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D4.2
Sensing algorithms for TVWS operations

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Abstract:
This deliverable D4.2 describes the sensing techniques that are used to identify primary users (PMSE and DVB-T) in the TV White Spaces as required for the first regulatory scenario in COGEU reference architecture. The analysis of the simulation results leads to the selection of a specific spectrum sensing technique for integration with the COGEU TVWS prototype transceiver.

Keyword list:
Sensing, broadcast, PMSE, DVB-T, simulations, algorithms;
Executive Summary

This deliverable D4.2 ‘Sensing algorithms for TVWS operations’ addresses the specification of sensing algorithms for reliable detection of PMSE (Programme Making and Special Events, e.g., wireless microphones devices) and DVB-T transmitters. After the evaluation of the simulation results, the most promising sensing algorithms will be further integrated into the COGEU transceiver (WP5) and the final COGEU demonstrator (WP7). The key conclusions of this deliverable are summarized here.

Sensing of PMSE devices

- In Europe there is no single PMSE system with standardized wave form. Therefore only blind detection techniques are investigated, i.e. without a priori information regarding the PMSE signal features.
- Considering a typical Wireless Microphone (WM) system with analogue FM modulation, the Energy Detector (ED) performs well in silent speaker conditions where the FM signal is a pure sinusoidal wave. However the performance degrades for soft and loud speaker modes because the energy of the FM signal is spread over the 200 kHz bandwidth. We observe that the performance of the ED degrades considerably in the presence of noise uncertainty, and that the degradation is dependent on the amplitude of the acoustic signal.
- Covariance (CAV) and Eigenvalue (MET-BCED) based methods require little information on the signal and have some immunity to synchronization error, fading and multipath, noise uncertainty, and unknown interference. They are particularly adequate when signals are highly correlated, as in the case of PMSE signals. The performance of such detectors was measured against the ED and has shown significant performance gains. These methods overcome the noise uncertainty problem and perform globally better than ED in both AWGN and Rayleigh faded channels. Moreover, the performance is maintained for all WM operation modes. This makes them a more suitable choice for the detection of WM signals, even if the computational complexity is increased.
- Simulation results showed that the Covariance Absolute Value detection algorithm (CAV) and the Maximum Eigenvalue to Trace detection algorithm (MET) are good candidates to be integrated in the COGEU TVWS transceiver. The proposed methods can sense a PMSE signal with SNR= -17 dB in a Rayleigh channel, considering 100 ms sensing time, 90% probability of detection and 10% probability of false alarm.
- The computation of exclusion areas around an incumbent PMSE system is important when booking channels for professional PMSE services using the COGEU geo-location database. Interference analysis based on CEPT-SEAMCAT tool show that for 1% probability of interference, the protection area around a PMSE receiver should have a radius above 10 km. A LTE UE transmitting at 23 dBm (max. power) in the same TV channel is consider as secondary user.
- A preliminary version of a testbed for performance analysis of PMSE sensing algorithms under real conditions in the field was developed using Labview™ and USRP2-SDR platform.

Sensing of DVB-T transmitters

- Since DVB-T has a set of standardized waveforms, the detection methods can rely on the signal features for a correct identification of such signals. The first version of the DVB-T sensing algorithm uses the data for the Cell ID which are provided by the TPS decoder. Due to the very rugged modulation scheme for the TPS (Transmission Parameter Signalling), the decoder provides the correct data even under very difficult reception conditions. The decoding of the Cell ID as implemented in the test receiver requires two complete OFDM frames, each consisting of 68 OFDM symbols. In the case of an 8K mode and a Guard Interval of 1/4, the acquisition time of the data alone is about 230 ms. The processing time for the consecutive steps adds up to 0.9 sec to 1.2 sec for the complete processing of the Cell ID information.
- The second version of the DVB-T sensing algorithm uses a stripped-down version of the TPS decoder. For the identification of a DVB-T signal, only the FFT size and the Guard Interval length are determined. This new algorithm can derive the required information from a much smaller sample, and the Matlab™ simulations indicate that a maximum of 30270 I/Q samples (i.e. approx. 2 % of the data required for the first sensing algorithm) provide the same degree of accuracy, leading to a short detection time of DVB-T signals.
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1 Introduction

This deliverable describes the sensing techniques that are used to identify primary users in the TV White Spaces as required for the first regulatory scenario in COGEU reference architecture (where both geo-location database access and spectrum sensing are required for the protection of incumbents, see D3.2 [1]). The incumbent systems considered are PMSE and DVB-T.

The sections on sensing of PMSE signals give an overview of the theoretical approach of sensing algorithms for PMSE devices and provides the results of Matlab™-based simulations. The different approaches to sensing from a literature review are introduced and analysed. They make use of different information available to the sensing probe. Blind detection works without any information on the source signal or the noise power. Another type of sensing makes use of information on the noise power without assumptions on the PMSE signal. The most sophisticated approach uses feature detection by applying the knowledge on the source signal and the noise power.

The analysis of the simulation results leads to the selection of a specific spectrum sensing technique for integration with the COGEU TVWS prototype transceiver.

A SEAMCAT-based study gives new results on the possible coexistence between PMSE systems and secondary users WSDs (White Space Device), by computing exclusion areas around PMSE receivers for co-channel operation. The focus is on the maximum power that a WSD could use in relation to its distance from a PMSE system.

The preliminary version of a testbed for performance analysis of PMSE sensing algorithms under real conditions in the field is described and gives an impression of its capabilities. The tool is resident on a PC and is based on the USRP2 hardware prototyping platform to which a GPS receiver is attached.

The section on sensing techniques for DVB-T signals includes a description of the main features of the DVB-T signal to illustrate how the different properties of the OFDM signal can be utilised to develop fast and reliable sensing algorithms.

Power level sensing by an energy detector estimates the signal power in the channel and compares that estimate to a threshold. It is introduced as a benchmark against which the results of other sensing techniques can be measured.

The current sensing algorithm for DVB-T, which is based on the decoding of the Cell ID, is described with the focus on its implementation on the test receiver platform developed by Rohde & Schwarz.

The main features of the improved sensing algorithm as currently under development are described and illustrated by simulation results. These simulations are Matlab™-based and form an integral part of the equipment that is to be implemented and integrated in the COGEU final demonstrator (WP7).

The deliverable is structured as follows:

In Chapter 2 an extensive evaluation of the spectrum sensing literature is presented and a set of candidate techniques is then evaluated using Matlab™.

In Chapter 3 proceeds with a coexistence study between PMSE and WSD using SEAMCAT. The focus of the calculations is on computing exclusion areas around PMSE receiver for different scenarios.

In Chapter 4 we present a tool for performance analysis of PMSE-sensing algorithms in real conditions. This tool is designed to measure the performance of different sensing algorithm for PMSE devices in the field.

In Chapter 5, the properties of a DVB-T signal, relevant for the current sensing algorithm and its improved version, are described. For the improved algorithm that is currently under development, simulation results are given to illustrate its capabilities.
2 Sensing techniques for PMSE detection

COGEU WSDs must not interfere with primary users, such as PMSE devices (e.g. narrow band signals from wireless microphones). As proposed in the reference architecture for COGEU in D3.2 [1], a spectrum sensing technique is required in the spectrum commons regime (unlicensed use of TVWS). Note that in the spectrum trading regime COGEU considers that PMSEs are moved to safe harbour channels, protected by the geo-location database, so PMSE sensing is not required. This deliverable assumes the spectrum commons regime.

In the next sections we present an overview of sensing algorithms suitable for PMSE detection, underlining the pros and cons of each technique. We then exclude a number of techniques and proceed to evaluate the most promising ones, by means of Matlab simulations, using metrics such as probability of detection, probability of false alarm and minimum SNR detected for a given observation time. The simple energy detector will be used as a benchmark. By analysing those simulations in view of our objectives, we then recommend a spectrum sensing technique for integration with the COGEU TVWS prototype transceiver (Task 5.2 - Integration of Sensing and Database Access Mechanisms).

2.1 Overview of sensing techniques suitable for wireless microphones detection

Sensing algorithms can be classified into three main categories, based on the amount of information used for sensing purpose:

I. Requires information on noise power only (noise dependent detection). These detection algorithms do not make any assumption on wireless microphones signal characteristics. Basically these techniques are detecting random signal in noise. They do not need prior information on the signal but on the other hand, very accurate information on noise statistics is necessary in order to obtain reliable detection performance.

II. Requires both source signal and noise power to be known (feature detection). These algorithms employ knowledge of structural and statistical properties of primary user signals in the decision-making.

III. Requires no information on source signal or noise power (blind detection). These detection algorithms rely on a statistical analysis, using covariance or eigenvalue matrix to identify the properties of a signal. These methods are independent of the noise power.

In the following subsections, we present an overview of several algorithms in each category, which can be used to sense Wireless Microphones (WM). We also outline the benefits and drawbacks of each approach.

2.1.1 Noise dependent detection

2.1.1.1 Energy Detector (ED)

Energy detector (ED) based approach, also known as radiometry or periodogram, is the most common way of spectrum sensing because of its low computational and implementation complexities [3]. In addition, it is more generic (as compared to methods given in this section) as receivers do not need any knowledge on the primary user’s signal. The signal is detected by comparing the output of the energy detector with a threshold, which depends on the noise floor. Some of the challenges with energy detector based sensing include selection of the threshold for detecting primary users and the inability to differentiate interference from primary users and noise.

Other energy detectors, using different schemes to detect WM, are described in several research papers, and have been simulated and experimentally tested. In [4], the ED algorithm exploits the non-linear model of the speech in a FM signal, with improved results as compared with the traditional ED. However, the implementation of such algorithm in a TVWS scenario relies on filter bank based technique to split the analysis of the wideband signal (8 MHz DVB-T channel) into sub bands, in order to increase the accuracy of the algorithm. More filter bank techniques with ED for wideband channels have been compared and simulated in IEEE 802.22 WRAN scenario [5].
Pros:
- Simple detector.
- Optimal for random signal in noise when noise variance is known.
- Detects all kinds of signals.
- Does not need prior information on primary signals.
- Robust to unknown multipath fading.

Cons:
- Highly susceptible to noise variance uncertainty.
- Cannot distinguish between different kinds of waveform.

2.1.1.2 Wavelets Transforms (WT)

In [6], this method is used for detecting local irregularities, or edges, in the Power Spectrum Density (PSD) of a wideband channel. Once the edges, which correspond to transitions from an occupied band to an empty band or vice versa, are detected, the powers within bands between two edges are estimated. Using this information and edge positions, the frequency spectrum can be characterized as occupied or empty in a binary fashion. The assumptions made in [6], however, need to be relaxed for building a practical sensing algorithm, by using sub-Nyquist sampling. Assuming that the signal spectrum is sparse, sub-Nyquist sampling is used to obtain a coarse spectrum knowledge in an efficient way. Multi-resolution spectrum sensing is achieved by changing the basis functions, by adjusting the wavelet's pulse width and carrier frequency. Hence, fast sensing is possible by focusing on the frequencies with active transmissions after an initial rough scanning.

Pros:
- Fast method

Cons:
- High sampling rate needed.
- Complexity

2.1.2 Feature detection

2.1.2.1 Matched-Filtering (MF)

This method is known as the optimum method for detection of primary users when the transmitted signal is known [7]. The main advantage of matched filtering is the short time to achieve a certain probability of false alarm or probability of miss detection as compared to other methods that are discussed in this section. In fact, the required number of samples grows as $O(1/SNR)$ for a target probability of false alarm at low SNRs for matched-filtering. However, matched-filtering requires cognitive radio to demodulate received signals. Hence, it requires perfect knowledge of the primary users signalling features such as bandwidth, operating frequency, modulation type and order, pulse shaping, and frame format, which makes the implementation complexity of sensing unit impractically complex. Another disadvantage of match filtering is large power consumption, as various receiver algorithms need to be executed for detection.

Pros:
- Fast method
- Good accuracy

Cons:
- Primary signal must be known and demodulated
- Large power consumption
- High sampling rate needed
- Complexity
- Susceptible to synchronization error

2.1.2.2 Cyclostationarity Feature Detection (CYFD)

This is a method for detecting primary user transmissions by exploiting the cyclostationarity features of the received signals [8]. Cyclostationary features are caused by the periodicity in the signal or in its statistics like mean and autocorrelation or they can be intentionally induced to assist spectrum sensing. Instead of power spectral density (PSD), cyclic correlation function is used for detecting signals present in a given spectrum. The cyclostationarity based detection algorithms can differentiate noise from primary users signals. This is a result of the fact that noise is wide-sense stationary (WSS) with no correlation while modulated signals are cyclostationary with spectral correlation due to the redundancy of signal periodicities. Furthermore, cyclostationarity can be used for distinguishing among different types of transmissions and primary users.
Pros:
- Distinguish between different types of signals
- Robust to noise uncertainty
- Can distinguish between different waveforms
- Can detect and classify primary users

Cons:
- High sampling rate needed
- Large number of samples needed.
- Affected by channel uncertainty
- Susceptible to synchronization error
- Lack of common standard in PMSE manufacturer providing the information that can be used in advance by feature detection

2.1.2.3 Waveform-Based Detection (WBD)

This method explores the fact that known patterns are usually utilized in wireless systems to assist synchronization or for other purposes. Such patterns include preambles, midambles, regularly transmitted pilot patterns, spreading sequences, etc. A preamble is a known sequence transmitted before each burst and a midamble is transmitted in the middle of a burst or slot. In the presence of a known pattern, sensing can be performed by correlating the received signal with a known copy of itself [9]. This method is only applicable to systems with known signal patterns, and it is termed as waveform-based sensing or coherent sensing. In [9], it is shown that waveform-based sensing outperforms energy detector based sensing in reliability and convergence time. Furthermore, it is shown that the performance of the sensing algorithm increases as the length of the known signal pattern increases.

