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闭环加速度计CMOS接口电路

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摘要：采用高压18 V CMOS集成电路工艺，设计了一种开关电容闭环加速度计接口电路芯片。芯片电路中包括开关电容型电荷敏感放大器，PID控制电路以及相关双采样电路。采用相关双采样技术并用大面积PMOS晶体管作前级放大器输入级来消除放大器的$1/f$噪声、失调电压及KT/C噪声；用高环路增益及静电力平衡技术消除后级电路的$1/f$噪声、电荷注入和时钟馈通。在相同电极的条件下，利用电荷检测与静电力反馈时域分离法，有效地消除了驱动馈通的影响。设计的芯片采用18 V电源电压供电，闭环加速度计刻度因子为420 mV/g，噪声密度为10 $\mu g/\sqrt{\text{Hz}}$，芯片面积为15.2 mm^2。

关　键　词：接口电路；闭环加速度计；开关电容；惯性传感器
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CMOS interface circuit for closed-loop accelerometer

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Abstract: An 18 V switched-capacitor CMOS interface circuit for the closed-loop operation of a capacitive accelerometer is designed. The circuit consists of a switched-capacitor charge sense amplifier, a PID control circuit and a correlated double sampling circuit. The effects of the $1/f$ noise of the amplifier and the offset voltage of an op-amp, as well as the kT/C charge noise from the parasitic capacitor are suppressed, by taking large area PMOS transistors at the Charge Sensing Amplifier (CSA) as an input stage and using a Correlated Double Sampling (CDS) technique and the $1/f$ noise, charge injection and clock feedthrough effects in the back-end circuits are eliminated by the technologies of high loop gain and force feedback. Moreover, the strong driving feedthrough is avoided by separating the drive and sense operations in the time domain by using the same electrodes. The designed complete chip with an area of 15.2 mm^2 is fabricated in a 2 μm two-metal and two-poly n-well CMOS and operated by a single 18 V supply, which can offer a measuring sensitivity of 420 mV/g and a noise of floor of 10 $\mu g/\sqrt{\text{Hz}}$ in closed-loop.

Key words: interface circuit; closed-loop accelerometer; switched capacitor; inertial sensor

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1 Introduction

Capacitive accelerometers have advantages such as zero static biasing currents and excellent thermal stability. Further, by using bulk micromachined devices with a large seismic mass, very high sensitivity can be reached. For the accelerometer, the force feedback is attractive, for it can offer the potential of wide dynamic range, high linearity. In the past few years, there have been many reports on low-power and small form-factor micro-g closed-loop accelerometers[1-9]. In this work, we present a force-rebalanced medium form-factor accelerometer (15 g) from a single 18 V supply. A wide variety of methods can be used for capacitance readout such as charge sensitive amplifiers, charge redistribution techniques, impedance measurements, RC-oscillators, or direct coupling to a field effect transistor gate. The most common readout circuit for micromachined differential capacitive sensor is based on a fully differential charge sensitive amplifier. In this way the voltage step is applied at the common node, and the charge difference is integrated in a fully differential integrator. This circuit is rather complex to implement since a fully differential operational amplifier with internal common node feedback is required. Further, an additional operational amplifier is required to control the common mode level at the differential amplifier inputs[10].

Our pendulous silicon accelerometer operates in normal atmosphere and is designed for non-peaking (low-Q) response with a sensitivity of 0.5 pF/g (static capacitance = 10 pF). The topology suggested in this paper is a differential version of the charge transfer method. It can be used with the capacitor configuration illustrated in Fig. 1 without any performance reduction and ultimately limited by the mechanical noise. The architecture only requires single ended operational amplifiers, transmission gates and capacitors. Further, a structure of high loop gain and force feedback is designed for suppression of $1/f$ noise of amplifier, charge injection and clock feedthrough effects in the back-end circuits. The strong driving feedthrough is avoided by separating the drive and sense operation in the time domain, while using the same electrodes.

2 Interface Overview

The sensing structure of pendulous silicon accelerometer is a second order dynamic system whose equation of motion can be written as:

$$J\frac{d^2\theta}{dt^2}+B\frac{d\theta}{dt}+K\theta=K_\alpha a(t)\ , \qquad (1)$$

In this equation J is the moment of inertia, B is the damping torque coefficient, K is the pendulum elastic restraint, K_α is the pendulosity, θ is the angular displacement of pendulum and $a(t)$ is the applied acceleration. To improve a linearity and a dynamic range, a servo accelerometer with electrostatic force-balancing has been designed as shown in Fig. 1.

