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Passive Radar using COFDM (DAB or DVB-T) Broadcasters as Opportunistic Illuminators

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1. Introduction

This chapter is not dedicated to improve DVB-T (Digital Video Broadcasters-Terrestrial) reception in critical broadcasting conditions. Our purpose is to explain and illustrate the potential benefits related to the COFDM (Coded Orthogonal Frequency Division Multiplex) waveform for passive radar application. As we'll describe, most of the benefits related to COFDM modulation (with guard interval) for communication purpose, could be derived as advantages for passive radar application. The radar situation considered is the following: the receiver is a fixed terrestrial one using COFDM civilian transmitters as illuminators of opportunity for detecting and tracking flying targets. The opportunity COFDM broadcasters could be either DAB as well as DVB-T ones even in SFN (Single Frequency Network) mode for which all the broadcasters are transmitting exactly the same signal. Such application is known in the literature as PCL (Passive Coherent Location) application [Howland et al 2005], [Baker & Griffiths 2005].

This chapter will be divided into three main parts. The first ones have to be considered as simple and short overviews on COFDM modulation and on radar basis. These paragraphs will introduce our notations and should be sufficient in order to fully understand this chapter. If not, it is still possible to consider a „classical" radar book as well as some articles on COFDM like [Alard et al 1987]. More specifically, the COFDM description will outline the properties that will be used in radar detection processing and the radar basis will schematically illustrate the compulsory rejection of the „zero-Doppler" paths received directly from the transmitter or after some reflection on the ground.

Then the most important part will detail and compare two cancellation filters adapted to COFDM waveform. These two filters could be applied against multipaths (reflection on ground elements) as well as against multiple transmitters in SFN mode. In this document, no difference will be done between SFN transmitters contributions and reflections on fixed obstacles : all these zero-Doppler paths will be considered as clutter or propagation channel. Obviously, these filters will be efficient also in a simple MFN (Multiple Frequency Network) configuration. Most of the results presented below concerns experimental data, nevertheless some simulations will also be used for dealing with some specific parameters.

2. Principle of COFDM modulation

As mentioned in the introduction, the purpose of this paragraph is just to briefly describe the principle and the main characteristics of the COFDM modulation in order to explain its

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advantages even for radar application. For further details, it's better to analyse the reference [Alard et al 1987], however for radar understanding this short description should be sufficient.

2.1 Basis principle

In a COFDM system of transmission, the information is carried by a large number of equally spaced sinusoids, all these sub-carriers (sinusoids) being transmitted simultaneously.
These equidistant sub-carriers constitute a "white" spectrum with a frequency step inversely proportional to the symbol duration.
By considering these sub-carriers:

$$f_k = f_0 + k/T_s \tag{1}$$

with T_s corresponding to symbol duration.
It becomes easy to define a basis of elementary signals taking into account the transmission of these sinusoids over distinct finite duration intervals T_s:

$$\psi_{j,k}(t) = g_k(t - jTs) \text{ with } \begin{cases} 0 \le t < T_s & : \quad g_k(t) = e^{2i\pi f_k t} \\ elsewhere & : \quad g_k(t) = 0 \end{cases} \tag{2}$$

All these signals are verifying the orthogonality conditions:

$$j \neq j' \;\; or \;\; k \neq k' : \int_{-\infty}^{+\infty} \psi_{j,k}(t)\psi^*_{j',k'}(t)dt = 0 \qquad and \qquad \int_{-\infty}^{+\infty} \left\|\psi_{j,k}\right\|^2 dt = T_s \tag{3}$$

By considering the complex elements $\{C_{j,k}\}$ belonging to a finite alphabet (QPSK, 16 QAM,…) and representing the transmitted data signal, the corresponding signal can be written:

$$x(t) = \sum_{j=-\infty}^{+\infty} \sum_{k=0}^{N-1} C_{j,k}\psi_{j,k}(t) \tag{4}$$

So the decoding rule of these elements is given by:

$$C_{j,k} = \frac{1}{T_s}\int_{-\infty}^{+\infty} x(t)\psi^*_{j,k}(t)dt \tag{5}$$

Remark:
From a practical point of view this decomposition of the received signal on the basis of the elementary signals $\psi_{j,k}(t)$ could be easily achieved using the Fourier Transform over appropriate time duration T_s.

2.2 Guard interval use

In an environment congested with multipaths (reflections between transmitter and receiver), the orthogonality properties of the received signals $\psi_{j,k}(t)$ are no longer satisfied.

In order to avoid this limitation, the solution currently used, especially for DAB and DVB, consists in the transmission of elementary signals $\psi_{j,k}(t)$ over a duration $T_s^{'}$ longer than T_s. The difference between these durations is called guard interval. The purpose of this guard interval is to absorb the troubles related to the inter-symbols interferences caused by the propagation channel. This absorption property needs the use of a guard interval longer than the propagation channel length. Then, we just have to "wait for" all the contributions of the different reflectors in order to study and decode the signal on a duration restricted to useful duration T_s.

The transmitted signal could be written:

$$x(t) = \sum_{j=-\infty}^{+\infty} \sum_{k=0}^{N-1} C_{j,k} \psi_{j,k}^{'}(t) \tag{6}$$

$$\text{with } \psi_{j,k}^{'}(t) = g_k^{'}(t - jT_s^{'}) \text{ with } \begin{cases} -\Delta \le t < T_s & : \quad g_k^{'}(t) = e^{2i\pi f_k t} \\ elsewhere & : \quad g_k^{'}(t) = 0 \end{cases} \tag{7}$$

Nevertheless the decoding rule of these elements is still given by:

$$C_{j,k} = \frac{1}{T_s} \int_{-\infty}^{+\infty} x(t) \psi_{j,k}^{*}(t) dt \tag{8}$$

with $\psi_{j,k}(t)$ always defined on useful duration T_s while signal is now specified (and transmitted) using elementary signals $\psi_{j,k}^{'}(t)$ defined on symbol duration $T_s^{'} = T_s + \Delta$.

This decoding rule means that even when signals are transmitted over a duration $T_s^{'} = T_s + \Delta$, the duration used, in reception for decoding will be restricted to T_s. Such a "cut" leads to losses equal to $10 \log T_s^{'} / T_s$ but allows easy decoding without critical hypothesis concerning the propagation channel. In practice, this truncation doesn't lead to losses higher than 1 dB (the maximum guard interval Δ is generally equal to a quarter of the useful duration Ts).

The guard interval principle could be illustrated by the figure 1.

The previous figure illustrates the main advantage of guard interval truncation: by "waiting" for all the fixed contributors, it's easy to avoid signal analysis over transitory (and unstationary) time durations.

Considering the parts of signal used for decoding (so after synchronisation on the end of the guard interval related to the first path received), the received signal in an environment containing clutter reflectors could be written as:

$$jT'_s \le t < jT'_s + T_s \quad : \quad y(t) = \sum_{k=0}^{N-1} H_{j,k} C_{j,k} \psi_{j,k}(t) \tag{9}$$

The propagation channel for the symbol j after the guard interval could be "summarized" with only one complex coefficient per transmitted frequency ($H_{j,k}$) as, during this portion of studied time, all the reflectors were illuminated by the signal $C_{j,k} \psi'_{j,k}(t)$ alone.

