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A 30ppm <80nJ Ring-Down-Based Readout Circuit for Resonant Sensors

H. Jiang, Z.Y. Chang and M.A.P. Pertijs

Electronic Instrumentation Laboratory, Delft University of Technology

Delft, The Netherlands

Abstract—This paper presents an energy-efficient readout circuit for micro-machined resonant sensors. It operates by briefly exciting the sensor at a frequency close to its resonance frequency, after which resonance frequency and quality factor are determined from a single ring-down transient. The circuit employs an inverter-based trans-impedance amplifier to sense the ring-down current, with a programmable feedback network to enable the readout of different resonant sensors. An inverter-based comparator with dynamically-adjusted threshold levels tracks the ring-down envelope to measure quality factor, and detects zero crossings to measure resonance frequency. The excitation frequency is dynamically adjusted to accommodate large resonance frequency shifts. Experimental results obtained with a prototype fabricated in 0.35 μm standard CMOS technology and three different SiN resonators are in good agreement with conventional impedance analysis. The prototype achieves a frequency resolution better than 30 ppm while consuming less than 80 nJ/meas from a 1.8 V supply, which is $7.8 \times$ less than the state-of-the-art.

Index Terms—MEMS resonant sensors readout, energy efficiency, ring-down measurement, frequency tracking technique, inverter-based front-end.

I. INTRODUCTION

Resonant sensors are widely applied for various physical measurements, such as pressure sensing, mass detection, oil-quality measurement and gas sensing. Generally, in these systems, a change in the physical quantity of interest leads to a shift in the resonance frequency f_{res} of a mechanical resonator. By equipping the resonator with an appropriate capacitive or piezoelectric transducer, f_{res} can be detected and digitized by a readout circuit. Micro- or nano-mechanical resonators are an attractive choice for energy-constrained applications, because they can be operated at very low energy levels. For instance, piezoelectrically actuated SiN resonators have been reported that can detect ppm-level concentrations of volatile organic compounds at power levels below 100nW [1]. For such resonators, an energy-efficient readout circuit is required, since the overall energy consumption is readily dominated by the circuit.

Resonant sensors are typically read out using an oscillator-based circuit [2 – 4]. This is illustrated in Fig. 1a, in which the resonator is modeled by a resonant circuit (consisting of R_m , C_m and L_m), representing the mechanical resonance, in parallel with a feed-through capacitance C_p . A sustaining amplifier, typically a trans-impedance amplifier, senses the current produced by the resonator and feeds back a drive signal that ensures 360° phase shift around the loop, so that oscillation is maintained at f_{res} . A comparator translates the drive signal to logic levels, after which a counter digitizes its frequency relative to a reference clock.

Oscillator-based readout circuits tend to be relatively power hungry [2, 3]. This is mainly because the bandwidth of the sustaining amplifier must be substantially higher than f_{res} , in order to prevent the phase shift that it introduces in the feedback loop from moving the oscillation frequency significantly away from f_{res} . Moreover, it is difficult to design oscillator-based readout circuits for piezoelectric resonators with relatively large parasitic feed-through

capacitance, because this capacitance, rather than the mechanical resonance, will determine the phase shift at f_{res} . In this case, the effect of C_p needs to be compensated for, for instance using a bridge configuration with a replica device driven out-of-phase [2], which will increase area and power consumption. Finally, oscillator-based readout circuits do not readily provide information about the resonator's quality factor Q , while this can also provide valuable information, for example, about surface binding or pressure changes, which relate to damping and energy-loss effects [5].

