginia were as high as 50% in the UHF (channels 14-69) bands and 77% in the VHF (channels 2-13) bands. While the occupancy in the TV bands for these locations were the highest observed, there was still an opportunity for reuse through cognitive radio.

Much of the motivation to operate in the TV bands comes from the desirable qualities of the primary signals, including a fixed modulation scheme, maximum bit rates and bandwidth locations. In 2009 the FCC instated the *Advanced Television*

Systems Committee (ATSC) standards for digital television. This same standard is set to be instated in Canada in 2011. Within the ATSC, digital television must use *8 vestigial sideband modulation* (8VSB) for terrestrial signals (see Appendix B for more).

After the 2004 report, the IEEE 802.22 working group on Wireless Regional Area Networks (WRAN) was formed. The IEEE 802.22 working group is currently developing a standard for using white spaces in the VHF and UHF TV frequency bands. In the IEEE 802.22 overview by Stevenson *et al.* [34], they state

> *The developing IEEE 802.22 standard will allow broadband access to be provided in sparsely populated areas that cannot be economically served by wire-line means, or other wireless solutions at higher frequencies, by using cognitive radio techniques to allow operation on a non-interfering basis in the VHF/UHF TV broadcast bands. This will increase the efficiency of utilization of that spectrum, and provide large economic and societal benefits.*

In the draft of the 802.22 standard, two methods for determining spectral allocation are proposed. The first is a centralized method, with GPS-enabled base stations communicating with central servers which would respond with the available bands for the base station's location. The second is a local sensing scheme, where base stations would sense their own radio environment and make independent decisions. The centralized method is preferred for fixed-location devices, while dynamic sensing is preferred for portable devices. The centralized case requires the development of policies and guidelines to be effective while the case for portable devices relies on the applications of spectral analysis to develop effective detection procedures.

2.3 Detecting White Space

The detection of white spaces has been an ongoing field of interest for several years. One of the first detectors used in this pursuit was the energy detector. First proposed for detecting Gaussian signals by I. Jacobs in 1964 [14, 43], the energy detector attempts to determine the presence of a zero-mean signal x,

$$\left(\mathbb{E}\{x\} = 0, \qquad \mathbb{E}\{x^2\} = P\right), \tag{2.1}$$

in the presence of additive white noise by evaluating the average power of the sample signal within a frequency band. The test statistic

$$\frac{1}{N}\sum (Y(n))^2, \tag{2.2}$$

where $Y(n)$ is the received signal, follows a central chi-squared distribution when no signal is present and a non-central chi-squared when a signal is present. The threshold for the test statistic is dependent on the probabilities of Type I and II error desired and the sample size. The threshold is given by these two formulas:

$$P_{fa} = Q\left(\frac{\gamma - N\sigma_w^2}{\sqrt{2N\sigma_w^4}}\right) \tag{2.3}$$

$$P_d = Q\left(\frac{\gamma - N(\sigma_w^2 + \sigma_x^2)}{\sqrt{2N(\sigma_w^2 + \sigma_x^2)^2}}\right), \tag{2.4}$$

where γ is is the threshold for the test statistic, N is the number of samples, σ_w^2 and σ_x^2 are the noise and signal variances and the Q function is the tail probability of the standard normal distribution. There are several limitations to the energy detector which must be considered. To meet Type I and II error constraints the sample size N must be large enough for the central limit theorem to hold and a bandpass filter must be applied; both of these will increase the sensing time required. As well, the energy detector cannot perform well in low signal-to-noise environments. The number of samples required is a function of the signal-to-noise ratio (SNR) and error probabilities,

$$N = 2\left[\left(Q^{-1}(P_{fa}) - Q^{-1}(P_d)\right) SNR^{-1} - Q^{-1}(P_d)\right]^2. \tag{2.5}$$

When the SNR is low (SNR $\ll 1$) the function will not meet the error probability requirements desired given N samples.

Tandra and Sahai [35, 36] showed that in the presence of noise uncertainty, which occurs in most practical situations, the energy detector fails to detect when

$$SNR < 10 \cdot \log_{10}\left[(10)^{x/10} - 1\right] \text{dB}, \tag{2.6}$$

where x is the noise uncertainty in dB. This limit is known as the *SNR wall*. This demonstrates that the energy detector is not a practical method and raises several important design issues for future detectors.

Another class of detectors that theoretically do not encounter SNR limitations in the presence of noise uncertainties are the class of *coherent detectors*. Coherent detection uses known characteristics of possible signals to make detection easier. Many modulation schemes use pilot tones and we can use the knowledge of the location of these pilots to produce a matched filter to increase the signal-to-noise ratio and determine the presence of the signal of interest (see Appendix A). The process of using a matched filter averages the noise while maintaining the power of the pilot. When averaged, the noise variance becomes σ^2/N and asymptotically any uncertainties in the noise variance become negligible. Unfortunately in *practical* situations we are restricted to finite sample sizes and the benefits of coherent detection are limited. Also, in the presence of multipath fading (see Appendix C), coherent detection is only possible when the multiple realizations of the signal are coherent, known as a *coherence block*, imposing a limit for the sample size and lowering the robustness of coherent detection towards noise uncertainties.

In this situation, combining the signal coherently in each block and then using the energy detector over multiple coherent blocks is the optimal coherent detector. The test statistic is

$$T = \frac{1}{N}\sum\frac{1}{\sqrt{N_c}}\sum\Big(Y(n)\hat{X} - p(nN_c + k)\Big) \tag{2.7}$$

where N_c is the number of samples in each coherent block. This detector can be thought of as a combination of two detectors. The signal is coherently combined within each block. The signal strength is boosted by a factor of N_c while the noise power does not change. Then, the boosted signal is detected at the receiver by an energy detector. Since we are reduced to an energy detector, our final detector is still non-robust to noise level uncertainties. However, this method is an improvement and the SNR wall is lowered by a factor of $10\log_{10} N_c$ dB.

Another procedure attempts to estimate the spectrum directly via the power spectral density (PSD). In the past, when attempting to estimate the PSD, the choice of method would require us to deal with the trade-off of bias and variance. The bias of the power-spectrum estimate of a time series is due to the side-lobe-leakage phenomenon, and to reduce this effect we window (or taper) the time series. However, by windowing the time series we increase the variability of the estimate due to the loss of information resulting from a reduction in the effective sample size.

This trade-off was resolved for stationary time series in 1982 when David J. Thomson developed the *Multitaper Spectral Estimate* (multitaper) [37, 38] (see Section 3.2). This technique allows for reduction in bias without the corresponding increase in variance, and is the preferred method for estimation of power spectra.

Another advancement that came courtesy of the multitaper was the *Harmonic-F* *test* (see Section 3.3). Because the Harmonic F test does not rely on the amplitude of a signal for detection, the uncertainties within noise estimates have no effect on the performance of this detector. However, while the F-test is robust against noise floor uncertainties, there may be other forms of uncertainty to which it is not robust. As an example, in Sahai *et al.* [29], the authors discuss the uncertainty created by multiple secondary users working within a geographic region. They present the result that the uncertainty of the *presence* of other secondary users would cause the variance of the noise to be unknown. This is similar to noise floor uncertainties, but since other secondary users are sending signals, they would show up on an F-test as a signal, making the F-test not a fully robust detector.

To reduce the effects of multiple secondary user uncertainties, the authors propose implementing a sensing *Medium Access Control* (MAC) where all other secondary users within a region will stop transmitting when one secondary user is sensing. This idea does not appear practical due to the loss of transmission time and coordination needed between secondary users. A system could operate a sensing MAC when in a fixed secondary network, optimized for minimal wait times, but is not a viable solution for mobile systems. Another concern with the proposed sensing MAC is the opportunity for antagonistic cognitive systems to exploit the sensing times and gain more of the available spectrum.