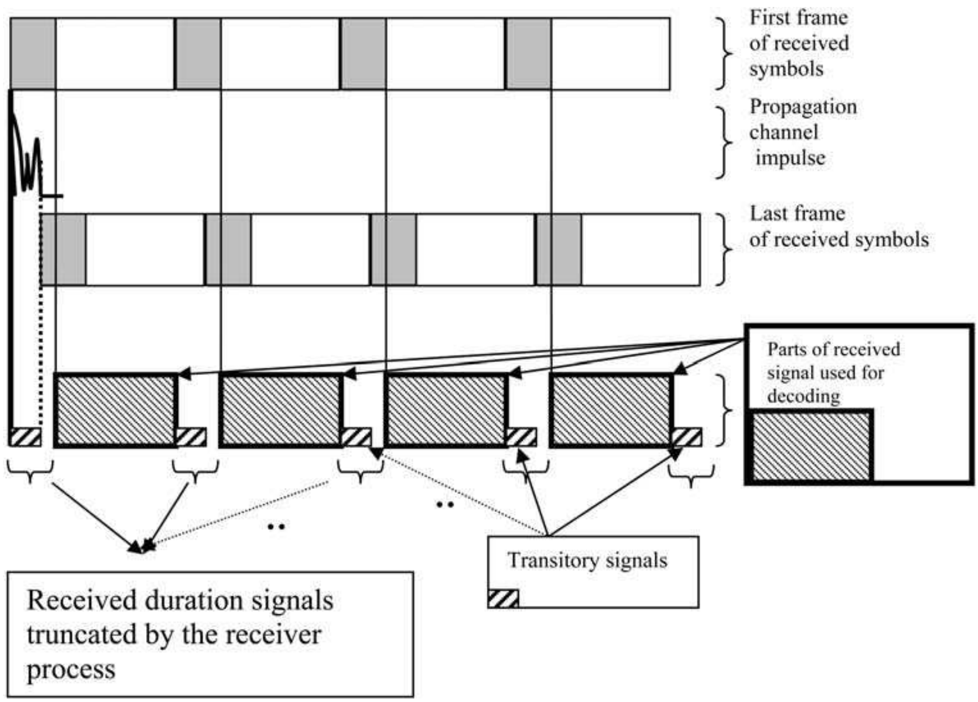

Fig. 1. Guard interval principle

Remark:

COFDM Waveform (with guard interval principle) can support superposition of different paths "without troubles". Such a property also allows a particular mode in a multiple transmitters configuration: all the transmitters can use simultaneously the same code and the same carrier frequency. This specific mode is called SFN (Single Frequency Network). In the rest of this chapter, there will be no difference considered between a multipath or a SFN transmitter. Furthermore, the propagation channel considered will include all the coherent paths, that means multipath on ground clutter as well as SFN transmitters.

2.3 Demodulation

The purpose of this paragraph is not to explain the demodulation principle well described in the DVB norm or in articles [Alard et al 1987 for example] for differential decoding when phase modulation is used.

Whatever considering optimal demodulation or differential one for phase codes, the decoding principle is based on estimating the transmitted codes using the received signal:

$$Y_{j,k} = H_{j,k} C_{j,k} + N_{j,k} \tag{10}$$

where $N_{j,k}$ represents a gaussian noise:

The knowledge of the channel impulse response $H_{j,k}$ and of the noise standard deviation $\sigma_{j,k}^2$ can be used for the coherent demodulation. This optimal demodulation consists in maximising over the $C_{j,k}$ the following relation:

$$\sum_j \sum_k \mathrm{Re}\left(Y_{j,k} H^*_{j,k} C^*_{j,k} / \sigma^2_{j,k}\right) \tag{11}$$

In order to simplify this demodulation, it's possible to perform differential demodulation instead of coherent demodulation for QPSK codes. This differential demodulation assumes propagation channel stationarity and consists in estimating the channel response from the previous symbol:

$$H_{j,k} \cong \frac{Y_{j-1,k}}{C_{j-1,k}} \tag{12}$$

This differential demodulation is particularly interesting for its simplicity. The 3 dB losses due to this assumption have to be compared to the practical difficulties encountered for the coherent demodulation implementation.
As a small comment, the differential demodulation doesn't estimate directly the elements of code $C_{j,k}$ but only the transitions between $C_{j-1,k}$ and $C_{j,k}$. However, for phase codes, like BPSK (or QPSK) the transition codes remains phase codes with two (or four) states of phase. In practice, such a differential demodulation just consists in Fourier transforms and some differential phase estimations (according to four possible states).
The most important conclusion dealing with these two possible demodulation principles is the following: using the received signal, it is possible to obtain and reconstruct an ideal vision of the transmitted one. In communication domain, this ideal signal is used for estimating the information broadcasted while for radar application this ideal signal will be used as a reference for correlation and could be also used for some cancellation process. For these radar applications, it is important to notice that this reference is a signal based on an ideal model. Furthermore, the decision achieved during the demodulation process has eliminated any target (mobile) contribution in this reference signal.

2.4 Synthesis

The COFDM signal has interesting properties for radar application such as:

- it is used for DAB and DVB European standard providing powerful transmitters of opportunity.
- the spectrum is a white spectrum of 1.5MHz bandwidth (1536 orthogonal sub-carriers of 1kHz bandwidth each) for DAB and 7.5 MHz for DVB-T
- the transmitted signal is easy to decode and reconstruct
- this modulation has interesting properties in presence of clutter : it is easy to consider and analyze only some parts of received signal without any transitory response due to multipaths effects.

3. Radar detection principle

3.1 Introduction

The principle of radar detection using DAB or DVB-T opportunistic transmitters will be classically based on the correlation of the received signals with a reference (match filter).
In the case of a transmitter using COFDM modulation, the estimation of the transmitted signal (reference) is easy to implement in order to ensure capabilities of range separation and estimation.

However, as the transmitted signal is continuous, we have to take a particular care of the ambiguity function side lobes for such a modulation. Firstly, we'll just verify that these side lobes related to the direct path (path between the transmitter and the receiver) are too high in order to allow efficient target detection and then we'll describe an adaptive filter whose purpose is to cancel all the main zero-Doppler path contributions and ensure efficient detection for mobile targets.

For limiting some specific correlation side lobes observable with the DVB-T signals, it is possible to consider the following article [Saini & Cherniakov 2005]: their analysis lead to a strong influence of the boosted pilot sub-carriers. The main suggestion of this article is to limit this influence by weighting these specific sub-carriers proportionally to the inverse of the „boosted level" of 4 over 3.

3.2 Radar equation example

In a first approach, that means excepting the specific boosted sub-carriers mentioned above for DVB-T, the COFDM modulation ambiguity side lobes can be considered as quite uniform (in range-Doppler domain) with a level, below the level of direct path, given by the following figure:

$$-10\log_{10}(MN) \tag{13}$$

where M designs the number of symbols (considered for correlation) and N the number of sinusoids broadcasted.

The next figure presents the exact ambiguity function (left part of the figure) for a COFDM signal with 100 symbols and 150 sinusoids per symbol, we can observe that the secondary lobes are roughly - 42 dB below the main path (except for low Doppler and range lower than the guard interval: here 75 kilometres). Under some assumptions (right part of the figure: Doppler rotation neglected inside one symbol)), we can consider, in some restricted range-Doppler domain (especially for range lower than the guard interval), a lower level of side-lobes. However, this improvement, related to an "optimal" use of the sub-carriers orthogonality, remains not enough efficient in an "operational" context so we'll don't discuss such considerations in this paper.