Pros:
- Short measurement times
- Accuracy
- Low complexity

Cons:
- Susceptible to synchronization error
- Primary signal must be known
- Most of PMSE systems are analogue

2.1.3 Blind detection

2.1.3.1 Covariance / Eigenvalue Based Detection

Covariance based detection (CBD) [10-15] exploits the fact that the statistical covariance matrices of received signal and noise are usually different, thus the distinguishing property can be used to detect whether the primary user exists or not. The covariance-based detections directly use the elements of the covariance matrix to construct detection methods, which can reduce computational complexity, as compared to other blind algorithms. Eigenvalue based detection (EBD) are based on the analysis of eigenvalues of the covariance matrix [16-22].

Pros:
- Requires no information on channel source signal or noise power
- Immune to noise uncertainty
- No synchronization needed
- Same detection method for all signals (DVB-T, WM)
- Same threshold for all signals (the thresholds is independent on the signal and noise power)

Cons:
- Moderate Complexity.
- Sensitivity to the presence of unknown correlated interferences

2.1.4 Comparison of PMSE detection techniques

In this section, a number of methods for PMSE detection have been presented and their features summarized. Each of them has advantages and drawbacks. Table 1 summarize the most important and differentiating features of the presented methods for use in the case of the COGEU sensing module. The first column describes each feature. A letter ‘Y’ is placed in a cell if the respective feature is fulfilled for the corresponding algorithm. If not, the letter ‘N’ is written. This table does not describe all parameters of presented algorithms.

We have rejected some algorithms so that only the most promising ones are tested. First, we can reject the Cyclostacionary Feature Detector (CYFD) due to the primary signal characteristics we are going to sense: The majority of the PMSE systems are currently using FM modulation to transmit signals; When a WM is set to silent mode (no voice signal), only the frequency carrier is transmitted, and no features can be detected since they are nonexistent.
Table 1 – Comparison of the most important features for all investigated sensing algorithms

<table>
<thead>
<tr>
<th>Feature</th>
<th>Algorithm</th>
<th>ED</th>
<th>WT</th>
<th>MF</th>
<th>CYFD</th>
<th>WBD</th>
<th>CBD</th>
<th>EBD</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sensitive to noise uncertainty</td>
<td></td>
<td>N</td>
<td>Y</td>
<td>Y</td>
<td>N</td>
<td>Y</td>
<td>Y</td>
<td></td>
</tr>
<tr>
<td>Sensitive to interference uncertainty</td>
<td></td>
<td>Y</td>
<td>Y</td>
<td>N</td>
<td>N</td>
<td>Y</td>
<td></td>
<td>Y</td>
</tr>
<tr>
<td>Sensitive to synchronization errors</td>
<td></td>
<td>N</td>
<td>N</td>
<td>Y</td>
<td>Y</td>
<td>Y</td>
<td>N</td>
<td>N</td>
</tr>
<tr>
<td>Low computational complexity</td>
<td></td>
<td>N</td>
<td>N</td>
<td>Y</td>
<td>N</td>
<td>N</td>
<td>Y</td>
<td>Y</td>
</tr>
<tr>
<td>High accuracy</td>
<td></td>
<td>N</td>
<td>N</td>
<td>Y</td>
<td>N</td>
<td>Y</td>
<td>N</td>
<td></td>
</tr>
<tr>
<td>Independent of the primary signal</td>
<td></td>
<td>Y</td>
<td>Y</td>
<td>N</td>
<td>N</td>
<td>N</td>
<td></td>
<td>Y</td>
</tr>
<tr>
<td>Low implementation complexity</td>
<td></td>
<td>Y</td>
<td>N</td>
<td>N</td>
<td>N</td>
<td>Y</td>
<td></td>
<td>Y</td>
</tr>
<tr>
<td>Signal differentiation</td>
<td></td>
<td>N</td>
<td>N</td>
<td>Y</td>
<td>Y</td>
<td>N</td>
<td></td>
<td>N</td>
</tr>
<tr>
<td>Suitable for PMSE signals</td>
<td></td>
<td>Y</td>
<td>Y</td>
<td>Y</td>
<td>N</td>
<td>Y</td>
<td></td>
<td>Y</td>
</tr>
</tbody>
</table>

Waveform-Based Detection (WBD) is robust, because of the coherent processing that comes from using deterministic signal component. However, there should be a priori information about the primary users. Due to the lack of a common standard for PMSE devices, this method is not suitable.

The Matched Filtering (MF) was also rejected; just like the WBD method, primary signal features should be known in advance, and demodulated. This causes significant computational complexity for this method that cannot be accepted by the COGEU demonstrator.

The Wavelet Transform (WT) method was also rejected. The high sampling rate needed to process the signal increases the complexity and ultimately the power consumption, which is a limited resource in a battery-powered device.

The methods chosen for further investigations are:

- Covariance and eigenvalue based methods

  These methods overcome the noise uncertainty problem and can even perform better than energy detection when the signals to be detected are highly correlated, as in the case of PMSE signals in a TV channel. These methods also require little information on the signal or the channel and have some immunity to synchronization error, fading and multipath, noise uncertainty, and unknown interference.

- Energy detector

  Although traditional ED is unlikely to be used in practice, it does give a lower bound on sensing performance, since it is likely that other complex detectors will outperform the ED. The performance of more complex detectors should be measured against the ED to evaluate if the higher complexity and sophisticated techniques results in significant performance gains.
2.2 Description of the proposed sensing algorithms

This section describes the details of the sensing algorithms previously selected, their test statistics and the methodology to compute the detection threshold for each case.

2.2.1 Hypothesis test

As already described in COGEU D3.1 [23], the sensing problem can be generally formulated as a binary hypothesis testing problem,

\[ H_0: x[n] = u[n] \]
\[ H_1: x[n] = s[n] + u[n], \quad n = 1, 2, ..., N_s \]  

(2.1)

where \( H_0 \) and \( H_1 \) are the hypotheses expressing the absence and presence of the wireless microphone (WM), respectively, and \( N_s \) is the number of samples. The terms \( s[n] \) and \( u[n] \) are sampled versions of the WM signal \( s(t) \) and the noise \( u(t) \) present in the system, respectively. Both signals \( s(t) \) and \( u(t) \) are modelled and described in details in section 2.4.

There are two ways to design hypothesis tests: Bayesian and frequentist (or classical). In the Bayesian setup, the prior probability of each hypothesis occurring is assumed known. However, it is not reasonable to assign an a priori probability in this particular application, since we don’t know the probability of a PMSE signals being present in a particular place or time. In such cases we need a decision rule that does not depend on making assumptions about the a priori probability of each hypothesis. Here the Neyman-Pearson criterion offers an alternative to the Bayesian framework. The Neyman-Pearson criterion is stated in terms of certain probabilities associated with a particular hypothesis test, such as the probability of false alarm (\( P_{fa} \)) and the probability of detection (\( P_d \)) [39].

The WM signal is detected by comparing the output \( d \) of the sensing algorithm, with a decision level (threshold level - \( TH \)). Figure 1 illustrates the decision process.

![Figure 1 – Sensing model diagram](image)

Figure 1 – Sensing model diagram

Depending on the sensing technique, \( d \) is given by a test statistic that will be described for each algorithm implemented. Moreover, a detection threshold \( TH \) is determined based on the given probability of false alarm (\( P_{fa} \)) and is also dependent on the sensing algorithm. Depending on the test statistics, the threshold can be formulated from the formulations for probability of detection (\( P_d \)) or probability of false-alarm (\( P_{fa} \)) as follows,

\[ P_{fa} = P(TH > d|H_0) = \int_{d_{min}}^{\infty} f_0(t) \, dt \]  

(2.2)

\[ P_d = P(TH > d|H_1) = \int_{d}^{\infty} f_1(t) \, dt \]  

(2.3)

where \( f_0(t) \) and \( f_1(t) \) are the probability density functions (PDF) under the hypotheses \( H_0 \) and \( H_1 \), respectively. Thus, \( P_{fa} \) and \( P_d \) represent the two degrees of freedom in a binary hypothesis test, and do not involve a priori probabilities of the hypotheses. Depending on the detection algorithm, \( TH \) is computed analytically or by using heuristic methods. More details on the computation of \( TH \) are given in section 2.2.5.
2.2.2 Energy Detection

The energy detector (ED) is one of the simplest kinds of detectors and estimates the signal power in the channel and compares that estimate to a threshold [3]. A signal is assumed to be present if the test statistic is above the threshold and vice versa. The test statistic is computed as,

\[
d = \max \left( \frac{1}{N_s} \sum_{n=0}^{N_s-1} x[n] \cdot x^*[n] \right) = \max \left( \frac{1}{N_s} \sum_{n=0}^{N_s-1} |x[n]|^2 \right)
\]  

Moreover, wireless microphone signals manifest as a group of tones that can span 200 kHz range in frequency domain. By sufficiently averaging over time, the wireless microphone signals can stand out even at low signal levels. Due to this property, spectrum sensing can be also performed in the frequency domain, detecting the maximum peak of the estimated power spectral density (PSD) of the received signal [3] [24-26],

\[
d = \max \left( \frac{1}{N_s} \sum_{n=0}^{N_s-1} |\text{FFT}(x)|^2 \right)
\]  

Instead of the FFT in equation (2.5), other well-known spectrum estimation method, such as the Welch periodogram, can also be used [25].

2.2.3 Covariance-based detection

In this section, we present a selection of covariance-based detection algorithms, which were proposed in [10-11] [16] and will be briefly described here. The statistical covariance matrices of signal and noise are generally different. Thus this difference is used in the proposed methods to differentiate the signal component from background noise, i.e. this technique is based on measuring the whiteness or correlation level of the covariance matrix. In practice, there are only a limited number of received signal samples. Hence, the detection methods are based on the sample covariance matrix,

\[
R_x[N_s] = \begin{bmatrix}
\lambda[0] & \ldots & \lambda[L-1] \\
\vdots & \ddots & \vdots \\
\lambda[L-1] & \ldots & \lambda[0]
\end{bmatrix},
\]  

where

\[
\lambda[l] = \frac{1}{N_s} \sum_{m=0}^{N_s-1} x[m]x[m-l], \quad l = 0,1,...,L-1,
\]  

are the sample autocorrelations of the received signal \(x[n]\) and \(L\) is the smoothing factor. Test statistics are constructed directly from the entries of the sample covariance matrix and generally are given as

\[
d = F_1(r_{nm})/F_2(r_{nm})
\]  

where \(F_1\) and \(F_2\) are two functions and \(r_{nm}\) are the elements of the sample covariance matrix \(R_x\). There are many ways to choose the two functions. We present four special cases in the following:

1. Covariance Absolute Value detection (CAV) [10-11] [16]. The test statistic is,

\[
d_{CAV} = \sum_{n=1}^{L} \sum_{m=1}^{L} |r_{nm}| / \sum_{m=1}^{L} |r_{mm}|
\]  

2. Covariance Frobenius Norm detection (CFN) [11]. The test statistic is,

\[
d_{CFN} = \sum_{n=1}^{L} \sum_{m=1}^{L} |r_{nm}|^2 / \sum_{m=1}^{L} |r_{mm}|^2
\]
3. Maximum auto-correlation detection (MAC) [16]. The test statistic is,
\[ d_{MAC} = \max_{m \neq n} |r_{nm}|/\sum_{m=1}^{L} |r_{mm}| \]  
(2.11)
The presence of the signal is decided by comparing the ratio from eq. (2.9) to (2.11) with a threshold.

2.2.4 Eigenvalue-based detection

To use eigenvalue-based sensing algorithm, we need to compute the eigenvalues of the sample covariance matrix (2.6), \( \lambda_1 \geq \lambda_2 \geq \cdots \geq \lambda_L \geq \lambda_0 \). We found several methods in the literature to implement test statistics, based on the sample covariance matrix eigenvalues. We present four special cases:

1. Maximum to minimum eigenvalue detection (MME) [16-17] [27-28]. The test statistic is,
\[ d_{MME} = \frac{\lambda_1}{\lambda_L} \]  
(2.12)

2. Energy to minimum eigenvalue detection (EME) [17]. The test statistic is,
\[ d_{EME} = \frac{\sum_{n=0}^{N-1} |x[n]|^2}{\sum_{n=0}^{N-1} \lambda_L} \]  
(2.13)

3. Arithmetic to geometric mean (AGM) [16]. The test statistic is,
\[ d_{AGM} = \frac{1}{L} \sum_{m=1}^{L} \lambda_m / \left( \prod_{m=1}^{L} \lambda_m \right)^{1/L} \]  
(2.14)

4. Maximum Eigenvalue to Trace detection [16] [18], also called Blindly Combined Energy Detection (MET_BCED) [22]. The test statistic is,
\[ d_{MET_BCED} = \frac{\lambda_1}{\lambda_L} \]  
(2.15)

2.2.5 Threshold computation

2.2.5.1 Energy detector algorithm:

When noise \( u[n] \) is AWGN with variance \( \sigma_u^2 \), the threshold \( TH \) can be derived analytically:
\[ TH = \sigma_u^2 \left( 1 + \sqrt{2Q^{-1}(P_{fa})} \right) / \sqrt{N_s} \]  
(2.16)
where \( Q^{-1} \) is the inverse Q function.

If \( u[n] \) includes also interference from other signals, the statistic of the interference is unknown, and the threshold can be determined by a heuristic method. This was accomplished using the following methodology:

- Compute the peak PSD value of the sensed channel, when no primary signal is present (noise only)
- Repeat the measurement \( N \) times and create a histogram of the results.
- Compute the complementary cumulative density function (CCDF).
- Search in the CCDF, for the threshold value associated with the desired probability of false alarm.
Figure 2 – ED algorithm: a) Histogram and b) CCDF of the test statistic $d$ in the presence of noise only. The red dot in b) represents the threshold for a $P_{fa}$ of 10%.

As an example, the histogram for 1000 simulations of an AWGN noise signal, and associated CCDF, are given in Figure 2 a) and b), respectively.

The estimation of the noise PSD is crucial for the performance of ED algorithms as the determination of the threshold is strictly dependent on the variance of the noise, as expressed in (2.16). In certain situations, the exact value of this variance is difficult to obtain and thus the influence of the noise uncertainty on the ED of signals can significantly deteriorate the performance of the sensing algorithm [3]. We perform a similar investigation in section 2.3.3.4 and compare with other sensing algorithms.

