DIGITAL PROCESSING FOR NON-STATIONARY NARROW-BAND INTERFERENCE SUPPRESSION IN FADING CHANNELS

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Abstract - The suppression of narrow-band interference has long been a concern in Direct-Sequence Spread Spectrum (DS-SS) Communication Systems. Recently, transform domain processing (TDP) approaches have been used to suppress intentional interference. This paper extends previous work done in the area of TDP utilizing non-linear rank order filters, namely conditional median filters (CMFs). The paper confronts problems incurred in Rayleigh distributed fading channels. A solution is suggested by introducing a normalized adaptive median filter. This adaptive approach considers each received bit independently and uses a normalization metric to compensate for fading. The paper also presents bit error rates (BER) for a DS-SS receiver employing the proposed approach. Using Monte-Carlo simulations the BER of the fading channel was determined for various jammer and fading scenarios, and results are presented.

1. INTRODUCTION

Direct-Sequence Spread Spectrum receiver systems have an inherent capability of suppressing narrow-band interferers. By increasing the processing gain (PG) of a DS-SS system the probability of intentional signal corruption is greatly reduced [1]. However, power from the undesired signal is still present, and it can still produce significant bit errors. It has been shown in references [2,3] that by adding an anti-jam processing stage at the receiver, the BER can be lowered significantly. Adaptive filter techniques have been investigated and have proven to suppress narrow band interference successfully. However, time domain filtering techniques are time consuming because of the adaptation processing time.

In recent years, research efforts in the area of surface acoustic wave (SAW) devices and charge coupled devices have lead to the development of real-time signal processing techniques in the frequency domain. In these techniques, the signal to be processed is first Fourier transformed in real-time, and the processing takes place in the transform domain. The time-domain result of the processing is obtained by an inverse Fourier transform. Fig. 1 shows a functional block diagram of the TDP unit. Operating in the transform domain allows for easier detection and efficient suppression of unwanted spectral components.

In the past, TDP techniques have included excising or soft limiting the undesired signal components by such techniques as notch filters or amplifier saturation. Notching the frequencies where the jammer is located also excises the signal though, and soft limiting the signal requires complex hardware [4]. Recently, there has been research in the area of TDP using conditional median filtering (CMF) techniques to suppress undesired large and narrow impulsive spectral components [5].

These techniques, in addition of being very simple to implement, they also offer important advantages over traditional notching or similar techniques. A first advantage of this technique is that the center frequency and bandwidth of each of the intentional interference do not have to be determined as the filter automatically adapts to these parameters. Due to these properties non-stationary interference can be easily suppressed. Another important advantage is that in the absence of interference the signal degradation is minimal. It has been shown that the BER of a DS-SS system using CMF based suppression is lower than when the traditional notch filter is used unless notchting is done at the exact center frequency and bandwidth of the jammer. However, in practice it is unrealistic to know the precise frequency of the jammer.

2. THE CONDITIONAL MEDIAN FILTER

The motivation for using filters for interference suppression is that in the transform domain narrow-band interference appears as large and narrow impulsive components. Since smaller impulsive components may also be present, the problem then becomes to suppress the large and narrow impulses while preserving the smaller ones, which might represent important information about the signal. This problem is similar to the problem of suppressing impulsive noise from an image while preserving the finer detail, and this suggested the idea to use filters traditionally used in image processing. Median filters (MFs) are well-known for their ability to suppress narrow impulses while preserving sharp signal transitions. However, MFs do not preserve fine detail, and a number of other detail preserving median-based filters have been proposed. One such filter which has been very successful for this type of image processing is the so called Conditional Median Filter (CMF) [6]. Based upon a threshold condition filtering of samples that have values close to their median is avoided by the CMF. The defining equation of the CMF is:

\[ y_{CMF} = \begin{cases} m_i & \text{if } |x_i - m_i| > C_i \\ x_i & \text{otherwise} \end{cases} \]  