Another possible method uses known signal classification techniques to determine the presence of secondary or primary signals. In Palicot and Roland [25] the authors studied European wireless communications and proposed several parameters for classifying signal types. The primary parameter they chose for classification was the channel bandwidth of the signal. They state that the bandwidth will discriminate almost all system types but, in the case where it does not, usually involving single wide band signals (like TV) versus local multiple-point distributions, using the guard band interval would be enough to make a determination.

While the guard band interval would be enough to determine if a secondary user was present in a UHF or VHF band, we will also need to distinguish between analog and digital TV signals. Since they have similar structures, including bandwidth and guard intervals, more detailed classification is needed. Palicot and Roland mention the idea of using the shape of the spectrum to categorize the modulation. Using the positioning of the pilot tone in digital TV signals, which can be found using the F-test, the determination of the TV signal type can be made. The effectiveness of the classification methods given by Palicot and Roland are limited by signal quality. In environments where the SNR is low the bandwidth or guard intervals are difficult to define (see Section 3.3).

As mentioned earlier, an issue that can present when taking measurements in non-ideal environments is *multipath interference*. In very severe cases of multipath interference, a *hidden terminal problem* may occur. A hidden terminal problem happens when a spectrum sensing system fails to detect a transmitter that is in close proximity due to environmental issues. This can be caused by multipath interference, shadowing from a structure like a mountain or building, or high penetration loss when a sensing node is indoors. The hidden terminal problem is an important concern in cognitive radio due to the unforeseen effects it can have on proprietary systems.

Several methods to ensure that no hidden terminals occur during sensing have been proposed. Many of them, including those proposed by Cordeiro and Challapali [4], Zhang *et al.* [44], Chiang *et al.* [3] and Cabric *et al.* [2], rely on cooperation between cognitive radio systems. For mobile systems this is an undesirable action due to the added computation time and need for control channels for secondary systems within which to communicate. Other concerns with cooperation are the aforementioned possibility for exploitation by antagonistic systems and reliance on proximity of other secondary systems.

2.4 Grey Space

In an effort to avoid the hidden terminal problem while not relying on cooperation between cognitive systems, operating in grey space may be desirable. Grey space, as defined by Haykin [11], is *a section of spectrum that is partially occupied by low power signals and is exploitable for reuse.* Whether it be through orthogonal frequency placement or power control, when transmitting within grey space one can cause *minimal* interference to the signals in the band. This minimal but present interference can be regulated, unlike the possibly far worse interference that can occur from a hidden terminal problem.

To ensure the interference is minimal, a strict definition of grey space must be developed. In the 2004 FCC proposal for cognitive radio, the Grade B contour acts as the geographic distinction between grey space and black space. For GPS-equipped devices in centralized systems this geographic cut-off would be acceptable. In the case where local spectrum sensing is the only option, the idea of a protected radius determined by power estimates may be more feasible. The idea of a protected radius for possible receivers, as defined by Sahai *et al.* [29], links distance from the transmitter to observed power. For a known communications system there is a limit on power loss before a signal is no longer decodable by the intended receivers. In TV systems the bound on decodability for residential receivers is a SNR at or above 15dB. All receivers that experience less signal loss are within the radius of decodability. In simple examples a circle centered at the transmitter is used to describe this space but in practical situations a contour similar to the FCC grade B contour would be realistic.

To ensure the receivers inside the protected radius are not affected by an increase in interference from secondary users, the secondary transmitter must account for the power of the transmissions it plans to send and increase the amount of expected signal loss accordingly. The secondary signal must be far enough from the primary radius to incur a minimum path loss so as to avoid effecting the primary receivers. For a secondary system at this distance, the received primary signal must experience an amount of extra signal loss. This extra signal loss is the expected amount for the primary signal traveling the distance outside the protected radius that secondary system must be. The contour for this increased power loss is known as the no-talk radius. In Figure 2.1 we see two examples of no talk radii for differing power levels. The right hand plot shows a secondary transmitter (ST), of high power, located at the no talk radius. Left left hand plot shows a lower power transmitter at its no talk radius. The lower power secondary transmitter is located closer to the primary transmitter, while the high powered secondary transmitter is required to be farther away.

While the rate of decodability is known, developing the extra power loss relies on an assumption of a *path loss model*. Many path loss models exist and choosing the proper model for the situation can be difficult. Some models, like the Okumara or Hata models [1], are designed for certain terrain types and frequencies, in addition

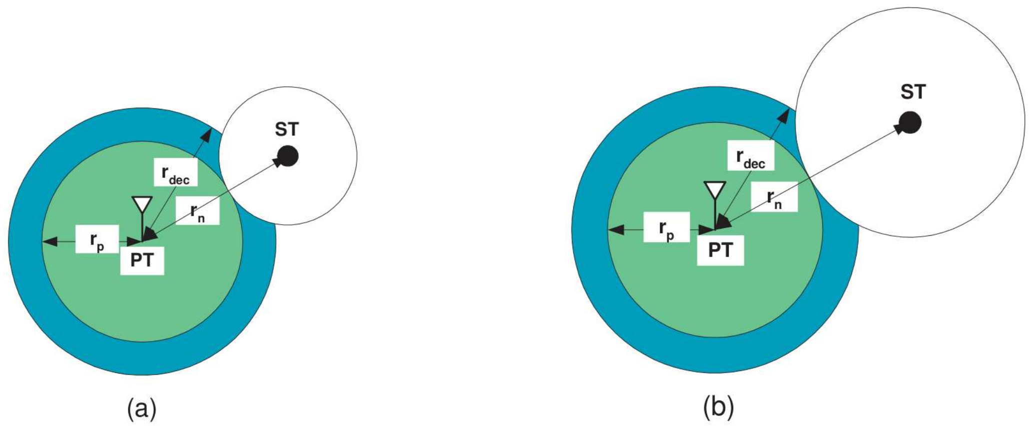

Figure 2.1: Two examples of a no talk radius, from [29].

to requiring information about the primary system that may not be known to the secondary system, like transmitter tower height or distance from the secondary sensing node. These models would be relevant in situations where this information is available but are not robust enough to be used in a mobile cognitive radio system with no centralized information.

Other models are very simplistic: the log distance, polynomial, and Friis [1] models all require less information but do not predict the path loss as well as the more complicated models. These models still require knowledge of the location of the transmitter and receiver, making them less than ideal for mobile systems. For robustness, optimistic models should be used for estimating path loss for grey space detection to avoid the underestimation in ideal atmospheric and geographic conditions.

The other important factor for determination of the no-talk radius is the planned power of transmission by the secondary system. Currently the IEEE are interested in allowing license-exempt devices to operate as secondary systems. This group of devices includes wireless regional area networks and low powered point-to-point or point-to-mobile systems like wireless microphones. There are no plans for extension into cellular phone technologies in the current IEEE 802.22 standard but this is an area for future research. The use of cognitive radio systems for wireless microphones is practical due to the distinct and stable modulation of their signals and low power requirements. The IEEE 808.22 standard allows for many different types of low power networks to operate cognitive radio systems and as technology improves many more of our current terrestrial networks will transfer to cognitive technologies, making the need for reliable detectors more pressing.

Chapter 3

Design Study

With the FCC [28] and IEEE [34] going forward with regulations for the reuse of the UHF TV band by cognitive radio, many white space detectors for the UHF band have been proposed. In an effort to work in conjunction with white space detection, the detection of grey space can also be useful. Reuse of grey space is desirable for its ability to avoid hidden terminals by only operating in partially occupied bands. Also as cognitive radio systems become commercially viable, the number of secondary users will increase and the flexibility of transmitting in grey and white space would help to increase the maximum occupancy for a given region.

The objective of our detector is to find grey space within a portion of the the UHF TV band, $470 - 698$ MHz, that is partially occupied by a digital television signal (DTV). The size of the UHF band tested is open to change but we recommend a minimum four-channel width to ensure a valid estimate of the noise.