We'll just end this COFDM ambiguity function considerations by the following expression ($\phi(\tau,\nu)$ represents the (range, Doppler) ambiguity function).

$$\phi_{left}(\tau,\nu) = \int_{\mathrm{T_{integration}}} \mathrm{s}_{\mathrm{received}}(t) s^{*}_{reference}(t-\tau) e^{-i2\pi\nu t} dt \tag{14}$$

$$\phi_{right}(\tau,\nu) = \sum_{j=0}^{J-1} e^{-i2\pi\nu\left(j+\frac{1}{2}\right)T_s^{'}} \int_{jT_s^{'}+\Delta}^{(j+1)T_s^{'}} s_{received}(t)\, s^{*}_{reference}(t-\tau) dt \tag{15}$$

where the coherent integration time $T_{\mathrm{integration}}$ is equal to $T_{\mathrm{integration}} = J\left(T_s^{'}\right) = J(T_s + \Delta)$

The signal of reference is obtained using differential decoding principle.

The two previous expressions illustrate that the "right" correlation is equal to the "left" one under the assumption that Doppler influence is negligible inside each symbol duration. Furthermore, equation (15) illustrates that range correlations are just estimated over useful signal durations for which all the sub-carriers are orthogonal until the effective temporal support (function of the delay) remains exactly equal to useful duration T_S.

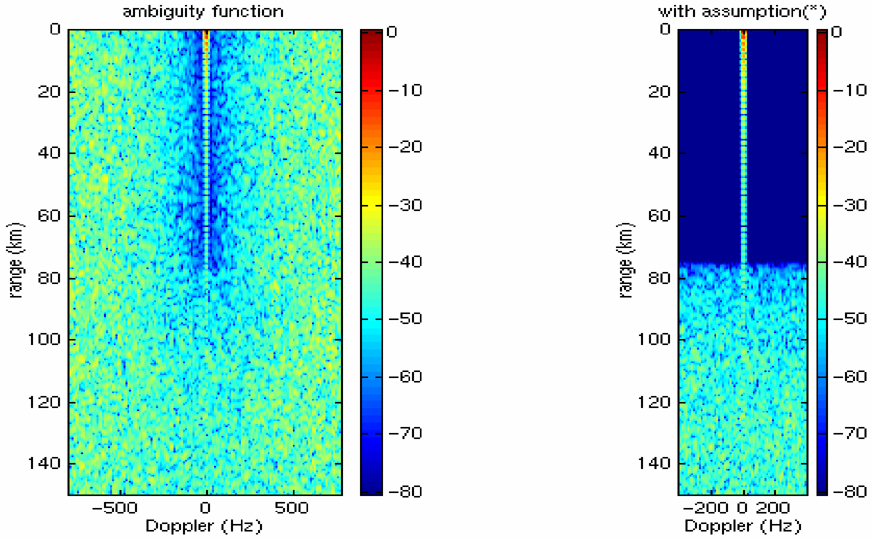

Fig. 2. COFDM ambiguity side lobes
(*) The Doppler rotation inside one symbol is neglected (right figure)

This property implies the lower level of side-lobes (visible on previous figure) for delays lower than guard interval length as using expression (15) there are no sub-carriers interferences in this range domain.
As our main purpose is to focus on the adaptive filter and not on radar equation parameters (coherent integration time, antenna gain and diagram,…), we'll don't discuss more in details on these radar equation parameters. We'll just consider: " as DAB or DVB-T waveforms are continuous, the received level of main path is always high and the isolation provided by side-lobes is not sufficient in order to allow detection."
As the side-lobes isolation (eq 13) is equal to the correlation gain (product between bandwidth and coherent integration time): when we receive a direct path with a positive signal to noise ratio (in the bandwidth of the signal), such a received signal allows reference estimation but its side-lobes will hide targets as these side lobes will have the same positive signal to noise ratio after compression (whatever coherent integration time we consider).
This phenomenon is schematically represented on next figure. Finally, observing this schematic radar equation, it's obvious that an efficient zero-Doppler cancellation filter is required as the targets are generally hidden by zero-Doppler paths side lobes.

3.3 Synthesis

This short description on radar principle had the only objective to prove the compulsory cancellation of the zero-Doppler paths in order to allow mobile target detection.
Only short overview on the correlation hypothesis and adjustments (for example for the boosted DVB-T pilots carriers) were given in order to be able to focus on the cancellation filter in the next part.

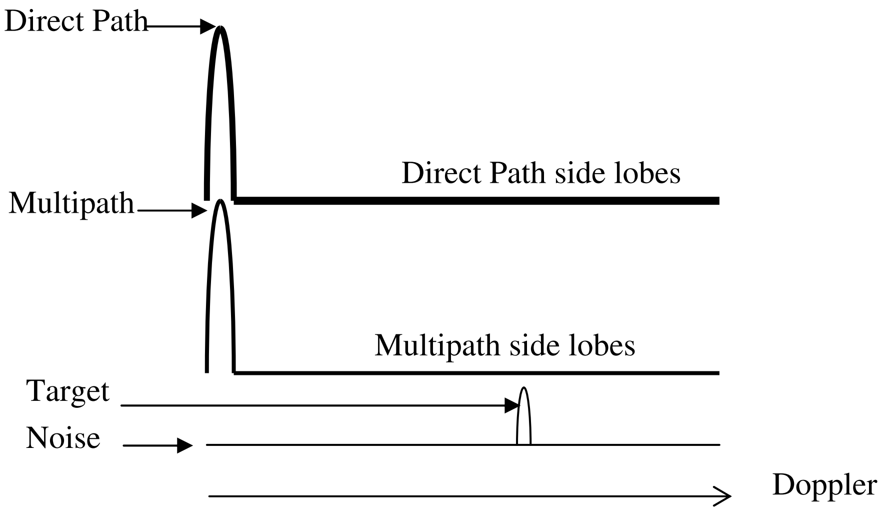

Fig. 3. Schematic radar equation (target hidden by side-lobes).

4. Detection principle

The purpose here is to present two approaches for the adaptive cancellation filter after a schematic description of the whole detection process.
The detection principle is divided into four main tasks described below:

- the first part consists in the transmitter parameters analysis (like carrier frequency, sampling frequency) and a "truncation" of the received signal in order to process only on stationary data
- the second part consists in estimating (by decoding) the reference signal that will be used for correlation
- the third part is more a diagnostic branch in order to allow a finest synchronisation for the direct path and consequently for the target echoes delays. This branch is also used for the propagation channel characterisation.
- The fourth part is related to the target detection and parameters estimation (Bistatic Doppler, bistatic range and azimuth).

This part dedicated to the target detection will be described in details in the following paragraphs.

5. Adaptive filter

5.1 Introduction

Before analyzing the filter itself, it seems important to remind the following elements.

- COFDM waveform allows the specific mode called SFN for which all the transmitters in a given area are broadcasting the same signal.
- From a global point of view, the level of COFDM side lobes is lower than the main path from the product (Bandwidth x integration time). As this product is also equal to the coherent gain over the integration time, a path with a positive signal to noise ratio in the

bandwidth of the signal (so typically most of the SFN transmitters direct paths) will have side lobes with the same positive signal to noise ratio after coherent integration.

