Attenuation of Narrowband Interference in Wideband Signals to relax ADC requirements

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***Abstract*—Narrowband interference can increase dynamic range demands on wideband receiver systems, when this receiver operates under terrestrial conditions and has to capture a relatively weak desired signal. In a software-radio approach, the Analog to Digital Converter (ADC) in the receiver has to accommodate this increase in dynamic range. This work discusses a receiver topology that relaxes the ADC requirements by attenuating the narrowband signal before A/D conversion takes place. The autocorrelation properties of the narrowband signal allow prediction of this signal using digital signal processing. Using this property a flexible receiver system is possible. The performance of the system under various circumstances is determined by means of simulations.**

***Index Terms*—Interference Attenuation, Analog to Digital Converter, Dynamic Range, Linear Prediction, Ultra Wideband**

I. Introduction

SOFTWARE-RADIO based receiver systems can capture any signal of any type within the Nyquist band of the ADC, given the power of that signal is adequate. This flexibility is achieved by using digital demodulation and downmixing, the implementation of which can be easily modified digitally. A drawback of this topology is the stringent requirements on the ADC [1]. If this system has to cope with a large signal band (1 GHz range) the samplerate of the ADC becomes large. Such a fast ADC has an inherent accuracy limitation that is related to device technology [2] and will dissipate significantly more power for each bit of additionally required accuracy [1]. If a wideband software-radio based receiver system is utilized under terrestrial conditions, the antenna may capture unwanted narrowband interfering signals (e.g. a nearby located GSM base station) having more signal power than the desired signal (see figure 1). The increase in signal power requires a larger dynamic range of the ADC, leading to a significant increase in power consumption. To alleviate this problem, the interferers can be attenuated before A/D conversion takes place [3].

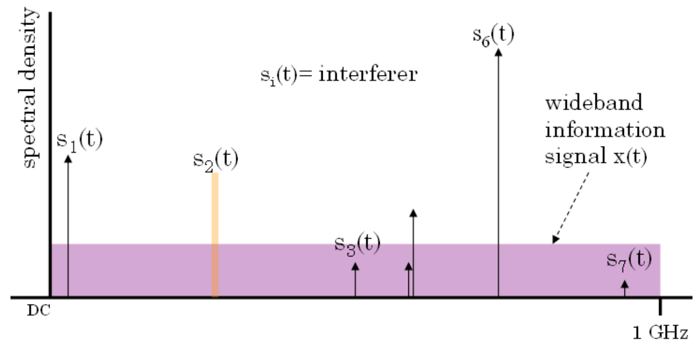

Fig. 1. Example spectrum received by the antenna of a wideband radio system. The desired wideband signal (purple) can be contaminated with unwanted narrowband interfering signals.

This paper discusses an ADC frontend system that exploits the narrow bandwidth of the interfering signals to estimate its future sample values. This facilitates an efficient implementation an ADC that handles these types of signals. The goal of this paper is to analyze the 'predictability' of interferers, so that the feasibility of the system can be assessed.

Section II will discuss the system topology and mixed signal system level issues relevant to design a suitable implementation. Section III will discuss the linear prediction theory used to make a prediction system. Various scenarios of interfering signals at various frequencies will be treated. Section IV will provide simulation results.

II. System level considerations

A. 2-step prediction ADC Topology

Figure 2 shows the front-end system-level scheme of the receiver. The input signal is fed into 2 branches; one branch encompasses a coarse ADC, a prediction system and a DAC. This branch creates an analog signal that is coarsely equal to the analog input, at the subtraction point. The delay caused by the coarse ADC and DAC is compensated by the prediction system such that both signals at the subtraction point are aligned in phase. The output at the subtraction point is fed into the 'main ADC', which is the fast high resolution data

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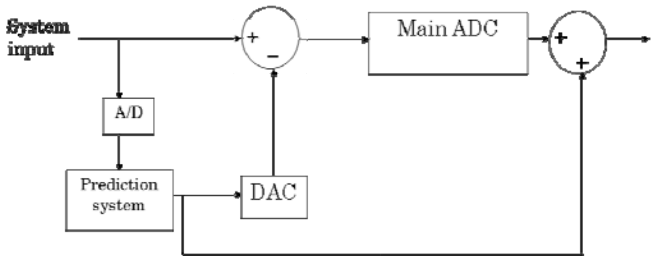

Fig. 2. System-level overview of ADC front end system.

converter that has to capture the desired wideband signals.

In [3], the phase alignment of the two branches is realized by using an analog delay line in the direct path rather than exploiting the predictive nature of the interferer. This approach requires a linear delay line with a constant group delay over a wide bandwidth which may create implementation difficulties.

