Analysis and Characterization of a Monobit Receiver for Electronic Warfare

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Detection of simultaneous signals and real-time operation are basic characteristics in broadband electronic warfare (EW) receivers. A digital channelized monobit receiver represents an attempt to achieve both features at the expense of low instantaneous dynamic range. This paper presents a detailed theoretical and experimental analysis, and characterization of the performance of this promising receiver composed of a monobit analog-to-digital converter (ADC) and a filter bank based on a monobit implementation of the discrete Fourier transform (DFT).

I. INTRODUCTION

The proliferation of electronic signals in modern combat environments requires the use of sophisticated electronic warfare (EW) receivers. Desirable characteristics of EW receivers include wideband frequency coverage, high sensitivity and dynamic range, high probability of intercept, simultaneous signal detection, frequency resolution, and full real-time operation. A classic receiver which accomplishes these requirements is a channelized receiver [1] which separates signals according to their frequencies.

Advancements in analog-to-digital converters (ADC) technology and in the speed of digital processors have made it possible to design relatively wideband digital channelized receivers. The use of digital channelization in comparison with the analog approach allows to improve the imbalances between filters, which is one of the fundamental problems in analog receivers.

However, broadband digital channelized receivers, mainly based on discrete Fourier transform (DFT) related processing, are computation intensive and yet not suitable for real-time applications in spite of the revolution of DSPs and FPGAs speed. In an attempt to improve the real-time operation, parallel processing can be considered. Another possibility is the reduction of the computational complexity of the signal processing algorithms by the simplification of the operations, e.g., avoiding complex multiplications in the calculations.

This is the philosophy in the monobit channelized receiver described in several US patents [2, 3] and papers [4]. As pointed out in [4], there are two possibilities in order to avoid multiplications in the calculation of the DFT: a single-bit digital representation of the input signal [2], which is equivalent to the use of a hard limiter, or a monobit representation of the kernel of the DFT [3]. Both schemes are possible, and it is even possible to use both in the same processing algorithm.

The optimum scheme in terms of number of operations for the DFT is the fast Fourier transform (FFT). An FFT algorithm without multiplications is only possible with a monobit kernel. The monobit representation of the input signal also allows a reduction of the computational complexity by decreasing the required number of bits in each adder.

The work presented here focuses on the theoretical and experimental evaluation of the performance, capabilities, and limitations of this receiver for the detection of multiple signals for radar applications. In Section II we describe the system to be analyzed.
Section III is devoted to studying in detail the monobit DFT and FFT algorithms. Section IV and V present the characterization of this system in terms of false alarm probability and detection probability. Dynamic range of the monobit receiver and its influence on the detection of multiple signals are discussed in Section VI. Section VII presents the experimental performance of this receiver for a continuous wave-linear frequency modulation (CW-LFM) signal. The interest of this signal is due to its inherent low probability of interception (LPI) characteristic and its common use in radar systems. Some conclusions are drawn in Section VIII.

II. DESCRIPTION OF THE SYSTEM

The system considered here is depicted in Fig. 1. The RF front end is not included. The receiver uses a one-bit ADC followed by a filter bank represented by a monobit DFT or FFT. Finally, a decision is made using the module of the different outputs of the DFT or FFT. The one-bit ADC is the digital version of an ideal limiting amplifier. It is named as a hard limiter in the literature. Thus, it is not necessary to use a limiting amplifier in the RF front end to limit the input signal to the ADC to a constant level.2 In the following sections different important characteristics of the behavior of this receiver are analyzed.

A lab prototype of this monobit receiver was implemented by using a commercial module composed of an ADC SMT320 of 12 bits (we use only the sign bit) and a DSP from TI: TMS320C40. The experimental set-up is depicted in Fig. 2.

III. MONOBIT DFT AND FFT

The concept of monobit DFT can lead to different implementations where the kernel function \( e^{j\phi} \) is rounded to 1, -1, j or -j through the function \( G(e^{j\phi}) \). The following function is used here [4]:

\[
G(e^{j\phi}) = \begin{cases} 
1 & \text{if } -\frac{\pi}{4} \leq \phi < \frac{\pi}{4} \\
 j & \text{if } \frac{\pi}{4} \leq \phi < \frac{3\pi}{4} \\
-1 & \text{if } \frac{3\pi}{4} \leq \phi < \frac{5\pi}{4} \\
-j & \text{if } \frac{5\pi}{4} \leq \phi < \frac{3\pi}{4}
\end{cases}
\]  

(1)

Due to the use of (1), the property of the \( N \)-point DFT that assured that if the input sequence was real-valued the output sequence verified \( X^*(k) = X(N-k) \) does not apply any more, as it is not always true that \( G(e^{j\phi}) = G^*(e^{-j\phi}) \).