Such considerations imply that the adaptive filter has to cancel efficiently all zero-Doppler contributors and not only the direct path. The two following filters considered here are fully adapted to the COFDM modulation and requires only a small array elements for the receiving system despite some other solutions sometimes developed [Coleman & Yardley 2008]. Furthermore, all the antennas (and related receivers) are used for the target analysis and detection: no additional hardware complexity and cost is added due to the zero-Doppler cancellation filters.

5.2 Adaptive filter principles

5.2.1 Cancellation filter using a receiving array

This first cancellation filter considers a small receiving array constituted by a set of typically four or eight receiving antennas: all these antennas will be used for the target analysis [Poullin 2001a]

Considering the signals over the different antennas of the receiver system, the zero-Doppler received signals for antenna i and symbol j (index k corresponds to the frequency) can be expressed as follows:

$$S_j^i = \sum_k H_{j,k}^i C_k^j \exp(j\, 2\pi k \frac{t_j}{T_s}) + N_{j,k}^i \tag{16}$$

$$\text{for } t_j \in \left[jT_s^{'} + T_O + L, (j+1)T_s^{'} + T_O \right]$$

with: $H_{j,k}^i$: complex coefficient characterizing the propagation channel for symbol j, antenna i and frequency k. We'll see an explicit expression of such a coefficient some lines below, this expression will consider a specific simple configuration.

T'_s : is the transmitted duration (per symbol)

T_o : corresponds to the first path time of arrival

L: designs the propagation channel length (delay between first path and last significant one including multipaths (echoes on the ground) as well as SFN paths).

$N_{j,k}^i$ designs the contribution of the noise (symbol j, antenna i and frequency k)

If the propagation channel length is lower than the guard interval, the previous expression will be valid for a duration longer than the useful one $T_s = T_S^{'} - \Delta$. So it will be possible to consider this expression over durations T_s for which all sub-carriers are orthogonal between each other.

Generally, we could consider stationary propagation channel over the whole duration of analysis (coherent integration time for radar) and so replace expression $H_{j,k}^i$ by H_k^i

Finally considering the received signals over:

- the appropriate signal durations: for each transmitted symbol over T'_s, we just keep signal over useful duration T_s. (defined by the first path received and the guard interval).
- the appropriate frequencies: over that specific durations, the composite received signals always verify the sub-carrier orthogonality conditions even in multipath (and SFN) configuration.
- the receiver antenna array.

It's possible to synthesise the propagation channel response over the receiver array with a set of vectors

$$\left\{(H_k^1,...H_k^i,...H_k^N) / k = 1,..,K : \text{frequency , i} = 1,...,\text{N: number of antennas}\right\} \tag{17}$$

where N is the number of elements in the receiver system.
So for each frequency k, it is possible to cancel the "directional vector" $\mathbf{H}_k = (H_k^1,...H_k^i,..H_k^N)^t$ using classical adaptive angular method based on covariance matrix as it can be seen below:
Considering for each frequency k the covariance matrix (with size related to the number of antenna) given by

$$R_k = E\left(\mathbf{H}_k\mathbf{H}_k^H C_k^j C_k^{j*} + \sigma_k^{j2} I\right) \tag{18}$$

$$\text{So } R_k = \mathbf{H}_k\mathbf{H}_k^H + \sigma_k^2 I \tag{19}$$

Consequently, when we'll apply the weightings related to the inverse of Rk for each frequency k, it appears weighting coefficients related to:

$$R_k^{-1} \approx \frac{1}{\sigma_k^2}\left(I - \frac{\mathbf{H}_k\mathbf{H}_k^H}{\mathbf{H}_k^H\mathbf{H}_k}\right) \tag{20}$$

which is the orthogonal projector to $\mathbf{H}_k = (H_k^1,...H_k^i,..H_k^N)^t$: propagation channel response vector at the sub-carrier k.
Remark:
This remark is just to give an explicit expression of a typical propagation channel response H_k^i (k: frequency, i antenna) in the particular case of two receiver antennas with a main path in the normal direction and a multipath characterized by its angle of arrival (θ). The normal path received on the first antenna is considered as reference. Under these hypothesis, the propagation channel responses could be written as:

$$\begin{aligned} H_k^1 &= (1+\alpha\exp(j\phi)\exp(-2\pi\, jf_k\,\tau)) \\ H_k^2 &= (1+\alpha\exp(j\phi)\exp(-2\pi\, jf_k\,\tau)\exp(j\,2\,\pi\, d_{12}\sin(\theta)/\lambda)) \end{aligned} \tag{21}$$

where d_{12} designs the distance between the two antennas and λ is the wavelength
$\alpha\exp(j\phi)$ represents the difference of reflectivity between main and multi-path (and τ is the delay between main path and multipath referred to antenna 1).
It is quite clear that H_k^1 and H_k^2 will quickly fluctuate according to frequency k due to the term $\exp(-j2\pi f_k\tau)$. Furthermore, for a given frequency the term $j2\pi d_{12}\sin(\theta)/\lambda$ implies different combinations of the two paths for the antenna.

5.2.1.1 *Example of cancellation efficiency on experimental data*

The filter implemented in order to cancel the zero-Doppler contributions was using four real antennas and the adaptive angular cancellation for each transmitted frequency as described previously. The transmitter was a DAB one and the correlation outputs in range-

Doppler already illustrate the cancellation capabilities that could be read on the cut along the Doppler axis for which the zero level reference corresponds to the receiver noise.

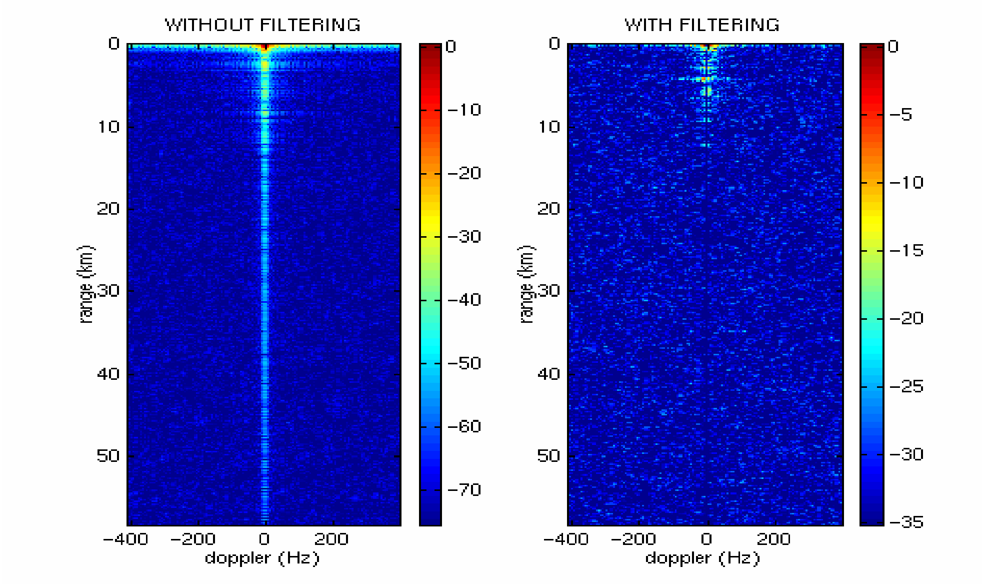

Fig. 4. Correlation output without cancellation filter (left) and with cancellation filter (right)