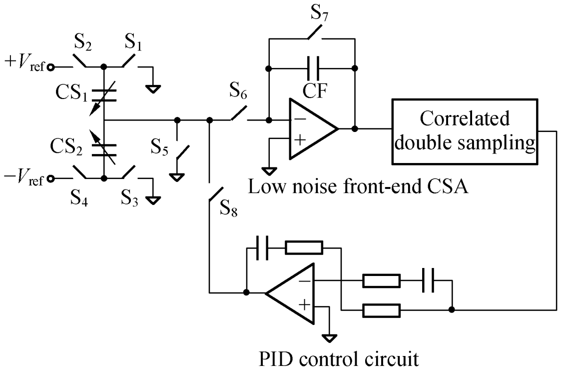

Fig. 1 Block diagram of major building blocks of implemented circuit

Fig. 2 shows the clock diagram of feedback control circuit. Each cycle starts with the end of the forcing period and begins with the first several sensing phases. Each cycle includes two phases: sensing phase and forcing phase. A structure of high loop gain and force feedback is proposed to get a high SNR. It suppresses the effects of the $1/f$ noise and offset voltage of the op-amp, as well as the kT/C charge noise from the parasitic capacitor by using CDS (Correlated Double Sampling) technique and implementing large area PMOS transistors at the CSA (Charge Sensing Amplifier) input stage. It also reduces charge injection and clock feedthrough effects. This scheme functions even with large parasitic capacitances between the sensor and the interface circuit.

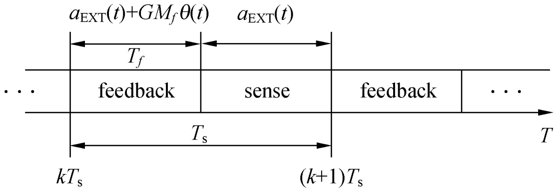

Fig. 2 Clock diagram of feedback control circuit

3 Static and Dynamic Analysis

Since the mechanical system is linear, the exact discrete time model of the sense element over one complete period is given by:

$$\theta(kT_{\mathrm{S}}+T_{\mathrm{S}}) = \int_0^{KT_{\mathrm{S}}+T_{\mathrm{S}}} K_a a_{\mathrm{EXT}}(\tau)h(t-\tau)d\tau + \int_0^{KT_{\mathrm{S}}+T_{\mathrm{S}}} GM_f\theta(\tau)h(t-\tau)\mathrm{d}\tau \ , \quad (2)$$

Where T_{S} is the length of time between sampling ends, M_f is the moment coefficient (Nm/V), $a_{\mathrm{EXT}}(\tau)$ is the applied acceleration, G is the coefficient of converting angular displacement into voltage (V/rad). Using equation (1), the response function of sensing structure $h(t)$ is given by:

$$h(t) = \sum_{k=1}^{2} a_k \mathrm{e}^{p_{\mathrm{K}}t}u(t) \ , \quad (3)$$

Equation (2) may be rewritten as:

$$\theta(kT_{\mathrm{S}}+T_{\mathrm{S}}) = \int_{KT_{\mathrm{S}}}^{KT_{\mathrm{S}}+T_{\mathrm{S}}} F(\tau)h(kT_{\mathrm{S}}+T_{\mathrm{S}}-\tau)d\tau + \int_0^{kT_{\mathrm{S}}} F(\tau)h(kT_{\mathrm{S}}+T_{\mathrm{S}}-\tau)d\tau \ , \quad (4)$$

Where

$$F(\tau)=K_a a_{\mathrm{EXT}}(\tau)+GM_f\theta(\tau) \ , \quad (5)$$

For $t>kT_{\mathrm{S}}$, the angular displacement may be found by combining this result with equation (1) to yield:

$$\theta(kT_{\mathrm{S}})=C_1\mathrm{e}^{p_1kT_{\mathrm{S}}}+C_2\mathrm{e}^{p_2kT_{\mathrm{S}}} \ , \quad (6)$$

$$\int_0^{kT_{\mathrm{S}}} F(\tau)h(t-\tau)d\tau = C_1\mathrm{e}^{p_1t}+C_2\mathrm{e}^{p_2t} \ , \quad (7)$$

Initial condition of equation (6) is given by:

$$\theta'(kT_{\mathrm{S}})=C_1p_1\mathrm{e}^{p_1kT_{\mathrm{S}}}+C_2p_2e^{p_2kT_{\mathrm{S}}}=0 \ , \quad (8)$$

Using equations (6), (7), and (8), it can be found:

$$\int_0^{kT_{\mathrm{S}}} F(\tau)h(kT_{\mathrm{S}}+T_{\mathrm{S}}-\tau)\mathrm{d}\tau = \left(\frac{-p_2}{p_1-p_2}\mathrm{e}^{p_1T_{\mathrm{S}}}+\frac{p_1}{p_1-p_2}\mathrm{e}^{p_2T_{\mathrm{S}}}\right)\theta(kT_{\mathrm{S}}) = C_n\theta(kT_{\mathrm{S}}) \ , \quad (9)$$

Because sampling frequency is much greater than signal frequency, the equation (4) may be rewritten as:

$$\theta(kT_{\mathrm{S}}+T_{\mathrm{S}})-C_n\theta(kT_{\mathrm{S}})=K_a a_{\mathrm{EXT}}(kT_{\mathrm{S}})h_a+GM_f\theta(kT_{\mathrm{S}})h_f \ , \quad (10)$$

Where

$$h_f = \sum_{k=1}^{2} -\frac{a_k}{p_k}\left[\mathrm{e}^{P_K(T_{\mathrm{S}}-T_f)}-\mathrm{e}^{P_KT_{\mathrm{S}}}\right] \ , \quad (11)$$

$$h_a = \sum_{k=1}^{2} -\frac{a_k}{p_k}\left[1-\mathrm{e}^{P_KT_{\mathrm{S}}}\right] \ , \quad (12)$$