2.2.5.2 Covariance based algorithm:

The threshold for methods CAV and CFN can be set using random matrix theory [16]. Here we present the expressions using theoretical derivation for two cases:

1. Covariance absolute value detection (CAV). The threshold is [10],

$$TH_{CAV} = \frac{1 + (L - 1) \sqrt{2} \pi}{1 - Q^{-1}(P_{fa}) \sqrt{2} N_s}$$

2. Covariance Frobenius norm detection (CFN). The threshold is [11],

$$TH_{CFN} = \frac{L + N_s + 1}{N_s + 2 - Q^{-1}(P_{fa}) \sqrt{8 N_s + 40 + \frac{48}{N_s}}}$$

The threshold is related to the number of samples $N_s$, the smoothing factor $L$ and the required $P_{fa}$, but not related to noise power, which gives this method robustness against noise uncertainty.

To the best of our knowledge, there is no analytic expression available to compute the threshold for the MAC method. Thus, in a similar process than ED, the computation of the threshold for MAC, and also for CAV and CFN, can be based on a heuristic method. Figure 3 shows the results for the same conditions as in the previous example. Note that the threshold is now a dimensionless value.
2.2.5.3 Eigenvalue based algorithm:

Analytical expressions to compute the threshold, for each eigenvalue based sensing method (except AGM), are given in the literature:

1. **Maximum to minimum eigenvalue detection (MME).** The threshold is,

\[
TH_{\text{MME}} = \frac{(\sqrt{N_s} + \sqrt{L})^2}{(\sqrt{N_s} - \sqrt{L})^3} \left( 1 + \frac{(\sqrt{N_s} + \sqrt{L})^2}{(N_sL)^{\frac{3}{2}}} F^{-1}_{11}(1 - P_{\text{fa}}) \right)
\] (2.19)

Where \( F^{-1}_{11} \) is the inverse cumulative distribution function of the Tracy-Widom distribution [17] of order 1.
2. Energy to minimum eigenvalue detection (EME). The threshold is [17],

\[
TH_{EME} = \frac{N_s}{(\sqrt{\frac{N_s}{c}} - \sqrt{E})^2} \left(1 + \sqrt{\frac{2}{N_s}} Q^{-1}(P_{fa})\right)
\] (2.20)

3. Blindly combined energy detection (BCED). The threshold is [21],

\[
TH_{MET_{BCED}} = \left(\frac{\sqrt{N_s} + \sqrt{L}}{N_s}\right)^2 \left(1 + \frac{(\sqrt{N_s} + \sqrt{L})^2}{(N_sL)^{\frac{1}{6}}} F_2^{-1}(1 - P_{fa})\right)
\] (2.21)

Where \(F_2^{-1}\) is the inverse cumulative distribution function of the Tracy-Widom distribution [17] of order 2.

The threshold is related to the same variables as in the covariance-based methods: It can be pre-computed based on \(N_s\), \(L\) and \(P_{fa}\), irrespective of signal and noise power, which gives this method immunity to noise variance uncertainty.

As we proceeded previously with ED and CBD methods, we present in Figure 4 the histogram and CCDF for the threshold computation of all EBD methods here studied (EME, MME, AGM and MET-BCED).
Figure 4 – a) Histogram and b) CCDF of the test statistic $d$ in the presence of noise only for EME, MME, MET-BCED and AGM algorithms. The red-cross indicates the threshold for $P_{fa} = 10\%$. 
2.3 Simulation Environment Description

2.3.1 PMSE system model

As primary users (PU), most PMSE devices use analogue frequency modulation (FM) to transmit information between a wireless microphone (WM) and a wireless receiver. FM continues to be the preferred choice; due the nature of the application (voice transmission) that imposes tight specifications such as continuous transmission and very low delay [23]. Although a new generation of wireless microphones are using digital modulation, they are currently used only in ISM bands (2.4 GHz) [29] and they are not investigated in this report.

A FM signal [30] is generally described by,

\[ s_{FM}(t) = A_c \cos \left[ 2\pi f_c t + 2\pi \int_0^t m(u) du + \theta \right], \tag{2.22} \]

where \( \theta \) is a random phase uniformly distributed on \((0,2\pi)\) and \( m(t) \) is the transmitted voice signal. It is zero-mean and its amplitude \( |m(t)| \leq 1 \). The parameters \( A_c \) and \( f_c \) are carrier amplitude and carrier frequency, respectively. The constant \( \Delta f \) is the frequency deviation of an FM modulator, representing the maximum departure of the instantaneous frequency of the FM signal from the carrier frequency \( f_c \).

For simulation purposes and for the remaining of this section, signal \( s_{FM}(t) \) is represented by its corresponding complex baseband signal \( s(t) \),

\[ S(t) = S_i(t) + j S_q(t), \tag{2.23} \]

where the In-phase and Quadrature terms are \( s_i(t) \) and \( s_q(t) \), respectively,

\[
\begin{align*}
S_i(t) &= \frac{1}{\sqrt{2}} \cos \left( 2\pi \Delta f \int_0^t m(u) du \right) \\
S_q(t) &= \frac{1}{\sqrt{2}} \sin \left( 2\pi \Delta f \int_0^t m(u) du \right) \tag{2.24}
\end{align*}
\]

Document [31] suggests three WM operating conditions to test sensing algorithms:

1. **Silent:**
   - The WM is silent, \( m(t) \) is a 32 kHz sinusoid signal and the FM deviation factor is \( \pm 5 \) kHz.

2. **Soft Speaker:**
   - MW is a soft speaker; In this situation, \( m(t) \) is modelled as a 3.9 kHz sinusoid signal with the FM deviation factor being \( \pm 15 \) kHz.

3. **Loud Speaker:**
   - MW is a loud speaker, \( m(t) \) is a 13.4 kHz sinusoid signal and FM deviation factor is \( \pm 32.6 \) kHz.

Amplitude of signal \( m(t) \) is \( A_m = 1 \) for all cases. According to Carlson’s Rule [32], the 90% (one-sided) bandwidth is given by,

\[ B_{90\%} = (1 + \beta) f_m \tag{2.25} \]

where \( \beta = A_m \Delta f / f_m \) is the modulation index. Table 2 presents a summary of the parameters of each operating conditions.

<table>
<thead>
<tr>
<th>Operating Mode</th>
<th>( A_m )</th>
<th>( f_m ) (kHz)</th>
<th>( \Delta f ) (kHz)</th>
<th>( \beta )</th>
<th>( B_{90%} ) (kHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Silent</td>
<td>1</td>
<td>32</td>
<td>5</td>
<td>0.16</td>
<td>37</td>
</tr>
<tr>
<td>Soft</td>
<td>1</td>
<td>3.9</td>
<td>15</td>
<td>3.85</td>
<td>19</td>
</tr>
<tr>
<td>Loud</td>
<td>1</td>
<td>13.4</td>
<td>32.6</td>
<td>2.43</td>
<td>46</td>
</tr>
</tbody>
</table>
Figure 5 shows the PSD of the various noise-free WM signal models. We can see from this figure that the power of the WM signal concentrates within a frequency band, which is close to 200 kHz for all operational modes. Moreover, there are apparent peaks contained in the PSDs of the various WM signal models that are more pronounced for silent mode.

![Figure 5](image.png)

Figure 5 – PSD of WM signals operating in a) Silent mode, b) Soft speaker and c) Loud speaker.

### 2.3.2 Characterization of the radio path environment

The two environmental conditions to be simulated are:

1. **Outdoor, Line-Of-Sight (LOS):**
   WM system is used in an outdoor environment where a LOS transmission path between transmitter and receiver exists. Therefore, channel is modelled as AWGN with zero mean and variance $\sigma^2_n$.

2. **Indoor, Rayleigh Faded:**
   WM system is used indoors. Because the distance between transmitter and receiver is short (tens of meters), a single-path Rayleigh fading channel is good enough to model the indoor channel. Therefore, a flat fading channel is used. Moreover, the maximum speed of the user is assumed to be $v = 0.6$ m/s (walking velocity). At this speed, and a maximum carrier frequency of 790 MHz (upper limit of the frequency band of interest for COGEU), the Doppler shift,

   $$f_{dop} = f_c \frac{v}{c}, \quad (2.26)$$

   is computed to be 1.58 Hz ($c$ is the velocity of light). Because the maximum Doppler shift is very small, the Doppler effect can be ignored. Hence, this channel is a single-path time-invariant (flat fading) channel [26]. We assume that mobile secondary users (UE) will provoke more interference to PMSE than fixed BS, since they can be located inside the same room as a wireless microphone receiver. Thus, BSs are not considered in this particular study.

![Outdoor and Indoor Scenarios](image.png)

Figure 6 – Scenarios geometries for PMSE sensing simulations.
2.3.3 Metrics to evaluate sensing algorithms

Several sensing schemes were introduced for PMSE detection in previous sections. Each scheme may have a different performance in a different scenario. It is therefore important to compare and choose the best scheme for a given scenario. In this section, we present important performance evaluation metrics.

2.3.3.1 SNR regime

In deliverable D3.1 (section 5.3.3) [23], COGEU sets the requirements for WM autonomous sensing. A minimum detection threshold of -126 dBm over a 200 kHz bandwidth is necessary to avoid causing interference to WM from TVWS devices. This value accounts for body loss and hidden terminal margin. Correspondingly, the required SNR at the TVWS sensing receiver can be calculated based on the receiver’s noise figure (NF). USRP’s NF is 8 dB [33] in the COGEU frequency band of interest (622-790 MHz). Considering that the thermal noise power spectral density (PSD) is -174 dBm/Hz, the TVWS receiver’s sensitivity over 200 kHz is,

\[-174 + 10 \log_{10}(200 \times 10^3) + 8 = -113 \text{ dBm}\]  \(2.27\)

Hence the TVWS receiver needs to detect signals with SNR at,

\[-126 + 113 = -16 \text{ dB}\]  \(2.28\)

The value -16 dB is used in this report as a reference value for minimum performance for sensing algorithms. Moreover, the SNR regime that we investigate is -32 dB to -10 dB. This is justified by the fact that reliable detection under these lower limits is difficult to obtain in the corresponding cases. For higher SNRs than the upper limits, the signals are detected with a high degree of certainty for almost all available sensing algorithms.

2.3.3.2 False alarm and detection probabilities

Minimum requirements are $P_d$ higher than 90% and $P_{fa}$ lower than 10% [23]. Performance of a specific detection technique is usually characterized by plotting:

1. The probability of detection curve as a function of SNR for a given false alarm probability (between 0% and 100%) and sensing time.
2. The Receiver Operation Characteristics (ROC) curves, namely a plot of $P_d$ vs. $P_{fa}$, given a sensing time.

2.3.3.3 Sensing time

The choice of the available sensing time is crucial for the performance of the system. It is important that vacant frequency bands are quickly detected so that they can be used efficiently. If sensing time is too long, the data transmission duration reduces, thereby reducing throughput of TVWS devices.

As stated in COGEU D3.1 [23], sensing requirements specify up to 2 second for detection time. This interval includes not only local sensing, but also other tasks. In our evaluation, we have chosen to run simulations with sensing time of 100 ms. However, the influence of the sensing time is analysed in detail for all sensing algorithms in section 2.4.4, from 8 ms to 500 ms.

For ED, sensing is made by successive averages of the spectrum. During the sensing time, a number $N_{avg}$ of FFT with size $FFT_L$ are computed. So, the sensing time $T_D$ is the product of the number of average with the computation time for each FFT (without buffer overlapping),

\[T_D = N_{avg} \cdot (FFT_L \cdot T_s)\]  \(2.29\)

where $T_s = 1/ f_s$ is the sampling time. A study is conducted to evaluate the optimum values of $N_{avg}$ and $FFT_L$ for a constant sensing time.

Other sensing algorithm that do not require spectral averages, but are dependent on a recorded sample of the signal (such as covariance or eigenvalue based methods), are evaluated with a number of samples $N_s$ corresponding to the total sensing time, in order to be compared with the ED algorithm.
2.3.3.4 Noise uncertainty
The noise level may change with time, which yields noise uncertainty. There are two types of noise uncertainty: receiver device noise uncertainty and environment noise uncertainty. The receiver device noise uncertainty comes from non-linearity of components and time-varying thermal noise in the components [11]. The environment noise uncertainty may be caused by transmissions of other users, either unintentionally or intentionally. Because of noise uncertainty, in practice, it is very difficult to obtain the accurate noise power.

Some of the effects of the noise uncertainty on ED techniques have been partially described in [3]. We conduct a similar study to investigate the robustness of the PMSE sensing algorithms proposed, against noise uncertainty between 0.5 dB and 2 dB.

2.3.4 Simulation chain to validate sensing algorithms
The software environment chosen for the simulation is MATLAB with SIMULINK. Different blocks (signal and channel generations; detection algorithm) are implemented so that they can be freely combined using predefined interfaces. This framework allows us to combine different WM signals with a variation of channel models and generate a signal received by the sensor. Suitable sensing algorithms can be applied and evaluated in terms of the metrics described in Section 2.3.3.

Figure 7 – Scenarios evaluation framework based on SIMULINK.

We have implemented a general FM signal generator. By choosing the parameters accordingly we are able to simulate all three PMSE operating conditions described section 2.3.1.

The channel models implemented are AWGN and Rayleigh channels. The Rayleigh channel corresponds to Non-line-of-sight (NLOS) situation encountered indoors. To model thermal noise variance and control SNR, an AWGN block is also included.