The detector is a combination of the harmonic F-Test, an energy detector and coherent detection. The steps for our grey space detector are as follows: data filtering;

multitaper spectral estimation; classification through use of the harmonic F-Test; secondary classification through bandwidth estimation; noise estimation; in-band power estimation; signal-to-noise ratio estimation; and signal placement.

3.1 Data Filtering

To properly evaluate the spectrum of interest we need to first use an *intermediate frequency* (IF) filter to phase shift our spectrum from the $470 - 698$ MHz band to baseband, 0 MHz, with the shift depending on our channels of interest. IF filters mix the observed signal with a known carrier signal through a local oscillator. This produces a signal with similar attributes as the observed signal that is phase shifted to the difference in frequency of the observed and carrier signals. The process of mixing signals and shifting the observed signal to the difference in frequencies is calling *heterodyning*. IF filtering is usually performed in the analog stages of signal processing and can cause unwanted amplitude modulation. Since the use of IF filtering is currently effective and practical in commercial radio and cellular phone systems, IF filters can be easily developed for the down-phasing of our observed data in a commercial setting.

Television signals use 6 MHz bands or *channels* of spectrum to transmit their data. To ensure that we are observing only the frequencies inside our channel of interest, we must remove all other possible signals through the process of filtering. There are many kinds of filters that have been designed with differing applications in mind, but the obvious choice for this application is a *band-pass filter*. A band-pass filter is a combination of low- and high-pass filters. Low- and high-pass filters are both types of *brick wall filters*. They remove only the frequencies below or above an assigned frequency. In our case the 6 Mhz channel is our band of interest. In realistic cases some frequencies just outside of the channel are not fully attenuated, this issue is known as side lobe leakage and occurs because an inverse Fourier transform is used to create a filter in the time domain and the *finite* summation of sinusoidal functions cannot produce a discontinuous function, like our filter design. The error in developing a square portion due to discontinuity issues is known as Gibbs Phenomenon [27].

Although not ideal, the combination of a low- and high-pass filter to isolate the channel of interest is precise enough for most data. TV data helps to minimize side-lobe leakage by having guard bands on both edges of each 6 MHz channel. A guard band is a section of spectrum that is filtered during modulation to be of low power and cause minimal interference to any adjacent signals. The power of a guard band within 8 VSB modulation is filtered to be .02 percent of the average power of the flat middle portion of the signal. The side lobe leakage is negligible in situations where the SNR is under 37 dB with respect to the average power of the flat portion of the signal. In this situation the power of the noise is greater than that of the guard band's allocated power. Most channels operating terrestrial digital signals that are not in close proximity will be operating under a SNR of 37 dB (see Chapter 4). This allows us to filter our data with little fear of side lobe leakage. In the case of a strong channel (SNR $\geq$ 37 dB) being found occupying an adjacent channel in a later stage of the detector, the in band noise levels should be increased accordingly. In our study of DTV stations surrounding Kingston we did not find a signal originating from outside the city that had a SNR above 37 dB.

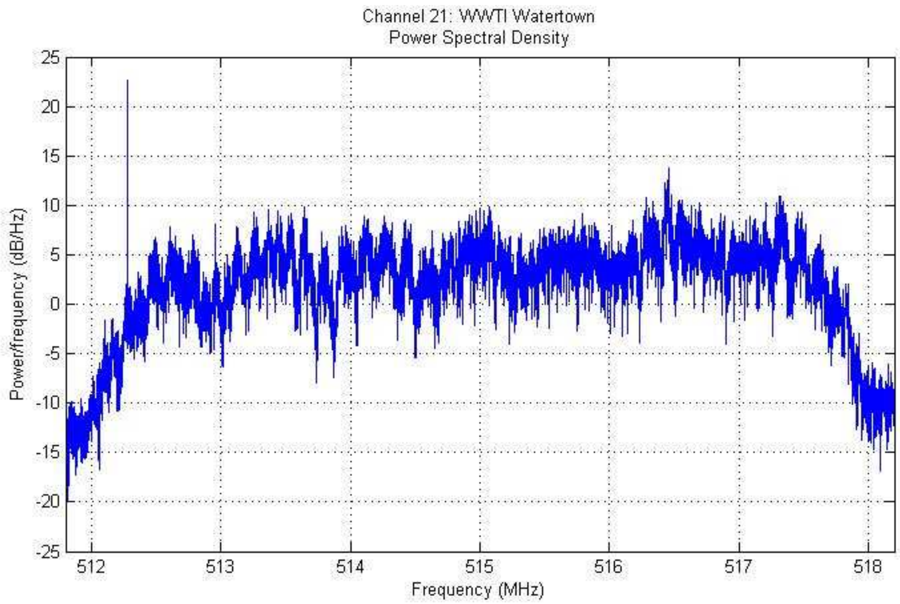

Figure 3.1: An observed realization of 8-VSB modulation.

3.2 Multitaper Spectral Estimate

After filtering the data and isolating the band of frequencies of interested, we need to transform the data from the time to the frequency domain for further analysis. This transformation is known as spectral estimation. Many methods of spectral estimation exist. The fundamental idea is to use the Fourier transform of the time series and to multiply by a sequence of weights (a *window* or *taper*). The choice of window is where most methods differ.

For most methods, depending on the choice of windows, there is a trade off between the bias and the variance of the estimate. This trade off is resolved when using the Multitaper Spectral Estimation method (MTM). The MTM was designed by David J. Thomson and allows for bias control without the corresponding increase in variance. The MTM is similar to many other methods in that it uses windowed

Fourier transforms of the time series to produce the spectral estimates. However, it differs through use of an *orthogonal family* of windows instead of a single choice. This orthogonal family consists of a group of discrete prolate spheroidal sequences (DPSS, or *Slepians*). A particularly attractive property of the Slepians is that the Fourier transformations of the windows have maximal possible energy within a given $(-W, W)$ frequency band.

The method is as follows: define a time-bandwidth product NW, with N the number of data points, and W a user-set bandwidth parameter. Given NW, compute $K \approx 2NW - 1$ Slepian sequences of length N, and use these as windows for K Fourier transformations X_k, called the *eigenspectra*. The initial naive spectral estimate is then formed as

$$\overline{S} = \frac{1}{K} \sum_{k=0}^{K-1} |X_k(f)|^2, \tag{3.1}$$

the average of the K eigenspectra. More advanced techniques are available, such as adaptively weighting the eigenspectra to provide superior bias control. Note that the MTM assumes local stationarity of the signal. As we are using parameters $NW = 4.0$ and $N = 524288$ (after zero-padding) we require stationarity only on a local band of 15.3μHz. As the coherence bandwidth of signals similar to TV were found to be a minimum of 20 KHz [6], our assumption of local stationarity is met.

We will use the MTM on the filtered data using $NW = 4.0$ and $K = 7$.

3.3 The Harmonic F test for known signals

After performing the MTM we use an important extension of multitaper theory also developed by David J. Thomson, the *harmonic F-test* (F-test). The purpose of the

F-test is to find line components in the spectrum that derive from periodic signals. We do this by attempting to determine the amplitude of a line component in the spectrum given by the our time series. The F-test can easily detect well defined signals in our series, like DTV pilot carriers and analog luma (visual) carriers. The F-test does not require signals to be of high power for detection. This makes the F-test desirable when attempting to find low powered signals. The F-test tests the null hypothesis that no line components exist at a given frequency. The statistic for the F-test follows a F (2,2K-2,P) distribution and is computed by:

$$F(f) = (K-1)\frac{|\hat{B}(f)|^2 \sum_{k=0}^{K-1} |U_k(0)|^2}{\sum_{k=0}^{K-1} |X_k(f) - \hat{B}(f)U_k(0)|^2}, \tag{3.2}$$

$$\hat{B}(f) = \frac{\sum_{k=0}^{K-1} U_k(0) X_k(f)}{\sum_{k=0}^{K-1} |U_k(0)|^2}, \tag{3.3}$$

$$X_k(f) = \sum_{t=0}^{N-1} v_t^{(k)}(t) e^{-2i\pi f t}, \tag{3.4}$$

$$U_k(-f) = \epsilon_k V_k(f) e^{-2i\pi f(1/2(N-1))}, \tag{3.5}$$

where X_k is the eigencoefficient, ϵ_k is equal to 1 for k even and i for k odd, the U_k are the Slepian-defined real eigentapers, and the V_k the more modern notation of the complex eigentapers. Note that the complex eigentapers are more convenient for actual computations, as they are the Fourier transforms of the time-domain Slepians. If the F statistic is above $F(2, 2k-2, p)$ we will reject the null hypothesis. For our data we use $F(2, 12, p)$.