Three possibilities of implementing the monobit DFT have been considered, which lead to different filter banks.

1) Replace directly the original kernel function with \( G(\cdot) \). This implementation is called monobit DFT. Using the definition of the \( N \)-point DFT:

\[
X(k) = \sum_{n=0}^{N-1} x(n)e^{-j(2\pi k/N)n}.
\]  

(2)

The implementation of the monobit \( N \)-point DFT is

\[
X(k) = \sum_{n=0}^{N-1} x(n)G(e^{-j(2\pi k/N)n}).
\]  

(3)

This is the most direct and the slowest implementation.

2) Implement the DFT using the decimation in time FFT algorithm [5], and replace the coefficients \( e^{j\phi} \) by \( G(e^{j\phi}) \). It is called monobit FFT by decimation in time.

3) Implement the DFT using the decimation in frequency FFT algorithm [5], and replace the coefficients \( e^{j\phi} \) by \( G(e^{j\phi}) \). It is called monobit FFT by decimation in frequency.

Both decimation algorithms are based on replacing the coefficients \( W_n^k = e^{-j(2\pi k/N)} \) by their monobit representation \( G(e^{-j(2\pi k/N)}) \) in the branches of the typical butterfly computation of these algorithms [5].

The use of the function \( G(\cdot) \) modifies the coefficients of the filters in the three implementations, and so their frequency responses. However, the filters obtained from both monobit FFT implementations are nearly equal. Figs. 3 and 4 compare the frequency responses for the filters of channels 3 and 4, respectively, implemented by the original 32-point DFT and a monobit 32-point FFT. The average of the highest sidelobe level is 9–10 dB below the mainlobe independent of the implementation and the number of filters in the filter bank. For filters \( k = i \cdot N/8 \)

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2D. S. Pek, et al. [4] implemented a monobit receiver using a 2-bit ADC. The limiting amplifier was necessary to provide a constant level at the input of the ADC.
Fig. 3. Filter response for channel 3 from original 32-point DFT and from monobit 32-point FFT (- -).

(i = 1, 3, 5, . . .) the difference between the mainlobe and the highest sidelobe has the minimum value for any implementation: 7.5 dB.

Windowing cannot improve this result, as can be seen in Fig. 5 where a Hamming window has been used. The impulse response for the k output of the original N-point DFT with window w(n) can be expressed as

\[ h_k(p) = w(N - p) e^{j(2\pi k/N)p}. \]  

(4)

The impulse response for filter k and a monobit N-point DFT is

\[ h_k(p) = w(N - p) G(e^{j(2\pi k/N)p}). \]  

(5)

w(·) defines the filter in base band and G(·) is responsible for the translation of the spectrum of w(·) to the frequency \(2\pi k/N\), and the modification of the sidelobes. This result is independent of w(·). Identical considerations apply for FFT monobit implementations.

It has been found that there are always two different kinds of channels for the different implementations of the monobit DFT or FFT. They are defined as follows.

Type 1 Channels. The kernel equals 1 or -1 for both the original and the monobit implementations. As the input signal is real valued, the output of this kind of channels is also real. The only channels of this kind are channel 0 (DC component) and channel N/2.
Fig. 5. Filter response for channel 4 from original 32-point DFT and from monobit 32-point FFT (---). Hamming window used in both cases.

Fig. 6. Relative losses for monobit 64-point DFT.

(high-frequency component). Quantization error due to the function $G(\cdot)$ is zero as the coefficients of the filters have not changed.

**Type 2 Channels.** The coefficients of the rest of the channels verify that half of them are real valued (1 or $-1$) and half of them are imaginary ($i$ or $-i$). Channels $N/4$ and $3N/4$ also verify that their coefficients have not changed due to the use of $G(\cdot)$. The output of a type 2 channel is a complex number.