where, \( x_i \) and \( y_i \) are the input and output data sequences, \( m_i \) is the median value of an odd sized sliding window of length \( N \), and \( C_i \) is a variable threshold. The CMF will not affect signal components that do not meet the threshold condition in (1), but any impulse in the local window narrower than \( \frac{N-1}{2} \) samples will be suppressed [11] provided that it is also large enough to surpass the threshold condition in (1).
Interference suppression is accomplished as depicted in the block diagram of Fig. 1 by filtering the magnitude of the Fourier transform using a CMF. With this approach the local median essentially provides an interference-free signal floor estimate which is used as a reference to detect the presence of any interference. The parameter \( N \) defines the maximum bandwidth of impulsive components that may be suppressed provided that the impulse’s amplitude exceeds the parameter \( C \) from the reference floor. By using the median as a reference it is possible to detect and suppress interference of lower levels than the simple transform clipping of [3]. In addition, this technique is sensitive not only to amplitude, but to bandwidth as well.

With this technique, the position and precise bandwidth of each interferer is not important; they only have to be sufficiently large and narrow. Also, since there is no adaptation time after the transform is generated, the technique is suitable for fast and accurate suppression of non-stationary interference. Also, multiple interference is automatically handled. Furthermore, when interference is not present processing of the signal is limited only to spectral components that may occasionally exceed \( C \). Finally, since MF hardware is available, real-time implementations are feasible.

An example of CMF filtering is presented in Fig. 2a. Fig. 2b is a time domain representation of a 32 chip sequence that is the transmitted signal; it also shows the effects of the addition of four jammers at unknown frequencies. The FFT of the original and corrupted PN sequences is shown in Fig. 2b and 2c. Notice that the frequency where the jammer was located is now “smoothed” instead of being null, which occurs in the case of notch filtering. The recovered time domain sequence is compared to the original sequence in Fig. 2c.

3. THE NORMALIZED CMF (NCMF)

When the received signal power is constant the threshold in (1) can be a constant, but if the received signal power varies a need exists for adaptation of the threshold to the signal strength. This results from the fact that the predetermined threshold, \( C \), will be optimum at only one value when the signal strength remains constant. When the signal power increases the value \( |m_i - \chi| \) could conceivably grow larger resulting in a larger variance neighboring the spectral components, thus causing an action to filter when no filtering is needed. Moreover, when the signal strength is low and a relatively low powered jammer is present it becomes difficult to detect the jammer.

Experimentation has shown that the probability distribution function (pdf) of the Fourier magnitude spectral components of a PN signal is well approximated by a Rayleigh distribution. The median value of a random variable that is Rayleigh distributed is approximately 1.5 times the mean value. If the current input sample is compared to the expected value of the PN signal’s spectrum and then normalized by the expected value of the input sample a dynamic approach of determining the threshold is established. When considering fading channels this method should eliminate the need for a priori knowledge of the power associated with the received signal. It is proposed that the normalized CMF (NCMF) to be defined by the following equation:

\[
Y_{\text{NCMF}} = \begin{cases} 
 m_i & \text{if } \frac{|x_i - m_i|}{m_i} > C \\
 x_i & \text{otherwise} 
\end{cases}
\]

where \( C \) is a normalization parameter.

4. SIMULATION MODEL

The model used for the DS-SS system is rather typical and was also used in [1-4]. The received baseband signal is of the form \( y(t) = a(t)b(t)\chi(t) + J(t) + n_r(t) \), where \( c(t) = \pm 1 \) is a PN sequence of \( L \) chips, \( b(t) = \pm 1 \) is the information bit, and \( n_r(t) \) is AWGN of two sided spectral density \( \frac{N_0}{2} \). Fading is modeled as a multiplicative amplitude term \( a(t) \) which takes random values in time with a Rayleigh pdf. The interference is a tone jammer of the form \( J(t) = A \cos(2\pi f_o t + \theta) \), where \( f_o \) is the frequency offset from the signal carrier, \( A \) is a constant amplitude, and \( \theta \) is uniformly distributed in \((0,2\pi)\). In reality even stationary interference will exhibit some short-term frequency jitter. To provide a more realistic simulation, stationary interference was modeled having \( f_o \) uniformly distributed in an interval extending \( \pm \frac{\Delta f}{2} \) around a nominal value while non-stationary interference was modeled with \( f_o \) uniform within the signal bandwidth (main lobe). Assuming perfect synchronization, the signal \( r(t) \) is sampled at twice the chip-rate and processed by the TDF one bit at a time using a 64-point FFT. To facilitate the FFT computation the processing gain was chosen to be \( L = 32 \). A functional block diagram of the channel is shown in Fig. 3.