To use this test for detection, we want to determine the presence of either the DTV pilot tone or analog luma carrier. The DTV pilot carrier is typically located 303.4 kHz above the base band. However, it can be offset by 12.7 kHz, 22.7 kHz or 32.7 kHz higher than its normal value depending on whether an analog station is in the channel below it and offset -10 kHz, 0 kHz or $+10$ kHz respectively. By checking for high F values in the band (300 kHz, 340 kHz) we can ensure we do not miss a pilot signal when present. Similarly, the luma carrier for an analog TV signal is located 1.25 MHz above the base band. If either of these areas have high F values (above 100) we are almost certain to have found a TV channel.

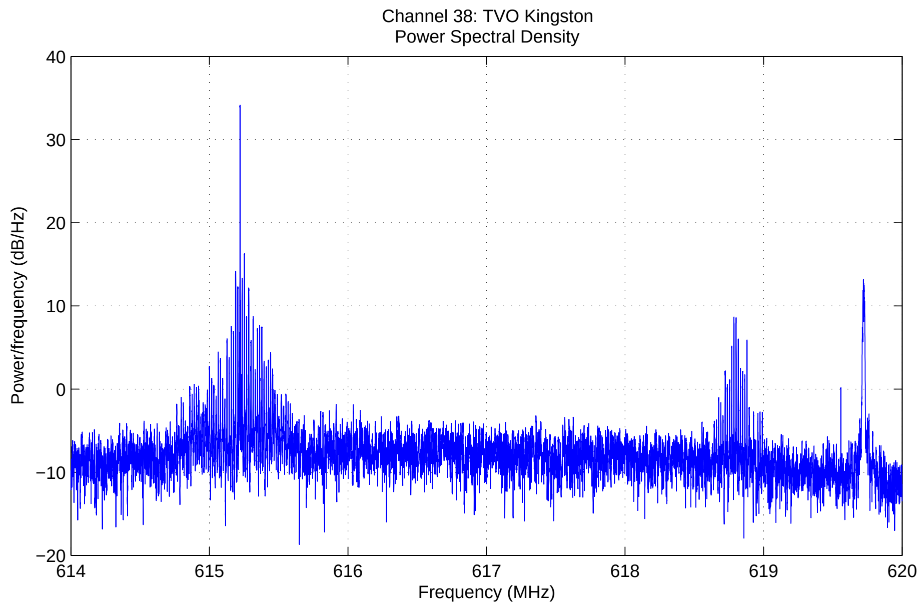

Figure 3.2: An observed realization of analog modulation.

To ensure that the signals are not unlicensed devices that are transmitting in these channels at the carrier frequencies, auxiliary classification must be performed. For digital signals with high SNR, evaluating the bandwidth of the signal will suffice in most cases. The bandwidth of the flat amplitude section (the head) of the DTV signals is 5.3 MHz starting at the pilot frequency. To calculate the bandwidth of a DTV signal, we determine where the guard bands are. Finding when the power of the signal begins to decrease determines the location of the guard bands. To account for randomness and minor modulation the data should averaged in 50 to 100 kHz blocks when trying to find the drop in power. Unfortunately, the amplitude modulaiton caused by the IF filter makes determination of the bandwidth of our signal becomes difficult. Since precision IF filters do exist and can be designed for the grey space detector in a commercial system, this issue will become moot.

In the case of low SNR signals, we cannot determine the bandwidth of our signal using the decrease in power in the guard bands since this is masked by the noise floor. Using more advanced spectral evaluation involving multiple measurements and coherences for classification is possible. The need for either an extra antenna or longer sensing time makes this method less desirable. The necessity of auxiliary classification can be avoided through regulation. Not allowing secondary signals to transmit at the pilot carrier frequency would remove any confusion concerning DTV classification. For auxiliary classification of an analog signal, performing an F-test to check the existence of the chroma carrier at 4.83 MHz above the base band would show the existence of an analog signal. Using these characteristics we can define our channel to contain either a digital TV signal or analog TV signal, or to be unoccupied.

3.4 Noise floor estimation

Once the channels are classified, using a robust estimate of location for the amplitude of the channels not occupied by TV signals will give an accurate estimate of the background noise for the occupied channels. In order to avoid possible unlicensed wireless systems in these bands, a robust estimate for the amplitude should be used. A 25% trimmed mean taken across all of the unoccupied channels will give an accurate result. In the event one wanted to devise a detector to test channel by channel individually, an estimate of the systematic noise and expected channel noise should be calculated before detection begins. When a channel is found unoccupied by a TV signal, the noise estimate should be updated to reflect the average amplitude found in that band combined with previous noise floor estimates. As more cognitive systems begin operating in the UHF band, further investigation should be done on detecting the secondary users in non-TV occupied bands to ensure the noise floor is accurate. After the regulations determine what communications systems can operate using cognitive radio, development of detectors for classifying the secondary signals present can begin. Using similar methods as found here for DTV, detection and avoidance of other secondary systems can be performed.

3.5 Signal to noise ratio

After updating the noise floor estimate, estimating the average amplitude for the DTV channels is needed to determine if the channel is fully occupied. Taking a trimmed mean of the head of the DTV channel will provide an accurate measure of the signal strength. Reporting both the signal strength and noise floor estimates and ensuring both are in the same units, such as dBm, we find the SNR to be,

$$dB_{SNR} = dB_{Signal} - dB_{Noise}. \tag{3.6}$$

To determine if the Signal is classified as grey space or black space depends on the choice of SNR limit. The Threshold of Visibility (TOV) for the DTV signal is 15dB [7]. To ensure that our transmissions are not affecting any receivers that are at or above the TOV, we need to set our limit for grey space distinction lower. Our limit depends on the power and distance to be serviced by our cognitive device. To ensure quality of service to the primary users, regulations on these limits should be made. For this academic exercise, an extra SNR drop of 10 dB is required. This is equivalent to 2.5 km when operating at 530 MHz and having a gain of 10 dB on both the transmitter and receiver, following Friss attenuation [1].

Using the model-specific SNR limit we can then state if the DTV channel is partially or completely occupied. We will avoid attempting to reuse analog signal. Analog will not be available commercially in Canada starting in August of 2011. With cognitive radio systems not available currently, developing a system to reuse analog now would be impractical. We will instead avoid analog channels for reuse until they cease operation.

3.6 Signal transmission placement

After finding a suitable channel for utilization, in-band transmission placement should be made. To avoid the need for auxiliary classification and confusion for other cognitive systems, not transmitting at the classification frequencies is necessary. While analog is still available, for minimal out-of-band interference, transmitting between the two analog carriers at around 3.03MHz above base band is best. As more cognitive users begin operating, breaking up TV bands into subsections for reuse would be useful. Developing regulations for signal type and placement would make the sensing of secondary systems transmitting currently in reusable spectrum easier. One method is to divide the channel into bands (+.4MHz, +1.2MHz), (+1.3MHz,+4.7MHz), (+4.9MHz,+5.7Mhz). This way we avoid the DTV pilot and the analog carriers. Dividing these bands into 100KHz sub-blocks for secondary allocation is a simple method for redistribution of narrow band signals.

