Experimental Validation of a Tuning Algorithm for High-Speed Filters

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Abstract

We report here the results of some laboratory experiments performed to validate the effectiveness of a technique for the self tuning of integrated continuous-time, high-speed active filters. The tuning algorithm is based on the application of a pseudo-random input sequence of rectangular pulses to the device to be tuned and on the evaluation of a few samples of the input-output cross-correlation function which constitute the filter signature.

The key advantages of this technique are the ease of the input test pattern generation and the simplicity of the output circuitry which consists of a digital cross-correlator.

The technique allows to achieve a tuning error mainly dominated by the value of the elementary capacitors employed in the tuning circuitry. The time required to perform the tuning is kept within a few microseconds. This appears particularly interesting for applications to telecommunication multi-standard terminals.

The experiments regarding the application of the proposed tuning algorithm to a baseband multi-standard filter confirm most of the simulation results and show the robustness of the technique against practical operating conditions and noise.

1. Introduction

Continuous-Time (CT) filters are widely used in signal processing but generally require a tuning system to keep under control their frequency performance.

Basically, tuning an integrated active filter consists in adjusting some parameters of its frequency response (e.g.: cut-off frequency, in-band ripple, static gain, etc.) to their nominal values within fixed tolerances. In particular, in CT filters the time-constants are defined by uncorrelated components (gm/C or RC).

The values of these constants may substantially differ from the nominal ones especially in deep sub-micron CMOS implementations, owing to the large spread in process parameters which characterize these technologies. But the need of compensating for component variations from their nominal values arises also from other effects connected to aging, temperature, etc.

Present tuning techniques can be grouped into two basic schemes: master-slave and self-calibration arrangements [1-5]. They can be compared in terms of their accuracy which strongly depends on the tuning circuit implementation. A key feature of the complete filtering systems is their implementation in scaled-down technologies. In fact, it is possible to compensate for the poor matching characteristics of their internal devices by using low-cost, small-area, low-power consumption digital blocks. This results in mixed signal tuning systems, where the number of bits of the control word affects the achievable system accuracy, whereas in pure analog tuning systems this could be limited by typical analog circuit non-idealities (gain, offset, etc.).

Another important issue concerns the input pattern to be used, which is strictly correlated to the kind of algorithm embedded in the tuning procedure. In pure analog systems, a single tone or a dc voltage are commonly used. The use of a single tone results in a measure of the effective filter frequency response at a given frequency. However, measurement noise can affect the tuning accuracy and besides that, a high resolution ADC is needed to sample the output. On the other hand, the use of a dc-voltage does not include the measurement of a frequency response and it is strongly dependent on dc errors like offset, etc. In addition, in both cases the accurate generation of the pattern signal could be critical and results in an increased system complexity.

We refer here to a recently proposed tuning approach, based on the use of a pseudo-random test pattern input signal and on the evaluation of a few samples of the cross-correlation between the pseudo-random test pattern and the corresponding filter response [6-7]. This information is used to properly adjust the values of the time constants which affect the cut-off frequency of the filter through an array of capacitors which can be switched on or off by means of MOS switches.

The proposed technique offers several advantages over more commonly adopted ones. First, the pseudo-random input pattern signal can be generated by a very simple circuit in a small die area. Moreover, it is robust against

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typical measurement errors (noise, clock jitter, etc) thanks to the averaging properties of the cross-correlation function. Last, the circuitry required for the evaluation of the cross-correlation samples needed by the tuning scheme is very compact [6-7].

The aim of this paper is to describe a set of laboratory experiments performed to assess the validity of the proposed technique in a real world operating environment, taking into account some practical limitations such as: the number and the minimum value of the elementary capacitances used to perform the tuning, the maximum achievable sampling frequency of the filter response, the finite length of the pseudorandom sequence, the width of each rectangular pulse, the effects of background noise, etc.