On the next figure, it could be seen that the level of the main path before cancellation had a signal to noise ratio (in the bandwidth of 10 Hz (related to the 100 milliseconds of integration) of 110 dB and the corresponding side lobes (for range lower than the guard interval which is equal to 75 kilometres) were still 35 dB above the noise level. After cancellation this residual spurious was only 7 dB above noise level. So the residual level of spurious could be considered as -103 dB below the main path

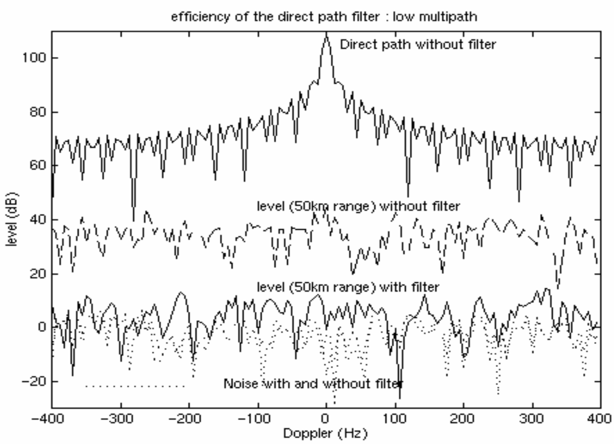

Fig. 5. Comparison of correlation outputs with and without cancellation filter .

5.2.1.2 *Specific case of filter efficiency*

The next figure corresponds to a target crossing the zero-Doppler axis. The trial configuration corresponds to the VHF-DAB transmitter analyzed just previously and six "snapshots" delayed from 1.5 seconds each are presented

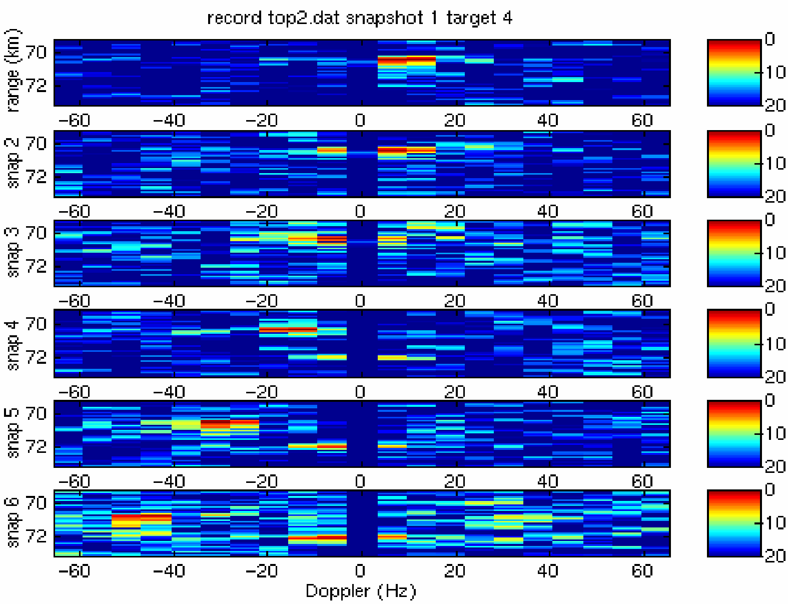

Fig. 6. Example of Results

This example of detection clearly shows the efficiency of the filter against multipath (reflector with null Doppler): when the target crosses the zero Doppler axis it is considered as an element of the clutter, so such a target is (during that time) fully coherent with clutter and main path and its contribution is integrated into the filter coefficients estimation, so such "mixed" coefficients reject both clutter and zero-Doppler target.

Such a result implies that even low multipath (clutter element whose signal to noise ratio is much lower than zero (dB) for the filter coefficient learning phase) could be filtered using such adaptive technique as "they are carried by the main path". From a schematic point of view, the filter detects a level of interference related to $(A + \Delta A)^2$ even if ΔA^2 is negligible with respect to the noise level (A correspond to the main path level and ΔA to the multipath).

5.2.1.3 *Example of target detections after filtering in SFN mode*

The first figure represents the output of the correlation filter (match filter) without zero-Doppler cancellation filter. Such a process allows the analysis of the main fixed echoes generally corresponding to the main transmitters in a SFN configuration.

The different transmitters are identified using "a priori" knowledge of the multi-static configuration while the multipath was located and identified using several receiver

locations and triangulation. During that experiment, receiver noise level was high: the receiving system used was an existing "generic" one and not a specific receiver defined for passive DAB application. The figure 9 is normalised according to this high receiver noise.
These results were obtained in a DAB-SFN configuration with numerous broadcasters and a two receiver antennas.

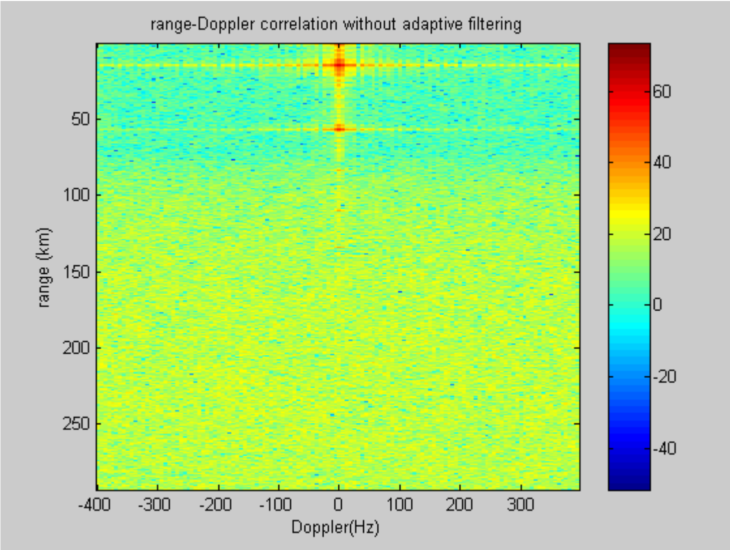

Fig. 7. Range Doppler correlation without zero-Doppler cancellation filter

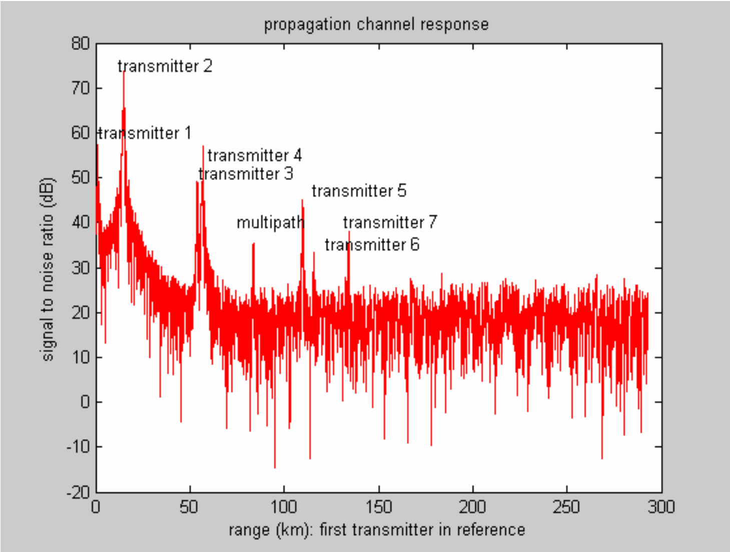

Fig. 8. Propagation channel response (analysis of correlation at zero Doppler: no filtering)