Once analog signals are no longer available, avoiding only the pilot carrier frequency would be important. Setting up sub blocks within each channel that can be reused is still an efficient method for placement. To properly create a placement scheme for cognitive radio, knowledge of secondary signal types allowed would allow proper sizing of sub bands to allow for more modulation methods. The currently proposed scheme in the IEEE 802.22 standard is for OFDMA with 48 sub-channels across the 6 MHz band for white space [5]. The proposed system would allow for QPSK, 16-QAM and 64-QAM modulation schemes and 1/2, 2/3 and 3/4 encoding methods for transmission. This variety gives flexibility to the types of signals sent and data rates that can be achieved. The development of methods and regulations for spectral reallocation are an ongoing project for both FCC and IEEE. After the 802.22 standard is enacted, secondary signal detection and signal placement schemes will be possible.

Chapter 4

Results

4.1 Experiment overview

In an effort to test our grey space detector on real world data, we decided to sample the RF environment at Queen's university. In order to obtain accurate data we purchased a Channel master 3671 crossfire antenna designed for digital television systems. This is a directional antenna with a 10dBi gain [1]. For robust detection a omnidirectional antenna would be required. This would ensure that no channels are classified incorrectly.

To obtain the best results possible we placed our antenna on the roof of Jeffery Hall [26]. This will cause less attenuation from obstructions like buildings and vehicles. The downside of roof mounting the antenna is that the low powered interference that would have been attenuated at street level is now unobstructed. The roof placement will therefore produce higher power signals but have more incidences of low power

[1] dBi is the measure of forward gain of an antenna compared to a theoretical isotropic antenna, which uniformly distributes energy in all directions.

local interference.

To maximize the gain of our antenna we needed to align the antenna in the direction of the TV transmitter of interest. There are more than 25 TV transmitters within a 200 km radius of Kingston [26]. To ensure we observed DTV signals, we took data sets with our antenna aligned towards three US cities: Watertown, Rochester and Syracuse.

Figure 4.1: Kingston and surrounding areas, from [26].

After situating our antenna, we needed to design an IF filter to down phase our data for evaluation. David J. Thomson designed and built the IF filter we employed. The IF filter has a tunable carrier frequency for adjustable heterodyning. This allows us to easily change the frequency band of interest. The IF filter also employs an analog bandpass filter to remove most of the unwanted out of band interference. Our band of interest was 50 MHz, as this is the maximum bandwidth possible when sampling at 100 MHz for real data. We know the channels that broadcast in each city and set the carrier frequency to different values to ensure we observe these channels. Since this filter is not of a professional grade, amplitude attenuation does occur. We also incurred amplitude modulation due to a reflection in the coaxial cable. The reflection was caused by running both a spectrum analyzer and the analog to digital converter. The shape of the IF filter is shown here.

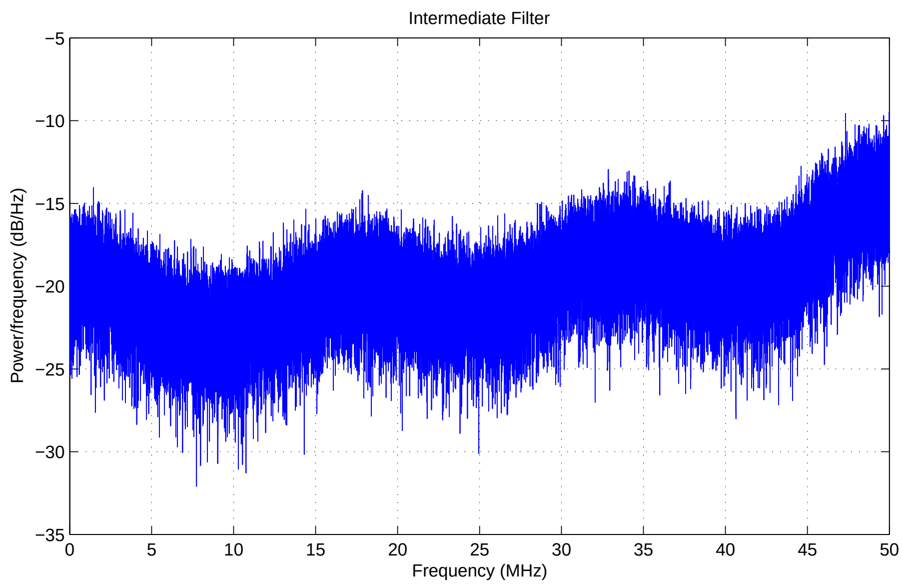

Figure 4.2: Attenuation from coaxial reflection and IF filter.

Once we have down phased our signals we run the signal into an analog-to-digital converter. We sampled the data at 100 MHz for 3 ms, giving us $300,000$ data points for each data set. We collected five data sets, testing different locations and frequencies. The data sets were collected within minutes of each other to ensure similar atmospheric and environmental conditions were observed for each data set. The weather was sunny with little clouds during our data collection. We chose these weather conditions because they have minimal atmospheric ducting and path loss.

The digital grey space detector was coded in Matlab. The detector determined the classification of all of the channels found within the 50 MHz simultaneously. The detector used the channels where no pilot or luma tones were found to estimate the background noise within the 50 MHz band. Depending on the frequency band of interest either 7 or 8 channels would be present. The detector had four possible classifications for each channel; no TV data, grey space, black space or analog signal.

4.2 Experimental Results

The detector was given all five data sets and produced the following results.

Frequency (MHz)	500	506	512	518	524	530	536	542
Classification	Grey	No TV	Black	No TV	No TV	Black	Black	No TV

Table 4.1: Data Set#1: Syracuse, 500-550 MHz.

Frequency (MHz)	500	506	512	518	524	530	536	542
Classification	No TV	No TV	Black	No TV	No TV	Grey	No TV	No TV

Table 4.2: Data Set#2: Watertown, 500-550 MHz.

Frequency (MHz)	602	608	614	620	626	632	638	644
Classification	No TV	No TV	analog	No TV	No TV	Black	No TV	No TV

Table 4.3: Data Set#3: Watertown, 600-650 MHz.

Frequency (MHz)	530	536	542	548	554	560	566
Classification	Grey	No TV	No TV	No TV	Analog	No TV	Analog

Table 4.4: Data Set#4: Rochester, 525-575 MHz.

Frequency (MHz)	626	632	638	644	650	656	662	668
Classification	No TV	Black	No TV	No TV	Grey	Grey	No TV	Grey

Table 4.5: Data Set#5: Rochester, 625-675 MHz.

For the first data set four DTV signals were detected. The signals were found at 500 MHz, 512 MHz, 530 MHz and 536 MHz. Examining the spectrum estimate, the four detected DVT signals have well defined pilot tones visible on the left edge of the band in the multitaper spectral estimate. The pilots also have large spikes on the F-test. All of the spikes are above the 99.996% p value level. This level corresponds to the appropriate p value for rejecting the null for our data set's size [39]. There are two small spikes on the right hand side of the MTM that are also visible on the F-test. These frequencies do not correspond to any relevant carrier locations and there is no amplitude modulation around them. This indicates they are either low powered unlicensed signals or some form of interference. Using a trimmed mean ensures that those spike do not affect our noise floor estimate.

