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Domain Selective Interference Excision and Energy Detection of Direct Sequence Signals

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ABSTRACT

We study the performance of energy detection of direct-sequence (DS) signals when domain-selective interference excision precedes the energy detection. A simple method for choosing the proper domain is proposed. A forward interference excision algorithm is used in the selected domain. The performance of the algorithm is simulated. Simulation results show that in the scenarios studied the domain selective excision performs well.

1. INTRODUCTION

Direct-sequence (DS) signals are widely used in spread-spectrum (SS) applications like code-division multiple-access (CDMA) or positioning. In addition to communication signals, these environments may contain also different kinds of interference, either intentional or unintentional. In signal detection, the goal is to determine if unknown signals are present. A particular goal is to determine if an unknown DS signal is present when there can co-exist an interfering signal (typically narrowband). In this case, the conventional energy detector decides that only one signal is present [1].

There exists a vast number of possible interference excision methods, e.g., the consecutive mean excision (CME) algorithms [2, 3]. In our previous paper concerning this issue [1], the consecutive mean excision (CME) algorithm [2] was used to suppress narrowband interferences in the frequency domain. The more powerful version of the CME algorithm, called the forward CME (FCME) algorithm, was introduced in [3]. The both CME algorithms are based on iterative computation of the excision threshold. The advantage of using blind excision algorithms, such as the CME algorithms, is that they limit the effects of the noise level estimation error [4] to the detection stage. This work studies the intercept receiver that first uses the FCME algorithm to excise interfering signals before energy detection. Some interferences are easier to detect in the time domain than in the frequency domain. Those include all short duration signals like impulses. Due to this reason we propose a domain selection based interference excision, where the adequate domain is selected automatically.

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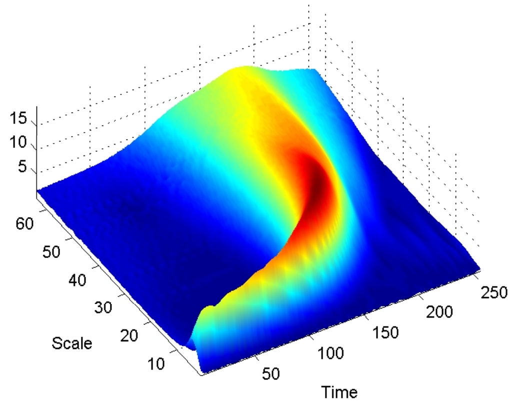

Fig. 1. Continuous wavelet transform of a chirp signal.

2. SYSTEM MODEL

The received signal is assumed to be

$$r(k) = s(k)e^{j(\omega_s k+\theta)} + i(k) + n(k), \qquad (1)$$

where $s(k)$ is the DS signal, ω_s is the frequency offset with respect to the center frequency of the detector, θ is a random uniformly distributed carrier phase, $i(k)$ is an interfering signal and $n(k)$ is a zero-mean signal modelling noise. The noise is assumed to be a white Gaussian process with one-sided power spectral density N_0. The N_0 is assumed to be known to the receiver. Ideal low-pass filters are assumed to be used before sampling the down-converted radio frequency signal. The variances of the real and imaginary parts of the noise signal are $\sigma^2 = N_0 W_S/2$, where $W_S = 1/T_S$ is the noise-equivalent bandwidth and T_S is the sampling interval. Note that the optimal sampling rate depends on factors that are usually unknown (such as the pulse shape of the DS signal). The detection period is assumed to include N_S samples with indices $k = 1, 2, \cdots, N_S$. The detection period is split into C non-overlapping segments of N_{FFT} samples. For simplicity, it is assumed that $N_S = CN_{\text{FFT}}$.

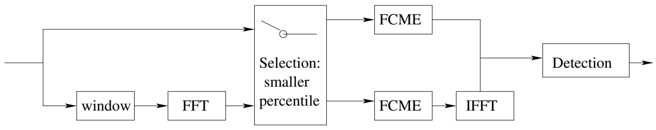

Fig. 2. Receiver structure. The selection is made between the time and frequency domain before interference excision.

The signal-to-noise ratio (SNR) per bit is

$$\gamma = E_b/N_0 = (ST_C N_C)/N_0, \quad (2)$$

where E_b is the energy of the DS signal per bit, S is the signal power, T_C is the chip duration and N_C is the number of chips per bit. Interference power is

$$I \equiv \xi S = (\xi\gamma)N_0/(T_C N_C), \quad (3)$$

where ξ is the average interference-to-signal ratio (ISR), i.e. $\xi = I/S$.