The paper is organized as follows: a short description of the tuning algorithm, together with an analysis of some practical implementation issues, is reported in Section 2. Section 3 deals with the description of the measurement set-up and with the application of the technique to a high-speed multi-standard UMTS filter for telecommunication applications. Finally, section 4 shows the results of a number of laboratory experiments performed on the prototype system described in section 3, implemented by an FPGA and a filter test board.

2. Essentials of the tuning algorithm

For a Linear Time Invariant (LTI) circuit, the tuning parameters are directly related to its impulse response h(t) and the tuning process can be accomplished through the knowledge of a suitable approximation of h(t) [8]. In particular, it has been shown [9] that the main circuit specifications can be related to a limited number of samples of h(t).

Using a single Dirac's pulse approximation to evaluate the impulse response of the filter would result in a circuit response easily corrupted by noise since the energy associated to a short pulse is severely limited by the input linearity range of the circuit [9]. This can be overcome by using a signal with the same white noise spectrum as input stimulus. A pseudorandom pulse sequence of suitable length features this property, i.e. it has auto-correlation function $R^{xx}(t)$ which is a single pulse $\delta(t)$ [10].

In a typical implementation of the cross-correlation algorithm, x(t) is a finite length sequence of L rectangular pulses of constant width Δt and whose amplitude can assume a positive or negative value with same probability.

In the real case, the finite length of the sequence introduces a tail in the auto-correlation function $R^{xx}(t)$ of the input stimulus and this affects the accuracy of the estimated h(t), especially for large values of t. As a consequence the width of the single pulse Δt and the length L of the sequence must be carefully chosen, depending on the bandwidth of the circuit and on the accuracy of the estimation of h(t) required by the tuning operation [9]. Moreover, the power density spectrum of $R^{xx}(t)$ exhibits the first zero at $f_0=1/\Delta t$, thus, for the accurate tuning of a filter with cut-off (LP) or central (BP) frequency f_c, f_0 should be chosen conveniently greater than f_c. We usually assume $f_0 \cong 5f_c$, so that the width of each pulse of the pseudo-random sequence is: $\Delta t \cong 1/5f_c$.

In practice, a Linear Feedback Shift Register (LFSR) with a suitable number of stages is used to generate the pseudo-random sequence.

The practical implementation of the cross-correlation algorithm introduces other sources of approximation in the estimation of h(t). In the typical scheme used for the on-line tuning, shown in fig. 1, the filter is embedded between a DAC and an ADC. Although this general scheme could appear somewhat complicated to be realized in an integrated system, in practice it can be greatly simplified making it very attractive. First of all, the pseudo-random pattern signal generator delivers a two-level bit-stream, thus a one-bit DAC, implemented by two switches connecting the filter input node to either of two reference voltages, is employed. Secondly, for the output signal, a very low resolution ADC may be employed as the quantization errors are averaged by the cross-correlation operation. Even for a moderate length input sequence, a 1-bit ADC, i.e. a simple comparator, can be employed. This is of particular relevance when tuning high speed filters where a high sampling rate is needed.

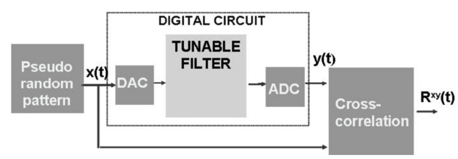

Fig. 1. Evaluation of the cross-correlation function.

Concerning, the output sampling frequency f_s, since both the input sequence and the filter response are sampled signals and the number kL of samples ($k=f_s/f_0$) can only be a finite number, in the expression:

$$R^{xy}(m) = \lim_{L\to\infty} \frac{1}{kL} \sum_{n=0}^{L-1} x(n)y(n+m)\,, \qquad (1)$$

the higher the number L, the better the estimation of R^{xy}. Of course, the number kL of output samples to be used to evaluate the cross-correlation R^{xy} is strictly related to the output sampling frequency f_s. Thus the sampling frequency of the ADC should be greater than $1/\Delta t$ due to the need of getting a suitable number of significant samples of R^{xy}. In other words, the higher the value of k, the closer the spacing between adjacent R^{xy} samples.