The signal that is subtracted in the analog domain is added again in the digital domain, much like the operation of a 2-step ADC. Therefore, as with the 2-step ADC, the accuracy of this system is determined by the accuracy of the DAC. If the DAC is accurate enough (i.e. the difference between the actual output voltage and its digital representation is smaller than half an LSB step of the main ADC) the frontend does not significantly limit the accuracy. Any deterministic errors caused by the coarse ADC (e.g. comparator offsets), prediction system (e.g. minor prediction inaccuracies) or DAC (limited quantization accuracy) are eliminated at the summation point situated after the main ADC, as long as the voltage swing at the main ADC input remains within the input range of the ADC. Incorporating a slight overrange in the ADC relaxes the accuracy requirements of the prediction branch.

The advantage of this topology lies in the fact that the additional bits of quantization required for the increase in dynamic performance do not cause additional latency of the ADC. Compared to this frontend, the latency caused by the coarse ADC and DAC in a normal 2-step ADC necessitate a longer hold time for the sampled voltage (because the total quantization time is longer). If a certain samplerate is required, that means the interleaving factor has to be increased (increasing the power consumption) or both the coarse and fine quantization circuitry of the ADC should work faster (thus dissipate more power). This increase in power consumption is not needed if the latency of the coarse stage is negated by the prediction system.

III. LINEAR FIR PREDICTION

A. Analytical approach

The narrow bandwidth of the interfering signals relative to the samplerate of the A/D system, and the higher power of the interferer relative to the desired wideband signals create correlation between adjacent signal samples that can be used to predict the momentary value of that signal at a certain time in the future [4]. This implies that a weighted average of the previous samples can provide a prediction estimate of the signal one or several samples into the future. Such a weighted average can be generated by using a Finite Impulse Response (FIR) filter. A FIR type filter only has zeros; if one coefficient is nearly zero, it can be 'shut off' to save power without risking instability.

The Mean Squared Error (MSE) is a measure for the quality of a prediction estimate. This error depends on the FIR filter order, the relative bandwidth and relative power of the interferers, the amount of interferers and their centre frequencies, the accuracy of the coarse ADC and DAC, the properties of the desired wideband signal and the amount of samples the predictor has to look 'ahead' (i.e. the prediction depth).

Because of the many factors involved with the prediction accuracy, numerical simulations are used to provide insight into the possibilities but first the mechanism behind linear signal prediction is described using a simplified approach.

The prediction error of a FIR predictor can be defined as

$$e(n) = s(n) - \hat{s}(n) \tag{1}$$

In this equation e(n) is the error signal, s(n) is the signal to be predicted and $\hat{s}(n)$ is the prediction of s(n). Assuming the prediction filter is of N^{th} order and has to predict d+1 samples ahead, the response of the FIR filter can be written as

$$\hat{s}(n) = \sum_{k=1}^{N} a(k)s(n-k-d) \tag{2}$$

The error e(n) has a total integrated energy of

$$E = \sum_{n=-\infty}^{\infty} e(n)^2 = \sum_{n=-\infty}^{\infty}\left[s(n) - \sum_{k=1}^{N} a(k)s(n-k-d)\right]^2 \tag{3}$$

To obtain a(k), which are N unknown variables, the derivative with respect to every filter coefficient a(i), 1≤i≤N is should be taken and set to zero. This will give a unique solution as the squared error is a parabolic function and therefore only has one global minimum.

$$\frac{\partial E}{\partial a(i)} = \frac{\sum_{n=-\infty}^{\infty}\left[s(n) - \sum_{k=1}^{N} a(k)s(n-k-d)\right]^2}{\partial a(i)} \tag{4}$$

$$= 2\sum_{n=-\infty}^{\infty} s(n)s(n-i-d) - 2\sum_{k=1}^{N}\sum_{n=-\infty}^{\infty} a(k)s(n-k-d)s(n-i-d) = 0$$

Because we integrate over an infinite range, the delay factor d can be left out of the right side of (4):

$$\sum_{n=-\infty}^{\infty} s(n)s(n-i-d) - \sum_{k=1}^{N} a(k)\sum_{n=-\infty}^{\infty} s(n-k)s(n-i) = 0 \tag{5}$$

This can be written in terms of autocorrelation, assuming for simplicity that s(n) is a Wide Sense Stationary (WSS) process:

$$\sum_{k=1}^{N} a(k)R_{ss}(k-i) = R_{ss}(i+d),\ 1 \le i \le N \tag{6}$$

This result can also be obtained using the orthogonality principle [5] that states that the prediction error is minimal when it is orthogonal to the predictor input data. When solving these N equations the filter coefficients a(k) will be found. If d=0 (i.e. the predictor predicts the next sample) (6) reduces to the normal Wiener-Hopf equations [5]. (6) can be written in matrix form:

$$\mathrm{R}_{\mathrm{N}} \cdot \vec{a} = \vec{r}_d \tag{7}$$

In (7) R^N is the N*N sized autocorrelation matrix, $\vec{a}$ is a vector containing all N filter coefficients a(k) and $\vec{r}_d$ is the autocorrelation vector, starting at a time lag d. Solving for $\vec{a}$:

$$\vec{a} = \mathrm{R}_{\mathrm{N}}^{-1} \cdot \vec{r}_d \tag{8}$$

This method therefore needs matrix inversion to obtain a(k).