Figs. 6 and 7 show the characteristics of the filter bank synthesized through the monobit 64-point DFT and the monobit 64-point FFT by decimation in time, respectively. The monobit 64-point FFT by decimation in frequency looks like very much the monobit 64-point FFT by decimation in time. These figures represent the relative losses for the mainlobe of the different filters of the filter bank. The normalization factor is the maximum amplitude response of one of the filters of the original DFT.

These figures demonstrate that the filters of the filter bank have different maximum gains, and that these gains vary with the implementation of the filter bank. Table I summarizes the range of variation for the losses at the maximum of the filters. Filters $k = i \cdot N/4$ ($i = 0, \ldots, 3$) are not considered. Besides, it is necessary to point out that the frequency for the
Fig. 7. Relative losses for monobit 64-point FFT by decimation in time.

TABLE I

Range of Variation of Relative Losses at Maximum of Filters for Different Monobit Implementations

<table>
<thead>
<tr>
<th>Implementation</th>
<th>Length</th>
<th>Relative Losses</th>
</tr>
</thead>
<tbody>
<tr>
<td>DFT</td>
<td>64</td>
<td>0.69–0.91</td>
</tr>
<tr>
<td></td>
<td>128</td>
<td>0.69–0.91</td>
</tr>
<tr>
<td></td>
<td>256</td>
<td>0.69–0.91</td>
</tr>
<tr>
<td>FFT decimation in time</td>
<td>64</td>
<td>0.69–1.47</td>
</tr>
<tr>
<td></td>
<td>128</td>
<td>0.69–1.79</td>
</tr>
<tr>
<td></td>
<td>256</td>
<td>0.69–2.09</td>
</tr>
<tr>
<td>FFT decimation in frequency</td>
<td>64</td>
<td>0.69–1.47</td>
</tr>
<tr>
<td></td>
<td>128</td>
<td>0.69–1.81</td>
</tr>
<tr>
<td></td>
<td>256</td>
<td>0.69–2.09</td>
</tr>
</tbody>
</table>

maximum of the filters is slightly displaced compared with implementation based on the original DFT.

IV. FALSE ALARM PROBABILITY

The 1-bit ADC fixes the power at its output independently of the input power. Therefore, the false alarm probability is independent of the input power noise. As it was shown in the previous section, there are two different kinds of channels in the receiver. Their behavior is analyzed separately.

In a type 1 channel, the filter coefficients are all equal to 1 or −1. Besides, the samples of the input signal are also 1 or −1. After multiplying them, a vector with \( k_+ \) elements equal to 1 and \( k_- \) elements equal to −1 is obtained. The sum of these elements renders \( k = k_+ - k_- \), which is always an even number between \(-N\) and \(N\) for a monobit \(N\)-point DFT (FFT) with \(N = 2^b\), \(b\) a natural number. Besides, \(k_+ = N + k/2\), \(k_- = N - k/2\).

After some reasoning based on combinatorial theory, the analytical expression for the probability of getting an output \(k\) can be computed

\[
P(k) = \binom{N}{N-k} \cdot \frac{1}{2^N} = \binom{N}{N+k} \cdot \frac{1}{2^N}.
\] (6)

After a linear detector at the output of each filter, an even value \(k' = |k|\) with values \(0 \leq k' \leq N\) is obtained. The probability of getting a specific \(k'\) is the sum of the probabilities of getting \(k = k'\) and \(k = -k'\) except for \(k' = 0\). In short

\[
P(k') = \begin{cases} \binom{N}{N-k'} \cdot \frac{1}{2^N} & \text{if } k' = 0 \\ \binom{N}{N-k'} \cdot \frac{2}{2^N} & \text{if } 0 < k' \leq N \end{cases}.
\] (7)

If a threshold \(T_h\) is chosen at the output of the linear detector, the false alarm probability is obtained as

\[
P_{fa}(T_h) = \sum_{k \geq T_h} P(k')
\] (8)

being \(P_{fa}\) a staircase function as it is clear in Fig. 8.

In a type 2 channel the output is a complex number. Its real and imaginary parts are both even numbers within the range \([-N/2, N/2]\). Following the same steps used for a type 1 channel, the probabilities of getting the real part of the complex number equal to \(k_r\) and its imaginary part equal to \(k_i\) are

\[
P_r(k_r) = \binom{N}{2} \cdot \frac{1}{2^{N/2}} = \binom{N}{2} \cdot \frac{1}{2^{N/2}}
\] (9)

and

\[
P_i(k_i) = \binom{N}{2} \cdot \frac{2}{2^{N/2}} = \binom{N}{2} \cdot \frac{2}{2^{N/2}}.
\]
Fig. 8. False alarm probability for both type 1 channels and type 2 channels and a 256-point monobit DFT (FFT). Simulated $P_{\text{fas}}$ down to $10^{-5}$ are obtained by Monte Carlo with $10^7$ independent trials.

