# ADAPTIVE SCHEME FOR OVERSAMPLED FRONT ENDS

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## ABSTRACT

A computationally efficient adaptive filtering scheme for oversampled A/D converters is discussed. The decimating digital filter is constructed as a combination of a sinc decimator and an adaptive predictor. A reduced-rank adaptive algorithm with two adaptive parameters is proposed for this purpose, avoiding the complexity of the commonly used full-rank algorithms. The characteristics of the adaptive filter are considered, and a comparative example of data signal processing is shown.

Keywords: Adaptive filters, multirate filters, oversampled converters.

## 1 INTRODUCTION

Oversampling techniques are commonly used in A/D and D/A converters, such as sigma-delta converters [1]. Most of the signal processing in oversampled converters is done in the digital domain, including anti-alias filtering. Thus, very little analog circuitry is needed, but the digital filters operate at a relatively high speed, and typically occupy a major portion of the chip area of the integrated converters. Multistage, multirate filters are typically used, taking advantage of computationally efficient structures to carry out decimation from the full oversampling rate to some intermediate sampling rate. Especially the averaging filter, with sinc function type frequency characteristics is popular as the first decimation stage.

The use of an adaptive predictor as the digital filter in oversampled all-digital sensors has been proposed [2]. An adaptive filter has the benefit of being capable of tracking narrowband signals corrupted by wideband noise. However, operating an adaptive filter at the full oversampling rate causes a large computational burden. Sigma-delta modulation has also been proposed for implementing the computations in an adaptive filter [3], avoiding actual multipliers but needing high-speed bit streams.

In this paper, we propose to construct the decimation filter for oversampled converters as a cascade of a fixed first decimation stage and a computationally efficient adaptive predictor, operating at the intermediate sampling rate. The filter structure and adaptation algorithm are described in Section 2 of this paper, and the frequency response is given in Section 3. An example of processing a modem signal is shown in Section 4.

## 2 ADAPTIVE DECIMATION FILTER

The structure of the proposed decimation filter is shown in Fig. 1. The sampling rate is lowered in two stages. The first decimation stage is a conventional averaging filter, consisting of $K$ running averagers, with sinc function type amplitude responses. Sampling rate is lowered by $M_1$ to the intermediate rate. In order to facilitate the use of a reduced-rank adaptive algorithm, the intermediate sampling rate is chosen to be relatively high, i.e., higher than four times the Nyquist rate. Such a choice also minimizes the baseband droop caused by the $\text{sinc}^K$ filter. The second filter stage is an adaptive predictor with prediction step $p$, followed by decimation to the output sample rate. The predictor operates in the line enhancer configuration, however, instead of the traditional LMS or RLS algorithm [4], a simplified adaptive algorithm is used.

The algorithm is a dual-parameter version of the so-called general parameter (GP) algorithm [5-8]. The filter output is computed as

$$y(n) = \sum_{k=0}^{N/2-1} [g_1(n) + h(k)]x(n-k)+$$

* This work was supported by the Academy of Finland under grant 51849.

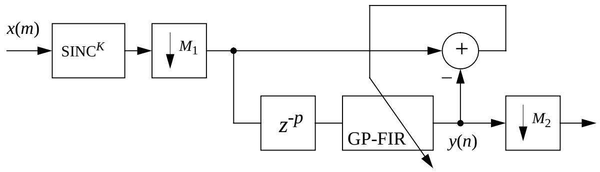


Figure 1: Structure of the decimation filter.

$$\sum_{k=N/2}^{N-1}[g_2(n)+h(k)]x(n-k), \qquad (1)$$

where the $h(k)$'s are the coefficients of a fixed basis filter, and the two adaptive parameters $g_1(n)$ and $g_2(n)$ are updated as follows:

$$g_1(n+1) = g_1(n)+\gamma[r(n)-y(n)]\sum_{k=0}^{N/2-1}x(n-k), \quad (2)$$

$$g_2(n+1) = g_2(n)+\gamma[r(n)-y(n)]\sum_{k=N/2}^{N-1}x(n-k). \quad (3)$$

$\gamma$ is a gain parameter, and $r(n)$ is the reference input. In many cases it is possible to operate without a fixed basis filter, i.e., we can set $h(k) = 0$ for $k = 0, 1, \cdots, N-1$. Using recursive running sum-based structures [8], the implementation of the dual-parameter adaptive filter then needs altogether only five multiplications.