Using $R_{ss}(\tau)$ to find a(k) implied the assumption of a WSS process. In reality, this is not always the case. If the 2nd moment statistics of the process change over time, a(k) have to be 'updated' over time as well to ensure a good prediction performance. This updating always lags behind because in order to obtain the new statistic properties (i.e. find a good estimate of $R_{ss}(m)$) many samples of s(n) are needed. This is explained in the next section.

B. Obtaining the autocorrelation of s(n)

Predicting a summation of M unmodulated sinewaves at arbitrary frequencies theoretically can be achieved with no prediction error, as long as the prediction filter order M > N [5] and the autocorrelation function $R_{ss}(m)$ of the incoming signal is known perfectly. This however is not possible when only a limited time-frame of the input signal is included in the calculation of $R_{ss}(m)$. The following section illustrates this effect. A sampled single tone input limited to P samples can be written as:

$$s(n) = \cos(An + \varphi_1)u(n)u(P - n) \tag{9}$$

where u(n) is the unit step function, A is the angular frequency and φ is an arbitrary phase constant. The autocorrelation of this bounded signal $\hat{R}_{ss}(m)$ can be written as

$$\frac{(P - m)\cos(Am)}{2P} + \frac{1}{2P}\sum_{n=0}^{P-m}\cos(2An + 2\varphi_1 + Am), \ \mathrm{m}\geq 0 \quad (10a)$$

$$\frac{(P + m)\cos(Am)}{2P} + \frac{1}{2P}\sum_{n=-m}^{P}\cos(2An + 2\varphi_1 + Am), \ \mathrm{m}<0 \quad (10b)$$

(11b) is symmetrical with respect to (11a) for m around m=0. The triangular scaling factor (P-m)/P can easily be cancelled in a predictor system. For m≥0, the error in the autocorrelation can be written as (m≤0 is equivalent)

$$Z_{ss}(m) = \hat{R}_{ss}(m) - R_{ss}(m) = \frac{1}{2(P - m)}\sum_{n=0}^{P-m}\cos(2An + 2\varphi_1 + Am) \tag{11}$$

Using a summation formula, the variable n is eliminated:

$$Z_{ss}(m) = \frac{\sin(-Am + 2\varphi_1 + A(2P + 1)) + \sin(-Am - 2\varphi_1 + A)}{4(P - m)\sin(A)} \tag{12}$$

Because $Z_{ss}(n)$ affects both R_N^{-1} and r_d in (8), the resulting coefficients are affected as well. If P is chosen large, this contribution is very small (as Z scales down with P). This means a larger time frame to record $R_{ss}(m)$ results in a more accurate prediction of s(n).

If additional sinewaves are added to s(n), it becomes harder to obtain all the statistics of the input signal within the limited timeframe. s(n) can have very different shapes within the timeframe P depending on the phases of the sinewaves.

An additional design parameter is the required accuracy of the filter taps and the quantization of the signal s(n-d) running through the filter. A higher quantization accuracy causes the filter to dissipate more power, but also allows the filter output to be more accurate.

IV. Simulation Results

A typical spectrum of a 100 MHz wide information signal (this is the desired signal) and a narrowband signal is shown in figure 4, and forms the basis of the simulations. The white noise floor power density is set to -50 dB with respect to the 100 MHz wide signal. This signal stays at the centre frequency of 850 MHz while the narrowband interferer is shifted through the Nyquist band to create different scenarios of the input signal. This example is one of many scenarios possible; the goal of the simulations therefore is not to find the minimum performance of the predictor under all circumstances, but to obtain an estimate of the prediction performance and evaluate how it depends on the nature of the incoming interferer.

The goal is to find an efficient FIR filter (as few taps and quantization as possible) to efficiently predict this signal several samples into the future. The exact amount of samples is based estimations of the latency of the coarse ADC and DAC, as well the latency the FIR filter has on its own [7].

For the abovementioned scenarios of s(n), the quantization of the filter taps was set to 6 bits, the filter input signal was quantized to 4 bits and the DAC is also set to 4 bits. The goal of this system is to reduce the voltage swing at the main ADC input by 75% and have the FIR filter predict the incoming signal 7 samples into the future.