2.3.4.1 Simulation Scenarios
We evaluate the implemented sensing algorithms in the following scenarios:

- **Scenario I:**
  Primary signal: Wireless microphone, bandwidth: 37 kHz (silent mode)
  Channel model: AWGN
  Sensing time: 100 ms

- **Scenario II:**
  Primary signal: Wireless microphone, bandwidth: 37 kHz (silent mode)
  Channel model: Rayleigh, frequency-flat (“single path”), maximum Doppler shift: 1.58 Hz
  Sensing time: 100 ms

Other scenarios follow the same framework, with the difference being the WM operational mode, being changed from ‘Silent mode’ to:
- ‘Soft Speaker operation’ (bandwidth: 19 kHz), **scenarios III to IV**,
- ‘Loud Speaker operation’ (bandwidth: 46 kHz), **scenarios V to VI**;

In Table 3, we summarize all the scenarios that will be used to test the performance of sensing algorithms.
Table 3 – Scenarios for simulation

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Primary signal</th>
<th>Operation mode</th>
<th>Bandwidth</th>
<th>Channel model</th>
<th>Sensing time</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>Wireless Microphone</td>
<td>Silent</td>
<td>37 kHz</td>
<td>AWGN</td>
<td></td>
</tr>
<tr>
<td>II</td>
<td></td>
<td></td>
<td></td>
<td>Rayleigh “flat fading”</td>
<td>100 ms</td>
</tr>
<tr>
<td>III</td>
<td></td>
<td></td>
<td></td>
<td>AWGN</td>
<td></td>
</tr>
<tr>
<td>IV</td>
<td></td>
<td>Soft Speaker</td>
<td>19 kHz</td>
<td>Rayleigh “flat fading”</td>
<td></td>
</tr>
<tr>
<td>V</td>
<td></td>
<td>Loud Speaker</td>
<td>46 kHz</td>
<td>Rayleigh “flat fading”</td>
<td></td>
</tr>
</tbody>
</table>

2.3.4.2 Simulated Algorithms
The algorithms chosen in Section 2.1.4 were evaluated to confirm features described in the literature and clarify the knowledge on the performance of chosen solutions for suitability of use for WM detection in a COGEU environment. The algorithms to be evaluated are:

- Energy detector (ED): as described in section 2.2.2
- Covariance Based Detection algorithms (CAV, MAC and FCN): as described in section 2.2.3
- Eigenvalue Based Detection algorithms (EME, MME, AGM and MET-BCED): as described in section 2.2.4

2.4 Simulation results and discussion
In the following subsections we present a series of simulations, which evaluate the performance of the sensing techniques in scenarios where PMSE devices are used. The first set of simulations evaluates the probability of Detection vs SNR regimes, in the presence of white noise and flat fading channels. Next, we investigate the Receiver Operating Curves (ROC) using different sensing algorithms. These simulations enable us to compare the robustness of each method against different propagation conditions.

In the next subsection we investigate the dependence of sensing methods with time, and compare the necessary time interval to attain a minimum percentage of successful detections. In the following subsection we push the performance evaluation of sensing algorithms to stressful limits by introducing noise uncertainty.

2.4.1 Simulator parameters
Common simulation parameters are:

- **Detector**: Neyman Pearson detector with constraint on false alarm probability.
- **Number of realizations to simulate the probability of detection**: 500
- **Primary signal**: PMSE signal (silent mode, soft and loud speaker)
- **Channel**:
  - **AWGN**
  - **Rayleigh**: frequency-flat ("single path"), maximum Doppler shift: 1.58 Hz
- **Sensing time**: 100 ms
- **SNR Regime considered**: -32 dB to -10 dB
- **Sampling frequency**: 1 MHz

Next, we present assumptions and parameters specifically for individual detectors.

2.4.1.1 Energy Detector
Parameters:

- FFT length is 1024 samples.
- Spectral window overlap is 128 samples.
2.4.1.2 Covariance / eigenvalue based detectors
Parameters:
- Signal sample length ($N_s$) is 100000 samples. This value corresponds to the ratio between the sensing time and the sampling frequency.
- Smoothing factor ($L$) is 12.

2.4.2 Probability of Detection vs. SNR regime
This section presents the results of probability of detection vs. SNR curve for $P_{fa} = 10\%$.

2.4.2.1 AWGN channel
Scenario I:

![Figure 8 – Prob. detection vs. SNR regime, for silent mode.](image)

Scenario III:

![Figure 9 – Prob. detection vs. SNR regime, for soft speaker operating mode.](image)
Scenario V:

![Figure 10](image)

Figure 10 – Prob. detection vs. SNR regime, for loud speaker operating mode.

Observations from Figure 8 to Figure 10:

1. The ED performance is dependent on the WM operational mode. The ED has the highest probability of detection in the SNR regime of interest, when WM is in silent mode.
2. For all CBD and EBD algorithms, the performance is independent of the WM operational mode.
3. For soft speaker and loud speaker WM operations, BCED has the highest probability of detection in the SNR regime of interest.
4. For CBD methods, CAV is the algorithm with higher detection performance.
5. Apart from EME, all the other methods have similar detection performance.

2.4.2.2 Rayleigh channel

Scenario II:

![Figure 11](image)

Figure 11 – Prob. detection vs. SNR regime, for silent mode.
Scenario IV:

![Graph showing Probability of detection vs. SNR regime for soft speaker operating mode.]

Figure 12 – Prob. detection vs. SNR regime, for soft speaker operating mode.

Scenario VI:

![Graph showing Probability of detection vs. SNR regime for loud speaker operating mode.]

Figure 13 – Prob. detection vs. SNR regime, for loud speaker operating mode.

Observations from Figure 11 to Figure 13:

1. It is observed that the performance of all detectors degrades considerably under Rayleigh fading.

2. All the other observations made for Figure 8 to Figure 10 also apply here.

**General Observations from Pd vs. SNR Curves:**

1. Detection performance of all detectors degrades in the presence of Rayleigh fading.

2. ED has the highest probability of detection when a WM is in silent mode. From Figure 5, we can verify that in silent mode, the signal energy is only distributed in few discrete spectral components, as compared with the other two profiles where the energy is spread across many more spectral components. This results in much higher spectral components in those few frequency bins, and therefore ED is able to detect the signal even at low levels.
3. Eigenvalues and covariance-based methods performance is independent of the WM operational mode. They have generally better results than ED detectors when the WM is in use (soft or loud speaker).

4. MET-BCED for eigenvalues based methods and CAV for covariance-based methods are the best choice, since they have the highest probability of detection of all the blind detectors tested.

2.4.3 **Receiver Operating Curves (ROC)**

In this section, we present ROC curves \( (P_d \text{ vs. } P_{fa}) \) for SNR = -16 dB for all scenarios (I to VI).

2.4.3.1 **AWGN channel**

Scenario I:

![ROC Curve](image)

Figure 14 – Probability of detection vs. Probability of false alarm, for a WM in silent mode.

Scenario III:
Figure 15 – Probability of detection vs. Probability of false alarm, for soft speaker operating mode.

Scenario V:

Figure 16 – Probability of detection vs. Probability of false alarm, for loud speaker operating mode.

From Figure 14 to Figure 16, it can be observed that all detectors perform quite well for SNR = -16 dB for AWGN scenario, except for the EME detector. Comparative performance of the detectors is similar to $P_d$ vs. SNR curves.
2.4.3.2 Rayleigh channel

Scenario II:

Figure 17 – Probability of detection vs. Probability of false alarm, for a silent WM.

Scenario IV:

Figure 18 – Probability of detection vs. Probability of false alarm, for Soft speaker operating mode.
Scenario VI:

Figure 19 – Probability of detection vs. Probability of false alarm, for loud speaker operating mode.

From Figure 17 to Figure 19, it can be observed that the performance of all detectors deteriorates in presence of Rayleigh channel. Comparative performance of the detectors is similar to $P_d$ vs. SNR curves for Rayleigh channel.

2.4.4 Sensing Time

In this section, we investigate how much improvement can be obtained in the detector performance, when we increase the sensing time. We verify these dependencies for ED, CAV and MET-BCED methods.

2.4.4.1 ED algorithm

For the ED detector, we first analyse the benefit of a longer FFT compared to more averages, for a constant sensing period of 100 ms. These results are presented for AWGN and Rayleigh channel scenarios.

2.4.4.1.1 AWGN channel

Figure 20 – $P_d$ dependence on SNR regime and FFT length, silent mode WM.
From Figure 20 to Figure 22, we present the values of the probability of detection as a function of the FFT length (between 128 and 4096 samples) and SNR regime (from -32 dB to -10 dB), for silent mode, soft speaker and Loud speaker WM. It can be observed that ED performs better with long FFTs and less averages, compared with short FFTs and more averages. These results are justified by the fact that the whole 8 MHz DVB-T channel is sensed, and the small bandwidth representative of WM signal (200 kHz which is 40 times lower) is better detected is we use FFTs with high spectral resolution.
This behaviour is more evident for WM in silent mode, where the signal is composed by a few spectral lines, as seen in Figure 5a. We present the best and the worst combination between FFT length and Number of averages, in Figure 23, for all WM operational modes.

From the previous conclusions, we choose a set of parameters for ED and conducted a simulation to verify the influence of the sensing time in the performance of the detector. Sensing time is varied between 8 ms and 600 ms. With a FFT length of 1024 samples, we compute the necessary SNR regime to reach the minimum requirement of 90% probability of detection, constrained to a $P_{fa}$ of 10%.

From the results depicted in Figure 24, we can see that the requirements are different, depending on the WM operational mode. Even if ED method has good performances at very low SNR regimes and short sensing times when MW is silent, the performance degrades when the WM is effectively used (soft and loud speakers). Moreover, for a SNR regime of -16 dB (set previously in 2.3.3.1 as a minimum requirement), ED should sense the signal at least 16 ms to detect a Wireless microphone in all possible modes of operation.
2.4.4.1.2 Rayleigh channel

Figure 25 – $P_d$ dependence on SNR regime and FFT length, silent mode WM.

Figure 26 – $P_d$ dependence on SNR regime and FFT length, soft speaker operation.
The general observations from Figure 25 to Figure 27 for Rayleigh Channel are similar to those in case of AWGN Channel. However, if we compare the curves from Figure 20 to Figure 22, the performance of ED method is better for AWGN as compared to Rayleigh channel. The same observation is made to Figure 28, but now the difference between the worst case and the best case are less pronounced. From Figure 29, we verify degradation in the performance of ED method. Using the same example as in AWGN scenario (with a FFT length of 1024 samples), ED sensing time should be **256 ms** (sixteen fold increase) to attain the same -16 dB SNR.
2.4.4.2 Covariance and eigenvalue based algorithms

The results from the previous subsection, dealing with ED method, are here compared with CAV and MET-BCED methods. These two sensing methods used a single data sequence, with a number of samples ranging from 8000 to 500000 and a sampling rate of 1 MHz, in order to match the same sensing time used for ED, i.e., from 8 ms to 500 ms.

2.4.4.2.1 AWGN channel

![SNR regime vs. Sensing time, for P_d of 90%: a) CAV and b) MET-BCED models](image)

Observations from Figure 30:
- Sensing performance improves for both detectors with sensing time.
- WM s in silent mode and soft speaker operation are sensed with less time to reach a $P_d$ of 90%, than a WM in loud speaker mode. This is mainly due to the energy spread across a wider bandwidth in loud speaker mode, thus decreasing the energy of each harmonics of the FM signal.
- CAV should sense the signal at least 64 ms to assure a SNR regime of -16 dB for all WM operation modes.
- MET-BCED has a better performance, and should sense the signal during 32 ms to assure a SNR regime of -16 dB for all WM operation modes.
2.4.4.2.2 Rayleigh channel

Observations from Figure 31:
- Detection performance of both detectors degrades in the presence of Rayleigh fading.
- CAV should sense at least 160 ms (fourfold increase compared to AWGN channel) to assure a SNR regime of -16 dB for all WM.
- MET-BCED with sensing time 100 ms (fivefold increase compared to AWGN channel) assure the same -16 dB SNR regime for all operational modes.
- All other observations are similar to those made for Figure 30.

Table 4 summarizes the comparative results of the sensing time needed for each method.

<table>
<thead>
<tr>
<th>Channel</th>
<th>Sensing Technique</th>
<th>AWGN</th>
<th>Rayleigh</th>
</tr>
</thead>
<tbody>
<tr>
<td>ED</td>
<td></td>
<td>16 ms</td>
<td>256 ms</td>
</tr>
<tr>
<td>CAV</td>
<td></td>
<td>64 ms</td>
<td>160 ms</td>
</tr>
<tr>
<td>MET-BCED</td>
<td></td>
<td>32 ms</td>
<td>100 ms</td>
</tr>
</tbody>
</table>

2.4.5 Noise uncertainty

We described in Section 2.3.3.4 the sources of noise uncertainty. To simulate the influence of noise uncertainty on the sensing performance, we added an uniform distribution to the noise variance $\sigma^2_u$, in an interval (-B; B), where B is equal to 0.5 dB, 1 dB and 2 dB.

For blind algorithms that do not calculate the threshold based on the noise variance, this uncertainty should not influence the performance. So, we investigate and compare the performance and robustness for CAV, MET-BCED and ED detectors against noise uncertainty. The curves for ED with no noise uncertainty are included for comparison.
2.4.5.1 AWGN channel

Scenario I:

Figure 32 – Prob. detection vs. SNR regime

Scenario III:

Figure 33 – Prob. detection vs. SNR regime, for soft speaker operating outdoor.
Scenario V:

From Figure 32 to Figure 34, it can be observed that the performance of the ED degrades considerably in the presence of noise uncertainty, except when the WM is in silent mode. However, CAV and MET-BCED methods maintain their performance for all WM operation modes, and make them a more suitable choice for the detection purposes in face of noise level uncertainty, even if complexity is increased.

2.4.5.2 Rayleigh channel

Scenario II:
Scenario IV:

![Figure 36](image_url)

Figure 36 – Prob. detection vs. SNR regime, for soft speaker operating indoor.

Scenario VI:

![Figure 37](image_url)

Figure 37 – Prob. detection vs. SNR regime, for loud speaker operating indoor.

From Figure 35 to Figure 37, it can be observed that all detectors performance deteriorates in presence of Rayleigh channel with noise uncertainty, for every WM operational modes. The general observations for Rayleigh Channel are similar to that in case of AWGN Channel from Figure 32 to Figure 34.
2.5 Summary of the simulation results

In the previous section, three sensing algorithms were simulated for PMSE detection. All algorithms were investigated in similar use scenarios, suitable for the COGEU transceiver. The previously defined sensing metrics from 2.3.3 were used for this comparison.

