ABSTRACT
Efficient and reliable spectrum sensing plays a critical role in cognitive radio networks. This paper proposes a cooperative sensing scheme that detects the existence of a common signal component in the signals received by multiple geographically distributed radios. The scheme assumes that signals received by different radios display strong coherence if they have a common source. Detection of this coherence in a wireless environment is studied, especially when the transmitted signal is distorted by multipath channels.

Index Terms—Cognitive radio, coherence detection, spectrum sensing.

1. INTRODUCTION
Cognitive radio has recently emerged as a useful technology to improve the efficiency of spectrum utilization [1]. Traditionally, the spectrum is assigned by the Federal Communications Commission (FCC) to specific users or applications, and each user can only utilize its pre-assigned bandwidth for communication. This discipline causes some bandwidth to be overcrowded while some other bandwidth may be seriously under-utilized. The concept of cognitive radio aims at providing a flexible way of spectrum management, permitting secondary users to temporally access spectrum that is not currently used by legacy users. In this regard, the FCC has taken a number of steps towards allowing low-power devices to operate in the broadcast TV bands that are not being used by TV channels. To promote this development, IEEE has established the IEEE 802.22 Working Group to develop a standard for a cognitive radio-based device in TV bands.

A key challenge in the development of the IEEE 802.22 standard is that a cognitive radio should be able to reliably detect the presence of TV signals in a fading environment. Otherwise, the radio may use the frequency band that is occupied by a TV channel, and cause serious interference to the TV receivers nearby. Many sensing or detection schemes have been recently reported in the IEEE 802.22 community. These schemes can be classified into two categories: single-user sensing and cooperative sensing. Due to the large variations in the received signal strength that are caused by path loss and fading, single-user sensing has proven to be unreliable, which consequently triggered the FCC to require geolocation-based methods for identifying unused frequency bands. The geolocation approach is suitable for registered TV bands; however, its cost and operational overhead prevent its wide use in the opportunistic access to occasional “white spaces” in the spectrum. Cooperative sensing relies on multiple radios to detect the presence of primary users, and provides a reliable solution for cognitive radio networks [2–5]. In this paper, we consider the problem of detecting the presence of a common signal component from the signals received by multiple geographically distributed sensors. If the signals received by these radios exhibit strong cross-correlation (coherence), it has a high probability that the spectrum is being occupied. This sensing technique minimizes the amount of prior information required to perform spectrum sensing.

Throughout this paper, we adopt the following definitions and notation. The network consists of $M$ cognitive radios that are monitoring the frequency band of interest (see Fig. 1). The two hypotheses corresponding to the signal-absent and signal-present events are defined as:

$H_0$: target signal is absent (i.e., spectrum is vacant);

$H_1$: target signal is present (i.e., spectrum is occupied).

The performance of detecting $H_0$ against $H_1$ is measured by the probability of false alarm and the probability of miss detection. False alarm refers to the error of accepting $H_1$ when $H_0$ is true, while miss detection refers to the error of accepting $H_0$ when $H_1$ is true.

2. COHERENCE DETECTION – LINE-OF-SIGHT CASE

For the case of line-of-sight propagation, the signals received by different radios are attenuated and delayed replicas of the target signal. We consider the following hypothesis testing...
where $X$ lay, and amplitude of the target signal, receiver. To solve this problem, we notice that for fixed $N$
the estimate of $H$
with proper alignment for given above equation as

$$H = \arg \min_{\tau} \sum_{m=1}^{N} (X_m[n] - A_m S[n - \tau_m])^2,$$

where $N$ is the number of signal samples acquired by each receiver. To solve this problem, we notice that for fixed $\tau_m$, the estimate of $A_m$ is given by

$$\hat{A}_m = \frac{\sum_{n=1}^{N} X_m[n]S[n - \tau_m]}{\sum_{n=1}^{N} (S[n - \tau_m])^2}.$$

The estimates of $\tau_m$ and $S[n]$ are thus given by solving

$$\langle \hat{\tau}_m, \hat{S}[n] \rangle = \arg \max_{\tau_m, S[n]} \sum_{m=1}^{M} \left( \frac{\sum_{n=1}^{N} X_m[n]S[n - \tau_m]}{\sum_{n=1}^{N} (S[n - \tau_m])^2} \right)^2.$$

