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Digital estimation and correction of DAC errors in multibit ΔΣ ADCs

X. Wang, P. Kiss, U. Moon, J. Steensgaard and G.C. Temes

An adaptive digital algorithm is described for acquiring and correcting the errors of the feedback DAC used in a multibit ΔΣ ADC. The method is highly accurate, and is particularly useful for wideband ADCs, where mismatch error shaping becomes ineffective.

Introduction: The use of multibit quantisation in a ΔΣ ADC improves the accuracy and stability of the circuit, and hence is often used in state-of-the-art converters [1 – 3]. Since the linearity of the conversion is primarily limited by that of the feedback DAC, the inherent accuracy of the DAC (typically 10–11 bits) is usually insufficient. Mismatch error shaping can improve the DAC performance for relatively high values (say, 16 or higher) of the over-sampling ratio (OSR): it transforms harmonic distortion into filtered pseudorandom noise which is usually acceptable in the output. However, for wide signal bands it is hard to realise the circuit with sufficiently large OSR for effective error shaping, and other methods must be used. In earlier Letters, we proposed an online analogue calibration technique [4] and a mixed-mode technique [5]. A digital linearising technique, based on correlation operations, was also suggested recently [6] by Galton for DACs embedded in pipelined ADCs.

In this Letter, we describe a digital correction scheme based on filtering for acquiring and correcting the effects of the DAC errors in ΔΣ ADCs. It incorporates mismatch error shaping, but enhances its effect by error cancellation.

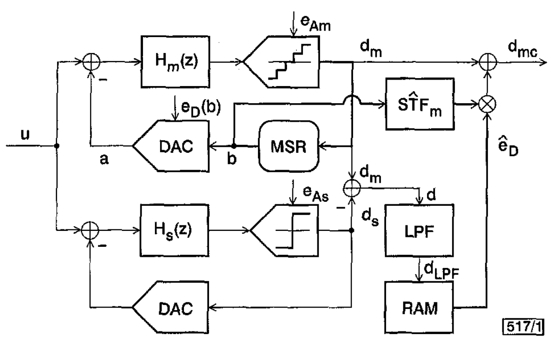

Fig. 1 *Multibit ΔΣ ADC with DAC error correction*

Principle of correction scheme: The block diagram of the self-correcting ΔΣ ADC is shown in Fig. 1. In the Figure, $H_m(z)$ and $H_s(z)$ are the loop filters of the main and self-calibration ΔΣ ADCs, and the block preceding the DAC in the main loop is a digital mismatch-shaping randomiser (MSR). Heavy lines indicate thermometer-coded digital data flow. The self-calibration ADC uses single-bit quantisation, and hence is inherently linear. The output d_m of the main loop contains the low-frequency input signal u, the ADC quantisation 'noise' e_{Am} highpass filtered by the noise transfer function of the loop, and the wideband DAC error 'noise' e_D shaped by the MSR. At low frequencies, only u and e_D contribute to the output. By removing u using the auxiliary single-bit ΔΣ ADC output d_s, it can be ensured that the DAC error e_D will dominate the output at low frequencies. (Note that bandpass filtering can also be used to achieve dominant e_D in a frequency range above the base band.)

Analysis in the z-domain shows that the input to the calibration lowpass filter (LPF) is given by

$$\begin{aligned} D(z) &= D_m(z) - D_s(z) \\ &= -STF_m(z)E_D(z) + NTF_m(z)E_{Am}(z) \\ &\quad - NTF_s(z)E_{As}(z) \end{aligned} \qquad (1)$$

where STF_m denotes the signal transfer function, NTF_m the noise transfer function of the main loop, and NTF_s the noise transfer function of the auxiliary loop. In eqn. 1, it was assumed that $STF_m = STF_s$, and hence that the terms containing $U(z)$ cancel exactly. However, the operation is not critically dependent on this assumption.