A key issue in the tuning algorithm is the choice of the set of samples of the cross-correlation function $R^{xy}(m_i)$ assumed as circuit signature. This, in general, requires a study of the sample sensitivities with respect to the filter specifications s_j which is beyond the scope of this paper and is reported in [10-11].

To perform the filter tuning, usually, only its cut-off frequency has to be adjusted. In this case a single sample of the R^{xy} function can be assumed as filter signature [10] and the cut-off frequency alignment can be done by using an array of binary-weighted capacitors which can be selected through a set of MOSFET switches. The capacitor array, whose total capacitance is indicated by C_{array}, consists of a fixed capacitance C_{off} and N binary weighted elements, the smallest of which has a value denoted δC. The capacitor array is addressed by a N bit digital code. The total capacitance value achieved for a given tune code is:

$$C_{array} = C_{off} + n \cdot \delta C , \qquad (2)$$

where n is an integer in the range $[0, (2^N-1)]$.

The procedure for the tuning of the cut-off frequency consists in evaluating the signature sample of the cross-correlation and comparing it with the corresponding expected value associated to the nominal filter behaviour. Then, on the basis of the result of this comparison, one δC at a time is added or subtracted until the minimum error between the actual and the nominal values of the signature is obtained.

3. Experimental set-up description

In order to find out more about the practical issues involved in the real implementation of the described tuning technique, an experimental set-up has been arranged, following the block diagram depicted in fig. 2. The diagram has been split into two parts: the upper one represents a board which includes the analog tunable filter and the interfaces needed to apply the pseudorandom sequence to the input of the filter and to convert its output in the digital domain. The bottom part of the diagram shows the main digital blocks which have been used for the estimation of the signature of the filter, including the LFSRs needed to generate the pseudorandom sequence and its suitably delayed replica, an up/down counter and a Finite State Machine (FSM) for the configuration of the tuning capacitors through an array of switches. The digital part has been implemented by means of an FPGA, which allows to easily vary, for instance, the number of stages of the LFSRs and, as a result, the length of the input sequence.

The benchmark filter selected for the experiments is a multistandard anti-aliasing filter for base-band receivers [12]. The filter tasks are the base-band anti-aliasing of the A/D sampling frequency, the filtering of the intermodulation interferers, necessary to reduce the A/D dynamic range, and the partial attenuation of the adjacent channel. For these applications, a typical low-pass, low-Q (<2) pole transfer function is required. Our benchmark

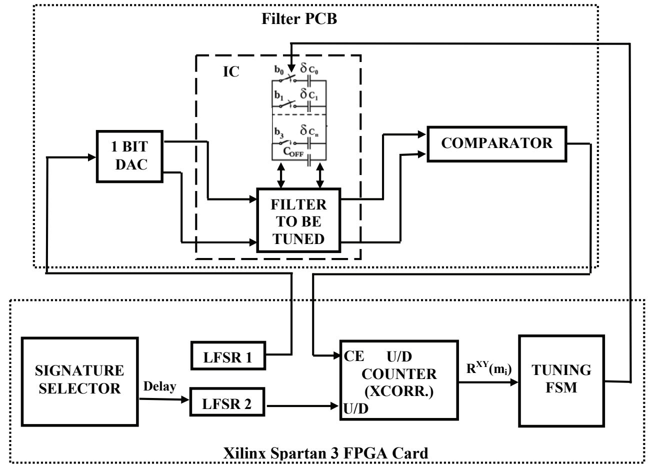

Fig. 2. Experimental tuning set-up.

filter has a 4th order low-pass Bessel transfer function with a 4 dB DC-gain and a cut-off frequency of respectively 2.11, 1 and 11MHz, according to the selected standard UMTS, GSM and WLAN. Other filter requirements (linearity, noise, etc.) do not concern this discussion and will not be addressed here. The basic filter building block is an RC-active, multipath biquadratic cell, thus the whole filter consists of the cascade of two multipath biquadratic cells in their fully differential form.