Besides, if there is only white noise at the input of the system, $k_i$ and $k_r$ are independent and the probability of obtaining the complex number $k_r + j \cdot k_i$ is

$$P(k_r + j \cdot k_i) = \left(\frac{N}{2}\right) \left(\frac{N}{2} - k_r\right) \frac{1}{2^N}. \quad (10)$$

The output of these channels is converted to a real number using a linear detector, and the false alarm probability for a fixed threshold $T_h$ can be computed as

$$P_{\text{fa}} = \sum_{|k_r + j \cdot k_i| \geq T_h} P(k_r + j \cdot k_i). \quad (11)$$

This mathematical expression can be applied to any channel of the three implementations as long as it is a type 2 channel: the false alarm probability remains constant independently of the maximum amplitude response of the specific channel. The demonstration is in Appendix A. As an example, theoretical and simulated false alarm probability as a function of the threshold for a monobit 256-point FFT (FFT) is shown in Fig. 8. The simulations are obtained by Monte Carlo with $10^7$ independent trials to simulate $P_{\text{fas}}$ down to $10^{-5}$. This figure makes clear the following facts.

1) Both functions ($P_{\text{fa}}$ for type 1 and type 2 channels) are staircase functions for any of the three implementations.

2) The number of different values of false alarm probability available is sensibly lower in a type 1 channel than in a type 2 channel.

3) The false alarm probability in type 1 channels is much higher than in type 2 ones for a fixed $T_h$. Therefore, the overall false alarm probability of the DFT (FFT)-based filter bank is higher than in type 2 channels. Type 1 channels correspond to the extreme frequencies of the band coverage of the digital receiver.

4) If a threshold $T_h > N$ for a type 1 channel or $T_h > N/\sqrt{2}$ for a type 2 channel is chosen, both the false alarm probability and detection probability are 0 because there are not any outputs with that magnitude.

5) If $T_{hN}$ is the threshold for a given $P_{\text{fa}}$ using a monobit $N$-point DFT (FFT), the threshold, $T_{hN'}$, for a monobit $N'$-point DFT (FFT) ($N' = 2^nN$) is $T_{hN'} = 2^{n/2}T_{hN}$.

V. DETECTION PROBABILITY

A figure of merit of this system regarding detection capabilities can be its losses compared with a system without digitalization and with the original DFT for a fixed detection probability and a given false alarm probability. Fig. 9 presents this figure of merit for $P_d = 90\%$ and different $P_{\text{fas}}$, different lengths for the monobit DFT, and centered sinusoids. Each point of the figure was calculated using Monte Carlo simulation with 5000 independent trials. These losses represent a reduction of the processing gain of the original $N$-point DFT: $G(dB) = 10\log N/2$.

The detection capabilities can change depending on the implementation of the filter bank and the frequency of the sinusoidal signal to be detected.

1) For centered sinusoidal signals, all channels of the monobit DFT have nearly the same response. Only channel $N/4$ has a better response, Fig. 6, with an additional gain of nearly 1 dB with respect to Fig. 9. This filter is equal to the ones obtained with the original DFT. A monobit FFT has different losses depending on the selected filter, Fig. 7. For example,
additional losses of 0.5 (1) dB appear for a 64 (256)-point monobit FFT, Table I.

2) Additional losses must be considered when the input signal is not at the central frequency of a filter. The impact on the detection probability is not simply an increase in the signal power at the input of the system to compensate the attenuation of the filter. The reason is the nonlinearity of the 1-bit ADC, which fixes the power at its output independently of the input power. As a consequence, a signal between two adjacent filters may not be detected for a fixed threshold because the power is spread between these filters, independently of the input power or the signal-to-noise ratio (SNR).