The value of \( \frac{\overline{E_b}}{N_0} \) (average bit energy) was varied by changing the noise variance and for each value the number of simulated bits was large enough \((\approx 10^3)\) for fading every bit) to ensure at least 95% confidence on the results [12]. Various levels of fading are modeled by changing the variance of \( \alpha(t) \). A correction to \( \frac{\overline{E_b}}{N_0} \) is needed because a change in the variance of the fading also affects the mean value and as a result the \( \overline{E_b} \) will change. The necessary correction is found as follows: A Rayleigh distributed random variable \( \alpha \) is generated from \( \alpha = \sqrt{X_1^2 + X_2^2} \) where \( X_1 \) and \( X_2 \) are two independent zero mean Gaussian deviates with variance \( \sigma^2 \). The pdf of \( \alpha \) is given by [12,13]:

\[
P_\alpha(x) = \frac{x}{\sigma^2} \exp\left(-\frac{x^2}{2\sigma^2}\right) \quad 0 < x < \infty
\]

The mean value is \( \overline{\alpha} = \sigma\sqrt{\frac{2}{\pi}} \) and the variance is \( \alpha^2 = \sigma^2\left(2 - \frac{2}{\pi}\right) \).

It can be proved that the value of \( \frac{\overline{E_b}}{N_0} \) in decibels at the presence of fading is given by:
\[ \frac{P_s}{N_o} = \frac{P_s}{N_o} + 0.5 + 10 \log(\sigma^2) \]  

where \( \frac{P_s}{N_o} \) is the signal to noise ratio before considering fading effects.

In the transform domain, the power associated with a single-tone jammer is not concentrated at one frequency; this is known as spectral leakage. Spectral leakage can be minimized by windowing the data before transformation by using one of the several window shapes available in the literature. Windowing reduces side-lobes of the jammer when leakage occurs, at the expense of widening its main lobe. In this paper, a sinusoidally shaped window was utilized because it was experimentally determined to produce the lowest BER.

6. EXPERIMENTAL RESULTS

Monte Carlo simulations were used to determine the BER of a DS-SS system using the proposed suppression scheme. The system model used is as described in the previous section. Through experimentation an empirical value of 1.5 for the parameter \( C \) was determined. This value was kept the same throughout the simulations. The fixed notch filter nulls two FFT bins around the nominal frequency of the stationary interference. This notch width is the smallest to ensure good suppression. Bit error rates were determined for various jamming scenarios and were plotted as a function of \( \frac{P_s}{N_o} \). Windowing was used for every case where a jammer was present.

Fig. 4 presents the BER results in the absence of jamming and in the presence of Rayleigh fading. As expected, no jammer produced the lowest BER with the NCMF approaching the performance of no jammer. Conversely, the fixed notch consistently degrades the performance. Fig. 5a is the same scenario as in Fig. 4, but in the presence of a stationary jammer with amplitude \( A = 6 \) V located at \( f_s = \frac{5}{2} \). It can first be observed that jamming processing greatly reduces the BER with the proposed method clearly outperforming the notch filter. In Fig. 5b the power of the transmitted signal has been increased by a factor of 4 in order to examine how well the NCMF adapts to a change in the signal power. It can be seen from Fig. 5b that the NCMF continues to outperform the notch, indicating that the NCMF adapts well to changing power.

Fig. 6 presents a scenario where a single non-stationary jammer is present. Note that the notch filter could not be utilized in this scenario, because a priori knowledge of the center frequency of the interfering tone is not known generally. It is evident from the low BER that the NCMF is capable of successfully suppressing non stationary interference as well.

7. REFERENCES


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Fig. 1 Block diagram of a transform domain processor
Fig. 2a Original and corrupted binary sequences

Fig. 2b Spectra of original binary sequence

Fig. 2c Spectra of corrupted binary sequence

Fig. 2d Result of a CMF applied to the signal of Fig. 2c

Fig. 2e Recovered vs. original binary sequences

Fig. 3 Channel model
Fig. 4 BER results without interference

Fig. 5b BER results with stationary interference and higher signal power

Fig. 5a BER results with stationary interference

Fig. 6 BER results with non-stationary interference