Transactions Brief

Multistage Bandpass Delta Sigma Modulators

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***Abstract*—A new architecture for oversampled delta sigma A/D conversion of high-frequency narrow band signals using cascaded low-order stages to obtain high overall order of noise shaping is described. The architecture involves using resonators in individual stages to suppress quantization noise at the resonant frequency. Many of the problems of previous single stage high-order architectures including poor stability, large component spread and design complexity are overcome by this new approach. Switched capacitor resonator circuit implementations and simulation results for a sixth-order modulator example are included. Estimates of the potential for 16-bit A/D conversion of 2.5 MHz signals exceed the 80 kHz capability of existing monolithic oversampled CMOS A/Ds.**

I. Introduction

Recently, oversampled delta sigma modulation has been extended from lowpass to bandpass signal frequency ranges [1]–[5]. This potentially enables oversampled A/D conversion of higher frequency signals than was previously possible in the lowpass case. With a lowpass modulator the maximum signal frequency that's converted is at a frequency of $F_s/(2R)$ where F_s is the modulator sampling rate and R is the oversampling rate, whereas a bandpass modulator can convert signals right up to frequencies of $F_s/2$. However, the bandwidth of the bandpass modulator is still limited to $F_s/(2R)$, the same as the lowpass modulator. Due to its ability to handle high-frequency narrow-band signals, the bandpass modulator is well suited for A/D conversion of IF signals that arise, for instance, in radios and modems.

Previous work on bandpass modulators has employed single-quantizer loop structures based on earlier work on high-order, lowpass modulators [1], [2], [5], [16]. These networks have the disadvantage of only conditional stability, and of large component spread and relatively high design complexity, i.e. design and realization of a high-order elliptic filter. An alternate architecture for lowpass modulators is cascaded connections of lower-order modulators [6]–[14]. If each stage, in this configuration, is second-order or lower, then the entire high-order modulator is unconditionally stable. In addition, the cascaded architecture tends to result in low component spreads, often with integer component ratios. A further advantage of cascaded networks is that higher overall resolution can be obtained, without sacrificing linearity, by simply using multi-bit quantization, instead of single-bit quantization, in the final stage [13], [14].

These significant advantages have motivated the work reported in this paper, namely the extension of lowpass cascaded modulators to the bandpass case. This new class of modulators is arrived at principally through the substitution of resonators for the integrators used in the lowpass modulators [3], [4]. Additional changes are needed in the transfer functions of the coupling networks that combine the separate outputs of the multiple stages into a single digital signal to handle the bandpass case. It is convenient to operate bandpass modulators at one quarter the sampling rate, i.e., $F_S/4$, since this allows for a very simple resonator to be used. It also allows single-loop structures of fourth-order to be implemented with unconditional stability. The following sections discuss various cascaded modulator networks and provide simulations results for a specific example.

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II. Cascaded Bandpass Modulators

The resonator used in these modulators has a z domain transfer function of

$$H_R(z) = \frac{z^{-2}}{1+z^{-2}} \tag{2.1}$$

This resonator has a pair of complex conjugate poles at $z = \pm j$ giving rise to resonance at $F_S/4$. In contrast, an integrator has a single pole at dc ($z = 1$).

In a lowpass cascaded modulator of order M, the noise-shaping frequency response is sinusoidal as shown in the following equation [10].

$$|H_Q(e^{j\omega T})| = g|2\sin(\omega T/2)|^M \tag{2.2}$$

where g is a scaling coefficient dependent on the architecture. In contrast to this, the noise-shaping frequency response for a cascaded $F_s/4$ bandpass modulator of order $2M$ is

$$|H_Q(e^{j\omega T})| = g|2\cos(\omega T)|^M \tag{2.3}$$

The cosinusoidal response arises here since the noise shaping zeros of $F_S/4$ bandpass modulators are at $\omega T = \pm\pi/2$ instead of at dc. It can be shown that the number of bits resolution of the bandpass modulator is

$$\text{Bits} = (M+1/2)\log_2 R - \log_2\left[\frac{g\pi^M}{\sqrt{2M+1}}\right] + \log_2(2^Q-1) \tag{2.4}$$

where Q is the number of bits in the quantizer of the final stage. This same expression applies for a lowpass modulator of half the order of the bandpass modulator [11].