It is next shown that by appropriately controlling the MSR operation, the DAC errors can be recovered from the sequence $d[n] = d_m[n] - d_s[n]$, and their effects can be corrected. The method involves a block-wise processing of the samples, with each block containing N samples. Typically, N is of the order of several hundreds or thousands, depending on the accuracy required.

During the processing of the first block of samples, all K unit elements in the DAC are used equally often by applying, for example, data-weighted averaging [3] in the MSR. At the end of the first period, after all N samples have been processed, the value $d_{LPF}(0)$ of the output of the LPF at DC is, to a good approximation, equal to $N \times e_{av}$, where e_{av} is the average value of the unit-element errors e_i, $i = 1 \ldots K$. This value is then stored in the RAM.

During the next N-sample-long period, every DAC output sample will be forced to contain the first unit element u_1. The other elements are used equally often, but less frequently than u_1. Let the number of usage of $u_2, u_3, \ldots, u_K$ be k_1. The final value of the LPF output $d_{LPF}(1)$ is then, to a good approximation,

$$d_{LPF}(1) \simeq (N - k_1)e_1 + k_1 e_{av} \qquad (2)$$

Since e_{av} is known from the results of the first block processing step, and k_1 can be found by counting, e_1 can be obtained from $d_{LPF}(1)$.

This process is then repeated for unit element u_2, then for u_3, etc., and $e_2, e_3, \ldots, e_K$ are obtained from $d_{LPF}(2)$, $d_{LP}F(3)$, ..., $d_{LPF}(K)$, respectively. After $K + 1$ periods, all DAC errors will have been acquired and stored in the RAM. If desired, the values of e_{av}, e_1, e_2, etc. may then be recomputed to account for error drift due, for example, to temperature effects.

Having obtained the estimates of the e_i for all unit elements, the output signal samples $d_m[n]$ may be corrected for these DAC errors. This is performed by finding the errors in each DAC output, and applying to the resulting error sequence the signal transfer function $\hat{STF}_m$. This gives the correction needed to cancel the DAC error in $d_m[n]$, as shown in Fig. 1, where $\hat{e}_D$ and $\hat{STF}_m$ are the digital estimates of e_D and STF_m, respectively.

The accuracy of the process is somewhat affected by the imperfect cancellation of $u[n]$ and the DC offsets in the outputs of the two ΔΣ ADCs in $d[n]$. However, both theory and simulations indicate that these effects are negligible if the ADCs use two or more op-amps in their loop filters. For optimum performance, correlated double sampling can be used in the input stages to reduce the DC offset and $1/f$ noise effects.

It is also possible to use a sorting algorithm to carry out the error estimation in a parallel, rather than block-wise serial, operation. In this algorithm, the error e_i is estimated by comparing the average DAC output error obtained for samples containing e_i with the average error obtained for samples which do not contain it.

Simulation results: Fig. 2 shows the computed results for a fourth-order ΔΣ ADC, designed using the Schreier MATLAB toolbox [7]. Block-wise operation was used, with $N = 1024$. $OSR = 32$ was assumed, and the unit-element errors were assigned randomly, with an RMS value of 0.1% for a K = eight-element DAC. The auxiliary ΔΣ ADC used a second-order one-bit circuit. The op-amps in both circuits had a DC gain of only 40dB, and a random capacitor mismatch (0.1%) was assumed.

Fig. 2*a* shows the output spectrum D_m under ideal DAC conditions; Fig. 2*b* shows it when the random DAC errors were included. Both the noise floor and the harmonics are unacceptably large for the ADC under these conditions. Fig. 2*d* illustrates the performance when only mismatch shaping was used to linearise the DAC. The harmonics disappear, but the noise floor is still too high for $f > f_S/100$, and hence the passband SNR is reduced to 97dB from the ideal value of 108.1dB. Finally, the output spectrum is shown in Fig. 2*c* under non-ideal conditions, but using the proposed correction algorithm. It is nearly identical with the ideal spectrum; the SNR is 108dB.