## 3 FREQUENCY RESPONSE

Setting the length of the averaging filters to $M_1$ and the $h(k)$'s to zero, the instantaneous frequency response of the filter, as seen from the input, is given by

$$D(e^{j\omega}, n) = A(e^{j\omega})G(e^{jM_1\omega}, n), \qquad (4)$$

where

$$A(e^{j\omega}) = e^{-j(M_1-1)\omega}\Big[\frac{1}{M_1}\frac{\sin(M_1\omega/2)}{\sin(\omega/2)}\Big]^K,$$

and

$$G(e^{jM_1\omega}, n) = [g_1(n)e^{-jM_1\omega(\frac{N}{4}-\frac{1}{2})}+$$
$$g_2(n)e^{-jM_1\omega(\frac{3N}{4}-\frac{1}{2})}]\frac{\sin(NM_1\omega/4)}{\sin(M_1\omega/2)}.$$

As an example of the resulting shape of the amplitude response, we consider the case where $K = 2$, $M_1 = 8$, $N = 32$, $p = 0$, and $x(m) = \sin(0.005\pi m)$ (expressed in terms of the oversampling rate). The two degrees of freedom allow complete adaptation to the single-frequency sinusoidal input. Following decimation by eight, the dual-parameter adaptive predictor adapts to unity gain at $\omega = 0.04\pi$, as seen in the amplitude response of Fig. 2(a). The oversampled equivalent is seen in Fig. 2(b). Combined with the $\mathrm{sinc}^2$ decimator, the overall amplitude response is the one shown in Fig. 3.

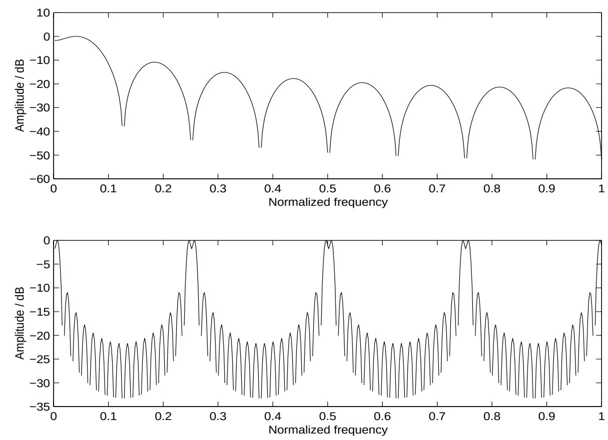


Figure 2: (a) Amplitude response of the adaptive predictor $G(z)$. (b) Amplitude response of $G(z^{M_1})$.

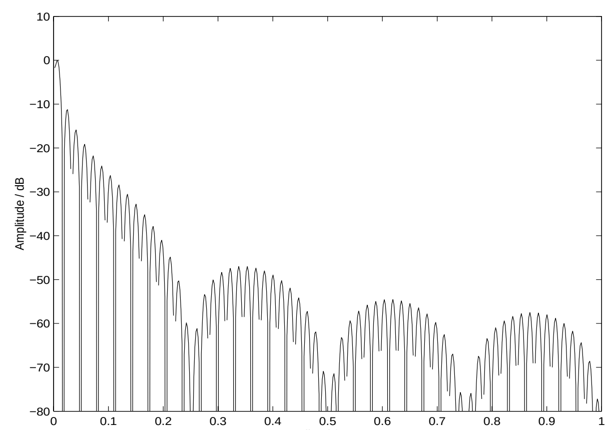


Figure 3: Amplitude response of the composite decimator ($M_2 = 1$).

Due to the characteristics of the frequency response (4), the length $N$ of the adaptive filter should be chosen such that the baseband width of the input signal does not exceed $2\pi/NM_1$. In other words, the oversampling factor should be at least $NM_1/2$. Another limitation of the structure is seen in the stopband attenuation. Regardless of the filter length, the first and highest peak of the stopband ripples never drops below -13 dB. The -13 dB value is reached when $g_1(n) \approx g_2(n)$. This peak is located at $6\pi/NM_1$, and is typically at -10 to -13 dB attenuation levels, depending on the adaptation, i.e., the values of $g_1(n)$ and $g_2(n)$. The rest of the stopband provides considerably higher attenuation. The width of the transition band from the passband edge to the stopband attenuation level is about $4\pi/3NM_1$.