With proper alignment for given $\tau_m$, we could rewrite the above equation as

$$\langle \hat{\tau}_m, \hat{S}[n] \rangle = \arg \max_{\tau_m, S[n]} \sum_{m=1}^{M} \left( \frac{\sum_{n=1}^{N} X_m[n + \tau_m]S[n]}{\sum_{n=1}^{N} (S[n])^2} \right)^2 = \arg \max_{\tau_m, S} \frac{s^T \hat{A}(\tau_1, \ldots, \tau_M) s}{s^T s},$$

where

$$s = [ S[1] \ S[2] \ \ldots \ S[N] ]^T,$$

$$\hat{A}(\tau_1, \ldots, \tau_M) = \sum_{m=1}^{M} \bar{x}_m(\tau_m) \bar{x}_m^T(\tau_m),$$

$$\bar{x}_m(\tau_m) = [ X_m[1 + \tau_m] \ \ldots \ X_m[N + \tau_m] ]^T.$$
Method 1 performs an exhaustive search and is computationally expensive, while Method 2 finds \( \tau_m \) for each radio one by one and might not be optimal. In Method 2, we first find the optimal \( \tau_2 \) to align the received signals from the 1st and 2nd radios. Then, we find the optimal \( \tau_3 \) to align the signal from the 3rd radio with the signals from the 1st and 2nd radios. We keep repeating this procedure until the \( M^{th} \) radio.

3. COHERENCE DETECTION — MULTIPATH CASE

For multipath propagation, the wireless channels are modeled as filters with finite impulse response (FIR). The signals received by each radio under \( \mathcal{H}_0 \) and \( \mathcal{H}_1 \) are represented by

\[
\begin{align*}
\mathcal{H}_0 : \{ X_m[n] = W_m[n], \quad m = 1, 2, \ldots, M, \} \\
\mathcal{H}_1 : \{ X_m[n] = (H_m[n] * S[n]) + W_m[n], \quad m = 1, 2, \ldots, M, \}
\end{align*}
\]

where \( * \) stands for the convolution operation and \( H_m[n], m = 1, 2, \ldots, M, \) are the impulse response functions of the propagation channels from the radio transmitter to the \( M \) receivers. Note that the line-of-sight case in Section 2 is a special case of the multipath case, where \( H_m[n] \) degenerates to the time-delay operators. By using matrix notation, we have

\[
\begin{align*}
\mathcal{H}_0 : \{ x_m = w_m, \quad m = 1, 2, \ldots, M, \} \\
\mathcal{H}_1 : \{ x_m = B_m s + w_m, \quad m = 1, 2, \ldots, M, \}
\end{align*}
\]

where

\[
\begin{align*}
x_m &= [X_m[1] \quad X_m[2] \ldots \quad X_m[N]]^T, \\
w_m &= [W_m[1] \quad W_m[2] \ldots \quad W_m[N]]^T, \\
s &= [S[1] \quad S[2] \ldots \quad S[N]]^T,
\end{align*}
\]

and \( B_m \) is defined at the top of next page and \( L \) is the length of the channel responses. The Log-Likelihood ratio (LLR) is

\[
\text{LLR} = -\frac{1}{2\sigma_W^2} \sum_{m=1}^{M} \|x_m - B_m s\|^2 + \frac{1}{2\sigma_W^2} \sum_{m=1}^{M} \|x_m\|^2.
\]

Let

\[
x = \begin{bmatrix} x_1 \\ x_2 \\ \vdots \\ x_M \end{bmatrix}, \quad B = \begin{bmatrix} B_1 \\ B_2 \\ \vdots \\ B_M \end{bmatrix}.
\]

The expression for LLR can be rewritten as

\[
\text{LLR} = -\frac{1}{2\sigma_W^2} \|x - Bs\|^2 + \frac{1}{2\sigma_W^2} \|x\|^2.
\]

Since \( B \) and \( s \) are unknown, their Maximum Likelihood estimate is given by

\[
\hat{(B, s)} = \arg \min_{B, s} \|x - Bs\|^2.
\]

We note that

(1) For fixed \( B \), the corresponding optimal estimate of \( s \) is given by

\[
\hat{s} = (B^T B)^{-1} B^T x.
\]

(2) For fixed \( s \), the corresponding optimal estimate of \( B_m \) can be found by

\[
\hat{h}_m = (S^T S)^{-1} S^T x_m, \quad m = 1, 2, \ldots, M,
\]

where

\[
\]

With the estimate \( \hat{B} \) of \( B \), the GLLR is given by

\[
\text{GLLR} = -\frac{1}{2\sigma_W^2} \|x - \hat{B}s\|^2 + \frac{1}{2\sigma_W^2} \|x\|^2
\]

\[
= \frac{1}{2\sigma_W^2} x^T \hat{B} (\hat{B}^T \hat{B})^{-1} \hat{B}^T x.
\]

The test is thus given by

Accept \( \mathcal{H}_1 \) if \( x^T \hat{B} (\hat{B}^T \hat{B})^{-1} \hat{B}^T x \geq \eta \),

Accept \( \mathcal{H}_0 \) if \( x^T \hat{B} (\hat{B}^T \hat{B})^{-1} \hat{B}^T x < \eta \),

for some predetermined \( \eta \). Based on (2) and (3), we have Algorithm 1 to find a suboptimal \( \hat{B} \) for the test.

