Higher Order Sigma-Delta Modulator Interfaces for MEMS Capacitive Sensors

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1. Introduction

MEMS capacitive sensors, such as pressure sensors, accelerometers and gyroscopes have been one of the most successful examples of micro-system technology. They are routinely used in many applications; the most important being automotive safety systems. These sensors are typically open-loop and have low to medium performance specifications. Recently, there has been an emerging requirement for higher performance sensors to be used in more sophisticated safety systems such as DSC (Dynamical Stability Control), active suspension and rollover sensing. Such sensors could also be used in navigation systems, virtual reality applications and everything related to so-called contextual awareness applications. To fulfil this requirement it appears a very promising approach to apply some more advanced control and electronic interface techniques to existing micromachined sensing elements, rather than adding complexity to the micro-fabrication process and/or the design of the mechanical sensing element itself. Incorporation of a sensing element in a force feedback control architecture is a promising approach to improve the sensors' performance, especially what linearity, dynamic range and bandwidth is concerned. Micromachined sensors with capacitive read-out are particularly suitable since a capacitor can be used to detect the movement of a membrane or proof mass and, concurrently, can exert an electrostatic force on the moveable part by applying a feedback voltage across it.

An elegant solution is to incorporate a capacitive sensing element in a sigma-delta modulator (SDM) control system. This yields a digital output signal that can directly interface to a standard digital signal processor. Such an approach has been successfully used for a variety of different sensors, mainly for inertial sensors such as accelerometers [1] and gyroscopes [2], but also for pressure sensors, flow sensors and fluxgate magnetic sensors. So far, MEMS capacitive sensors have mainly been incorporated in lower order SDM architectures, where the sensing element only acts as the loop filter. Since any mechanical sensing element can be modelled as a second order system comprising a mass, damping coefficient and spring constant, it can only provide noise-shaping up to order two. The best noise-shaping is achieved if the sensing element has a low spring constant, as this is equivalent to high gain at low frequencies. However, any practical sensing element will have a significantly lower gain at low frequencies compared to a purely electronic integrator. Therefore, using a mechanical sensing element as loop filter will always result in worse noise shaping compared to an electronic second order SDM. To provide higher order noise-shaping it is necessary to add additional electronic integrators. Higher order architectures, which are routinely used for purely electronic SDM analogue to digital converters, can be used with MEMS capacitive sensors. The challenge in applying such control systems to these sensors is that there is no access to the internal node in the micro-mechanical sensing element. Furthermore, micromachined sensors suffer from relatively high manufacturing tolerances. This makes the design of stable closed loop control systems difficult. Recently, there has been growing interest in higher order SDM interface for micromachined sensors [3, 4]. In this work, we present novel 5th and 6th order architectures for a typical bulk-mircomachined accelerometer sensing element.

2. Higher Order SDM Topologies

We have investigated various higher order SDM topologies up to order six, i.e. using three or four additional electronic integrators. Fig. 1 shows a distributed feedback network topology which is the simplest higher order topology having the fewest feedback path. M(s) is the transfer function of the sensing element, relating input force to a deflection x of the proof mass. This gives rise to a differential change in capacitance which is