A. Multiple Second-Order Resonator Cascades

The multiple second-order cascade structure, shown in Fig. 1, is derived from the lowpass multiple first-order cascade structure [7]–[10]. It has a z domain transfer function of

$$Y(z) = z^{-2N}X(z) + g_N(1+z^{-2})^N Q_N(z) \tag{2.5}$$

where $g_i = 1/(j_2\cdots j_i), h_i = 1/k_i, N$ is the number of stages and $Q_N(z)$ is the noise from the quantizer of the final stage. Scaling coefficients g_i and h_i are used to scale signal levels to avoid clipping. Since each stage is second-order, the order of the cascaded modulator is $2N$. An alternative multiple second-order cascade structure using the resonator signal R_i of the preceding stage as the input to each of the following stages instead of the quantization error signal is also feasible [8], [10]. This alternate approach, borrowed from lowpass modulators, simplifies the implementation somewhat.

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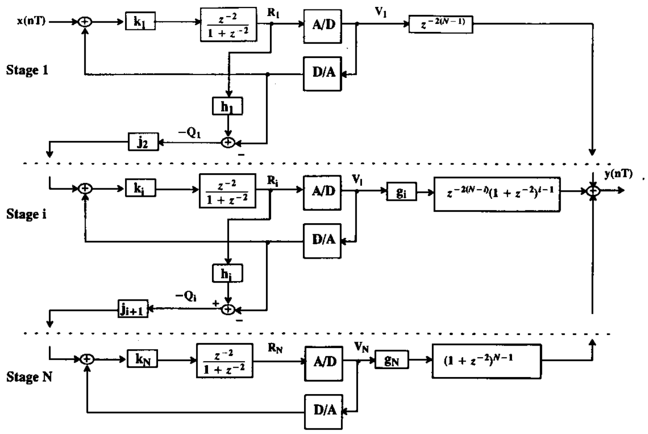

Fig. 1. Multiple second-order resonator cascade modulator.

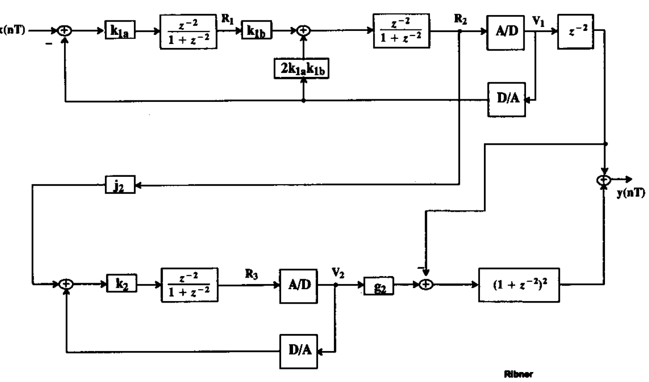

Fig. 2. Fourth-order second-order resonator cascade modulator.

B. Fourth-Order Second-Order Resonator Cascades

Fig. 2 shows a sixth-order bandpass modulators consisting of a fourth-order modulator followed by a second-order modulator. This modulator, derived from a lowpass second-order first-order cascade modulator [9]–[14], has the benefit of considerably lower sensitivity to component mismatch, and finite op amp gain [10]. The z domain transfer function of this bandpass modulators is

$$Y(z) = z^{-6}X(z) + g_2(1 + z^{-2})^3 Q_2(z) \tag{2.6}$$

where $g_2 = 1/(j_2 k_{1a} k_{1b})$ and $Q_2(z)$ is the noise signal from the quantizer in the second loop. In constrast to the modulator in Fig. 1, this modulator applies the resonator signal from the first stage directly as the input to the second stage. It is also possible to use the quantization noise signal of the first stage as the input of the second stage.

Since the first stage of these cascaded modulators is a single-quantizer-loop structure of fourth-order, the question of stability arises if the quantizer is two-level. Normally modulators above second-order are unstable with a two-level quantizer. This network, however can be viewed as a two-path network that maps z to $-z^2$ by simply inverting alternate samples [3]. Specifically, the bandpass delta sigma modulators here, with cosinusoidal noise shaping, are equivalent to two interleaved lowpass delta sigma modulators, with sinusoidal noise shaping, with interleaved outputs, alternately negated. Therefore the fourth-order, single-loop bandpass modulator is stable since a second-order lowpass modulator is stable.

C. Multiple Fourth-Order Resonator Cascades

It is also possible to develop high-order bandpass modulators using a cascade of N fourth-order modulators. This results in an overall order of $4N$. Fig. 3 shows one type of multiple fourth-order cascade in which the inverted quantization error of a given stage is used as the input to the following stage. In this configuration $g_i = 1/(j_2 \cdots j_i)$ and $h_i = 1/(k_{ia} k_{ib})$. Another implementation, where the resonator signal of a stage R_i is used as the input to the next stage, is also possible. From a practical standpoint, it's doubtful that versions beyond eighth-order, i.e., above two stages, will come close to the theoretical performance due to component mismatch, except at very low oversampling ratios.