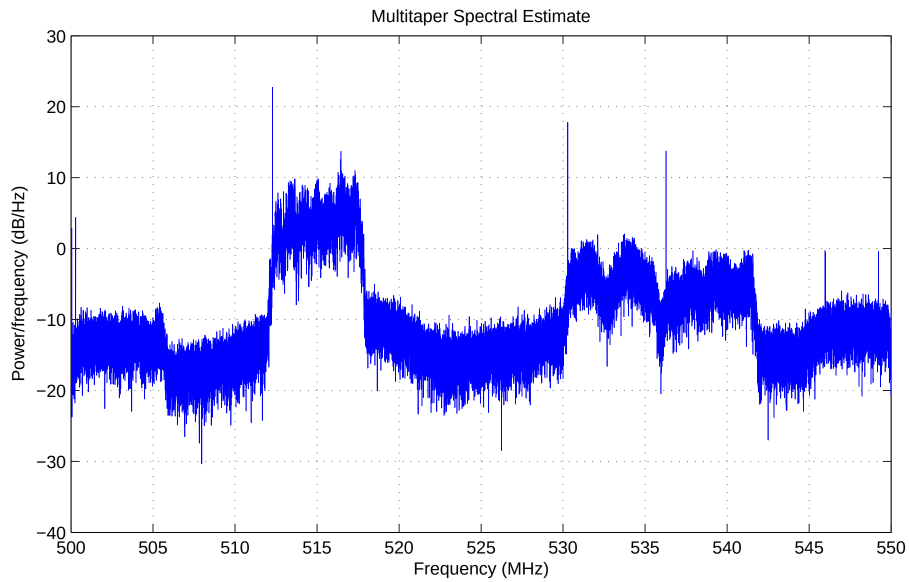

Figure 4.3: Data Set#1: Syracuse, 500-550 MHz.

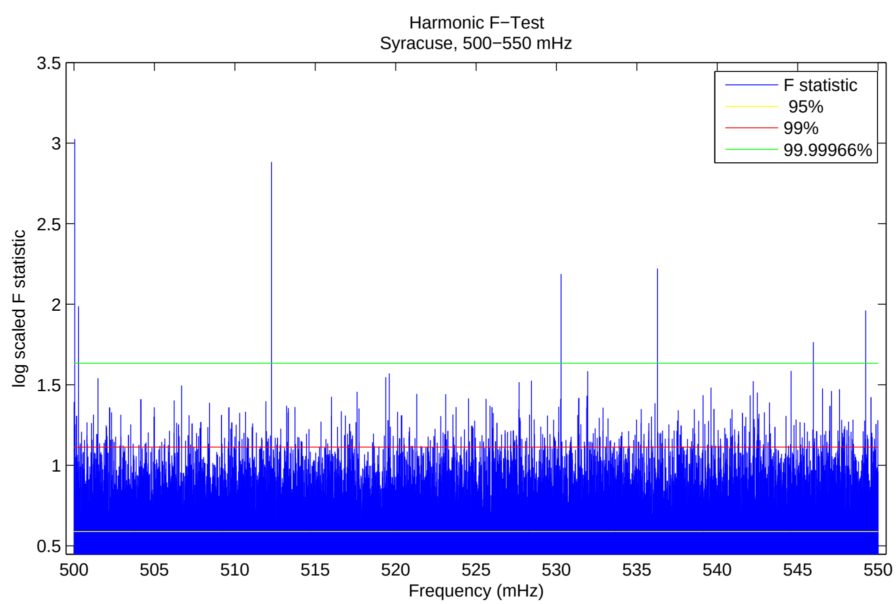

Figure 4.4: Data Set#1: Syracuse, 500-550 MHz.

There is visible amplitude attenuation across the 50 MHz band from the coaxial reflection and IF filter. The channel at 512 MHz is of greatest amplitude followed by the channels at 530 and 536 MHz with the lowest being found at 500 MHz. The detector classified only the channel at 500 MHz to be grey space. This indicates the detector is measuring the SNR correctly. Using a higher SNR cut off would cause the channels at 530 and 536 MHz to be classified as grey space as well. There are DTV transmitters in Syracuse operating at 500, 530 and 536 MHz [26]. There is also a DTV channel in Watertown transmitting at 512 MHz. The high power of the channel at 512 MHz is due to the relatively short distance between Kingston and Watertown of 75 km. Syracuse is farther away at 145 km and would be expected to have lower power accordingly. Choosing the Syracuse channel as grey space agrees with our interest in finding signals that are far away and less susceptible to interference.

The results of the second data set are very similar to those found in the first. The antenna is now facing Watertown and the only changes in the classification are those of the Syracuse channels. They are now found to be either grey space or non existent. Since we are using a directional antenna a change in power is to be expected. The change in results without changing the actual signals strength or location shows that using directional antennas is *not* an appropriate way to sense the RF environment for cognitive radio. Omni-directional antennas have less gain and when employed with the grey space detector, the SNR limit would need to be lowered.

In the third data set the detector finds two TV signals. The first is an analog signal at 614 MHz, which is the frequency at which Kingston's TVO channel operates. The three carriers of the analog signal are visible in the MTM estimate. The luma and chroma carriers also appear in the F-test. The DTV signal that is detected as black space at 614 MHz is being transmitted from Watertown. This again shows the relationship between power and distance to be valid for the grey space detector.

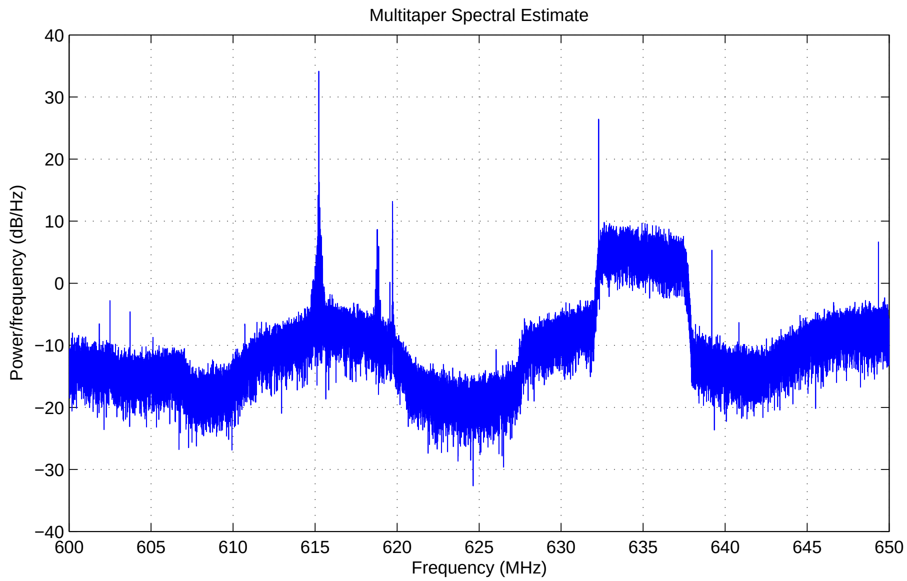

Figure 4.5: Data Set#3: Watertown, 600-650 MHz.

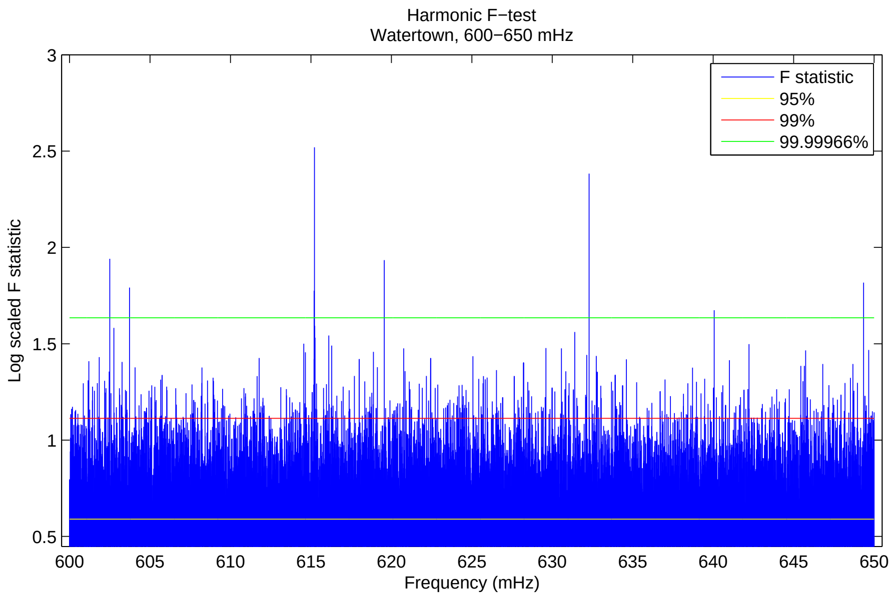

Figure 4.6: Data Set#3: Watertown, 600-650 MHz.