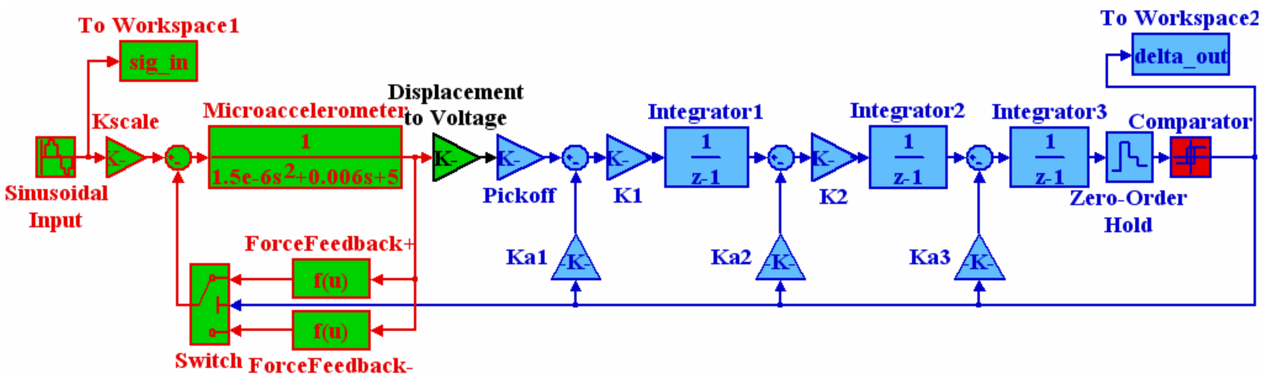

Fig.1: Simulink model of a capacitive micro-accelerometer incorporated in a fifth order, distributed feedback SDM architecture.

detected by a standard charge integrator. This chain of transductions can be lumped into a gain constant Kpo. Followed by the signal pick-off are three integrators with a gain constant each. The electrical signal is then digitised by a 1-bit quantiser which is sampled at a frequency fs much higher than the bandwidth of the sensor. Using the standard white noise approximation, the quantiser can be represented by white quantisation noise and a variable gain KQ. In the feedback path a voltage is applied giving rise to an electrostatic feedback force that nulls the motion of the proof mass. For small displacements, the feedback force can be assumed as constant and is represented by Kfb. The right choice of the various gain constants is crucial for the stability. An algorithm was published in ref [5] which allows to find gain coefficients K1...K4 which optimise stability and performance.

For frequencies considerably below the sampling frequency, the noise transfer function (NTF) can be approximated by:

$$\left|Q_n(z)\right|_{f \ll f_s} \approx \frac{\left|1-z^{-1}\right|^5 + \left|2-b_f-c_f\right|\left|1-z^{-1}\right|^4 + \left|(1-c_f)(1-b_f)\right|\left|1-z^{-1}\right|^3}{K_{fb}K_{po}K_f(1-a_f)\prod_{i=1}^{4}K_iK_Q + \left|(1-c_f)(1-b_f)\right|\prod_{i=2}^{4}K_iK_Q} \quad (1)$$

where K_f, a_f, b_f and c_f are gain, zero and poles of the transfer function of the sensing element using discrete representation, which is given by:

$$M(z) = K_f \frac{\left(1-a_f z^{-1}\right)z^{-1}}{\left(1-b_f z^{-1}\right)\left(1-c_f z^{-1}\right)} \quad (2)$$

This was obtained by taking the discrete representation of the s-domain transfer function of the mechanical sensing element:

$$M(s) = \frac{1}{ms^2 + bs + k} \quad (3)$$

where m is the mass of the proof mass, b the damping coefficient and k the mechanical spring constant. The analytical derivation of the dependency of the coefficients K_f, a_f, b_f and c_f on m, b, and k is somewhat cumbersome, and can be found in [6].

Using the Simulink model, simulations to determine the signal-to-noise ratio can be carried out. Fig. 2 shows a simulation result of the spectrum of the output bitstream. This simulation considers only quantisation noise, hence results in an unrealistically low noise floor. However, it is convenient for comparison with a second order SDM topology, it is

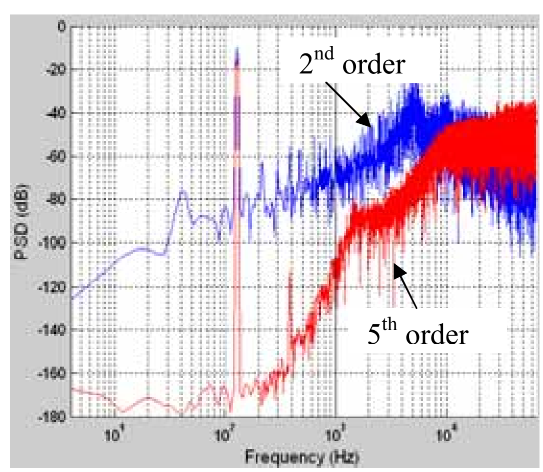

Fig. 2: Comparison of quantisation noise shaping between a second order and fifth order SDM architecture with a capacitive micro-accelerometer.