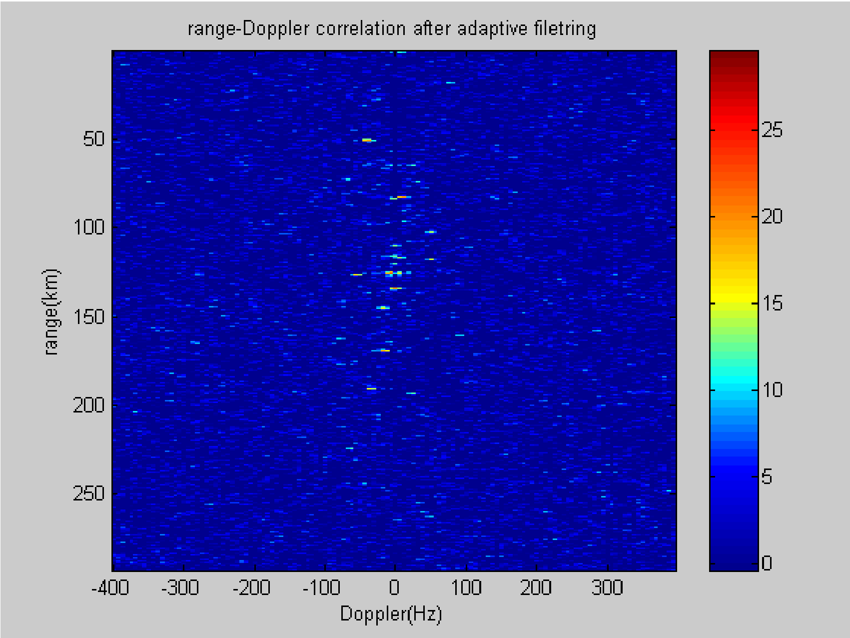

Fig. 9. Examples of mobile (non zero Dopplers) target detections after clutter cancellation

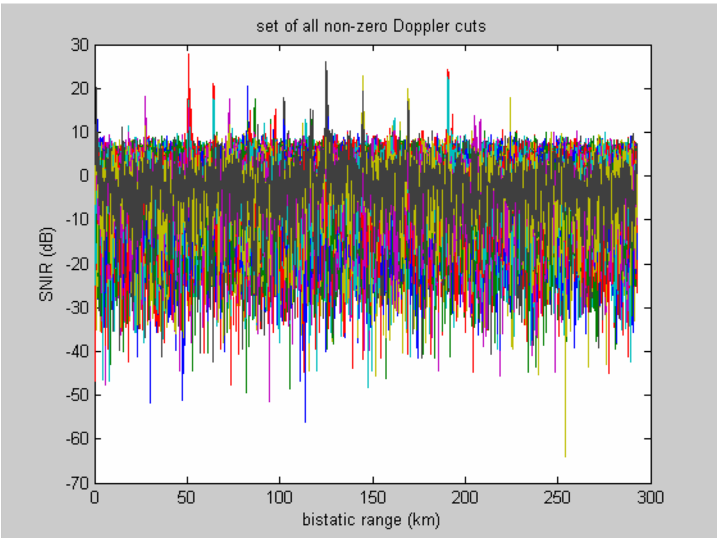

Fig. 10. Non-zero-Doppler cuts (of the range-Doppler correlation) after adaptive filtering

This figure corresponds to the previous multistatic situation with at least seven transmitters clearly identified on the propagation channel response. It becomes obvious that many mobile targets could be detected after zero-Doppler cancellation adapted to the COFDM-SFN configuration even using only two receiving antenna.

The superposition of all non-zero Doppler cuts is represented in order to give a "clear idea" of the detected targets (2-D images like the upper one with high number of pixels aren't always suitable for such a purpose). Furthermore, these cuts illustrate that after adaptive filtering the floor level corresponds to the (high) level of receiver noise.

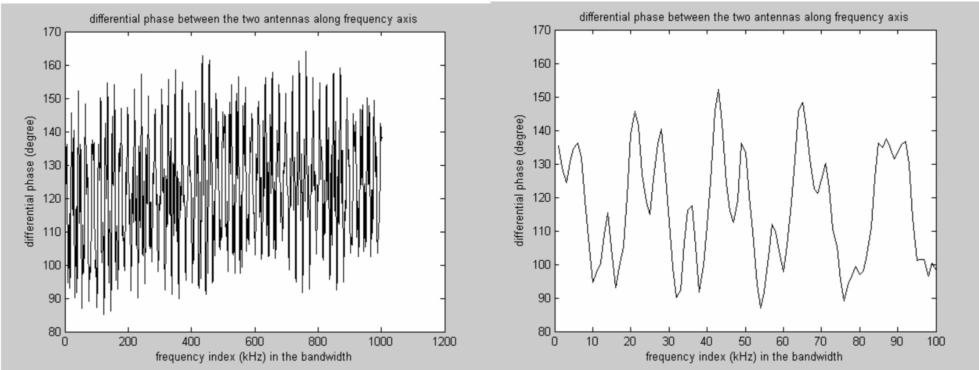

Fig. 11. Differential phase fluctuation between the two antennas along frequency axis.

Between the two distinct antenna, the important differential phase fluctuation along frequency axis in SFN mode (or high multipath configuration) is clearly illustrated on the previous figure.

These results show that, with COFDM modulation, it is possible to filter many SFN transmitters (or multipaths) with a small antenna array receiver. Nevertheless, the following principle: "bigger is your array, better are your results in terms of stability and narrow corrupted domain" remains true.

5.2.1.4 *Synthesis:*

In order to filter the clutter contributions (or the SFN transmitters), it is possible to consider the following algorithm described above involving time, frequency and angular domains:

- time domain: (cut of received guard intervals)
  this truncation ensure stationary durations for signal analysis with no time codes superposition
- frequency domain (analysis over the transmitted sub-carriers)
  The Fourier transform over the selected useful durations ensure signal analysis over stationary frequencies: no frequency codes superposition.
- "angular" domain: (adaptive beamforming for each frequency)
  the adaptive filter (for each transmitted sub-carrier) ensures clutter rejection. On that specific durations and frequencies, as all the clutter contributors are fully coherent, only one degree of freedom is necessary for adaptive cancellation of all the fixed echoes.

This filter has been successfully tested on real DAB signals and is currently tested using DVB-T broadcasters for which preliminary results seem encouraging.

This "angular" filtering applied for each transmitted sub-carrier will:

- lower all the zero-Doppler contributors as the set of H_k coefficients summarises all the clutter contributions.
- Be theoretically able to lower multiple transmitters using two antennas as only one degree of freedom is required for cancelling the zero-Doppler paths as long as the propagation channel length remains lower than the guard interval.

- Orthogonalise the received signals to a composite vector that doesn't correspond to a particular direction (see explicit expressions of H_k coefficient in equation (21)). This phenomenon is due to the full coherency of all the clutter contributors over that selected time durations and that frequency sub-carriers. This particularity also implies that only one degree of freedom (per frequency) is used for all clutter cancellation.
- Have to be applied for each transmitted frequency as the composite propagation channel vector fluctuates quickly in frequency domain. This fluctuation could be deduced from the explicit expression of H_k (equation 21) coefficient. Furthermore, this fluctuation was illustrated on an experimental example on figure 11.

The next paragraph will present another cancellation filter that will be less efficient in most of the situations. Nevertheless, its interest relies in the following capabilities: it requires only one real antenna in order to lower the different SFN contributions and it could be more efficient than the previous filter when the target is close to the composite directional vector of the zero-Doppler contributors.