As described above and outlined in fig. 2, the fine variation of the cut-off frequency of the filter is possible by means of configurable arrays of binary-weighted capacitors. The capacitor array can be configured by a N-bit digital code in order to compensate for the spread in the time constants resulting from the process parameter fluctuations. In our case 4 bits are available to address the capacitor array, i.e. the filter has 16 possible configurations corresponding to 16 different cut-off frequencies. The tuning procedure must be able to select the configuration corresponding to the cut-off frequency closest to the nominal one.

The filter structure requires the pseudorandom input sequence to be fully differential, with a suitable common mode value (1.25V) and must have a zero average value, therefore a simple DAC must be put between the digital pseudorandom sequence and the filter input. For this purpose a broad-band transformer, with a central tap on the output coil biased at 1.25V, has been used. The 1-bit ADC used to convert into the digital domain the filter response is a simple fast comparator placed at the fully differential output of the filter, followed by a flip-flop to synchronize the output bitstream with the system clock. A picture of the analog board which accommodates the chip containing the tunable filter, the transformer, the comparator and a buffer, used to monitor the analog response of the filter with an oscilloscope without introducing further parasitics, is shown in fig. 3.

As already mentioned, an FPGA has been used to implement the digital part of the experimental set-up, so

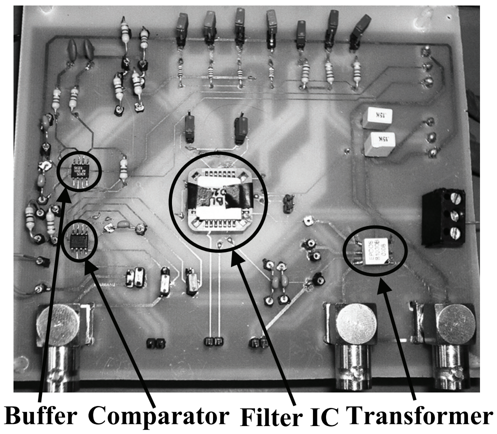

Fig. 3. Filter board.

as to provide maximum flexibility in the choice of the experimental parameters. Two variable length LFSRs provide respectively the pseudorandom sequence used as input stimulus of the filter and a delayed version of the same sequence, used to evaluate a single sample of the input-output cross-correlation function. An up/down counter is employed to calculate the selected R^{xy} sample. The UP/DOWN signal is the output of the delayed LFSR, whereas the counting input CE is the bitstream provided by the 1-bit ADC. In this arrangement, the counter output accumulates the product between the 1-bit digitized output and the delayed input, according to eq. (1). Notice that, with this solution, the delayed input sequence is equivalent to a succession of positive (UP/DOWN=1) and negative (UP/DOWN=0) values, thus preserving the zero average requirement for the input. The counter and the flip-flop which synchronizes the comparator output are operated at a clock frequency f_s which is a multiple of the LFSR clock frequency f_0. The ratio $k=f_s/f_0$ can be easily varied exploiting a PLL based facility which is available inside the FPGA for clock management purposes.

The signature selector block in fig. 2, which sets the delay between the two LFSRs, operates also as a control unit for the up/down counter, stopping the counting when the number of the filter output samples needed to evaluate the signature has been reached, thus avoiding further contributions due to the fact that the LFSR sequence used as input of the filter starts again after N clock cycles. The R^{xy} sample used as a signature of the filter is then compared to the nominal one and, on the basis of its actual value, the programmable capacitors are increased or decreased, according to the described tuning procedure.