Three interfering signal types are considered. These are impulses, sinusoidals and chirp signals. It is expected that the time domain excision is selected in the case of impulses and the frequency domain excision for the others. Impulses are a set of Dirac delta functions. The ratio of the number of corrupted samples to the total number of samples is called the contamination ratio. A discrete sinusoidal signal can be expressed as

$$i(k) = \sqrt{I} \exp\left[j(\omega_0 k + \theta)\right], \; k = 1, 2, \cdots, N_S, \quad (4)$$

where ω_0 and θ are the frequency-offset and the initial phase, respectively. We define ISR per sinusoidal (different sources) in this case. A discrete chirp signal is a frequency sweeping sine wave and it can be presented as

$$i(k) = \sqrt{I} \exp\left[j(\omega_0(k)k + \theta)\right], \quad (5)$$

where $\omega_0(k)$ is the time-varying frequency-offset. Fig. 1 shows continuous wavelet transform of a chirp signal. In the figure, the scale increases with time so that the instantaneous frequency is decreasing.

3. SELECTION OF TIME OR FREQUENCY DOMAIN EXCISION

The structure of the receiver is shown in the Fig. 2. The selection of the time or frequency domain is made independently in each segment. The decision is based on the magnitude squared time-domain and frequency-domain values. The frequency-domain values are calculated using the windowed FFT. In the selection process, first the frequency-domain energy is normalized to be the same as the time-domain energy (this is needed because windowing affects the energy) and the values are sorted in the ascending order. This sorting is required also in the FCME algorithm such that sorting can be reused. Sorting algorithms are discussed in [5]. The average complexity of Quicksort and Heapsort routines is similar than that of the FFT [5]. The 25 % percentiles are then calculated in the receiver. The domain with the smaller percentile is selected. This is based on the fact that the correct domain concentrates interference energy in a fewer number of samples, while in the wrong domain the interference energy is spread more widely (thus increasing the low percentiles).

The interference excision is performed to the values of the selected domain. Operation of the FCME algorithm is controlled by the CME parameter T_{CME}. It can be analytically found based on the desired clean sample rejection rate [2]. Typically rejection rates 10^{-3} and 10^{-2} are used. These correspond to T_{CME} 2.97 and 2.42, respectively. The smaller T_{CME} is, the more interference (and also signal and noise) is removed. The excision is performed independently in each block. The outputs of the excision are the cleaned signal, where the bins that exceed the FCME threshold are set zero, and the bin numbers of the zeroed bins. After the interference excision, energy detection is performed.

4. ENERGY DETECTION

The energy detector calculates

$$V = \sum_{t=1}^{C} \|\mathbf{Y}_t\|^2 = \sum_{t=1}^{C} \mathbf{r}_t^* \left(\mathbf{A}_t^* \mathbf{A}_t\right) \mathbf{r}_t = \sum_{t=1}^{C} \mathbf{r}_t^* \mathbf{G}_t \mathbf{r}_t, \quad (6)$$

where $\mathbf{r}_t$ is vector that contains the received signal samples corresponding to the segment t, $\mathbf{r}_t = \left[r((t-1)N_{\text{FFT}}+1) \cdots r(tN_{\text{FFT}})\right]^T$ and * is the conjugate transpose. If FFT excision is selected for the segment t

$$\mathbf{A}_t = \frac{\Phi^*}{N_{\text{FFT}}} \mathbf{P}_t \mathbf{\Phi} \mathbf{W}, \quad (7)$$

where $\mathbf{\Phi}$ is the Fourier transform matrix, $\mathbf{P}_t$ is an identity matrix except that the diagonal entries corresponding to the excised bins are set zero, $\mathbf{W}$ is a diagonal matrix with the window coefficients in the diagonal and $\mathbf{G}_t = \mathbf{A}_t^* \mathbf{A}_t$ are hermitian matrices. In the case of time domain excision in segment t, $\mathbf{A}_t = \mathbf{P}_t$.