## 4 COMPARATIVE EXAMPLE

Let us consider a data signal encoded in the 2B1Q (2 Binary 1 Quaternary) scheme. This signaling method is used, for example, on the ISDN Physical Layer in North America. The voltage level can be one of four distinct levels, -2.5 V, -0.833 V, 0.833 V, and 2.5 V. Thus, two bits are transmitted per baud. A simulated 2B1Q signal is shown in Fig. 4. The random symbol stream has been upsampled by 256 using a raised cosine filter with a rolloff factor of one, and superimposed white noise has been added in the signal.

### 4.1 Without Fixed Basis Filter

First we consider the case where $h(k) = 0$, $k = 0, 1, \cdots, N-1$. The front end signal processing consists of a second order sigma-delta modulator with single-bit quantization, a sinc$^2$ decimator with $M_1 = 8$, and a GP-based adaptive filter with $N = 32$. When the received noisy signal is passed through the front end, the resulting eye diagram is the one shown in Fig. 5. For comparison, the eye diagram given by a standard LMS adaptive filter of equal length is shown in Fig. 6, and the final amplitude responses of the two adaptive filters are shown in Fig. 7. Both algorithms fulfil the filtering task with nearly similar signal characteristics, but the GP-based algorithm needs only 15% of the multiplications of the full-rank LMS filter.

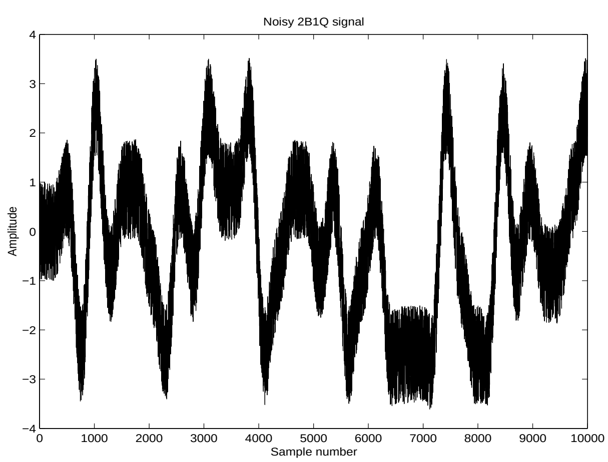


Figure 4: Noisy 2B1Q signal.

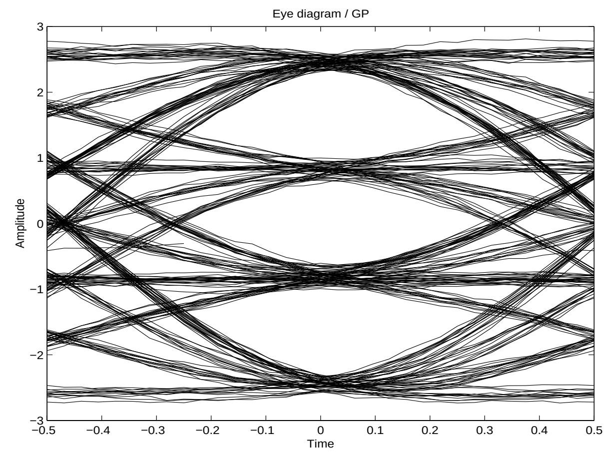


Figure 5: The eye diagram resulting from GP algorithm-based filtering.

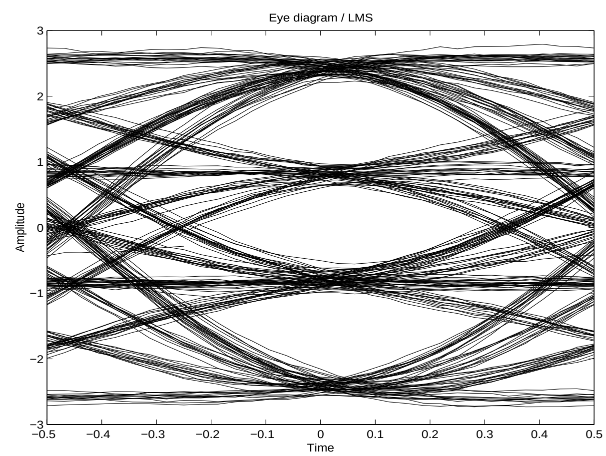


Figure 6: The eye diagram resulting from LMS algorithm-based filtering.