It is important to observe that the described experimental set-up does not allow to operate at very high frequencies, mainly due to the limited bandwidth of the connections between the analog board and the FPGA. On the other hand, the choice of an FPGA flexible implementation of the digital part makes possible to easily vary some of the crucial parameters of the tuning technique, such as the length of the pseudorandom input sequence, the width of the single input pulse and the output sampling frequency. In this way, among the other things, a deep investigation about the trade-off between the length of the input sequence and the quality of the achieved signature in terms of tuning efficiency can be carried out at no cost.

4. Experimental results

Using the system described in the previous section we performed some experiments to prove the effectiveness and the robustness of the tuning technique against the non-idealities introduced by the actual operating conditions, such as noise, clock jitter and distortion of the input pulses. The filter has been configured for the UMTS standard, which requires a nominal cut-off frequency f_c equal to 2.11 MHz. First, the frequency response of the filter has been measured for all the 16 possible configurations of the programmable capacitors, so that the configuration nearest to the nominal performance of the filter has been identified, which features a cut-off frequency of about 2.09 MHz. Next, some measurements have been carried out to verify the reliability of the behaviour of the whole system. Fig. 4 shows the analog response of the filter to the real pulse sequence applied to the input, whereas fig. 5 illustrates the square-waved behaviour of the comparator output.

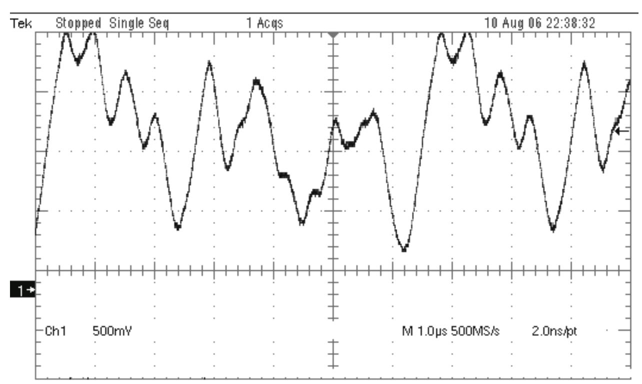

Fig. 4. Waveform of the filter output.

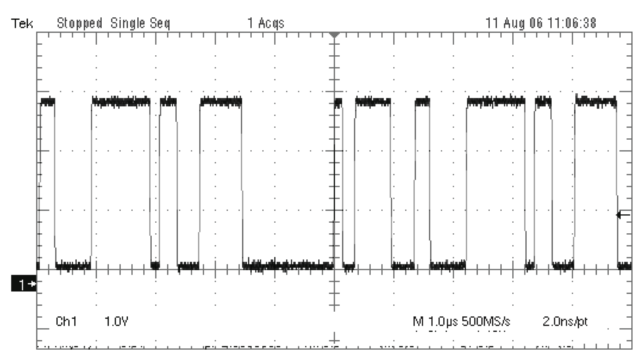

Fig. 5. Waveform of the comparator output.

With the "nominal" configuration selected, a comparison has been done between the simulated impulse response of an ideal version of the filter used in the experiments and the measured input-output cross-correlation functions obtained using two different lengths of the input sequence, respectively L=63 (6-bit LFSRs) and L=511 (9-bit LFSRs). In these measurements the width of the single pulse has been set to $\Delta t=1/5f_c$ and the factor k is equal to 6. Fig. 6 reports the results of the comparison between the normalized functions, showing that the system is able to correctly extract the R^{xy} function of the filter and that the experimental results are in agreement with the theory, since the longer sequence provides a more accurate estimation of the impulse response of the circuit as compared to the shorter one.