Table 5 shows the performance of different sensing techniques for a probability of false alarm of 10% and probability of detection of at least 90%, for a sensing time of 100 ms.

<table>
<thead>
<tr>
<th>Sensing Technique</th>
<th>Channel</th>
<th>Silent mode</th>
<th>Soft speaker</th>
<th>Loud Speaker</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>AWGN</td>
<td>Rayleigh</td>
<td>AWGN</td>
<td>Rayleigh</td>
</tr>
<tr>
<td>ED</td>
<td>-27 dB</td>
<td>-21 dB</td>
<td>-21 dB</td>
<td>-13 dB</td>
</tr>
<tr>
<td>CAV</td>
<td>-22 dB</td>
<td>-16 dB</td>
<td>-22 dB</td>
<td>-16 dB</td>
</tr>
<tr>
<td>MET-BCED</td>
<td>-23 dB</td>
<td>-17 dB</td>
<td>-23 dB</td>
<td>-17 dB</td>
</tr>
</tbody>
</table>

Based on the results from Table 3 and Table 4, we select CAV and MET-BCED as algorithms to be implemented in the COGEU TVWS transceiver in task T5.2, reported in D5.3.
3 Coexistence study between PMSE and WSD

SEAMCAT is a professional software tool for evaluation of interferences between radio systems. It was used by COGEU in Task 4.1 (TV White Spaces characterization) to study the coexistence between primary users of the UHF spectrum (DVB-T and PMSE), and secondary users based on LTE technology. The results are available in COGEU deliverable D4.1 [2]. This section is a follow-up of the previous research, now focused on PMSE coexistence with TV white Spaces devices: The objective is to compute exclusion areas around PMSE receivers to avoid interference, in different scenarios.

![Exclusion Area illustration around PMSE devices.](image)

It was shown in D4.1 [2] that an LTE UE (secondary transmitter) could use up to 2 dBm power at 300 m from a PMSE receiver, in a co-channel scenario. Due to the lack of a valid SEAMCAT propagation model for low height transmitter antennas (1.5 m) and for distances above 300 m (maximum distance for Extended Hata SRD model implemented in SEAMCAT), we didn't complete the study for longer distances and higher LTE UE's power. This chapter extends D4.1 exclusion area computation, we start presenting an adaptation of the Two-Ray ground reflection model to overcome the lack of a suitable SEAMCAT model. The chapter continues with the description of the coexistence scenario geometries and system parameters. Finally, we present the results and their implications in COGEU.

3.1 Simulation scenarios

We consider three deployment geometries, as depicted in Figure 39:

- **Scenario 1**: Primary and secondary user outdoors;
- **Scenario 2**: Primary user outdoor and secondary user inside a building;
- **Scenario 3**: Primary user and secondary user inside different buildings.
This study considers only LTE UE transmitters as a source of interference, due to their mobility and possible proximity with PMSE systems.

### 3.1.1 Propagation model

A two-ray ground reflection model takes into account the effect of ground reflection, and the antenna heights above it. This model, although simplistic, can be very well suited for analysis involving line-of-sight scenarios. Considering flat earth surface and angles of incidence with the ground close to grazing, then the reflection coefficient is close to -1 and the path-loss for an outdoor scenario is given by,

\[
L_{\text{outdoor-outdoor}}^{2\text{RayGround}} = 10 \log \left( 2 \left( \frac{\lambda}{4\pi d} \right)^2 \left[ 1 - \cos \left( \frac{4\pi h_t h_r}{\lambda} \right) \right] \right)
\]  

(3.1)

where \( h_t \) and \( h_r \) are the heights above ground of the transmitter and receiver antennas, respectively, \( \lambda \) is the signal wavelength and \( d \) is the total path length. An approximation of this model takes free space path loss of 20 dB/decade up to a cross-over distance \( d_c \), and 40 dB/decade thereafter. The cross-over distance is given by,

\[
d_c = \frac{4\pi h_t h_r}{\lambda}
\]

(3.2)

Both models are depicted in Figure 40.
The variation in path loss is achieved by applying a lognormal distribution, using a random variable $\sigma$ to model shadowing and location variability,

$$\text{Pathloss}_{\text{dB}} = L_{\text{outdoor-outdoor}}^{2\text{RayGround}} + \sigma_{\text{outdoor-outdoor}}^{2\text{RayGround}}$$ \hspace{1cm} (3.3)

### 3.1.1.1 Indoor-outdoor propagation

The use of the two-ray ground model for indoor-outdoor propagation introduces the following terms in the path-loss:

$$L_{\text{indoor-outdoor}}^{2\text{RayGround}} = L_{\text{outdoor-outdoor}}^{2\text{RayGround}} + L_{\text{we}}$$ \hspace{1cm} (3.4)

where $L_{\text{we}}$ is the attenuation due to external walls.

Uncertainty on materials and relative location in the building increases the standard deviation of the lognormal distribution:

$$\sigma_{\text{indoor-outdoor}}^{2\text{RayGround}} = \sqrt{\left(\sigma_{\text{outdoor-outdoor}}^{2\text{RayGround}}\right)^2 + (\sigma_{\text{add}})^2}$$ \hspace{1cm} (3.5)

### 3.1.1.2 Indoor-indoor propagation

When LTE UE transmitter and PMSE receiver are located in different buildings, the calculation is similar to the indoor-outdoor propagation mode but with doubled additional values:

$$L_{\text{indoor-indoor}}^{2\text{RayGround}} = L_{\text{outdoor-outdoor}}^{2\text{RayGround}} + 2 \times L_{\text{we}}$$ \hspace{1cm} (3.6)

where $L_{\text{we}}$ is the attenuation due to external walls.

The standard deviation of the lognormal distribution is also increased through the following equation:

$$\sigma_{\text{indoor-indoor}}^{2\text{RayGround}} = \sqrt{\left(\sigma_{\text{outdoor-outdoor}}^{2\text{RayGround}}\right)^2 + (2 \times \sigma_{\text{add}})^2}$$ \hspace{1cm} (3.7)

Values for the additional path-loss and the standard deviation are given in the following table:

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{\text{we}}$</td>
<td>10</td>
<td>dB</td>
</tr>
<tr>
<td>$\sigma_{\text{outdoor-outdoor}}^{2\text{RayGround}}$</td>
<td>5.5</td>
<td>dB</td>
</tr>
<tr>
<td>$\sigma_{\text{add}}$</td>
<td>5</td>
<td>dB</td>
</tr>
</tbody>
</table>
SEAMCAT presently doesn’t have a built-in model for a two-ray ground reflection approach, so we use the propagation plugin functionality to write the code in Java language. The plugin model was proposed to SEAMCAT as an enhancement (ticket #671).

3.2 Primary and secondary users specifications

The WM link is defined in such a way that a WM receiver is at the edge of the coverage area with the received signal equal to -95 dBm. With 10 dBm power emitted from the wireless microphone and using Extended Hata SRD propagation model, this corresponds to 100 m separation distance from the WM receiver. A summary of the specifications for the WM system is shown in Table 7.

Table 7 – PMSE link budget

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Units</th>
<th>Downlink</th>
<th>Comment</th>
</tr>
</thead>
<tbody>
<tr>
<td>Link bandwidth</td>
<td>kHz</td>
<td>200</td>
<td>Bandwidth occupied by link</td>
</tr>
<tr>
<td>Thermal noise density</td>
<td>dBm/Hz</td>
<td>-173.98</td>
<td>$kT_0$</td>
</tr>
<tr>
<td>Receiver noise figure</td>
<td>dB</td>
<td>6</td>
<td>NF</td>
</tr>
<tr>
<td>Noise power over link BW</td>
<td>dBm</td>
<td>-115</td>
<td>$P_n = kTB + NF$ plus any noise rise</td>
</tr>
<tr>
<td>Minimum SNR at cell-edge</td>
<td>dB</td>
<td>21</td>
<td>SNRmin for WM</td>
</tr>
<tr>
<td>Target &quot;mean&quot; received signal level</td>
<td>dBm</td>
<td>-94</td>
<td>$P_{target} = P_n + SNR$</td>
</tr>
<tr>
<td>EIRP</td>
<td>dBm</td>
<td>10</td>
<td>$P$</td>
</tr>
<tr>
<td>Mean wall loss</td>
<td>dB</td>
<td>5</td>
<td>$L_w$</td>
</tr>
<tr>
<td>Receiver Antenna Gain</td>
<td>dBi</td>
<td>2.15</td>
<td>$G_a$</td>
</tr>
<tr>
<td>Max allowed path loss</td>
<td>dB</td>
<td>101.15</td>
<td>$L_p = (P - L_w + G_a) - P_{target}$</td>
</tr>
<tr>
<td>WM Tx height</td>
<td>m</td>
<td>1.5</td>
<td>$H_t$</td>
</tr>
<tr>
<td>WM Rx height</td>
<td>m</td>
<td>1.5</td>
<td>$H_r$</td>
</tr>
<tr>
<td>Cell size</td>
<td>m</td>
<td>100</td>
<td>Extended Hata SRD</td>
</tr>
</tbody>
</table>

At the WM receiver, maximum co-channel interference permitted should be below -115 dBm, taking as a basis analogue FM WM systems. Figure 41 represents the absolute power level (in dBm) of maximum interfering signal, which might be tolerated by the receiver at a given frequency separation.

![Diagram showing maximum interference levels for a 200 kHz channel PWMS receiver [37].](image)

Table 8 – WSD LTE UE transmitter technical parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Units</th>
<th>Uplink</th>
<th>Comment</th>
</tr>
</thead>
<tbody>
<tr>
<td>Link bandwidth</td>
<td>MHz</td>
<td>5</td>
<td>LTE 5 MHz</td>
</tr>
<tr>
<td>In-block EIRP</td>
<td>dBm</td>
<td>23</td>
<td>$P$</td>
</tr>
<tr>
<td>Tx antenna height</td>
<td>m</td>
<td>1.5</td>
<td>$H_t$</td>
</tr>
</tbody>
</table>
Other parameters for WM systems and secondary LTE users are omitted in this section, since they were already been defined in D4.1 (section 2.2) [2].

### 3.3 Results and discussion

We compute for each scenario, the probability of exceeding a predefined Interference Criterion of $\frac{C}{(N+I)} = 20$ dB, varying the separation distance from 1 to 15 km. The results from Figure 42 shows that the worst case is from an outdoor-outdoor scenario, where both systems are in Line-of-Sight (LOS).

![Figure 42](image)

**Figure 42** – Probability of exceeding a predefined Interference Criterion of $\frac{C}{(N+I)}=20$ dB: a) Overall results b) Detail.

For 1% probability of interference (see Figure 42b), the protection area around a PMSE receiver should be defined with a radius above the values given in Table 9, for scenarios 1, 2 and 3.

<table>
<thead>
<tr>
<th>Scenario</th>
<th>Exclusion radius</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>1: outdoor-outdoor (LOS)</td>
<td>13.9</td>
<td>km</td>
</tr>
<tr>
<td>2: indoor-outdoor</td>
<td>9.2</td>
<td>km</td>
</tr>
<tr>
<td>3: indoor-indoor (separated buildings)</td>
<td>7.7</td>
<td>km</td>
</tr>
</tbody>
</table>

### 3.4 Summary

We proceed in this chapter, with the extension of a coexistence study between PMSE systems and secondary users WSDs initiated in D4.1 [2], using the interference analysis SEAMCAT tool developed by CEPT. We compute exclusion areas around PMSE receivers, for co-channel scenarios, when WSD are transmitting at maximum power, summarized in Table 9. The exclusion area is important for the booking of TV channels for professional PMSE systems using the COGEU geo-location database.
4 Testbed for validation of PMSE sensing algorithms in real conditions

Wireless microphones sensing is a mandatory task for COGEU spectrum commons model, as described in the requirement from D3.1 [23], since non-professional PMSE devices are not registered in a database and their location unknown. This section describes the initial development of a testbed, to validate the performance of sensing algorithms for PMSE devices in real scenarios.

4.1 Overview

The tool is based on USRP2 hardware prototyping platform, a GPS receiver and a PC as a host. The software is supported by Labview platform from National Instruments. Labview combines a graphical programming language with the capability to create user-friendly user interfaces, which makes it a preferred choice compared with other software solutions. This hardware and software combination was chosen due to the recent development of USRP drivers for Labview [38].

Figure 43 – Deployed system during the COGEU exhibition at FuNeMS 2011.

4.2 Software Application

The program is divided into a setup process and a sensing process, which are also graphically separated by two different interfaces, as depicted in Figure 44 and Figure 45, respectively. The following section will describe in detail all the important features of the tool.
Figure 44 – Setup interface.

Figure 45 – Sensing interface (with a presence of WM signals in Channel 47 and 61).
4.3 Features and configuration files

4.3.1 Setup interface

After the initial configuration stage starts a communication protocol between the host PC and the USRP, the program acquires GPS coordinates (if no GPS signal is available, coordinates can be inserted manually), queries a web-based database and displays a local map centered on the sensing device, represented by a communication tower, as shown in Figure 46. A circle is drawn around the tower, representing the intended coverage area of a secondary spectrum user (e.g. LTE BS).

![Google map with location of the sensing tool.](image)

Figure 46 – Google map with location of the sensing tool.

After a query from the geolocation database, a list of all available DVB-T channels is received and displayed as LED indicators with different colours: red means that the channel is already occupied by a DVB-T channel, and green means that the channel is free of DVB-T signals. Each time the user wants to reserve (book) a free channel for professional PMSE usage, he can do so by clicking on a free channel indicator that will change to a microphone symbol, as represented in Figure 47.

![Wireless Microphone Booking](image)

Figure 47 – Booking bar

From the same interface, the user can also define sensing parameters, such as detection threshold, sensing time and the sensing algorithm itself, as shown in Figure 48.
4.3.2 Sensing interface

After pressing the RUN button from the setup interface, the USRP starts to sense for signals on the free channels list indicated by the geolocation database. The tool automatically presents statistics on miss detection and false alarm measures for each TV channel sensed, and average values for all target channels.