The simulated detection probabilities for sinusoidal signals with random frequencies within the bandwidth of each filter are shown in Figs. 10 and 11 for a $P_{fa} = 10^{-6}$ for $N = 64$ and $N = 128$, respectively. The dispersion of the traces is originated by the different amplitude response of the filters. The most important result of these figures is the impossibility to obtain simultaneously a mean detection probability per channel of 90% for a fixed $P_{fa} = 10^{-6}$ by employing a monobit 64-point DFT-FFT. These curves are
Fig. 11. Detection probability for sinusoidal signals with random frequencies in band of each channel of 128-point monobit FFT. Each trace represents a channel. $P_{fa} = 10^{-6}$. Curves are compared with average response of sinusoidal signals centered in different channels (Centered frequency) of 128-point monobit FFT. 1000 independent trials used for results of this figure.

compared with the average detection probability for sinusoidal signals centered in different channels (label Centered frequency in these figures). It is clear that a monobit 128-point DFT-FFT can overcome this drawback, Fig. 11.

VI. MONOBIT RECEIVER PERFORMANCE

One of the most important requirements for a channelized receiver is the detection of simultaneous signals. Due to the high nonlinearity of the 1 bit ADC, there will be capture effect [6] and reduction of the instantaneous dynamic range (the ability to process concurrent signals of different amplitude [7]). Therefore, a detailed study of the dynamic range with a single signal and with two simultaneous signals (with the same power or with different power) has been performed. The study presented below will show that this receiver has a limited capability for simultaneous detection of several signals.

First of all, it is necessary to clarify the origin of the spurious responses in this receiver. The concept of digital channelized receiver is modified in the following two important aspects.

1) The one-bit ADC or hard limiter is an odd nonlinearity, and, therefore, odd harmonics of the input signal will be generated.

2) The synthesized filters through the monobit DFT(FFT) have sidelobes whose positions and amplitude response differ from the original DFT. In fact, the highest spurious sidelobe level can be only 7.5 dB below the mainlobe level; see Section III.

It is possible to predict the channels where the spurious can appear once the signal with the maximum power is detected, and to eliminate the spurious response by blanking: elimination of the odd harmonics of the detected input signals and the possible signals that appear due to the sidelobes of the filters. However, the main objective of blanking is not to improve the dynamic ranges (only the single signal dynamic range can be improved with the spurious signal prediction), but to reduce the probability that spurious signals can be detected and identified as real signals present in the analyzed environment (false targets). On the other hand, the procedure for spurious determination and blanking of the spectrum works properly for narrowband signals. However, if wideband signals are present in the environment, blanking will eliminate important parts of the spectrum with the corresponding loss of detection capability.

The main conclusions obtained after a detailed study on the performance of this receiver can be summed up briefly in the following.

Single Signal Test. Spurious responses generated by the nonlinearity of the ADC can be detected with a significant detection probability; see Fig. 12. Symbols (+) represent experimental results obtained with the experimental set-up, Section II. Depending on the frequency of the input signal, the influence of the spurious response can be negligible in terms of detection probability, Fig. 13.

The false alarm probability for the channels where neither the original frequency nor the spurious responses appear decreases with the power of the input signal due to the capture effect, Figs. 12 and 13.

Two-Signal Test: Two signals with the same power. The detection probabilities for both signals decrease compared with the detection probability of one signal with the same power, as can be seen in Fig. 14. This
Fig. 12. Simulated and experimental results (+) for detection probability with input signal centered in channel 8. Signal in channel 24 is spurious. 64-point monobit FFT using decimation in frequency, and $P_{fa} = 10^{-3}$.

Fig. 13. Simulated and experimental results (+) for the detection probability with input signal centered in channel 3. Signal in channel 1 is spurious. 64-point monobit FFT using decimation in frequency, and $P_{fa} = 10^{-3}$.

effect can be explained again by having in mind that power at the output of the ADC is constant and is spread among different channels. Using Parseval’s Theorem [5]:

$$\sum_{n=0}^{N-1} |x(n)|^2 = \frac{1}{N} \sum_{k=0}^{N-1} |X(k)|^2$$

where $\sum_{n=0}^{N-1} |x(n)|^2 = N$ is the energy of $N$ samples $x(n)$ at the output of the monobit ADC. Therefore, the sum of the squared outputs of the monobit FFT is constant, and independent of the number of signals presented at the input of the monobit ADC:

$$\sum_{k=0}^{N-1} |X(k)|^2 = N^2$$. The energy at the output of the ADC is distributed among the two input signals and the spurious generated in the ADC. Therefore, the power at the output of the filters of the monobit FFT is reduced, and consequently the detection probability.