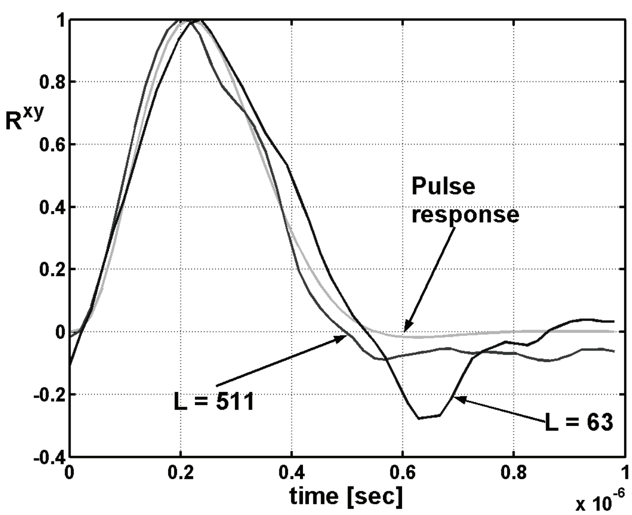

Fig. 6. Measured, normalized cross-correlation functions obtained with input sequences of different lengths.

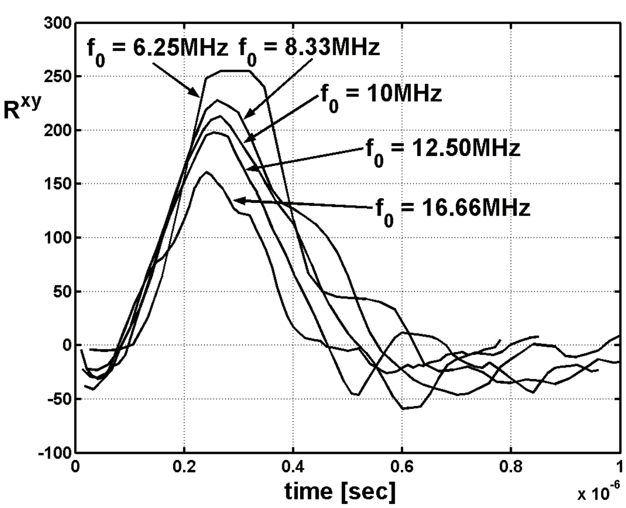

Fig. 7. Measured cross-correlation functions obtained with input sequences of the same length (L=63) and different f_0.

To assess the effect of the choice of a different value for the parameter Δt, input sequences with the same length L=63 but with different frequency f_0 have been used to extract the R^{xy} function of the filter. Fig. 7 reports the results of such measurements, showing that if a f_0 value too different from $5f_c$ is chosen, the cross-correlation function behaviour diverges from the filter impulse response, possibly producing unreliable filter signatures.

Setting $\Delta t \cong 1/5f_c$, L=63 and k=6, further investigations have been done to verify that the R^{xy} extracted by measurements can be used as an effective signature for the cut-off frequency of the filter. In fig. 8 the cross-correlation functions obtained with the mentioned operating conditions are reported for all the possible configurations of the filter, showing that several R^{xy} samples can be successfully adopted as a signature of the circuit, since they exhibit an appreciable variation when the configuration is changed. As a matter of fact, in our tuning scheme the effectiveness of a signature can be guaranteed only if its behaviour is monotonic with respect to the specification to be tuned, i.e. the cut-off frequency.

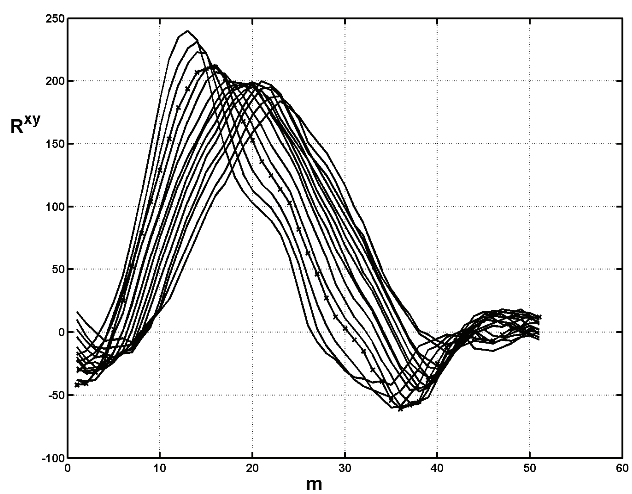

Fig. 8. Measured cross-correlation functions (L=63, f_0=10MHz, k=6).