5.2.2 Cancellation using a single antenna

This other zero-Doppler path cancellation filter could be obtained according two different approaches:

- An "angular" approach derived from the previous method but with the following adaptation: the cancellation is no longer achieved between several real antenna of the receiving array but between each real antenna and a sort of fictive one receiving the signal of reference obtained after decoding.
- A "temporal approach" using the classical Wiener Filter adapted to COFDM waveform.

We'll detail simply the "temporal" approach

Considering the decoded signal and a signal received on a real antenna it is possible to consider the Wiener filter:

$$z(t) = s_{received}(t) - \sum_{\tau=1}^{L} W(\tau) ref(t-\tau) \tag{22}$$

$$\text{with } \min_{w}\left(\left| s_{reçu}(t) - \sum_{\tau} w(\tau) ref(t-\tau) \right|^2\right)$$

Under that formulation, there is no specifity due to COFDM waveform and the similarity with the previous cancellation filter is not obvious.

So let us consider the spectral domain and the assumption that the length of the propagation channel and the corresponding Wiener filter length (here L) are lower than the guard interval.

So under that assumption, it is possible to consider the following expression

$$\sum_{\tau} W(\tau) ref(t-\tau) = \sum_{k=1}^{K} G_k C_k e^{j2\pi \frac{k}{T_u} t} \quad with \quad L_{channel} < \Delta \tag{23}$$

And as the signal received on one of the real antenna is:

$$S_{antenna\,1}(t_j) = \sum_{k=1}^{K} H_k^{antenna\,1} C_k^j e^{j2\pi \frac{k}{T_u} t_j} + \text{target}(t) + b(t) \tag{24}$$

It becomes clear that these two signals could be used to cancel the zero-Doppler paths using the same kind of algorithm than previously but with the important following modification:

- The previous adaptive angular filter was using only real antenna. All these antenna were containing the moving targets contribution as well as some "common" imperfections on the received signal.
- The new suggested filter is involving one real antenna with the target contributions and the signal of reference which is an ideal one. Furthermore, this ideal signal doesn't contain the targets contributions.

5.2.3 Comparison of the two filters

5.2.3.1 *Introduction*

The previous comments dealing with the main differences in the signals used at the input of the two zero-Doppler cancellation filters allow us to have the following observations:

- The filter involving only real antenna:
  - Requires several receiving antenna to cancel the zero-Doppler paths. Nevertheless a few antenna system could be sufficient as only one degree of freedom is required for cancelling all the zero-Doppler paths below the guard interval.
  - will be more robust to the imperfections as all the antennas suffer from that nuisances and the cancellation filter will be able (at least in a first order) to deal with most of these troubles
  - will have potential influence (and losses) on the targets contributions.
- The filter involving each real antenna and the signal of reference
  - Could be implemented using only one real receiving antenna
  - Will be more sensitive to all the defaults affecting the received signal according to the ideal model used for estimating the reference
  - Will have no influence on the targets contributions as these targets are no longer in the signal of reference obtained after decoding the received one.

In other words, the limitation of these two filters are not identical and it is quite obvious, that the filter involving only real antenna will be more efficient in terms of zero-Doppler cancellation. It is also evident that the consequences on some targets will be higher than with the filter involving real antenna mixed with the reference signal.

5.2.3.2 *Example of comparison on experimental data*

On the left figure, the adaptive filter was able to cancel efficiently the zero-Doppler paths but the losses on the target were too high due to the vicinity between the target directional vector and the one characterizing the zero-Doppler paths.

On the right figure, the adaptive filter involving one real antenna and the reference signal was also able to cancel efficiently the zero-Doppler path without destructive effect on the target. The small image of the target called ghost (fantôme) was due to a required correction in order to adapt the reference signal model to some imperfections occurring in the receiver system.

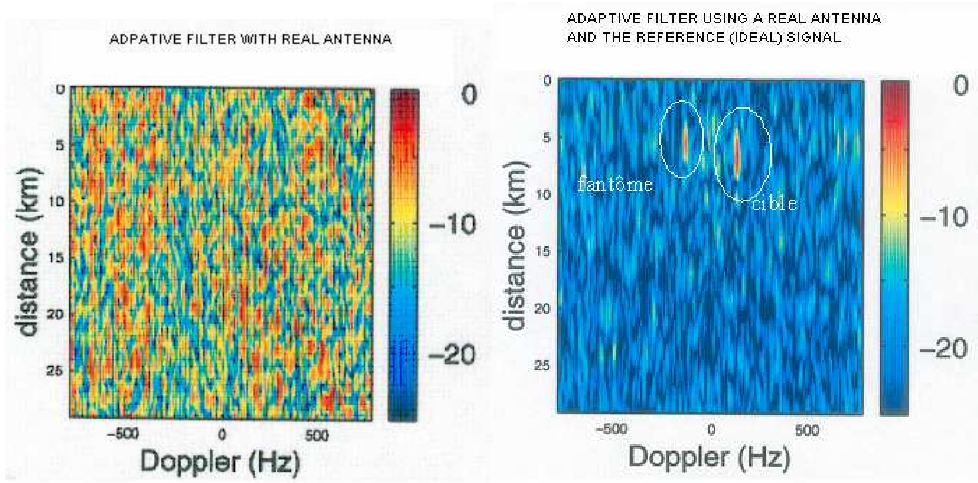

Fig. 12. Example for which the losses on the targets were too high with the adaptive filter involving only real antennas

Nevertheless, generally the method using the reference signal could not be as efficient (in terms of cancellation) as the one involving only real antenna.

5.2.3.3 *Example of limitations due to carrier frequency errors.*

This short paragraph is just to illustrate the higher sensitivity of the cancellation filter involving the ideal reference signal to one of the possible misfits between the received signal and this "ideal reference". This illustration will consider a non-corrected error of frequency between the receiver and the transmitter which corresponds to an error between the received signal and the ideal reconstructed reference.

In such a situation, it is possible to consider that this frequency error will lead, for the filter using the reference signal, to an additional interference due to the superposition of the different sinus cardinal functions:

The other filter, involving only real antenna, is less sensitive to such an error as it is the same error on all the antenna used for cancellation.

$$RSI_j = \frac{\left(H_j C_j\right)^2}{\left(\sum_{\substack{k=1 \\ k \neq j}}^{K} \left(H_k C_k \frac{\sin((k-j)\pi + \pi \Delta \nu T)}{(k-j)\pi + \pi \Delta \nu T} \right) e^{j\left(\pi(k-j) + \pi \Delta \nu T\right)} \right)^2} \qquad (25)$$

If we consider small errors on this frequency carrier, the expression above could be simplified using:

$$I_j = \left(\sum_{\substack{k=1 \\ k \neq j}}^{K} \left(H_k C_k \frac{\sin((k-j)\pi + \pi\Delta\nu T)}{(k-j)\pi + \pi\Delta\nu T} \right) e^{j(\pi(k-j)+\pi\Delta\nu T)} \right)^2$$
$$I_j \approx H^2 \left(\sum_{\substack{k=1 \\ k \neq j}}^{K} \left(C_k \frac{(-1)^{k-j} \Delta\nu T}{(k-j)} \right) \right)^2 \tag{26}$$

considering the average power of that perturbation:

$$H^2 C^2 \Delta\nu^2 T^2 \frac{\pi^2}{6} \leq E\left[I_j \right] \leq 2 H^2 C^2 \Delta\nu^2 T^2 \frac{\pi^2}{6} \tag{27}$$

Finally

$$\frac{1}{2\Delta\nu^2 T^2 \frac{\pi^2}{6}} \leq RSI_j \leq \frac{1}{\Delta\nu^2 T^2 \frac{\pi^2}{6}} \tag{28}$$

The following figure illustrates the influence of an error of 80 Hertz on simulated data. According to the level of the main path (80 dB including the gain of 50 dB for coherent integration time) considered and the useful duration time of 1 millisecond, the troubles due to a misfit between the transmitter frequency and the receiver one become to occur at 20 Hz.