It is necessary to point out that Parseval’s Theorem applies for monobit FFTs, both for decimation in time FFT and decimation in frequency FFT, but it is not valid for the monobit DFT used here; see Appendix B. Although (12) cannot be applied to the DFT monobit, the reduction of the detection probability is also observed when two signals with the same power are present.
Fig. 14. Simulated results for detection probability of two signals with same power centered in channels 8 and 16 respectively. 64-point monobit FFT using decimation in frequency, and $P_{fa} = 10^{-3}$. Detection probability for spurious signal in channel 24 and for single and centered sinusoid also included.

Fig. 15. Instantaneous dynamic range for two sinusoids centered in channels 16 and 21. Results are associated in pairs with constant SNR for signal in channel 16, and variable SNR for signal in channel 21 varies. 64-point monobit FFT using decimation in frequency. Threshold for $P_{fa} = 10^{-3}$.

Two-Signal Test: Instantaneous dynamic range. A fundamental feature in channelized receivers is the instantaneous dynamic range. One-bit digitalization results in a rather low instantaneous dynamic range. Fig. 15 shows the detection probability for two sinusoids centered in channels 16 and 21, respectively. The different curves are organized in pairs: for each pair, SNR of the signal labeled with Channel 16 is constant and SNR of signal in Channel 21 decreases from SNR of signal in Channel 16 to lower values. We have analyzed signal Channel 16 in a range of SNR from $-5$ dB to $10$ dB in steps of $1$ dB.

1) When both signals have the same power, the detection probability can be different due to the different response of the filters.

2) The capture effect is more significant when the signals are in filters with different amplitude response. Fig. 16 shows a similar analysis for two sinusoids centered in filters with the same amplitude response.

3) If we define the instantaneous dynamic range when $P_d = 10\%$ for the weaker signal, it is clear that the instantaneous dynamic range can be as low as $3$ dB (Fig. 15). The case presented in Fig. 16 is more
Fig. 16. Instantaneous dynamic range for two sinusoids centered in channels 8 and 24. Results are associated in pairs with constant SNR for signal in channel 8, and variable SNR for signal in channel 24 varies. 64-point monobit FFT using decimation in frequency. Threshold for $P_{fa} = 10^{-3}$. Filters for channels 8 and 24 have same maximum amplitude response.

Fig. 17. Simulated and experimental results (+) for instantaneous dynamic range. SNR for signal in channel 16 is 0 dB, and power for signal in channel 21 varies. 64-point monobit FFT using decimation in frequency. Threshold for $P_{fa} = 10^{-3}$.

favorable and the instantaneous dynamic range is higher than 10 dB.

4) An increase in the number of points of the monobit DFT(FFT) can slightly alleviate this problem.

Fig. 17 shows a comparison between measurements and simulations for signals in channels 16 and 21 and a 64-point monobit FFT: the dynamic range is 3 dB.

VII. DETECTION OF A CW-LFM SIGNAL USING A MONOBIT RECEIVER

The interest of a CW-LFM signal is due to its inherent LPI characteristic, which is a desired feature of radar systems. Radars using this waveform are already in use: A sweep of $\Delta f = 500$ MHz in $T = 1$ ms is already available in the new generation of LPI radars [8]. Given that our experimental set-up has a maximum sampling frequency of 41 MHz, we have analyzed this problem using a scaled version: A CW-LFM of $\Delta f = 10$ MHz in $T = 50$ ms. $^3$ This waveform preserves the processing gain $G_p$ of the original signal: $G_p = \Delta f \cdot T$.

Fig. 18 shows the measured detection probability for $P_{fa} = 10^{-6}$ and $P_{fa} = 10^{-3}$. Detection is declared $^3$A sampling frequency of 20.5 MHz was selected.
when at least one of the outputs of the FFT is higher than the threshold. For comparison purposes, the simulated average detection probability for a centered sinusoidal signal is included ($P_{fa} = 10^{-3}$). It is clear that there exist losses for the LFM signal compared with the centered sinusoid.

We have pointed out in Fig. 10 that the detection probability for sinusoids of random frequency within a channel is lower than for centered sinusoids. As a consequence of the modulation, the frequency of this signal changes, and successive captures show the signal in different filters or in different frequencies within each filter. Therefore, we can see this signal as a sinusoid with random frequencies within the bandwidth specified by $\Delta f$. If we compare the detection probability for the LFM signal with $P_{fa} = 10^{-6}$ in Fig. 18 to the detection probability for sinusoids with random frequency in Fig. 10, we can observe that the results for the LFM signal can be viewed as an average of the results of Fig. 10.