Figure 49 – Bar indicator for false alarm and miss detection of PMSE

Figure 50 shows the displayed information on the sensing algorithm, indicating the time used to sense each channel, the channel sensed, elapsed time since the beginning of the sensing process, and an estimate of the SNR. More details are given in section 4.3.2.1 on the computation of the SNR.
In Figure 51, we can see a series of bars indicating the status of each channel. A white tab indicates which channel is being sensed in each moment.

![Figure 51 - Status bar indicator](image)

The result of sensing is compared with a threshold. If sensed power is above threshold and there is booking of a WM in that channel, status is set to “occupied”, colour black, and if there’s no previous booking of WM, then status is set to “false alarm”, colour yellow. On the other hand, if the sensed power is below the threshold and there is a WM booking for that channel, status is “miss detection”, colour red, if there is no booking of WM, the status is “free”, colour green. Table 10 resumes the colour code used and their meaning.

<table>
<thead>
<tr>
<th>Colour</th>
<th>Status</th>
</tr>
</thead>
<tbody>
<tr>
<td>Red</td>
<td>Miss detection</td>
</tr>
<tr>
<td>Green</td>
<td>Free channel</td>
</tr>
<tr>
<td>Yellow</td>
<td>False alarm</td>
</tr>
<tr>
<td>Black</td>
<td>Occupied channel</td>
</tr>
</tbody>
</table>

Table 10 - Colour code used to identify sensing results

This methodology is based on a “double affirmative”, as already described in section 4.3 of deliverable D7.1 [34]. A channel is defined as free if both geolocation database and sensing results have the same affirmative decision.

4.3.2.1 SNR estimation

SNR in a channel is computed only if there is a WM booked for that channel. In order to estimate it, we implemented the following method:

1. Sum of the squared amplitude of all the samples from a channel that is free, where only noise $u[n]$ is present;
2. Sum of the squared amplitude of all the samples from a channel that is occupied by a wireless microphone signal plus noise, i.e. $x[n] = s[n] + u[n]$;
3. Computation of the SNR, in dB, using the following expression:

$$SNR = 10 \log \left( \frac{x[n] - u[n]}{u[n]} \right)$$  \hspace{1cm} (4.1)

This method considers that noise levels are the same for all channels, and that only one WM is present in a channel. Figure 52 shows the result of a simulation to verify the accuracy of the proposed method, where the exact SNR of an FM signal with AWGN noise (variance 1) is computed and compared with the proposed methodology.
From the results, we observe that the method is adequate to give an estimate of the SNR, and more accurate for higher SNR regimes.

4.4 Flowchart of the testbed for sensing algorithms validation

The sensing tool is able to showcase detection of PMSE devices combined with geo-location database access as described in D7.1. Figure 53 shows the flowchart of the procedure implemented.
4.5 Future developments

The implemented testbed is at a preliminary stage and needs enhancements to operate in real conditions. Some functionalities that will be integrated in the near future are:

- The capability to identify the central frequency and bandwidth of multiple WMs present in a TV channel. This feature is crucial for COGEU spectrum shaping and spectrum aggregation techniques described in D5.2 [35], allowing coexistence between WSDs and PMSEs systems in the same TV channel.
- Implementation of advanced blind sensing algorithms such as the Covariance Absolute Value (CAV) and the Maximum Eigen value to Trace detection (MET) selected by simulation analysis in section 2.5.
- Perform sensing campaigns in realistic scenarios such as auditoriums and classrooms (LOS and NLOS). The testbed will create automatic reports of detection statistics for comparative analysis of different algorithms. These developments will be reported in deliverable D5.3.
5 Sensing techniques for DVB-T detection

5.1 Description of the main features of the DVB-T signal

The DVB-T system is described in the European Norm EN 300 744 [36]. The first version was published by ETSI in March 1997, the current version 1.6.1 was published in January 2009. In the following subsection, the main features of the DVB-T signal are described, as they are relevant for the development of sensing techniques.

Since DVB-T has a set of standardized wave forms, the detection methods can take advantage of the embedded signal features in order to improve performance.

5.1.1 OFDM modulation

The DVB-T signal is constructed as an OFDM (Orthogonal Frequency Division Multiplexing) signal with 1705 carriers (2K) or 6817 carriers (8K). The 4K variant (3409 carriers) is only used for DVB-H systems. The DVB-T signal occupies one nominal broadcast channel in the RF range. The bandwidth of the broadcast channels is different for different world regions and can be 8 MHz, 7 MHz or 6 MHz. The actual bandwidth of a DVB-T signal is slightly lower than the channel bandwidth (for a 8 MHz broadcast channel the actual bandwidth is 7.61 MHz) to allow for a sufficient signal attenuation in the adjacent channels.

![Figure 54 DVB-T spectrum in an 8-MHz-channel (Source: SFU, -20 dBm, C/N = 30 dB) (Fig. 54)](image)

The spacing between the individual carriers is approx. 1.1 kHz for a 2K system and 4.5 kHz for a 2K system. Therefore, the 2K system is more resilient in a mobile environment and can be received at a higher speed because the tolerable Doppler shift is higher. On the other hand, a 8K signal is more resilient to impulse noise (if the 'native' interleaver is used which stores one symbol). The key parameter for the definition of the OFDM structure is the elementary period T from which the characteristics of the DVB-T signal are derived (Figure 55).
The OFDM carriers that transport the payload, i.e. the data of the television programmes etc., are modulated with QPSK, 16QAM or 64QAM constellations.

For the data of the transmitted DVB-T signal a frame structure is defined. One frame contains 68 OFDM symbols. The OFDM symbols have a duration between 924 µs and 1120 µs for the 8K mode, depending on the length of the guard interval. For the 2K mode, the symbol duration is a quarter of the given values. The duration of an OFDM frame varies accordingly between 62.8 ms and 76.2 ms, a superframe which contains four frames, has a duration of about 250 ms.

5.1.2 Pilot cells

The multi-carrier OFDM signal includes a number of carriers which are not used for the transport of payload information but are modulated with information that is a-priori known by the receiver, to assist it in frame synchronisation, frequency synchronisation, time synchronisation, channel estimation, transmission mode identification and reduction of phase noise.

These special carriers are called pilots and the modulation carried by one carrier (in this case by one pilot carrier) during one symbol is referred to in this context as a pilot cell. The pilots are modulated with reference information whose transmitted value is known to the receiver. The power level of these pilots is raised to a higher value than that of the carriers for the payload. There are two types of pilots: scattered or continual pilots. The continual pilots have the same position amongst all carriers in each OFDM symbol, i.e. their carrier index does not change from symbol to symbol. The scattered pilots are shifted from one OFDM symbol to the next in such a way that they occur in the same position for every fourth OFDM symbol. The modulation value of these pilots is derived from a Pseudo Random Binary Sequence (PRBS). The number of continual pilots is 177 in the 8K mode and 45 in the 2K mode. The first continual pilots in both cases are positioned at carrier number 0, 48, 54, 87, 141, 156, and 192, i.e. they are not equidistant.
5.1.3 TPS pilots und TPS information

Apart from the continual and scattered pilots and the payload carriers, an OFDM frame also contains so-called TPS carriers (TPS: Transmission Parameter Signalling). These carriers are not boosted in their power level. They carry information on the transmission scheme, the channel coding and modulation.

There are 17 TPS carriers in the 2K mode and 68 in the 8K mode. The modulation of the TPS carriers is DBPSK (Differential Binary Phase Shift Keying) which is an extremely robust modulation scheme. Each TPS carrier in an OFDM symbol carries the same information bit. The TPS information consists of 68 bit and is synchronised to the OFDM frame. It includes:
- 1 initialisation bit;
- 16 synchronisation bits;
- 37 information bits;
- 14 redundancy bits for error protection.

Of the 37 information bits, 31 were originally defined. Two more bits were later added for signalling of DVB-H signals.

The information bits transport the information on:
- QAM constellation
- Information on hierarchical modulation including the α value of the QAM constellation pattern
- Interleaver type
- Guard interval
- Inner code rates
- transmission mode (2K or 8K)
- frame number in a super-frame
- cell identification
- time-slicing information

5.1.4 Cell ID with unique transmitter identification

The Cell Identifier (Cell ID) consists of 16 bit. It is mapped to 8 of the 31 used information bits over two consecutive OFDM frames. The Cell ID is used to identify the transmitter cell from which the signal is received.

<table>
<thead>
<tr>
<th>TPS bit number</th>
<th>Frame number 1 or 3</th>
<th>Frame number 2 or 4</th>
</tr>
</thead>
<tbody>
<tr>
<td>s_{40}</td>
<td>cell_id b_{15}</td>
<td>cell_id b_{17}</td>
</tr>
<tr>
<td>s_{41}</td>
<td>cell_id b_{14}</td>
<td>cell_id b_{16}</td>
</tr>
<tr>
<td>s_{42}</td>
<td>cell_id b_{13}</td>
<td>cell_id b_{15}</td>
</tr>
<tr>
<td>s_{43}</td>
<td>cell_id b_{12}</td>
<td>cell_id b_{14}</td>
</tr>
<tr>
<td>s_{44}</td>
<td>cell_id b_{11}</td>
<td>cell_id b_{13}</td>
</tr>
<tr>
<td>s_{45}</td>
<td>cell_id b_{10}</td>
<td>cell_id b_{12}</td>
</tr>
<tr>
<td>s_{46}</td>
<td>cell_id b_{9}</td>
<td>cell_id b_{11}</td>
</tr>
<tr>
<td>s_{47}</td>
<td>cell_id b_{8}</td>
<td>cell_id b_{10}</td>
</tr>
</tbody>
</table>

Figure 59 Mapping of the Cell ID on the TPS bits as defined in [36]

5.2 Power level sensing

The power level sensing, i.e. the measurement of the total signal power of a DVB-T signal is used as a reference for the testing of the sensitivity of the sensing algorithms. Typically, DVB-T receivers provide full quality reception down to a signal level of about -80 dBm (for an 8K signal) provided the C/N values fulfills the requirements of the transmission mode in use, e.g. around 14 dB for 16QAM, CR 2/3.

Figure 60 DVB-T signal at -95 dBm (red lines: 8 MHz channel limits)

A spectrum analyser can measure the power of a received signal by integrating the selectively received power over the nominal signal bandwidth, in this case over 7.61 MHz. Figure 60 shows a screenshot of the spectrum of a DVB-T signal that has an overall power of -95 dBm. This is well below the level at which normal receiver operate, and only 3 dB or 4 dB above the system noise floor, i.e. C/N: 3 dB ... 4 dB). Even the test receiver cannot demodulate such a signal although it is capable of synchronising to the RF signal.
5.3 Description of current sensing algorithm on the DVB-T test receiver

The current implementation of the signal acquisition on the DVB-T test receiver ETL is based on an algorithm that proceeds through a number of subsequent synchronisation steps. It should be noted that in normal operational mode, the test receivers require the synchronisation of the MPEG2 Transport Stream decoder to the output of the demodulator stage, and switches to stand-by if this is not possible. This function can be disabled on the test receiver, and in this mode (synchronisation of MPEG decoder disabled, indicated by red 'MPEG' field in lower right corner of the screen in Figure 65) the test receiver synchronises only to the RF signal and then demodulates the TPS signal. Since the TPS signal is much more resilient than the actual DVB-T data stream, the TPS signal can be demodulated and decoded for much lower signal levels and CNRs.

The main steps of the DVB-T signal acquisition are listed in Table 11. The process is started by setting the demodulator of the test receiver to 'Reset' status.

<table>
<thead>
<tr>
<th>Name</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>RESET_SS</td>
<td>Reset at the beginning of the synchronisation process</td>
</tr>
<tr>
<td>DETECT_SS</td>
<td>Detection of signal power activated</td>
</tr>
<tr>
<td>CS_SS</td>
<td>Coarse Synchronisation activated</td>
</tr>
<tr>
<td>IFS_SS</td>
<td>Integer Frequency Synchronisation activated</td>
</tr>
<tr>
<td>LMOD4_SS</td>
<td>LMOD4 synchronisation activated</td>
</tr>
<tr>
<td>CE_SS</td>
<td>Channel Estimation activated</td>
</tr>
<tr>
<td>EQ_SS</td>
<td>Equaliser activated</td>
</tr>
<tr>
<td>TPS_SS</td>
<td>TPS decoder activated</td>
</tr>
</tbody>
</table>

Table 11 Steps during DVB-T signal acquisition

There are additional intermediate states which can be used for debugging and tracing of software incompatibilities. These states are not relevant to the Cell ID decoding path and therefore not described here.

5.3.1 Description of the relevant stages of the DVB-T signal acquisition

When the DVB-T receiver is switched on or when the DVB-T signal is connected to the instrument, the algorithm starts with a general reset. After the reset, the next step is the activation of the AGC (Automatic Gain Control) in the RF path. As soon as the signal is received within the optimal power range, the Coarse Synchronisation algorithm is activated.

The Coarse Synchronisation establishes the FFT size of the OFDM signal, i.e. if the signal received is an 8K, a 4K (e.g. for DVB-H) or a 2K signal, and the length of the Guard Interval.

The Integer Frequency Synchronisation (IFS) provides the correct adjustment of the OFDM carriers after the FFT (Fast Fourier Transformation) process. The algorithm is based on the continual pilots and the number of trials depends on the required frequency acquisition range. A rough estimate is obtained by dividing the acquisition range (e.g. 200 kHz) by the spacing between the sub-carriers (e.g. 1.1 kHz for an 8K DVB-T signal).

![Figure 61 Result for a correct FFT correlation](image)
For the results shown in Figure 61, the FFT output is shifted by one OFDM carrier over a range of -400 to +400 sub-carriers, and the FFT correlations are calculated by using the Continual Pilots. This corresponds to a carrier offset of more than 400 kHz for an 8K signal. Since the input signal had no carrier offset, the maximum of the correlation occurs at index number 0.