Equation (10) may be rewritten in the z-transform domain as:

$$\frac{\theta(z)}{a_{\mathrm{EXT}}(z)}=\frac{K_a h_a}{(z-C_n-M_fGh_f)} \ . \quad (13)$$

Equation (13) provides an intuitive understanding of factors affecting the performance. There is a problem of stability in the accelerometer because of discontinuous feedback force, which is different from a continuous feedback system. Although we use over-damped proof mass to make it stable, it is important to choose a suitable value of G for stability of the system. For $z=1$ in equation (13), the scale factor of the accelerometer is found:

$$\frac{V_{\mathrm{out}}}{a_{\mathrm{EXT}}}=\frac{GK_a h_a}{1-C_n-M_fGh_f} \ , \quad (14)$$

The scale factor of accelerometer is correlated with the length of feedback time T_f, which is different from that in continuous-time system.

4 Sensor performance parameters

Fig. 3 shows the hybrid packaged accelerometer and the interface chip in a 20-pin IC package. The chip is fabricated in the 2 μm two-metal and two-poly n-well CMOS process with an area of 15.2 mm^2.

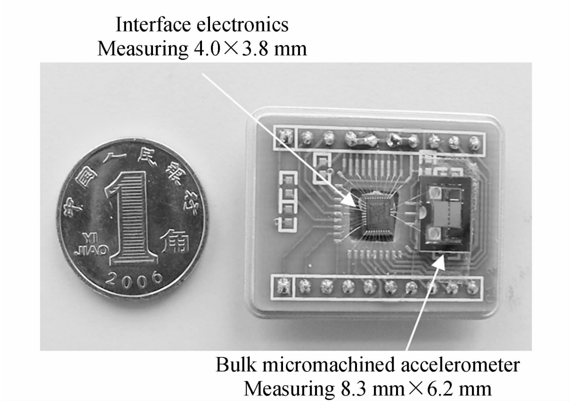

Fig. 3 Hybrid packaged accelerometer and interface chip in a 20-pin IC package

Tab. 1 shows performance parameters of the hybrid system. Test results show that a full scale acceleration of ±15 *g* and a signal bandwidth of 300 Hz are achieved. The complete module operates from a single 18 V supply and has a measured sensitivity of 420 mV/*g* with a noise of floor of 10 $\mu g/\sqrt{\text{Hz}}$ in the closed-loop operation.

Tab. 1 Performance parameters of hybrid system

CMOS readout electronics	
Sensitivity (V/pF)	10-50 (open-loop)
Resolution (aF)	<10
MEMS accelerometer	
Static capacitance (pF)	10
Sensitivity (pF/*g*)	0.5
MEMS device and interface circuit module	
Sensitivity (mV/*g*)	420
Full scale (*g*)	±15
Bandwidth (Hz)	300
Noise density ($\mu g/\sqrt{\text{Hz}}$)	10 (10-100 Hz)
Nonlinearity	0.12%

There are several noise sources affecting the overall system resolution of an accelerometer system. These noise sources can be classified in two main groups: mechanical and electrical. Mechanical noise is from the sensing structure design and environment. The electronic noise has different components including the front-end amplifier noise, switch resistance noise and sensor charge referencing voltage noise. The circuit has a calculated noise floor of 1 $\mu g/\sqrt{\text{Hz}}$ resulting in a capacitance resolution of better than 1 aF. The mechanical noise is approximately 5 $\mu g/\sqrt{\text{Hz}}$ by calculation. The output noise of the hybrid module is measured by using an HP35670 dynamic signal analyzer as shown in Fig. 4.

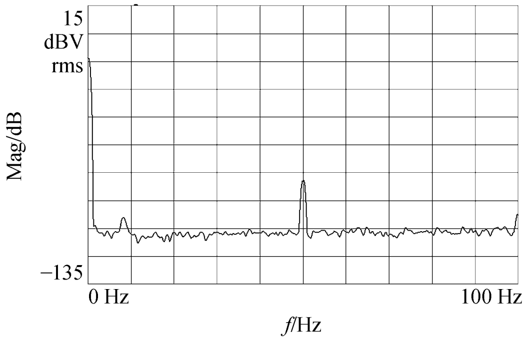

Fig. 4 Noise spectrum of hybrid system with a *z*-axis device, showing 10 $\mu g/\sqrt{\text{Hz}}$ noise floor

5 Conclusions

In this report, the interface circuit for capacitive closed-loop accelerometer applications has been presented. The chip is suitable for high accuracy requiring micro-gravity acceleration sensors and it can work in the closed-loop mode with only two changing capacitances. The circuit only requires single ended operational amplifiers and it can work without referential capacitances. The circuit operates at 100 kHz speed, and has an adjustable open-loop sensitivity between 10 and 50 V/pF with better than 10aF expected resolution, while occupying an area of 15.2 mm^2. This circuit achieves a dynamic range of 120 dB for 1Hz bandwidth.

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