The statistic V is compared to a detection threshold. If V is larger than the threshold, it is decided that a wideband signal is present in the received signal in addition to noise and possible interference. If V is smaller than the threshold, it is decided that there was only noise (and possible interference) present in the received signal. The problem is to select the detection threshold in such a way that the probability of false alarm (P_{FA}) is equal to (or smaller than) the desired value ($P_{\text{FA,DES}}$). The P_{FA} is the probability that V is greater than the threshold in the case where the received signal contains only noise (and possibly interference). The probability of detecting a signal, when it really is present, is called the probability of detection (P_{D}). The probability of miss is 1-P_{D}. The cumulants of the decision variable V can be calculated analytically. The problem of approximating probability density function (PDF) and cumulative distribution function (CDF) based on some cumulants is a well known problem. The threshold is based on the inverse of the CDF. One relatively simple method is the Cornish-Fisher algorithm, which can use a variable number of cumulants [6]. Note that the accuracy does not always increase with the increasing number of cumulants. The Cornish-Fisher algorithm gives the threshold directly based on the cumulants. It is based on inversion of the Edgeworth expansion of the CDF. We used thresholds based on the first six cumulants and found that the resulting thresholds are very accurate (practically they are exact). In the noise-only case the cumulants $\kappa_n(V)$ can be calculated with (cf. [7])

$$\begin{aligned}\kappa_n(V) &= \sum_{t=1}^{C} \kappa_n \left(\mathbf{r}_t^* \mathbf{G}_t \mathbf{r}_t\right) \\ &= \sum_{t=1}^{C} 2^n (n-1)! \sigma^{2n} \text{trace}\left(\mathbf{G}_t^n\right) \\ &= \sum_{t=1}^{C} 2^{n-1} (n-1)! \sigma^{2n} \text{trace}\left(\hat{\mathbf{G}}_t^n\right),\end{aligned} \tag{8}$$

where

$$\hat{\mathbf{G}} = \begin{bmatrix} \text{Re}(\mathbf{G}) & -\text{Im}(\mathbf{G}) \\ \text{Im}(\mathbf{G}) & -\text{Re}(\mathbf{G}) \end{bmatrix} \tag{9}$$

is a real-valued symmetric matrix ($\mathbf{G}$ is hermitian) of the equivalent real-valued quadratic form. The threshold is found by using cumulant vector $[\kappa_1(V) \cdots \kappa_6(V)]$ and $P_{\text{FA,DES}}$ with the Cornish-Fisher algorithm [6]. The threshold depends on the decisions of the excision algorithm (via the excision matrices $\mathbf{P}_t$). The threshold is based on assumption that the excised frequencies are independent of the noise realization in the detection interval. In practice this is an approximation. In the case of a small threshold parameter and no interference, P_{FA} will be slightly smaller than the required one since excision removes some of the strongest noise samples. It is also assumed that the excision removes the possible interference (with the excision matrices $\mathbf{P}_t$). This is justified because usually the interference is unknown and so it cannot be taken into account in the threshold calculation. Simulations confirmed that when the excision works well, P_{FA} will not rise too much. In [7], it was assumed that interference is perfectly known. This allows the exact calculating of the threshold (that takes into account the interference).

It is possible to obtain theoretical results with the numerical inversion of the characteristic function (or the moment generating function) of the quadratic form on the P_{D} and P_{FA} based on the knowledge of the DS signal and interference. These results cannot be applied in an actual intercept receiver due to the requirement that the signal has to be known. The signal and interference are usually random so the results should be averaged over all the possible realizations of the signal and interference.

When the time domain excision is used, the normalized measured energy after excision follows the chi-square distribution with the degrees of freedom being the number of non-excised samples multiplied by two. Accordingly, the exact detection threshold can be calculated (the Cornish-Fisher approximation does not need to be used) and it is

$$\eta = \Lambda^{-1}(1 - P_{\text{FA}}, 2(N_S - N_E)), \tag{10}$$

where Λ^{-1} is the inverse of the chi-square CDF, N_S is the number of samples in the detection interval and N_E is the total number excised samples. Again, the effect of strong noise samples is ignored. In practice, this is not a problem because only a small amount of the noise is excised (this depends on the T_{CME}).

5. SIMULATION RESULTS

The radiometer is a sensitive device. When the interference power is small, it is practically impossible for the excision algorithms to find the interference. In the simulations we assume that the interference power is reasonably high. We used ISR 20 dB or 30 dB. The signal to be detected is a DS signal with a square-root raised-cosine (SRRC) chip waveform. Note that if the DS signal level is rather high related to the noise level, the excision will also remove some of the strongest signal components, thus degrading performance. Random spreading codes are used. There are 64 chips per bit and 2 samples are taken per chip. The roll-off factor was 0.5 (the signal bandwidth per the receiver bandwidth is 75 %). If the signal bandwidth is very small compared to the receiver bandwidth, the interference excision will also remove most of a strong signal. In this case the task of detecting the signal is with the narrowband signal detection algorithm. The detection interval contains 8 bits (1024 samples), and the desired P_{FA} is 10^{-2}. The windowed FFT uses 4-term Blackman-Harris window. The segment length is 64, so there are 16 segments per detection interval. It is possible to use different CME parameter in each segment. We used the $T_{\text{CME}} = 2.42$ if the frequency-domain excision was selected. Simulations (not shown here) were also performed with $T_{\text{CME}} = 2.97$, but the false alarm probabilities were too high ($\geq 10^{-1}$). This is due to fact that too little interference is excised. If the time-domain excision was selected in a segment we used $T_{\text{CME}} = 2.97$ in that segment. This was done because detection performance is better with the higher CME parameter and simulations showed that false alarm probabilities did not rise too much.