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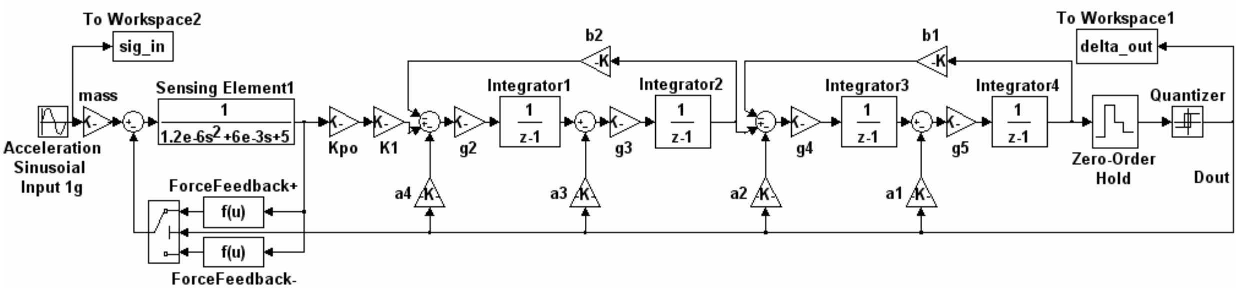

Fig. 3: 6th order topology with distributed feedback and local resonators.

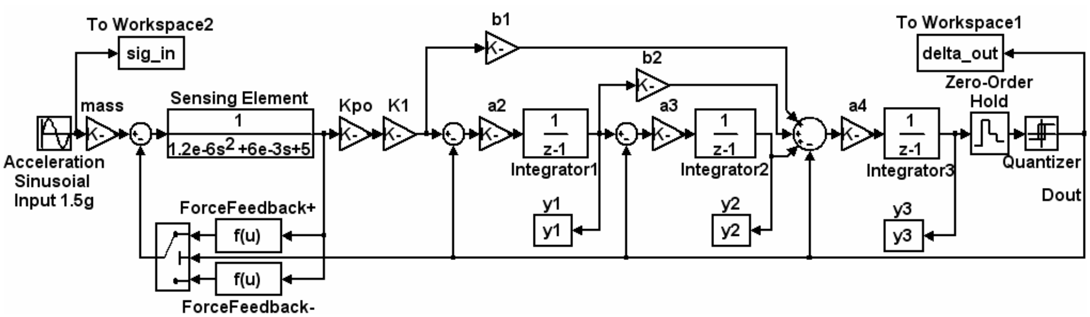

Fig. 4: 5th order topology with distributed feedback and feedforward topologies..

obvious that the quantisation noise floor has been drastically reduced. It should be noted that in a real system other noise sources such as from the electronic pick-off electronics would dominate. However, suppressing the quantisation noise in the baseband allows to decrease the oversampling ratio (OSR), and hence to run the circuit at a lower