III. Resonator Implementations

Implementation of precision $F_s/4$ resonators is of crucial importance to the success of any of these bandpass modulator networks. In its simplest form, this resonator can be implemented as a double analog delay in a negative feedback loop. This can be realized using sample and holds for the delay elements [5]. Another approach is to use 2-path integrators to implement the resonators, however, this approach is sensitive to op amp gain and/or capacitor mismatch depending on the specific circuit [3], [4]. The approach suggested here is to implement the resonator using two switched-capacitor (SC) integrators, coupled in a resonant loop [2]. Although the two-path approach is superior in accuracy of resonant frequency, the coupled

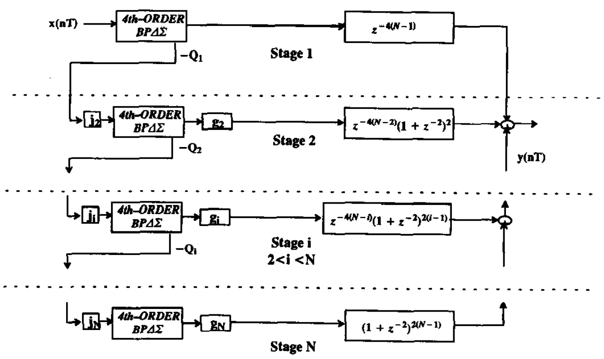

Fig. 3. Multiple fourth-order resonator cascade modulator.

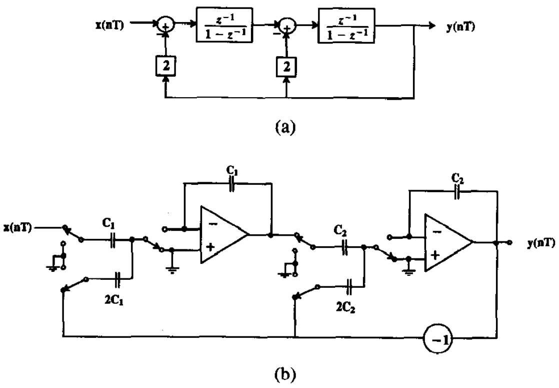

Fig. 4. Resonator using unit-delay integrators.

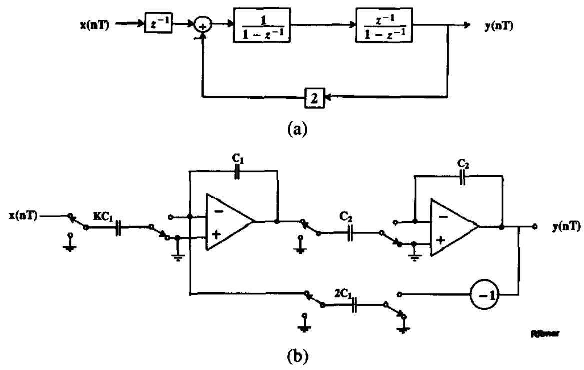

Fig. 5. Resonator using LDI biquad.

integrator circuit is expected to have better resonator quality-factor (Q) sensitivity to circuit nonidealities such as component mismatch, finite op amp gain and bandwidth. In narrowband applications the capacitor matching of the coupled integrator resonator needs to be good to avoid lowering the signal-to-noise ratio (S/N). The degradation from ideal performance due to a relative resonator frequency error of ε for the sixth-order modulator of Fig. 2 is to a first approximation

$$\Delta\text{MSB(dB)} = 10\log_{10}\left[1 + \left(\frac{2\varepsilon}{g_2}\right)^2 \frac{7}{3}\left(\frac{R}{\pi}\right)^4 \left(\frac{\sigma_{Q1}}{\sigma_{Q2}}\right)^2\right] \quad (3.1)$$

where σ_{Q1}/σ_{Q2} is the ratio of the quantization noise of the first stage to the second stage and is approximately equal to 1.5 over the passband. Two resonator approaches using SC integrators are described next.

A. Unit Delay Integrator based Resonator

A block diagram of a resonator using unit delay integrators is shown in Fig. 4(a), and a possible single-ended SC implementation is shown in Fig. 4(b). The signal inversion shown in the loop can be realized in a differential structure by reversal of signal connections. In a single-ended structure, the inversion can be realized by changing the clock phasing of one of the SC integrators to make it inverting.