The calculation for each positioning of the OFDM carriers is regarded as a trial. The correct frequency offset has been found if the calculation of the correlation shows a clearly expressed maximum. Figure 61 gives an example of the correct FFT correlation for an ideal channel, i.e. no additional noise, no multi-path propagation. In this case, the trial was using the actual frequency offset which can then be used to re-adjust the signal for the next steps.

![Figure 62 Results of the FFT correlation for two testcases FFT length 2048, TU6](image)

The results in Figure 62 were calculated for a 2K DVB-T signal which represents the most difficult case. On the left, the SNR of the simulated signal is ideal, on the right AWGN is added up to $SNR = 0 \text{ dB}$. In both cases, the decision is a robust one, the correct FFT correlation is clearly identifiable.

In the LMOD4 state, the receiver searches for the correct order of the scattered pilots pattern which is repetitive with a repetition period of four symbols. The correlation is calculated in the time domain, not for two consecutive symbol but for two symbols with a timely distance of four symbols. Through this process, the receiver identifies the index modulo 4 of the OFDM symbol within a DVB-T frame, i.e. if its structure is equivalent to symbol number 0, 1, 2 or 3 (a modulo4 reduction of the number of possibilities).

![Figure 63 Typical results of the LMOD4 correlation](image)

In an ideal channel the correlation calculation (Figure 63, left diagram) shows a very reliable result. The correlation function provides a value of approximately 1, the ideal value. The right diagram in Figure 63 shows the results in a channel with $C/N = 0 \text{ dB}$, a significant clock frequency offset and a Doppler shift (here 400 Hz for a 2K signal). Although the maximum of the correlation function is significantly lower than in an ideal channel, the correct position can be identified with sufficient confidence.
After this step has been concluded successfully, the Channel Estimation is carried out, followed by the frequency tracking and clock tracking. Several cases are considered, based on parameters derived during the previous steps. The equaliser uses the results from the channel estimation in the next step to correct the frequency response of the OFDM signal and deliver this signal to the TPS decoder.

Even under very difficult reception conditions, the TPS decoder provides the correct data. The example in Figure 64 shows on the left a constellation diagram at C/N = 0 dB. For such a signal no synchronised constellation is possible due to low C/N (red: nominal constellation points) but the TPS data are correct.

![Figure 64 Constellation plane and digitalised spectrum for 64QAM signal with C/N = 0 dB](image)

The right part of Figure 64 shows the sampled signal of the digitised OFDM signal after filtering, so that the additive noise is reduced outside the occupied bandwidth (consisting of 6817 sub-carriers). The curve is derived from the output of the FFT calculation (8K-FFT in this example).

![Figure 65 Indication of TPS information on the screen of the test receiver](image)

The TPS data displayed in the lower part of the DVB-T test receiver screen in Figure 65 include:

- Type of modulation (16 QAM non-hierarchical)
- FFT type (8k)
- Guard Interval length (1/4 of the duration of the useful part of the symbol)
- Code Rate for convolutional coder (2/3)
- Cell ID (1234)
- PTS Reserved bits (0, 0, 0, 0)
- INT (Interleaver for DVB-H, native)
- MPE FEC status (off)
- Time slicing status (off)
- Length Indicator (LI 1F i.e. 31 TPS bits used)

The decoding of the Cell ID as in the original algorithm used in the test receiver requires two complete OFDM frames, each consisting of 68 OFDM symbols. In the case of an 8K mode and a Guard Interval of 1/4, the acquisition time of the data alone is about 230 ms. This is based on the assumption that the starting point of the acquisition is arbitrary and therefore the data have to be received over a period that is equal to three OFDM frames to guarantee that the data set includes two complete OFDM frames. The processing time for the various steps of the DVB-T signal synchronisation and the time for updating the display buffer periodically have to be added to the acquisition time, resulting in an estimate of 0.9 sec to 1.2 sec for time from the availability of the DVB-T signal at the input of the test receiver to the display of the value of the Cell ID on screen (and the respective data being available at the remote control interface).

5.4 Description and simulation of the improved DVB-T sensing algorithm

5.4.1 Assumptions and simplifications

The design of an improved sensing algorithm started with the listing of the most important requirements: identification of a DVB-T signal in shorter time and/ or at a lower signal level/ lower C/N level.

The current sensing algorithm which relies on the Cell ID information, is implemented on a test receiver platform. The DVB-T test receiver needs to carry out all steps of the synchronisation in the process to demodulate and decode the DVB-T signal including the TPS signal. In comparison with the resources available in a White Space Device (WSD), the resources of the test receiver in terms of computing power and memory are plentiful.

An improved sensing algorithm for DVB-T signals that should be implementable also on a WSD platform has to be down-sized as much as possible and its FPGA footprint should be significantly smaller than that of the current sensing algorithm.

To achieve this, the pre-existing knowledge on the occupation of the TV band in the UHF range is used. For example, in Germany, all DVB-T signals are 8K signals, in the UK there are 2K and 8K signals (but no 4K signals) on air. The channel bandwidth is also known. In Europe, a UHF TV channel is 8 MHz wide, in the US 6 MHz. When sensing for DVB-T signals in the UHF TV band, the frequency range is pre-defined as 470 MHz to 790 MHz. In the cases considered in the COGEU project, the frequency range is limited further to 622 MHz to 790 MHz. In this frequency range, only DVB-T signals and signals from PSMEs are legitimately present.

Based on this knowledge, a test receiver or a WSD could significantly reduce the number of trials which are necessary to synchronise to a DVB-T signal. If a sensing algorithm was based on the same principles as used in the current sensing algorithm of the test receiver it would be sufficient to check the Coarse Synchronisation. If this is achieved it can be concluded that the signal is a DVB-T signal, and the Guard Interval is then also known.

This is much quicker than the complete synchronisation, demodulation and decoding of the TPS information which is necessary to obtain the Cell ID information.

To allow such a short-cut, the test receiver is being modified by implementing an additional interface that allows access to the data of the complex time domain samples at the frontend. These I/Q samples are sent to an external computer on which a newly developed software application is running. This software is based on a Matlab implementation and is stripped of all parts which are not required for the Coarse Synchronisation, and it makes use of the pre-existing knowledge on the mode of the DVB-T signal in a country.
The Coarse Synchronisation is described hereafter in more detail as it is being carried out by the improved sensing algorithm.

The newly developed sensing algorithm also provides a statistical evaluation of the measurement of the time required to establish that a DVB-T signal is received (8K or 2K, GI 1/4, 1/8, 1/16 or 1/32), the correct detections and the false alarms.

5.4.2 Requirements for the DVB-T sensing algorithm

The sensing algorithm shall work under the following conditions:

- The radio channel is unknown. Multi-path propagation, Doppler shift and added noise need to be considered.
- The frequency offset of the RF and of the system clock have not been compensated for at a previous stage of the signal processing.
- The FFT length $N_{FFT}$ and the length of the Guard Interval $N_g$ have to be established from the received DVB-T signal.
- A relative clock frequency error of at least 40 ppm has to be tolerated.

The likelihood for a correct identification of a DVB-T signal should be tested as a function of sensing time and signal level/noise level.

5.4.3 Principle functionalities of the DVB-T sensing algorithm

The Coarse Synchronisation is based on a time-domain procedure that makes use of the periodic repetition of the signal during the Guard Interval. The metric that is used to establish the beginning of the OFDM symbol is derived for an AWGN channel according to the Maximum-Likelihood (ML) estimation theory. An approximation of this metric is

\[
Metric_{comp}(i) = \frac{1}{N_g} \sum_{j=1}^{i+N_g-1} r(j+ N_{FFT}) \cdot r^*(j)
\]

\[
Metric(i) = |Metric_{comp}(i)|^2
\]  

where $r(i)$ is the I/Q input sequence. The search range

\[
i = [1, N_g \cdot N_{FFT}]
\]

is equivalent of the duration of one OFDM symbol. This is sufficient since a new OFDM symbol has to start during this period. A second OFDM symbol is required for the calculation of the metric value $Metric(i)$. This leads to a required observation interval of two OFDM symbols.

The estimation of the starting point of the OFDM symbol searches for the absolute maximum of the complex metric

\[
i_{\text{max}} = \arg\{\max\{Metric(i)\}\}\]

The fractional frequency offset is also estimated from the maximum metric since the frequency offset $\Delta f$ is equivalent to a phase shift.

\[
\left(\frac{\Delta f}{f_{bin}}\right)_{\text{fractional}} = \frac{\arg\{Metric_{comp}(i_{\text{max}})\}}{2\pi}
\]

where $f_{bin} = 1/T$ is the frequency difference between two adjacent sub-carriers and $T$ is the duration of the orthogonality interval. Phase shifts of multiples of $2\pi$ are not evaluated.
The relative clock error of $r(i)$ is
\[ \xi = \frac{T_{Rx} - T_{Tx}}{T_{Tx}}, \]  
(5.5)

Where $T_{Rx}$ is the duration of the orthogonality interval at the transmitter and $T_{Rx}$ at the receiver. These two periods are not exactly equal to $T$ since both transmitter and receiver use different reference oscillators.

The coarse estimation of the clock error also relies on the metric
\[
\text{Metric}_{\text{comp}}(i, \tilde{d}i) = \frac{1}{N_g} \sum_{j=1}^{i+N_g-1} r(j + \tilde{d}i + N_{FFT}) \cdot r^*(j)
\]
(5.6)

with $\tilde{d}i$ as an integer search parameter. A two-dimensional maximum search provides the correct estimate for the timing drift $\tilde{d}i$. This estimation works reliably even for large clock offsets (compared to the carrier spacing).

In Figure 66 the Coarse-Timing and the buffer management are illustrated. After the switching-on of the receiver or the connection of the input signal, the Coarse-Timing buffer is filled. Reception starts arbitrarily during an OFDM symbol. As soon as two OFDM symbols, i.e. $2 \cdot (N_g + N_{FFT})$ have been written into the Coarse-Timing buffer, its content is frozen and the following samples are written into the FFT buffer.

The $\text{Metric}(i)$ is then calculated for all samples $i = [1, N_g + N_{FFT}]$ of one OFDM symbol and the maximum is identified. The point in time for $i_{\text{max}}$ is the estimated start of the OFDM symbol. Figure 66 shows also a typical curve for the $\text{Metric}(i)$ as a function of $i$. An increased distance from $i_{\text{max}}$ leads to lower correlations and hence a lower value for the $\text{Metric}(i)$. The FFT start is set as $i_{\text{Start FFT}} = i_{\text{max}}$ i.e. at the earliest possible time.

During the Coarse-Timing procedures, $N_{FFT}$ and $N_g$ are not known yet to the receiver. These parameters need to be estimated from the received signal as well. For each possible FFT length (8K, 4K, 2K), the $\text{Metric}(i)$ is calculated in a trial-and-error algorithm for each possible Guard Interval length.
This algorithm starts always with the longest Guard Interval $N_{g,Tr} = \frac{N_{FFT}}{4}$ and then reduces it – if necessary – in the next step. As soon as the curve of the Metric(i) shows the correct triangular shape with the maximum $\text{Metric}_{\max} = 1$, the correct Guard Interval length has been found, i.e. $N_{g,Tr} = N_{g}$.

For a trial with a Guard Interval of double the correct length $N_{g,Tr} = 2N_{g}$, the curve of Metric(i) shows a plateau with a maximum value of $\frac{1}{2}$.

For a Guard Interval that has half the length of the correct one, the curve of Metric(i) also has a plateau, although the maximum is similar to the maximum of the correct trial $\text{Metric}_{\max} = 1$. In principle, the value $\text{Metric}_{\max} = 1$ can occur for the correct Guard Interval and all other that are shorter than the correct one. Since the channel properties may impact on the maximum of Metric(i), a threshold is introduced in the range $< 1.0$ to check for a sufficiently high value.

The FFT length is established with a similar correlation algorithm. It also uses a trial-and-error approach which leads to an unambiguous result since the unsuccessful trials with an incorrect FFT length result in $\text{Metric}_{\max} \approx 0$.

To economise on the time needed for the establishment of the FFT length and the Guard Interval, the calculations are done in parallel for all 12 Trials $(N_{FFT,Tr}, N_{g,Tr})$

\[
N_{FFT,Tr} = [2048, 4096, 8192] \\
N_{g,Tr} = N_{FFT,Tr} \cdot [1/4, 1/8, 1/16, 1/32]
\]

If the receiver has pre-existing knowledge of the FFT size, this leads to a reduction in the number of trials and decreases the necessary computing power.
The number of samples \( r(i) \) of the I/Q input sequence that are written into the memory space \([v_r]\) depends on the FFT length and the length of the Guard Interval:

\[
[v_r] = \text{nof}_{\text{Symbols, Tr}} \cdot (N_{g, Tr} + N_{FFT, Tr}) \\
= \text{nof}_{\text{Symbols, Tr}} \cdot N_{FFT, Tr} \cdot (N_{g, Tr} / N_{FFT, Tr} + 1) 
\]

(5.7)

Table 12 below shows the possible values of \([v_r]\) for all combination of FFT and GI (Guard Interval) in the DVB-T standard. The maximum of 30720 samples also defines the required buffer size.

<table>
<thead>
<tr>
<th>( N_{g, Tr} / N_{FFT, Tr} )</th>
<th>( N_{FFT, Tr} = 2048 )</th>
<th>( N_{FFT, Tr} = 4096 )</th>
<th>( N_{FFT, Tr} = 8192 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1/4</td>
<td>20480</td>
<td>25600</td>
<td>30720</td>
</tr>
<tr>
<td>1/8</td>
<td>18432</td>
<td>23040</td>
<td>27648</td>
</tr>
<tr>
<td>1/16</td>
<td>17408</td>
<td>21760</td>
<td>26112</td>
</tr>
<tr>
<td>1/32</td>
<td>16896</td>
<td>21120</td>
<td>25344</td>
</tr>
</tbody>
</table>

Table 12 Number of I/Q samples required for Coarse Synchronisation

### 5.4.4 Preliminary Matlab simulations

A number of Matlab simulations have been carried out to establish the reliability of the algorithm in terms of FFT length and Guard Interval length identification.