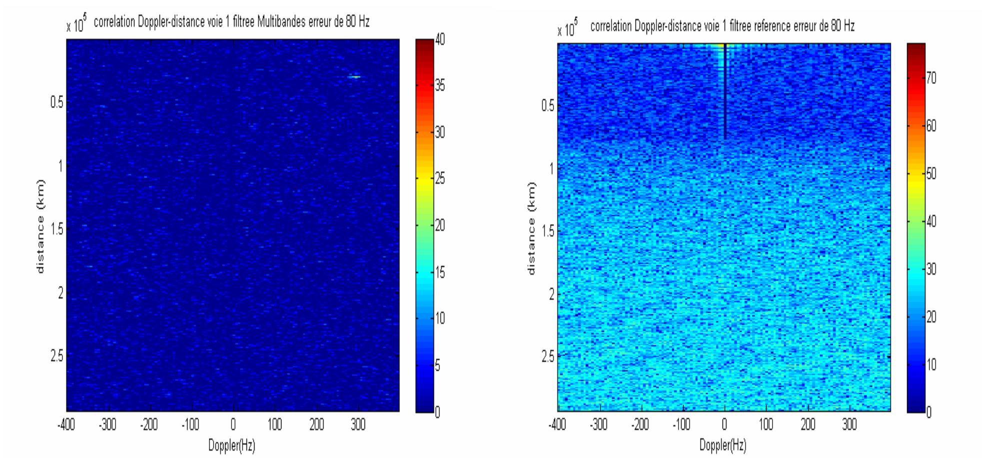

Fig. 13. Correlation output for the two cancellation filter described considered a 80 Hz error between transmitter and receiver (filter with real antenna only: left, filter with real antenna and ideal signal: right)

As illustrated on the following figures, the influence of such an error becomes to occur (according to our simulation parameters) at 20 Hz

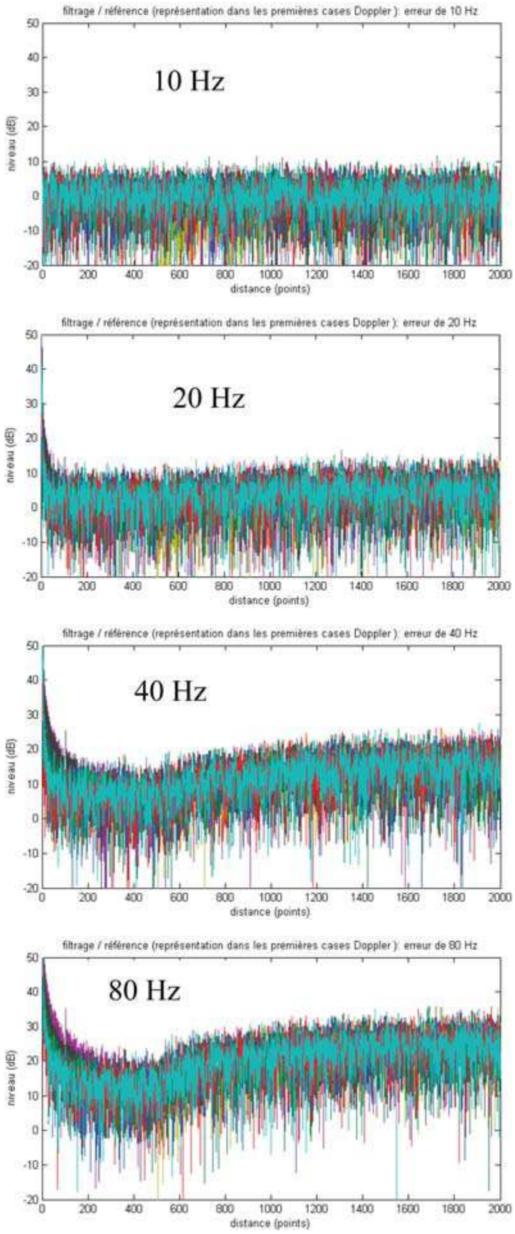

Fig. 14. Analysis of the frequency errors over correlation cut for the cancellation filter involving real antenna and ideal signal.

Of course, it is still possible to define and correct such an error, this example was just an illustration of the higher sensitivity of the second filter to the misfits. Nevertheless, this higher sensitivity will remain even for other kind of interferences that couldn't be corrected as easily as the frequency error...

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7. Conclusion

The COFDM waveform has a great robustness against propagation effects as according to some basic operations (synchronisation and truncation), under the hypothesis of a propagation channel length lower than the guard interval, it is still possible to analyse the received signals over the orthogonal basis of the transmitted sub-carriers even in an environment with numerous reflections.

Using such COFDM civilian broadcasters like DAB or DVB-T as opportunity transmitters for radar application leads to implement a compulsory efficient cancellation filter in order to remove all the main fixed (zero-Doppler) contributors and their corresponding multipaths. Such application is known as Passive Coherent Location: PCL.

Two specific cancellation filters were described in this chapter and illustrated on real data. Their main characteristics are the following:

- The two filters are using the properties of the COFDM modulation in order to "optimise" their efficiencies
- Under the assumption of a propagation channel length lower than the guard interval, only few antenna are necessary in order to lower all the fixed contributors as only one degree of freedom is required for such a cancellation.
- The first method requires a small receiving array (typically 4 or 8 antenna) while the second method could be applied even with one antenna but it implies a higher sensitivity to errors and misfits between the receiving signal and the ideal reconstructed reference
- In practice, these two methods can be complementary as the first one is more efficient for cancelling zero-Doppler paths but it could lower also the targets while the second one is less efficient (due to its higher sensitivity) from the cancellation consideration but it has no destructive effects on the targets.

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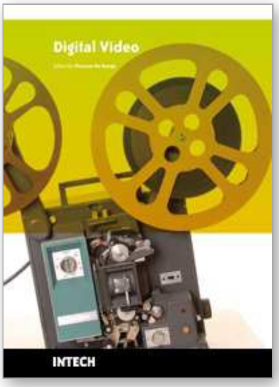

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This book tries to address different aspects and issues related to video and multimedia distribution over the heterogeneous environment considering broadband satellite networks and general wireless systems where wireless communications and conditions can pose serious problems to the efficient and reliable delivery of content. Specific chapters of the book relate to different research topics covering the architectural aspects of the most famous DVB standard (DVB-T, DVB-S/S2, DVB-H etc.), the protocol aspects and the transmission techniques making use of MIMO, hierarchical modulation and lossy compression. In addition, research issues related to the application layer and to the content semantic, organization and research on the web have also been addressed in order to give a complete view of the problems. The network technologies used in the book are mainly broadband wireless and satellite networks. The book can be read by intermediate students, researchers, engineers or people with some knowledge or specialization in network topics.

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