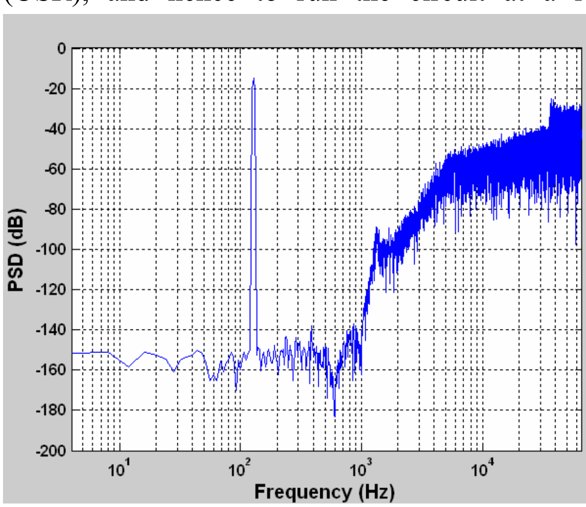

Fig. 5: Spectrum of the topology with distributed feedback and local resonators. The dynamic range is about 125dB.

sampling frequency leading to reduced power consumption and more relaxed speed requirements for the interface circuitry.

A range of other topologies are possible. For example, a sixth order, distributed feedback topology with two local resonators is shown in fig. 3, and a 5th order topology with distributed feedback and feedforward is shown in fig. 4. In comparison to the topology shown in fig. 1, the additional feedback or feedforward loops generate a pair complex zero in the NTF in the baseband. This leads to a further reduction of the in-band quantisation

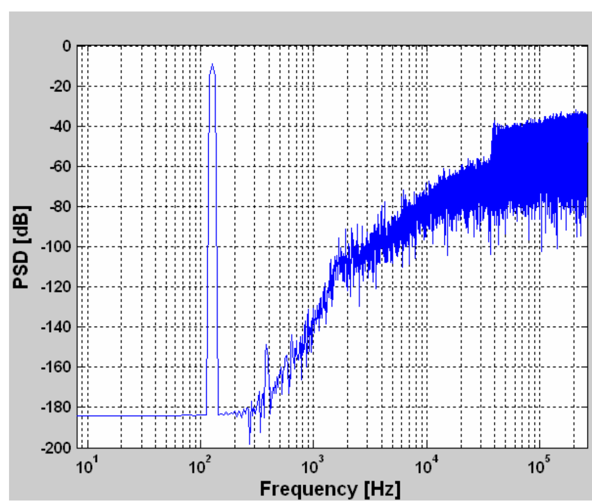

Fig. 6: Spectrum of the topology with distributed feedback and feedforward loops. The dynamic range is about 120dB.

noise. In fig. 5, a spectrum of output bitstream of the topology with distributed feedback and local resonators is shown. In this simulation, thermal noise from the electronic position measurement interface and Brownian noise from the mechanical sensing element has been considered as well. The OSR was 64 and a signal bandwidth of 1kHz was assumed. For the same conditions, fig. 6 shows the output signal spectrum of the topology with distributed feedback and feedforward loops. The dynamic range is similar, however the simulations indicate a larger input signal swing is possible. The improvement is about 6dB.

3. Measurement Results

To validate above approach in hardware, a printed circuit board (PCB) was realised with a 5th order distributed feedback topology (see fig. 2). The sensing element used was a standard bulk-micromachined, capacitive accelerometer that was put on the same PCB, together with a reference accelerometer. In fig. 6 a measured noise floor of about -90dB is shown; this is dominated by thermal and interference noise from the electronic interface electronics. The sampling frequency was only 125kHz. When compared to a second order loop with only the sensing element as a loop filter, the noise floor is reduced by at least 50dB. This agrees reasonable well with theoretically predicted drop of 60dB in noise floor (20dB for each additional integrator).

Fig. 3: Measured noise-floor of a fifth-order electro-mechanical SDM.

3. Conclusions

We have investigated various higher order SDM topologies to be used with a micromachined sensing element of a standard capacitive accelerometer. Achieving stability is a challenging task since it is not possible to feed back a signal to the internal node of the sensing element. Stability can be investigated by simulations using Matlab and Simulink. The noise shaping characteristic can been greatly improved compared to an architecture where the sensing element is used as a loop filter only. The approach presented in this paper can be easily adapted for other capacitive micromachined sensors.

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