To make sure that this requirement is fulfilled, among all the possible choices, the nominal R^{xy} zero-crossing sample has been assumed as a signature. On one hand, this choice simplifies the tuning procedure, since the measured signature must be simply compared to zero in order to decide if an elementary capacitor must be added or switched off in order to get closer to the nominal f_c value. On the other hand this particular sample of R^{xy} is in practice independent of the DC-gain of the filter, thus making the cut-off frequency adjustment insensitive to small variations of the DC-gain. Fig. 9 reports the behaviour of the chosen signature as a function of the filter configuration, for the cases L=63 and L=511. Notice that two adjacent points in the graph of fig. 9 correspond to a single elementary capacitor δC switched off or on, therefore the x axis of the graph represents also the cut-off frequency of the filter which monotonically increases. Fig. 9 demonstrates that the chosen signature can be effectively used to tune the filter in both cases L=63 and L=511, the only difference between the two situations being a better sensitivity of the signature in the case of the longer sequence. In our case L=63 is sufficient to conveniently tune the cut-off frequency and the only limitation to the accuracy of the performance achieved after the tuning procedure is related to the value of the elementary capacitor δC used, which cannot be made too small compared to parasitics related to the employed technology.

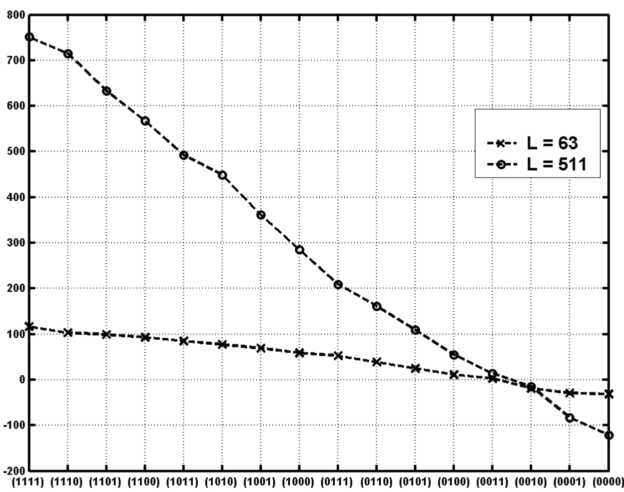

Fig. 9. Signature sample vs tuning configurations.

5. Conclusions

The use of a pseudo-random pattern to perform a start-up tuning of CT active filters has been experimentally validated on an actual high-speed multi-standard anti-aliasing filter.

The cross-correlation technique has demonstrated to be more effective than the direct application of a single short pulse in the estimation of the impulse response since the pseudorandom sequence has a greater associated energy which makes the filter response less sensitive to noise interference. Moreover, the amplitude of the individual pulses can be chosen so as to match the input linearity constraints of the filter, thus limiting also the effects of non-linearity distortion.

On the basis of the experiments performed, the technique has proven to be robust against typical measurement errors such as noise, clock jitter, etc., thanks also to the averaging properties of the cross-correlation function.

The overall tuning accuracy is essentially dominated by the value of the smallest capacitor employed, whereas the accuracy of the algorithm is essentially determined by the length of the pseudorandom sequence.

Concerning CT high-Q filters, more sensitive to component mismatch then low-Q ones, the tuning procedure involves the adjustment of individual parameters of each filter section. This is accomplished by using more R^{xy} samples as filter signature. Further simulation experiments show that high-Q CT filters call for higher accuracies achievable with longer input sequences.

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