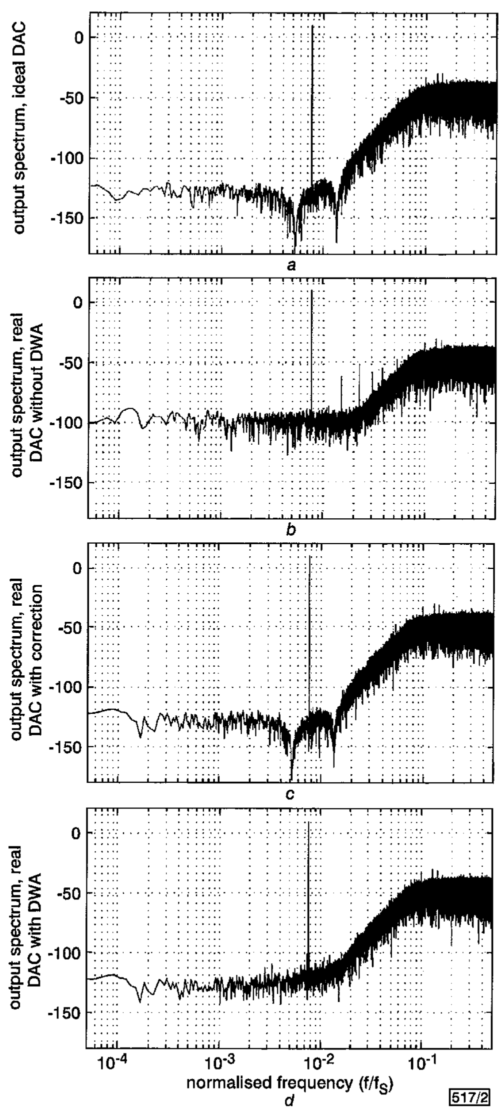

Fig. 2 *Output spectra of fourth-order ΔΣ ADC with eight-unit elements*

Computed for 2^{16} samples, $OSR = 32$, $A_u = -3.1$dB
a Ideal DAC (SNR = 108.1dB)
b Nonlinear DAC without correction (SNR = 71.1dB)
c Nonlinear DAC with proposed correction (SNR = 108.0dB)
d Nonlinear DAC with data-weighted averaging (SNR = 97.0dB)

Conclusions: A digital algorithm is described for the correction of the nonlinear distortion introduced by the internal DAC in a multibit ΔΣ ADC. The correction scheme does not rely on a high over-sampling ratio, and hence is particularly useful for wideband data converters which cannot apply mismatch shaping for suppressing the inband DAC errors. A simulation example verifies the high accuracy that can be obtained using the proposed technique.

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Fabrication and optical properties of neodymium-, praseodymium- and erbium-chelates-doped plastic optical fibres

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Neodymium, praseodymium and erbium ions are successfully incorporated into the core of deuterated polymer-based optical fibres. The spectra of three fibres have several strong absorption bands in the visible and infrared regions. The fluorescence lifetime (6.24μs) at 1060nm of Nd-chelate doped plastic optical fibre with an uncooled photodiode is obtained.

Introduction: The most widely used optical amplifiers in optical communications are erbium-doped fibre amplifiers (EDFAs), which have high gain, low noise and a lack of polarisation sensitivity [1]. Pumping by diode laser emitting at 980 or 1480nm, the generating gain has a broad range of wavelengths from 1520 to 1630nm. Unlike EDFAs, praseodymium-doped fibre amplifiers (PDFAs) operating at 1310nm [2 – 4], have been used as laboratory devices for nearly a decade and are only now available with sufficient reliability and reasonable prices to be used in telecommunication systems. This progress has been driven by the need to make better use of the available optical bandwidth at 1310nm, the second communications window, in currently installed singlemode fibre. Improvements in the quality of plastic optics have allowed them to proliferate into many areas dominated by glass optics. The major applications will include telecommunications. Graded-index plastic optical fibres (GI-POFs) [5, 6] are of particular interest for short-distance communications (premise networks, home networks and consumer electronics) because of their large core diameters, high bandwidths, and low losses at the communication