The prediction performance is defined by the number of samples that exceed the reduced voltage swing. Figure 5a shows the percentage of samples that fall outside of the allowed 25% of relative voltage swing, figure 5b shows the largest absolute signal value found in a test series of 60,000 samples (as specified above no value should be larger than 0.25).

It can be seen that the performance is very scenario dependant (varying orders of magnitudes). This can be reasoned as follows: for every frequency band that has a relatively high spectral density (i.e. the interferer and wideband signal frequencies) the filter should have an accurately defined phase response that approximates a negative group delay of 10 samples and an amplitude response of 1. The best fitting curve of a finite order FIR filter deviates relatively more from this response in certain cases (when e.g.

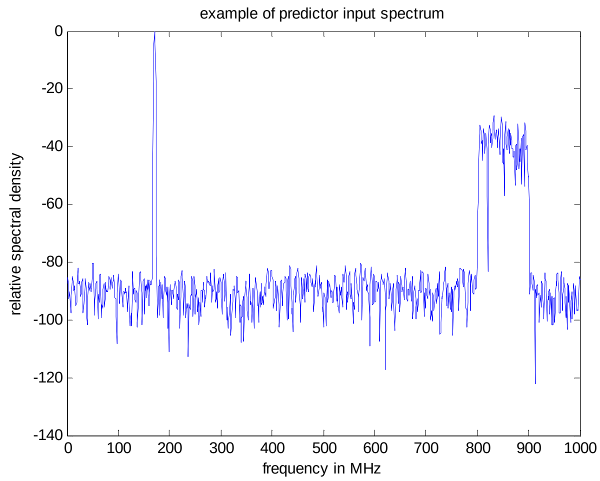

Fig. 4. Spectrum used for simulation (1 narrowband interferers and 1 wideband signal). Power difference between wanted signal and interference = 20 dB. The narrowband signal is shifted through the Nyquist band.

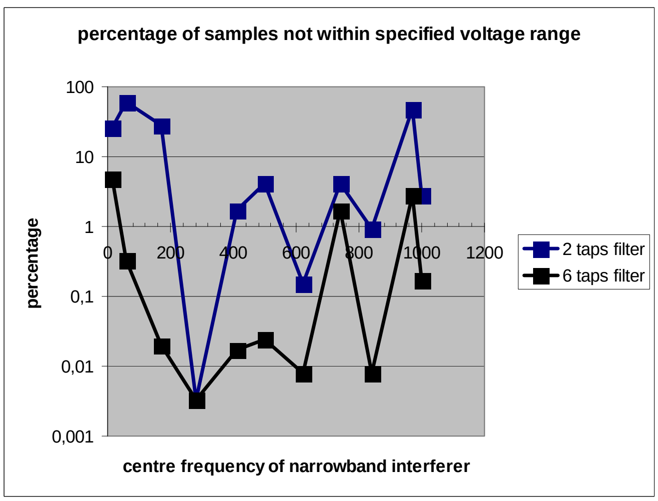

Fig. 5a. Prediction performance in terms of 'missed samples'

the interferer is either close to DC or the Nyquist frequency). If the FIR order N is higher, the fitting curve is more accurate, yielding a better performance. Furthermore a performance bottleneck is seen. This is caused by the relative power of wideband interference and noise (which cannot be predicted (well)) with respect to the narrowband interferer, as well as the quantization noise of the various components in the system.

When adding more filter taps and more bits of quantization accuracy for the tap coefficients, DAC and/or filter input, the predictor performance increases significantly at the cost of dissipated power (the filter has to run at a 2 Gs/s rate). Therefore a power optimum will be strongly dependant on the exact nature of the interfering signal.

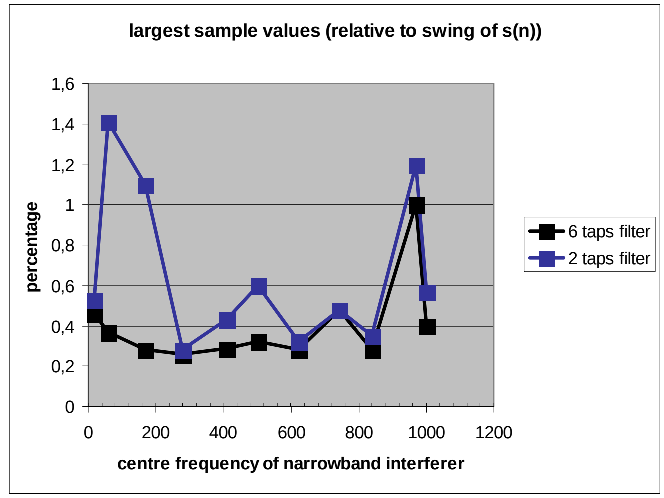

Fig. 5b. Prediction performance in terms of largest value (in a series of 60000 samples)

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