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A Switched Actuation and Sensing Method for a MEMS Electrostatic Drive

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Abstract—MEMS technology is being investigated to improve the performance, integration and cost of nanopositioning systems. The most basic MEMS fabrication processes produce designs etched into a single layer of silicon and electrostatic transduction is often seen as the most viable actuation and sensing technology. This work provides a method to utilize the same electrostatic drive for both actuation and sensing functions. By combining both functions into the one drive, an effective increase in actuator force and sensor sensitivity can be achieved. The sensor utilizes a sigma-delta type arrangement to create a displacement-to-digital converter. With the actuator composed of a switching amplifier, this allows the nanopositioner to be controlled directly from a DSP platform. This work outlines the design of the nanopositioning system, provides the system modeling and identification and characterizes the open loop performance of the nanopositioner.

I. INTRODUCTION

Nanopositioners are devices that actuate and sense motion at nanometer resolutions [1]. Nanopositioners have found applications in areas such as microscopy, data storage, microfabrication and optics [1]–[3]. The use of microelectromechanical system (MEMS) technology to fabricate nanopositioner designs has become of interest to attain improved performance [3]–[5]. MEMS are miniature mechanical systems fabricated using the technology employed in the production of integrated circuits [6]. In miniaturizing the nanopositioner with MEMS technology, researchers hope to produce devices with improved dynamics while at the same time reducing the cost of fabrication allowing the technologies incorporating nanopositioners to become more ubiquitous [3]. In addition the use of MEMS allows for the fabrication of more complex designs, in particular designs which utilize massive arrays of devices to perform tasks in parallel [4].

Actuation in MEMS utilizes a number of phenomena including electrostatic, piezoelectric, thermal and electromagnetic phenomena [7]. Electrostatic is arguably the most common method to implement actuation in MEMS because of its compatibility with any conductive material and low power consumption. Numerous MEMS nanopositioners utilize electrostatic actuation [3]–[5], [8], [9]. Compared to the other phenomena listed the disadvantage of electrostatic actuation is its low force output for the same die area and applied voltage.

Displacement sensing can utilize phenomena such as thermal [3], piezoresistive [3], [8] and electrostatic [9] phenomena. In designs using electrostatic sensing or no sensing at all [4], [5], the transduction efficiency is improved by utilizing the same drive for both the actuation and sensing functions. Called self-sensing actuation, this simplifies the mechanical design and increases the effective die space that can be dedicated to a single drive. When electrostatic drives are used to sense displacement, the capacitance of the drive is electrically read-out as the capacitance is a function of displacement [7].

Self-sensing electrostatic drives are implemented by time division multiplexing, whereby the drive is switched between actuation and sensing functions [10], [11], or spectral separation, whereby the actuation and sensing signals are separated by frequency [12]. An advantage of time division multiplexing is its suitability to be implemented with switching electronics. However, most implementations of this technique fully discharge the drive during the sensing phase. This reduces the maximum voltage that can be applied reducing the range and maximum control effort achievable in a nanopositioning application [10], [11]. Another issue commonly associated with self-sensing designs is that of electrical isolation. Simple microfabrication processes such as silicon-on-insulator produce devices where all drives share a common ground and are thus incompatible with a number of self-sensing techniques which requires each drive to be electrically isolated from one another [10]–[13].

A capacitive sensing method called charge balance is also suitable for implementation with switching electronics and has a natural affinity to be used in conjunction with time division multiplexing self-sensing [10], [11]. Charge balance, alternately charges a capacitor with a fixed voltage and discharges it onto a charge amp to read-out the measurement [10], [14].

Due to the similarity between charge balance capacitive sensing implementations and switched capacitor type integrators commonly used in $\Sigma \Delta$ modulators, a number of designs have modified the $\Sigma \Delta$ architecture to implement capacitance-to-digital converters [11], [15], [16] including specific applications in MEMS [17], [18]. By using the capacitance-to-digital conversion and a type of switching amplifier like a class D amplifier, this allows the entire nanopositioner to be controlled through digital signals. The utility of this digital interface is the ability to apply DSP platforms to directly control the nanopositioner without the need for additional analog conversion electronics.
The ideal mass electrostatic comb drive which is used for the sensing aspect damper system as shown in the equation below.