Modern EW equipments are not only interested in detection of signals but also in identification and classification of these signals. Time-frequency techniques have emerged as useful tools to overcome some of the limitations of original spectral analysis [9]. The most simple time-frequency technique is the short-time Fourier transform (STFT) which can be viewed as a collection of Fourier transforms of
successive pieces of the signal under analysis. We have analyzed the LFM signal with this technique, where the Fourier analysis has been performed through the monobit Fourier transform. In the present analysis it is supposed that there is no overlap between successive pieces of the signal. The results are presented in Fig. 19 for $P_{in} = 10^{-3}$ and SNR = 0 dB. From this analysis it is possible to obtain the modulation of the signal: slope and period. Therefore, the monobit DFT can be used as the kernel for more sophisticated techniques [10].

VIII. CONCLUSIONS

A monobit channelized receiver based on a one-bit ADC followed by a monobit DFT or FFT is analyzed both theoretically and experimentally. Simplifications in the operations of the FFT allow to improve real time operation. However, one-bit digitalization and monobit implementation of the DFT(FFT) result in detection losses compared with the original FFT and a rather low instantaneous dynamic range regarding the multibit approach. Other kernels for the FFT and a rather low instantaneous dynamic range in detection losses compared with the original DFT(FFT) are under investigation in order to improve these shortcomings.

APPENDIX A. CONSTANT NOISE POWER AT THE OUTPUT OF A MONOBIT DFT(FFT)

The DFT(FFT) is a linear transformation that can be interpreted as a filter bank whose impulse responses $h_k(n)$ verify that $\sum_{n=0}^{N-1} |h_k(n)|^2 = N$, $k = 0, \ldots, N-1$. This is valid for both the original DFT(FFT) and their monobit implementations defined in Section III.

Assuming that the input signal $x(n)$ to the filter bank defined by the monobit DFT(FFT) is white noise with zero mean and variance $\sigma_x^2$, its power density spectrum $\Phi_x(\omega)$ is

$$\Phi_x(\omega) = \sigma_x^2, -\pi \leq \omega \leq \pi. \quad (13)$$

Thus, the power at the output $y_k(n)$ for any of the filters of the filter bank is [5]

$$E[y_k^2(n)] = \frac{1}{2\pi} \int_{-\pi}^{\pi} \Phi_x(\omega)|H_k(\omega)|^2 d\omega = \frac{1}{2\pi} \int_{-\pi}^{\pi} \Phi_x(\omega) \cdot |H_k(\omega)|^2 d\omega.$$  

$$\quad (14)$$

$$E[y_k^2(n)] = \sigma_x^2, \frac{1}{2\pi} \int_{-\pi}^{\pi} |H_k(\omega)|^2 d\omega. \quad (15)$$

Using Parseval’s Theorem for Fourier transform of discrete signals:

$$E[y_k^2(n)] = \sigma_x^2 \sum_{n=0}^{N-1} |h_k(n)|^2 = \sigma_x^2 \cdot N.$$  

Given that $x(n)$ is the output of the one-bit ADC, $\sigma_x^2 = 1$, and the power at the output of any filter $E[y_k^2(n)] = N$. It is necessary to point out that type 1 and 2 channels have the same output power, but the amplitude probability distribution functions and so the false alarm probabilities are different.

APPENDIX B. COMMENTS ON PARSEVAL’S THEOREM FOR DFT AND FFT MONOBIT

Parseval’s Theorem for the original $N$-point DFT(FFT) is expressed as [5]

$$\sum_{n=0}^{N-1} |x(n)|^2 = \frac{1}{N} \sum_{k=0}^{N-1} |X(k)|^2 \quad (16)$$

where $x(n)$ is the discrete sequence and $X(k)$ its DFT. Using a matrix notation, the DFT(FFT) can be written as

$$X = A \cdot x. \quad (17)$$

The necessary and sufficient condition for Parseval’s Theorem is

$$A^{-1} = \frac{1}{N} A^H \quad (18)$$

where $A^H$ denotes the hermitian matrix of $A$. The condition of (18) is verified by the monobit FFT by decimation in time and the monobit FFT by decimation in frequency. However, the monobit DFT does not verify this condition.

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