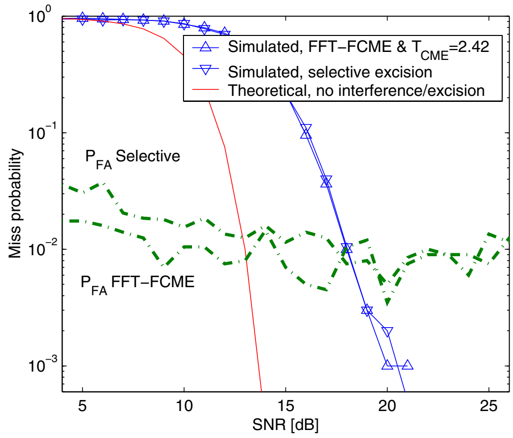

Fig. 3. Sinusoidal interference, frequency domain and domain selective interference excision, ISR is 20 dB per sinusoidal signal and the number of sinusoidal signals is 2.

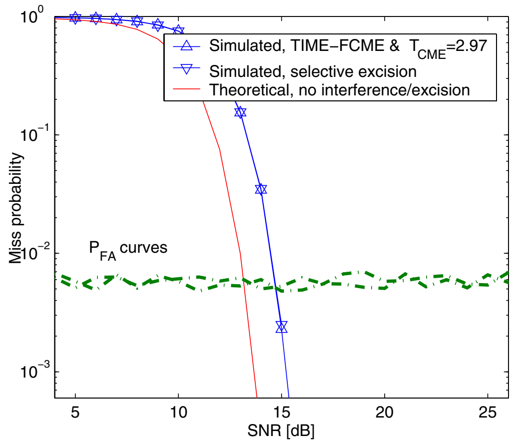

Fig. 4. Impulse interference, time domain and domain selective interference excision, ISR is 30 dB and 40 % contamination.

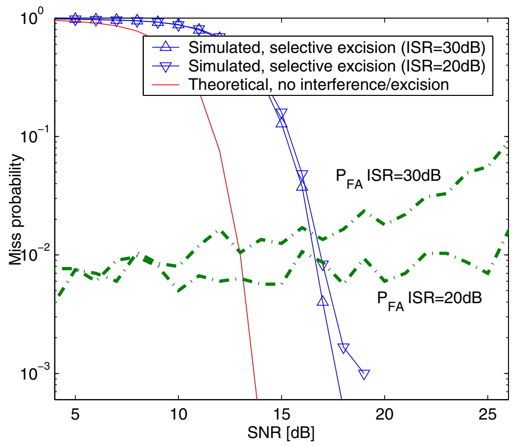

Fig. 5. Chirp interference and domain selective interference excision, ISR is 20 dB or 30 dB and the bandwidth swept during the detection interval is 100 % of the system bandwidth.

Figures 3– 5 present the theoretical P_D without interference or excision (calculated using the well known radiometer formulas, cf. [8]), the simulated P_D with interference and excision and the simulated P_{FA} with interference and excision (dash-dotted lines) vs. SNR. Fig. 3 shows that when the interference is sinusoidal the domain selective excision works almost as well as the excision that is performed in the correct domain (frequency). Fig. 4 shows that in the case of impulse interference, the selective excision works as well the excision performed in the correct domain (time). Note that the results are closer the theoretical results in Fig. 4, where the time-domain excision is used most of the time, than in Fig. 3, where the frequency domain is chosen. This is mainly caused by the windowing loss associated with the frequency-domain excision. Fig. 5 shows that chirp interference that sweeps through the whole system bandwidth during the detection interval can be excised with the domain selective excision. When the interference power is very large the P_{FA} increases.

6. CONCLUSIONS

The performance of energy detection has been studied when interference excision is performed before detection. We have proposed a simple method for choosing either the time-domain or frequency-domain excision. The results show that when the interference is sufficiently strong, the excision performs well and the probability of false alarm stays close to the required value.

7. ACKNOWLEDGEMENT

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