The parameters are: \( n \) the number of overlapping faces, \( \varepsilon \) the dielectric permittivity, \( g \) gap between comb fingers and \( V \) is the applied voltage. For one drive in the current design \( n = 1694, \varepsilon = 8.8541879 \times 10^{-12} \), \( g = 2 \mu m \) and \( h = 25 \mu m \).

The mechanical dynamics of a single axis of the nanopositioner is modeled as a lumped second order spring-mass-damper system as shown in the equation below.

The ideal mass \( m = 6.9752 \times 10^{-7} \) kg and the ideal stiffness \( k = 31.6875 \) N m\(^{-1}\).

The final aspect of the model is the capacitance of an electrostatic comb drive which is used for the sensing aspect of the design. The capacitance of the comb drive as a linear function of displacement is shown below.

\[
C = \frac{\varepsilon n h (x_0 + x)}{g} \tag{3}
\]

The only additional parameter to be introduced here is the initial overlap \( x_0 \) which is 13 \( \mu m \).

The purpose of the proceeding nanopositioner design is the control of displacement. The following model describes the mapping from voltage to displacement.

\[
\dot{x} + \frac{\omega_n}{Q} \dot{x} + \omega_n^2 x = \alpha V^2 \tag{4}
\]

To identify these parameters, the displacement of the nanopositioner was measured using a laser Doppler vibrometer. Figure 2 shows the static relationship between displacement and voltage and Figure 3 shows the small signal frequency response. From these measurements \( \omega_n = 8125 \) rad s\(^{-1}\), \( Q = 99.32 \) and \( \alpha = 0.231 \) m s\(^{-2}\) V\(^{-2}\).

### III. Design of the Self-Sensing Method

The basic topology of the nanopositioner system is shown in Figure 4. It comprises of a \( \Sigma \Delta \) actuator driver to convert the input control action \( u \) to a set of switching signals to
Actuator Circuit
Actuator Circuit

Fig. 4. The layout of the nanopositioning system. It comprises of an actuator driver that switches two switching amplifiers to apply a voltage to two opposing electrostatic drives. The sensor circuit and the driver convert the displacement of the drive to a switching signal using a Σ∆ algorithm. Two signals φ1 and φ2 coordinate the logic to alternately switch each system onto the electrostatic drives.

Fig. 5. A charge balance type capacitance sensor is utilize to sense the displacement of the nanopositioner. The electrostatic drives are arranged in as a switched capacitor resistor circuit and the current flowing across them is read into a Σ∆ circuit to digitize the signal. The switch IC is the TS5A23157, the op-amp is the OPA355 and the comparator is the TLV3501, all from Texas Instruments.

drive two switching amplifiers. The two amplifiers drive two opposing electrostatic drives which pull the nanopositioner’s central stage in the positive and negative directions. The sensor injects signals into the two drives to measure the displacement of the nanopositioner. The sensor and its driver form a Σ∆ arrangement to provide a displacement-to-digital type sensor.

The sensor and actuator drivers are implemented within an FPGA to allow for digital control of the system. For both systems to use the same electrostatic drives, time division multiplexing is used to perform self-sensing. To perform the appropriated switching, two phase signals are generated within the FPGA. The sensing phase signal φ1 is high when the sensor is connected to the electrostatic drive and the actuation phase signal φ2 is high when the actuator is connected. With the FPGA logic clocked at 200 MHz, the sensing phase is on for 50 cycles followed by a 70 cycle dead-time and the actuating phase is on for 50 cycles followed by a 30 cycle dead-time.

The following subsections describe the theory and operation behind the sensor and actuator circuits. The sensor circuit is described first as its principle of operation influences the architecture of the actuator circuit.

A. Sensor

The sensor circuit, shown in Figure 5, is a charge balance type capacitive sensor with a Σ∆ output stage to digitize the sensor signal. The low voltage sensing circuit is decoupled from the high voltages applied to the MEMS device via decoupling capacitors Cd. The sensor ensures one side of the decoupling capacitors are kept at a fixed low voltage, and the actuators are designed to supply a limited current to ensure over-current conditions do not allow this voltage to rise.