B. LDI Biquad Resonator

Another resonator implementation makes use of the lossless digital integrator (LDI) biquad structure without damping [15]. The block diagram shown in Fig. 5(a) consists of a two integrator loop with a total loop delay of z^{-1}. This structure is simpler and less sensitive to component mismatch than the preceding network. Error in the integrator gains here only result in an error in the resonant frequency. In the other circuit, the same types of errors can additionally lower the Q of the resonator. A possible single-ended switched-capacitor implementation for the LDI resonator is shown in Fig. 5(b). The earlier comments about signal inversion apply here equally well.

IV. Simulation Results

Simulation results for the sixth-order cascaded modulator using the fourth-order, second-order cascade structure of Fig. 2 along with the LDI resonator are shown in Fig. 6 for $j_2 = 1$, $k_{1a} = k_{1b} = k_2 = 1/2$ and $g_2 = 4$. This plot shows the S/N versus the input level at an oversampling ratio $R = 64$ for a sinusoidal input signal at a frequency of 0.25002 F_s (0 dB is full-scale). The solid line shows performance in the ideal case whereas the dashed line shows the result with a 0.2% resonator frequency error. The peak S/N of 94.5 dB reached in the ideal case drops by 5.3 dB with the mismatch. Over the entire curve the average drop in S/N is 3.2 dB close to the 2.8 dB predicted by (3.1) Similar simulations have been performed for third-order lowpass modulator using the analogous second-order first-order cascade structure and virtually identical ideal performance was obtained in [10, Fig. 10].

V. Conclusion

The modulators described here are a useful extension to cascaded modulators already widely used for lowpass modulators. These modulators extend the same benefits to bandpass delta sigma modulation their counterparts did to lowpass delta sigma modulation, namely unconditional stability, low component ratio spreads, and amenability for multi-bit quantization without loss of linearity. One disadvantage to these networks, in certain applications, may be the restriction of operation centered at one fourth the sampling rate. A future paper will report on an approach to extend cascaded modulators to operation at arbitrary center frequency. Finally, extrapolating from previous lowpass modulator results [12], it is estimated that 16-bit A/D conversion of narrowband 2.5 MHz signals at a conversion rate of 10 MHz. is feasible, exceeding the capability of conventional monolithic CMOS A/Ds.

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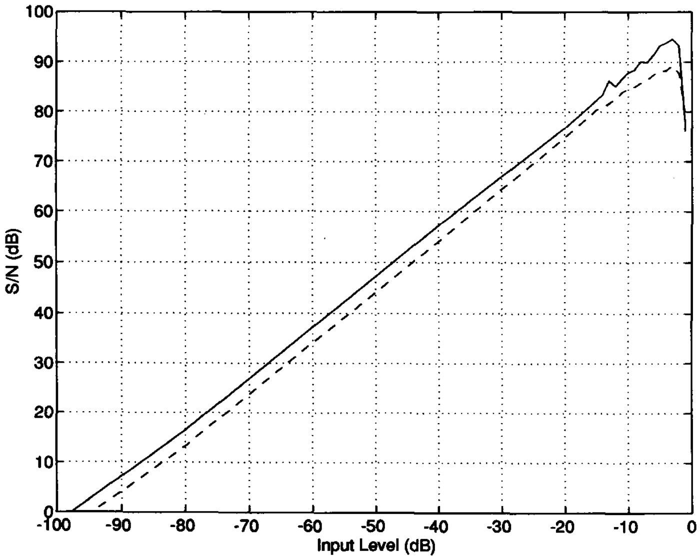

Fig. 6. S/N versus input level for sixth-order modulator of Fig. 2.

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Wide-Area Adaptive Active Noise Cancellation

A. J. Efron and L. C. Han

Abstract—**This paper describes an adaptive active noise cancellation (ANC) system that is based on an output-whitening approach and is an extension of the work of Graupe and Efron on single point adaptive ANC. The controller works by continually estimating the time-series parameters of the noise to be canceled and by forcing the cancellation network to follow the identified parameters, so that the output is whitened. To address the problem of wide-area cancellation, a system that consists of a weighted array of single point output-whitening cancellation systems is proposed. Each controller is chosen to cancel the local temporal effects of the noise, and the array weights are chosen to match the wavefront shape. A comparison of cancellations based on single-point and wide-area systems demonstrates the superiority of the wide-area system.**

I. Introduction

Active Noise Control (ANC) is a technique of quieting unwanted acoustic emission by using a secondary source to create sound that is equal in amplitude (volume), identical in frequency (pitch) and

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