These tests were based on some general assumptions:

1. For all simulations, the starting point coincides with the start of an OFDM symbol, i.e. the start of the OFDM symbol was determined in a previous step. Consequently, the maximum of the Metric is positioned at \( i_{\text{max}} = 1 \) for an ideal channel \((h(\tau) = \delta(\tau))\) and the correct FFT and GI length.

2. The average power of the received signal is normalised to 1. Therefore, the maximum of the Metric should be \( \text{Metric}_{\text{max}} \approx 1 \) for an ideal channel without AWGN.

3. The threshold for the identification of \( \text{Metric}_{\text{max}} \) was set to 0.6 in the following examples. It is indicated in the diagrams as a blue dotted line.

4. Since the Metric for trials with the incorrect FFT length yield zero, only the trials with the correct FFT length are depicted for the different Guard Interval lengths.

The conditions for Testcase 1 are

- \( N_g = 256 \), \( N_{FFT} = 2048 \)
- ideal channel \( h(\tau) = \delta(\tau) \)
- no AWGN, i.e. \( \text{SNR} = \infty \)
- no clock frequency offset

Figure 68 shows the Metric for GI = 1/4 in the upper left part. For the correct Guard Interval \( N_g = 256 \) (upper right diagram) the threshold is clearly exceeded and the maximum of the Metric occurs, as expected, for \( i_{\text{max}} = 1 \).
The lower part shows the respective results for shorter Guard Intervals (1/16, 1/32) which give much lower values for the Metric. The chosen GI length is therefore 1/8. The differences between the maximum of the Metric in all cases point to a very robust algorithm.

The second Testcase (Figure 69) is based on the same DVB-T signal but introduces strong AWGN and a significant clock frequency offset.

- added AWGN with $SNR = 0\, dB$
- clock frequency offset $\psi = 40\, ppm$

The Metrics for all four possible Guard Intervals show that only in one case the threshold, which was reduced to compensate for the impact of the added AWGN to $\text{Metrik}_{\text{max}} = 0.24$, is exceeded. It is an indication for the stability of the algorithm that even under such adverse conditions, the maximum of the Metric occurs at $i_{\text{max}} = 1$. The trials with incorrect Guard Interval lengths show significantly lower values for the Metric, the decision for the correct GI length is therefore unambiguous.
The next Testcase 3 is based on an 8K DVB-T signal with two propagation paths. In this case, the two paths have the same power level (i.e. two 0-dB-echoes) and the delay between them is equal to the length of the Guard Interval.

- \( N_g = 1024 \), \( N_{FFT} = 8192 \)
- worst case channel \( h(r) = \delta(r) + \delta(r - T_g) \)
- no AWGN \( SNR = \infty \)
- no clock frequency offset \( \psi = 0 \)

In this case, it is necessary to average over several symbols for the calculation of the Metric. If the calculation is only carried out for one symbol, the Guard Interval could be incorrectly estimated (Figure 70). This is mainly due to the artificial condition, that the second path has the same amplitude and the delay is exactly \( T_g \).

Therefore, the implementation considers at least \( n_{Symbols,Tr} = n_{AVG} + 1 = 3 \) OFDM symbols for a signal with the maximum FFT size \( N_{FFT} = 8192 \), see also Table 12.

The Metrics for more than one Guard Interval show results above the threshold. In this case it is possible to increase the number of symbols over which the Metric is calculated until the robustness of the decision is sufficient. For shorter FFT length, even more symbols can be written into the same buffer space.
In Testcase 4 the 3rd Testcase is repeated with two additional types of impairments:

- additional AWGN \( SNR = 10 \) dB
- clock frequency offset \( \psi = 40 \) ppm

Figure 71 shows the results of the four trials, and the conclusion is that the trial need repeated with more symbols for the averaging before the decision for the correct Guard Interval length can be made. The Matlab simulations point to an averaging over 6 to 8 sample sets of maximum 30270 I/Q samples being sufficient for a correct determination of the FFT size and the GI length, even under extremely difficult reception conditions. These intermediate results from the simulations are to be verified with the real-time implementation of the second DVB-T sensing algorithm.

In conclusion, the number of samples required by the second algorithm is much smaller than for the first algorithm (maximum about 10 %). This allows the second algorithm to identify a DVB-T signal during a shorter sensing time, or obtain results over a similar sensing time as the first algorithm, but then with higher confidence.

5.4.5 Description of implementation constraints

The improved sensing algorithm for DVB-T is implemented on an external computer that is directly connected to the DVB-T test receiver ETL. The I/Q samples from the ETL are directly sent through a newly developed interface to this computer. The Matlab simulation on this computer carries out all the required calculations that allow the decision, if a DVB-T signal is present at the input of the test receiver or not.

The reason for the implementation of the algorithm on the external computer lies in the design of the FPGA of the test receiver and the constraints of its use. The main constraints are:

- The FPGA contains all the different algorithms for the synchronisation, demodulation and decoding of the DVB-T signal and other types of signals, plus all measurement routines and statistical evaluations. It therefore needs to be secured against any attempt to gain access to its routines.
- The surplus of unused cells on the FPGA is very limited. Any modification could jeopardise the stability of any other function. It would therefore be necessary to test the complete functionalities after each modification. For reasons of economising on costs (this process can easily take several weeks), this is only done for new versions of the R&S commercial software/firmware.
• The newly developed interface for the handover of the I/Q samples to the Matlab application on the computer is only implemented on the instrument that will be integrated into the COGEU project's testbed, and makes use of internal test ports.

The creation of a work environment that is almost completely decoupled from the test receiver FPGA implementation, is a great advantage for the project work and allows a flexibility that could not be achieved otherwise.

5.4.6 Expected improvements of the enhanced sensing algorithm for DVB-T

The most important requirement for the enhanced sensing algorithm for DVB-T signal was the reduction of the sensing time. The current algorithm that checks the decoded information of the Cell ID, consists of the following main steps:

• Reset of the software state of the test receiver
• AGC initialisation and pulling of the RF signal to the correct frequency positioning
• Coarse Synchronisation with identification of FFT length and Guard Interval (including a significant number of trials)
• Integer frequency synchronisation to compensate for any RF offset
• LMOD4 synchronisation with identification of the starting point of the OFDM frame
• Calculation of the channel estimation
• Correction of the received OFDM symbols through the equaliser
• Decoding of the TPS bits

These calculations are based on the processing of at least two OFDM frames with 136 OFDM symbols (equivalent to 136 x (8192+2048) = 1392640 I/Q samples).

The enhanced sensing algorithm provides a short-cut insofar as it only carries out the first three steps of the list above, i.e.

• Reset of the software state of the test receiver
• AGC initialisation and pulling of the RF signal to the correct frequency positioning
• Coarse Synchronisation with identification of FFT length and Guard Interval (including a significant number of trials)

If the enhanced sensing algorithm can make use of pre-existing knowledge of the FFT size, e.g. 8K, this reduces the number of trials for the decision making process. The process is also sped up by the reduced number of samples that are necessary for the decision making, i.e. a maximum of 30720 samples.

In the version to be implemented on the project's testbeds, the new algorithm allows a detailed statistical evaluation of the average rate of correct detections, false alarms etc. under different test conditions, e.g. different signal power levels, different propagation environments, adjacent channel occupations and other more.
6 Conclusions

This deliverable D4.2 provides a description of the sensing methods for PMSE systems and DVB-T signals. It reports the status of the work in Task 4.3 ‘Sensing algorithms for monitoring and analysing the network conditions’.

In Chapter 2 we present an extensive description of the spectrum sensing literature for PMSE devices and then progress to evaluate a set of candidate techniques using Matlab™. Several methods that are suitable for DVB-T signal, namely Cyclostationary Feature Detector, Waveform-Based Detection or Matched Filtering, have no application for PMSE, due to the nature of FM signal and the lack of a common standard for PMSE devices. We observe that the performance of the ED degrades considerably in the presence of noise uncertainty, and that the degradation is dependent on the amplitude of the acoustic signal. It is however simple to implement.

Covariance (CAV) and Eigenvalue (MET-BCED) based methods require little information on the signal or the channel and have some immunity to synchronization error, fading and multipath, noise uncertainty, and unknown interference. They are particularly adequate when signals are highly correlated, as in the case of PMSE signals. The performance of such detectors was measured against the ED and has shown significant performance gains. These methods overcome the noise uncertainty problem and perform globally better than ED in both AWGN and Rayleigh faded channels. Moreover, the performance is maintained for all WM operation modes. This makes them a more suitable choice for the detection of WM signals, even if the computational complexity is increased.

In Chapter 3 we then proceed with the description of a coexistence study using SEAMCAT, between PMSE and TVWS devices. We compute the exclusion area around a PMSE receiver for different scenarios, when secondary users are portable devices using LTE based technology. The results show that, when TVWS devices are used with maximum power (23 dBm), PMSE receiver’s protection radius should be in the order of several km (from 7.7 km to 13.2 km).

In Chapter 4 we present a tool for performance analysis of PMSE-sensing algorithms in real conditions. This tool is meant to measure the performance of the different sensing algorithm for PMSE devices presented in this deliverable. Field tests will be conducted for indoor and outdoor scenarios. The algorithms with better performances will be integrated into the COGEU TVWS transceiver. The integration and implementation work in the real TVWS transceiver prototype using USRPs will be reported in D5.3 as planned in the COGEU DoW.

In Chapter 5, the first sensing algorithm currently implemented on the DVB-T test receiver is described. It is based on the decoding of the TPS information, in particular the Cell ID, and requires two complete OFDM frames (equivalent to approx. 1.3 million I/Q samples) for reliable decoding.

The improvements by the newly developed sensing algorithm are illustrated by preliminary simulation results. This second DVB-T sensing algorithm determines only the FFT size and the GI length, which can identify a DVB-T signal reliably, without decoding the TPS information. It can operate with a similar reliability as the first algorithm on a ration of the I/Q samples (between 2 % and 10 % of the samples needed for the first algorithm). The Matlab simulations indicate that the results of the DVB-T sensing are available in shorter time, or with higher confidence if sensing is applied over the same time.

The sensing algorithms for the DVB-T test receiver will be integrated and validated in real-time operation on the testbed for coexistence evaluation in Task 4.4. There the sensing tools will be used for the evaluation of system components.

The sensing tools are also to be part of the COGEU project demonstrator developed in WP7.
7 References


[36] ETSI EN 300 744 V1.6.1: Digital Video Broadcasting (DVB); Framing structure, channel coding and modulation for digital terrestrial television (2009-01).


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## 10 List of Abbreviations

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<th>Description</th>
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<tr>
<td>3GPP</td>
<td>3rd Generation Partnership Project</td>
</tr>
<tr>
<td>AGC</td>
<td>Automatic Gain Control</td>
</tr>
<tr>
<td>AGM</td>
<td>Arithmetic to geometric mean</td>
</tr>
<tr>
<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
</tr>
<tr>
<td>CAV</td>
<td>Covariance absolute value</td>
</tr>
<tr>
<td>Cell ID</td>
<td>Cell Identifier</td>
</tr>
<tr>
<td>C/N, CNR</td>
<td>Carrier-to-noise (ratio)</td>
</tr>
<tr>
<td>CNR</td>
<td></td>
</tr>
<tr>
<td>CEPT</td>
<td>Conference of European Postal &amp; Telecommunications</td>
</tr>
<tr>
<td>CFN</td>
<td>Covariance Frobenius norm</td>
</tr>
<tr>
<td>CR</td>
<td>Cognitive Radio</td>
</tr>
<tr>
<td>DVB-H</td>
<td>Digital Video Broadcasting - Handheld</td>
</tr>
<tr>
<td>DVB-T</td>
<td>Digital Video Broadcasting - Terrestrial</td>
</tr>
<tr>
<td>DTT</td>
<td>Digital Terrestrial Television</td>
</tr>
<tr>
<td>DTV</td>
<td>Digital Television</td>
</tr>
<tr>
<td>ED</td>
<td>Energy detector</td>
</tr>
<tr>
<td>EN</td>
<td>European Norm</td>
</tr>
<tr>
<td>ETSI</td>
<td>European Telecommunications Standards Institute</td>
</tr>
<tr>
<td>EU</td>
<td>European Union</td>
</tr>
<tr>
<td>FFT</td>
<td>Fast Fourier Transformation</td>
</tr>
<tr>
<td>GI</td>
<td>Guard Interval</td>
</tr>
<tr>
<td>LTE</td>
<td>Long Term Evolution</td>
</tr>
<tr>
<td>MAC</td>
<td>Maximum auto-correlation</td>
</tr>
<tr>
<td>MME</td>
<td>Maximum to minimum eigenvalue</td>
</tr>
<tr>
<td>MET</td>
<td>Maximum Eigenvalue to Trace detection, also called Blindly Combined Energy Detection</td>
</tr>
<tr>
<td>-BCED</td>
<td></td>
</tr>
<tr>
<td>ML</td>
<td>Maximum-Likelihood</td>
</tr>
<tr>
<td>OFDM</td>
<td>Orthogonal Frequency Division Multiplexing</td>
</tr>
<tr>
<td>PMSE</td>
<td>Programme Making and Special Events</td>
</tr>
<tr>
<td>PRBS</td>
<td>Pseudo Random Binary Sequence</td>
</tr>
<tr>
<td>PU</td>
<td>Primary users</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
</tr>
<tr>
<td>SDR</td>
<td>Software Defined Radio</td>
</tr>
<tr>
<td>TPS</td>
<td>Transmission Parameter Signalling</td>
</tr>
<tr>
<td>TU6</td>
<td>Typical Urban 6 paths (propagation profile)</td>
</tr>
<tr>
<td>TV</td>
<td>Television</td>
</tr>
<tr>
<td>TWWS</td>
<td>TV White Spaces</td>
</tr>
<tr>
<td>UHF</td>
<td>Ultra High Frequency</td>
</tr>
<tr>
<td>WLAN</td>
<td>Wireless Local Area Network</td>
</tr>
<tr>
<td>WM</td>
<td>Wireless Microphone</td>
</tr>
<tr>
<td>WP</td>
<td>Work Package</td>
</tr>
<tr>
<td>WSD</td>
<td>White Space Device</td>
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