The electrostatic comb drives are represented by the capacitances C1 and C2. To measure displacement, the capacitance of the comb drive can be measured as shown in Equation (3). Given the differential topology of the two comb drives, the capacitance of one drive increases with displacement while the other decreases. To remove the effect of the common mode capacitance and to increase the transduction efficiency, the difference in capacitance between the two opposing drives C∆ is considered and is proportional to the displacement. The parameter β is defined as the constant of proportionality.

\[
C_\Delta = C_1 - C_2 = \frac{2n c h}{y} x = \beta x \quad (5)
\]

Prior to executing a sensing phase, the actuator’s output changes to a high impedance state to prevent interference with the sensing signal. This leaves the decoupling capacitors in series with the electrostatic drives. The decoupling capacitors are 1 nF while the electrostatic drives are in the order of 10 pF. Therefore the series combination is dominated by the electrostatic drives. One drive is charged to the positive rail Vs and the other is charged to the negative rail −Vs. The supply rails are at ±1.65 V. When the sensing phase is executed the two drives are switched to ground and a current from the drives flows into the Σ∆ portion of the sensor. The average current that flows from the two drives is as follows.

\[
I_1 = -V_s f_s C_1 \quad I_2 = V_s f_s C_2 \quad (6)
\]

The switching frequency fs = 1 MHz. The currents I1 and I2 combine on the same node and the net current Is flowing into the Σ∆ circuit is proportional to the difference in capacitance C∆.

\[
I_\Delta = I_1 + I_2 = -V_s f_s C_\Delta \quad (7)
\]

The Σ∆ modulator is described by the model shown in Figure 6. First, the difference between the input signal and its quantized version is taken and the error is passed to a filter, in this case an integrator. Note that the op-amp integrator has a negative gain which is canceled by the negative sign of I_\Delta. The output of the integrator is quantized using a 1-bit analog-to-digital converter comprising of a comparator and a D flip-flop within the FPGA. This value is then fed-back through a switched capacitor resistor to produce the quantized version of I_\Delta. This loop acts to keep the sensor output y proportional to I_\Delta and thus displacement x. The values of the 1-bit signal y are taken to represent the values ±1. The transfer function from the displacement x to the average sensor output y is as follows.

\[
H_s(s) = \frac{y}{x} = \frac{\beta f_s}{C_1 s + C_f f_s} \quad (8)
\]
The parameters $f_s$ and $\beta$ are set, therefore the gain of the sensor is set by choosing an appropriate feedback capacitance $C_f$. $C_f$ is also the value of $C_{\Delta}$ that causes the integrator to saturate. With an estimated maximum deviation of 10 pF, $C_f$ is chosen to match. The integrator capacitor is chosen at 100 pF which sets a bandwidth of 15 915 kHz.

**B. Actuator**

Each axis of the MEMS nanopositioner is actuated by two opposing electrostatic drives. An advantage of this arrangement is that the quadratic nonlinearity of the electrostatic drives can be removed if the two drives are actuated differentially. In this design, the electrostatic drives are driven differentially around a bias of 25 V.

A class D type amplifier was considered appropriate to actuate the design. Class D amplifiers are efficient, simple and are directly controllable from a DSP. The mechanical bandwidth of the nanopositioner and the capacitive nature of electrostatic drives filter the switching harmonics produced. They are suited to implement a self-sensing actuator with the time division multiplexing method as the control can disconnect the amplifier from the electrostatic drive during the sensing phase.

The actuator for a single drive is shown in Figure 7. A $\Sigma\Delta$ modulator is used to perform the analog to digital conversion. The system input $u$ is a number in the range $[-1, 1]$ and the ideal amplifier maps the input to an actuator voltage $v_a$ in the range $[0, V_{cc}]$. The transfer function of the $\Sigma\Delta$ modulator is a delay of one sample period. For the RC filter, opening the half-bridge every cycle in effect creates a switched resistor RC filter. The switching reduces the average current flowing through the resistor creating a larger effective resistance in the RC filter decreasing its bandwidth. Let $D_a$ be the duty cycle for which the actuation is connected to the RC filter with resistance $R_a$ and thus the equivalent resistance of the switch resistor is $R_a/D_a$. The transfer function of the actuator is as follows.

$$H_a(s) = \frac{v_a}{u} = \frac{V_{cc}}{2} \frac{D_a}{R_a C_d s + D_a} e^{-s T_a}$$  \hspace{1cm} (9)

In the implemented system, $V_{cc} = 50 V$, $R_a = 3.3 k\Omega$, $C_d = 1 \mu F$, $D_a = 0.25$ and $T_s = 1 \mu s$. The gain of the system is 25 and the bandwidth is 11 938 Hz.