Evaluation of the 4^{th} and 5^{th} data sets continue to show proper detection and classification of TV signals. The signals classified as grey space are coming from Rochester and Syracuse, while the fully occupied signals are still coming from Watertown. In data set 5 we see that at low power the guard bands of the DTV signals are not distinguishable from the background noise. This shows how bandwidth evaluation is difficult and not an ideal candidate as an auxiliary characteristic in classification. The positioning of the pilot signals is enough for properly classifying a DTV signal in our data. When secondary users begin utilizing the TV bands without regulation pilot tones will not be enough.

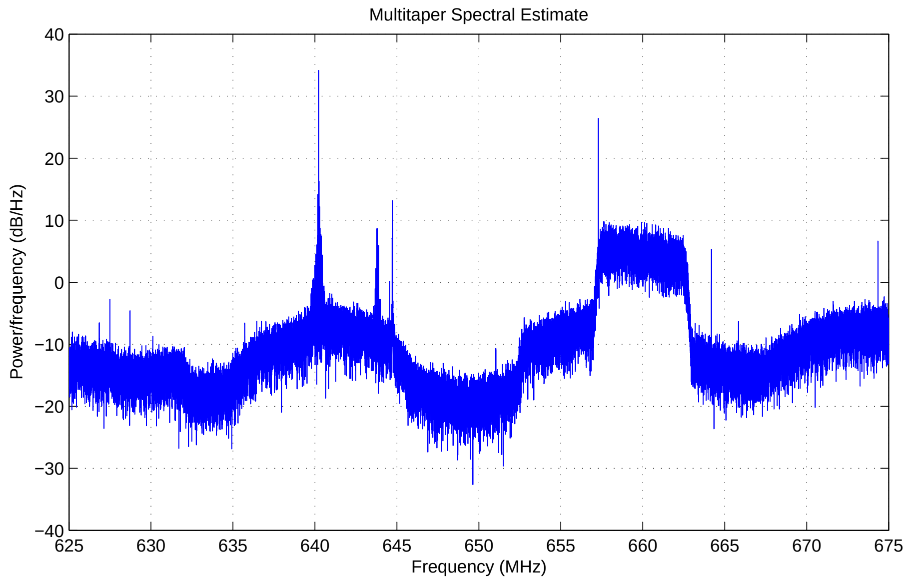

Figure 4.7: Data Set#5: Rochester, 625-675 MHz.

Chapter 5

Conclusion

5.1 Results

This study has shown that detection and classification of digital signals using the grey space detector designed is effective when dealing with real world data. The SNR limit may be lower than required for most situations but did prove effective at finding signals that are far from the antenna. The signals found to be grey space were all at least 145 km away from our antenna. From the examples of Grade B contours found in literature [21, 24], the Grade B contours ranged from $100-175$ km. Our limit is in the range of these finding and shows this method's applicability to the FCC proposal for spectral reuse.

The experiment showed the need for omni-directional antennas within cognitive radio systems. The use of antennas with directional gain can underestimate the SNR and cause the classification to be incorrect. Currently proposed classification characteristics of band width and guard band intervals [25] have been shown to be ineffective for low SNR signals. The use of these characteristics within the grey space detector is inappropriate. Without the use of multiple data sets to allow for coherence detection and other powerful statistics, classification relies on carrier frequency detection. Classification of digital and analog signals through the carrier frequencies proved effective. In the current RF environment only licensed systems operate within TV bands, making carrier tone mis-classification impossible. Once cognitive radio systems begin operation, without regulations on in-band transmission allocation, mis-classification is plausible.

To avoid mis-classification of carrier frequencies, regulations are required to keep cognitive radio systems from transmitting signals on or near the frequency locations for carrier tones. Currently this means avoiding the DTV pilot and the three analog tones. Once analog is no longer in use, the avoidance of only the DTV pilot will be needed. This requirement leaves much of the 6 MHz band available for utilization and flexible to many types of modulation.

The FCC's regulations and IEEE 802.22 standard are designed for implementation in rural areas. The data used in this experiment was taken from an urban center and shows the realistic extension of cognitive radio to urban environments. Urban environments experience more interference and receivers within them are more likely to incur hidden terminal problems. The fundamental avoidance of hidden terminals in grey space detection make it appealing for urban environments. When used in combination with a reliable white space detector, the grey space detector provides an increase in spectrum available for reuse. The grey space detector is a necessary tool in the effort to maximize spectrum use.

5.2 Future Work

From this study it is shown that grey space detection is practical. To ensure minimal interference by systems using grey space detection, rigorous testing of potential systems must be performed. Implementation of the system specifications to determine precise SNR limits is needed. Development of robust and accurate attenuation models for TV signals and secondary systems will further help define a precise SNR limit.

To minimize interference and help with classification, detailed regulation of grey space methods and uses need to performed. Regulation of modulation types, allocation schemes and pilot carrier avoidance are all needed to make grey space detection practical. Optimized code for grey space detection must be developed to allow the systems to operate as close to real time as possible. Also in an effort to get closer to real time transmission rates and minimal delays during sensing, the development of parallel operation of sensing and sending is needed. These improvements open the door for many more types of systems to use grey space detection, including live audio and video transmissions.

Appendix A

Coherent Detection

When the fraction of the total power to transmit a deterministic pilot tone is known, the detector can focus on that part alone. We assume all the potential signals are of the form,

$$Y(n) = \sqrt{\theta}X_p(n) + \sqrt{1-\theta}X_d(n). \tag{A.1}$$

where X_p is the known pilot tone, θ is the fraction of the total power allocated to it and X_d is a zero mean iid process.

A matched filter is used within the test statistic. The matched filter correlates a known signal with the observed signal. This is done in an effort to determine the presence of the known signal within the observed data. The matched filter asymptotically obtains the maximum SNR of all linear filters when in known noise [42]. If a signal $X = s + w$, where s is a known signal and w is additive noise, then the matched filter is,

$$Y_m(n) = \sum_{k=-\infty}^{\infty} A(n-k)X(k), \tag{A.2}$$

where,

$$A = \frac{1}{\sqrt{s^H E(vv^H)s}} E(vv^H)s. \tag{A.3}$$

The test statistic given by the matched filter is,

$$T(Y) = \frac{1}{\sqrt{N}} \sum_{n=1}^{N} Y(n)\hat{X}_p(n), \tag{A.4}$$

where, $\hat{X}_p$ is the unit vector in the direction of the pilot tone.

Coherent detection is only appropriate while the effects of multipath fading do not produce multiple independent realizations of the signal. Dependent realizations of a signal only exist for a finite period of time, known as a coherence block. When the realizations become independent, the signal will no longer be observed with correlated phase and amplitude [30]. This will cause the detector to make incorrect decisions by distorting the amplitude and phase of the pilot tone.