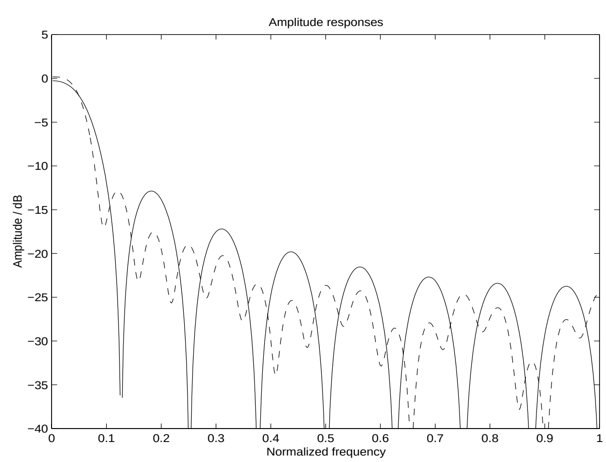


Figure 7: Amplitude responses of the GP filter (solid line) and the LMS filter (dashed line).

### 4.2 With Fixed Basis Filter

We repeat the experiment using a hybrid approach with a fixed basis filter, which means nonzero coefficients $h(k)$ in (1). The basis filter is designed as a least-squares lowpass filter with the bandwidth matching the symbol rate. The actual signal baseband is twice wider due to the rolloff factor of the raised cosine filtering. Thus, the basis filter is too narrowband to allow a high signal quality alone, but the adaptive extension provides compensation. The received eye diagram is shown in Fig. 8. Compared to Fig. 5, the eye opening is slightly larger. The amplitude responses of the composite filter and the basis filter are shown in Fig. 9.

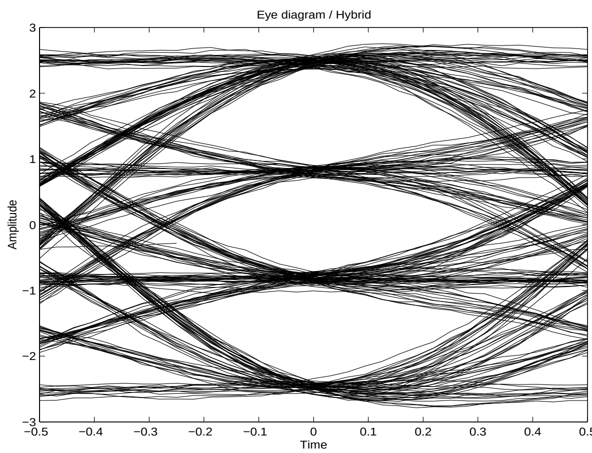


Figure 8: The eye diagram resulting from GP adaptive filtering with a fixed basis filter.

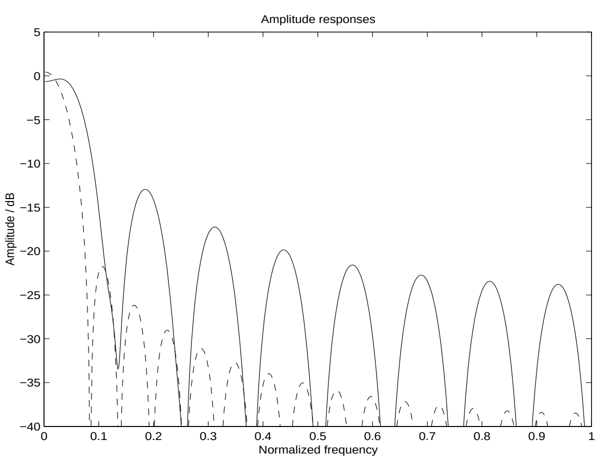


Figure 9: Amplitude responses of the composite adaptive filter (solid line) and the basis filter (dashed line).

Typically, the fixed basis filter is designed to be mostly responsible for the filtering task, and the adaptive part is used for tuning. To some extent, the GP algorithm can also provide signal equalization, especially for lowpass cases such as the decimation filters considered in this paper.

## 5 CONCLUSIONS

The computational simplicity of the dual-parameter GP algorithm allows one to construct lowpass type adaptive predictors for oversampled front ends even with a moderately high intermediate frequency. Due to the structural constraints of the approach, the selectivity is not as high as with a full-rank adaptive filter. However, the lowpass nature of the decimation filtering task is well suited for the GP type algorithm. For applications requiring higher resolution, the adaptive scheme can be supplemented by additional fixed stages for enhanced stopband attenuation and droop compensation.

### Acknowledgment

The author would like to thank H. Lampinen for the sigma-delta modulator model.

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