**IV. Preliminary Results**

The system was fabricated as shown in Figure 8. The first set of results are to show the correct operation of the time-multiplexed switching scheme. Figure 9 shows the actuator voltage $v_a$ on the opposing electrostatic drives while the input reference is set to 0. The switching cycle begins with the sensing phase which can be seen by a step of 1.65 V in both waveforms in opposing directions. After 250 ns the sensing phase ends and the voltages return to their previous value. After a dead-time of 150 ns the actuation phase begins. To protect the bridge circuits in the amplifier, a shoot through delay of 50 ns causes the delay in the amplifier response. The actuation phase lasts for 250 ns. After which there is a 350 ns dead-time to allow the actuator time to open-circuit.
Fig. 8. The fabricated nanopositioner system. The MEMS nanopositioner is seen in the lower right corner of the circuit board. Adjacent to the MEMS device is the sensing circuitry for the x and y axes. Only the operation of a single axis is demonstrated in this work. The boards stacked under the sensor comprise of the actuator circuits and the power supply. The system is connect to an development board housing the Xilinx Kintex 7 FPGA which controls the system.

Fig. 9. The voltage waveforms at opposing electrostatic drives over 1.6 sampling periods. These waveforms demonstrate the operation of the time-multiplexed self-sensing method. The square pulse at 0.2 µs and 1.2 µs is the sensing signal, while the ramp in the center of the waveform is the actuating signal. The dead-time between each phase is sufficient for each system to switch off before the other is switched on. The delay between the actuation phase and the actuation signal is attributed to a shoot-through protection in the amplifier and the time taken for the switches to turn on and off. Note the actuation and sensing signals are of the opposite polarity on each drive.

Fig. 10. The step response of the actuator. The system input \( u \) was stepped by a value of 0.7625. The measured response is the actuation voltage \( v_a \) at one of the electrostatic drives.

Fig. 11. The static characteristic of the nanopositioning system from the input \( u \) to the output \( y \). The system was actuator with a 0.5 Hz triangle wave of amplitude 0.75 while measurements were performed and the signals were filtered to remove the switching harmonics from the sensor signal. The linear fit has a gain of 0.2674 though a distinct nonlinearity is observed in the characteristic.

Fig. 12. The step response of the system from the input \( u \) to the output \( y \). The resonant motion of the MEMS device is clearly captured by the sensor.

To show the operation of the actuator, the system was stepped by \( u = 0.7629 \). The step response is shown in Figure 10. From the fitted exponential lag, the identified model of the actuator is as follows.

\[
H_a(s) = 22.85 \cdot \frac{66020}{s + 66020} \tag{10}
\]

The reduced gain compared to the theoretical gain of 25 is attributed to the voltage drop across the amplifier switches and the isolating diodes.

The measurements in Figure 11 and Figure 12 show the response of the entire system from the input signal \( u \) to the sensor output \( y \). Both measurements are filtered by a 4th order butterworth low pass filter with a cut-off frequency of 10 kHz. The measurements in Figure 11 show the static relationship from input to output. The linear fit has a gain of 0.2674, however, a distinct nonlinearity can be seen in the theoretically linear system. This effect is due to the isolating diodes in the amplifier. When reverse biased, these diodes have a junction capacitance that appears in parallel to the electrostatic drives from the perspective of the sensor. The junction capacitance is a nonlinear function of voltage and the current that flows due to the application of the sensing signal distorts the sensor output. The step response of the system is shown in Figure 12. The system is stepped by \( u = 0.7629 \) to achieve this response and it clearly shows the sensor has sufficient bandwidth to clearly capture the resonant motion of the MEMS device.
V. CONCLUSION
This work demonstrates the operation of a switched self-sensing method for a MEMS nanopositioner that employs electrostatic transduction. Actuator and sensor circuits have been designed to provide a fully digital interface to allow for easy control and the architecture of the actuator allows for fast switching despite the high voltages needed to drive the MEMS nanopositioner.

The current issue with the device is the nonlinearity that is observed in the static response of the system. This is caused by the junction capacitance of the isolation diodes used in the actuator circuit. Current work is being undertaken to eliminate this nonlinearity with architecture changes to the actuator circuit and appropriate choice of switches to minimize the non-ideal characteristics of semiconductor devices.

Upon addressing this issue future work will identify a linear model and feedback control will be implemented within the FPGA to allow the nanopositioner to be used within an atomic force microscope to perform imaging.

REFERENCES