4. SIMULATION RESULTS

The simulated network has \( M = 4 \) cognitive radios for spectrum sensing. The signal and noise are Gaussian distributed. The received signal variances \( \sigma_x^2 \) and noise variances \( \sigma_w^2 \) under \( \mathcal{H}_1 \) and \( \mathcal{H}_0 \) are listed in Table 1. The signal block length is \( N = 100 \). Fig. 2 shows the performance of the proposed detector with different threshold values for the line-of-sight propagation environment. Fig. 3 shows the performance of the proposed detector with different threshold values for the multipath propagation environment. The channel response is modeled as an FIR filter with length 4, and its taps are independently Rayleigh distributed with the total power normalized to be 1.
and signal variances are defined in Table 1.

\[ \mathbf{B}_m = \begin{bmatrix}
H_m[0] & 0 & 0 & \ldots & 0 & 0 & 0 & \ldots & 0 \\
H_m[1] & H_m[0] & 0 & \ldots & 0 & 0 & 0 & \ldots & 0 \\
H_m[2] & H_m[1] & H_m[0] & \ldots & 0 & 0 & 0 & \ldots & 0 \\
\vdots & \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & \ddots & \vdots \\
H_m[L-1] & H_m[L-2] & H_m[L-3] & \ldots & H_m[0] & 0 & 0 & \ldots & 0 \\
0 & H_m[L-1] & H_m[L-2] & \ldots & H_m[1] & H_m[0] & 0 & \ldots & 0 \\
\vdots & \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & \ddots & \vdots \\
0 & 0 & 0 & \ldots & H_m[L-1] & H_m[L-2] & \ldots & H_m[0] \end{bmatrix}_{N \times N} \]

(1)

Table 1. Simulated signal and noise variances.

<table>
<thead>
<tr>
<th>(m)</th>
<th>(\sigma_s^2)</th>
<th>(\sigma_n^2)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>-4 dB</td>
<td>0 dB</td>
</tr>
<tr>
<td>2</td>
<td>-5 dB</td>
<td>0 dB</td>
</tr>
<tr>
<td>3</td>
<td>-6 dB</td>
<td>0 dB</td>
</tr>
<tr>
<td>4</td>
<td>-7 dB</td>
<td>0 dB</td>
</tr>
</tbody>
</table>

Fig. 2. Achievable \(P_{FA}\) and \(P_{MISS}\) with the proposed algorithm for a line-of-sight propagation environment. The noise and signal variances are defined in Table 1.

Algorithm 1 Coherence Sensing in a Multipath Environment

0: Set \(i = 0\).
1: \(\text{repeat}\)
2: \(i = i + 1\).
3: Find the optimal \(\hat{s}_i\) for the previously obtained \(\hat{B}_{i-1}\) by
   \[ \hat{s}_i = (\hat{B}_{i-1}^T \hat{B}_{i-1})^{-1} \hat{B}_{i-1}^T \mathbf{B}_m \mathbf{x} \]
4: Find the optimal \(\hat{B}_{m,i}\) for the previously obtained \(\hat{s}_i\) by
   \[ \hat{h}_{m,i} = (\hat{S}_i^T \hat{s}_i)^{-1} \hat{s}_i^T \mathbf{B}_m \mathbf{x} \]
where \(\hat{B}_{m,i}\) is determined by \(\hat{h}_{m,i}\) according to (4), (1) and \(\hat{s}_i\) is determined by \(\hat{s}_i\) according to (5).
5: \(\text{until}\) there is no significant improvement in the objective function \(\|\mathbf{x} - \hat{B}_i \hat{s}_i\|^2\).
6: If \(\mathbf{x}^T (\hat{B}_i^T \hat{B}_i)^{-1} \hat{B}_i^T \mathbf{x} \geq \eta\), \(\mathcal{H}_1: \text{Target Signal is Present}\) is claimed; if \(\mathbf{x}^T (\hat{B}_i^T \hat{B}_i)^{-1} \hat{B}_i^T \mathbf{x} < \eta\), \(\mathcal{H}_0: \text{Target Signal is Absent}\) is claimed.

Fig. 3. Achievable \(P_{FA}\) and \(P_{MISS}\) with the proposed algorithm for a multipath propagation environment. The noise and signal variances are defined in Table 1.

5. REFERENCES