Appendix B

8 Vestigial Sideband Modulation

The current standard for digital television, the *Advanced Television Systems Committee* (ATSC), required terrestrial signals to use 8 vestigial sideband modulation (8VSB). 8VSB modulation is only one step in the transmission process for digital television (DTV) signals. The whole preparation process for the MPEG-II compressed DTV data is known as the 8-VSB exciter. To begin the exciter uses the MPEG-II sync byte to synchronize itself with the MPEG-II data block, This is done to ensure the data is properly processed within the exciter. After synchronization the data must be randomized. In order to create the flat portion of the signal within the 6 MHZ band, the data must follow a noise like pattern. When the data follows periodic patterns the RF energy of the signals will accumulate on certain frequencies. This creates sub-optimal frequency usage and would lower the rate of transmission. The MPEG-II bytes are changed using a known pseudo-random pattern that can be reversed by the DTV receivers.

Following randomization, the data is encoded with a forward error correction scheme. The 8VSB exciter uses Reed Solomon encoding. Reed Solomon encoding

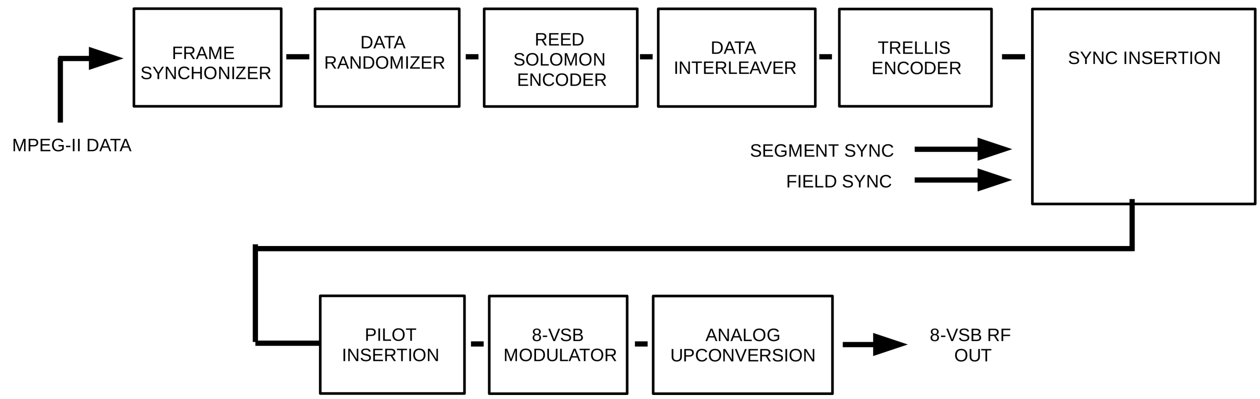

Figure B.1: Block diagram for the 8-VSB exciter, from [33].

evaluates the 187 bytes of the MPEG-II data block and creates a 20 bytes identification block that is attached to the end. The identification block is known as the Reed-Solomon parity bytes. These bytes are used by the receiver as an error check of the data packet. The receiver checks the parity bytes against the 187 byte block to see they match and if they do not match the receiver will attempt to find a similar 187 byte block that does match the parity bytes. If the receiver cannot find a 187 byte block that matches the parity bytes and has less than 10 bytes differing from the received 187 byte block the MPEG-II data block is discarded. This is done to avoid gross errors in transmission or errors in the parity bytes.

After the Reed-Solomon encoding, the data is interleaved. Interleaving takes the MPEG-II data packets from a 4.5msec time block, splits them up and combines them into new 207 byte blocks. This is done to ensure that if noise ruins one of these new blocks that only a fragment of the data from each original MPEG-II block is ruined. This makes it easier for the Reed-Solomon decoder to recover the original MPEG-II data blocks. The interleaving is done in a known pattern that is reversed in the receiver.

Another form of forward error correction is done next. Trellis encoding is performed to newly interleaved MPEG-II data packets. Unlike Reed-Solomon encoding, Trellis encoding treats the entire MPEG-II block as one element. Trellis encoding also does not work with static blocks with Reed-Solomon coding. It instead tracks the progress of a stream of bits with respect to time. The Trellis encoder first turns each bytes into four 2-bit words. The encoder examines the change from a 2-bit word to the next and then replaces the word with a 3-bit code describing the relative transition of the 2-bit word from the previous. These 3-bit words are then sent over the air as the 8 levels of the 8-VSB modulation. The Trellis decoder works similarly, tracking the transitions of the 2-bit words and using the transitions to reproduce the original data.

Prior to modulation several synchronization and pilot tones must be added to the signal. The 8-VSB exciter uses a pilot signal, segment sync and field sync. After the data has been Trellis encoded there are 828 8-level symbols. The segment sync is added to the front of the 828 symbols. It is comprised of 4 symbols that alternate between $+5$ and -5. These symbols are added to every 828 symbol block. The repetitive pattern of the sync allows the correlation circuits of the DTV receiver to determine the start of each symbol block. For the receivers to properly find the segment sync the against the background noise, the minimum SNR is 15 dB. If this is not attained the receiver cannot distinguish the sync segment from the pseudo-random symbol segments.

To avoid multipath issues, a field sync is added every 313 828-symbol blocks. A *field* is the name given to the combined 313 828-symbol blocks. The field sync is a 828-symbol block of alternating positive and negative values. The last 12 symbols are repeated from the previous block to allow the Trellis decoder to operate. The errors found in the alternating symbols are used in the receiver to adjust the ghost-canceling equalizer. The ghost-canceling equalizer is used to remove multipath interference issues. This allows the signal to still be decoded in poor reception areas and not increase the minimum SNR required from 15 dB.

Afterwards the pilot carrier is added. The pilot is created by having a small amplitude shift performed on the 8-level baseband signal. This will create a small residual carrier to appear at the base band of the spectrum. This carrier is used by the receiver as a marker on the spectrum for the DTV signal that does not change with respect to the data transmitted. This can be used to help fix phase shift issues.

The signal is then amplitude modulated onto an intermediate frequency (IF) carrier. The output is a double-sided signal with multiple side lobe copies of the center signal. Most of this spectrum is redundant and this signal is far too large for the chosen 6 MHz band. A Nyquist filter is used to remove the longest possible amount of redundant signal. The data rate of the DTV transmission after Trellis encoding is 32.28 Mbit/sec. This is an increase from the 19.39 Mbit/sec given by the orignal MPEG-II data block to the exciter. The symbol rate is $32\text{Mb}/3 = 10.72$ Million symbols/sec, since 3-bits are transmitted with 8-levels each. By the Nyquist theorem, the minimum frequency is $1/2 * 10.76\text{MHz} = 5.38\text{MHz}$. The signal is filtered down to 5.38MHz by removing the lower side lobes and narrow band filtering the upper sidebands. The signal is transmitted with 620 KHz of excess representing the guard bands.

The signal is then up-converted to its assigned channel using the regular method of heterodyning. The signal is run through a signal mixer that is attached to a local oscillator and then filtered to only broadcast the heterodyned signal at its assigned channel. The signal is now ready to be transmitted terrestrially. The DTV stations' antennas are directional in the horizontal plane. This removes unnecessary transmission upwards and increases the gain of the transmitter. The original MPEG-II is now available to receivers for decoding.

Appendix C

Multipath Fading

Multipath propagation or fading is an interfering effect that can occur to radio signals. Multipath propagation is caused by a radio signal taking two or more paths to the same receiver. Multipath has many environmental causes, including atmospheric ducting and reflection from the bodies of water, mountains or buildings. Multipath propagation can cause either constructive or destructive interference and can also cause phase shifting. The model for a signal that experiences mutlipath fading is,

$$X_m p(n) = \sum_{l=0}^{L-1} h_l(n) X(n-l), \tag{C.1}$$

where $h_l(n)$ are the multipath fading coefficients and L is the number of paths the signal experiences.

The coherence time for a multipath model is important when transmitting narrow band signals. The coherence time is the expected time over which the received signal is essentially invariant. When the received signal's autocorrelation function approaches zero the the signal is no longer coherent. The signal will no longer exhibit the same properties of phase and amplitude as at previous intervals. [30]. This time is essential to coherent detection.

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