As an alternative, ring-down based readout, illustrated in Fig. 1b, can be more energy-efficient, is capable of detecting both f_{res} and Q , and can operate in the presence of large parasitic capacitance [6 – 10]. The basic idea of ring-down measurement is to operate the resonator as a tuning fork: excite it and let it ring down at f_{res} . To excite the resonator, an excitation source drives the transducer in order to add sufficient energy to the desired resonant mode. When it is turned off, the sensor dissipates the energy while vibrating at f_{res} . This produces an exponentially-decaying current through the piezoelectric transducer, which is converted to a voltage by an I-V converter. From the zero crossings of this voltage, f_{res} can be extracted, while its envelope contains information about Q . Integrated readout circuits based on this technique have been reported [8, 9], and consume significantly less energy than oscillator-based readout circuits for similar sensors [2, 3]. These designs, however, require three or more ring-downs to measure both f_{res} and Q , while this information can, in principle, be obtained from a single ring-down, and thus at significantly lower energy consumption.

This paper describes the architecture, circuit realization and experimental characterization of an energy-efficient ring-down-based readout circuit for resonant sensors [10]. Unlike previous designs, it combines f_{res} and Q measurements in a single ring-down. It employs a comparator with a dynamically-adjusted threshold and a current-efficient inverter-

based front-end. A prototype realized in 0.35 μm standard CMOS technology achieves 1.6 $\times$ better resolution at 7.8 $\times$ lower energy consumption than prior ring-down-based designs.

This paper is organized as follows. In Section II, the architecture and the operating principle of the proposed ring-down-based readout circuit are described. Sections III and IV discuss the implementation details of the circuit and present a noise analysis, respectively. Section V describes the measurement results. The paper concludes in Section VI.

II. BASIC OPERATING PRINCIPLE

A. Target Sensors

The readout circuit described in this paper has been optimized for use with silicon-nitride (SiN) clamped-clamped beam resonators with integrated piezoelectric (aluminum nitride) transducers [1]. Their length ranges from 100 μm to 1000 μm , their width is between 20 μm and 100 μm , and the SiN thickness is about 500 nm. A typical impedance characteristic around f_{res} , measured using a bench-top impedance analyzer, is shown in Fig. 2. At frequencies close to the resonance frequency of approximately 536 kHz, the resonator can be modeled as an RLC tank (motional resistance R_m , inductance L_m and capacitance C_m) in parallel with a capacitor C_p that models the capacitance of the piezoelectric transducer and any parasitic capacitances. The readout circuit aims to measure f_{res} and Q of the RLC branch, where

$$f_{res} = \frac{1}{\sqrt{L_m C_m}}, \quad Q = \frac{1}{R_m} \sqrt{\frac{L_m}{C_m}} \quad (1)$$

The relative resolution for the measurement of f_{res} targeted in this work is 30ppm. The impact of C_p on the impedance is evident from the fact that the resonator exhibits only a small phase shift around resonance. This makes it difficult to implement an oscillator-based readout circuit for this type of resonators, unless the effect of C_p is compensated for [2, 3].

B. Readout Architecture

The basic architecture of the ring-down-based readout circuit is shown in Fig. 3. It consists of a switched excitation source, a trans-impedance amplifier (TIA), a comparator and a digital counter. There are two operating phases in the ring-down measurement: an excitation phase ϕ_{exc} and a ring-down phase ϕ_{ring} . During ϕ_{exc} , the sensor is driven by the excitation source V_{exc} , which has a frequency close to f_{res} , and starts oscillating at the driving frequency. Then, during ϕ_{ring} , the driving source is turned off and the sensor is connected to the readout circuit. Due to the energy that has been accumulated during the excitation phase, the sensor starts to produce an oscillating current at f_{res} , with an exponentially-decaying envelope [6]. This ring-down current is turned into a voltage V_{ring} by the TIA. The comparator detects the zero-crossings or threshold-crossings of this voltage, from which f_{res} and Q are derived.

C. Resonance Frequency Measurement

Zero-crossing detection and frequency counting are used to measure f_{res} , as illustrated in Fig. 4a. To detect the zero-crossings of V_{ring} , the threshold voltage of the comparator is set to V_{th0} , which corresponds to the average of V_{ring} . The resonance frequency is then extracted by counting the number of cycles M of a reference clock, with period T_0 , during a fixed number of N_{total} zero-crossings of the ring-down signal:

$$f_{\text{res}} = \frac{N_{\text{total}}}{MT_0} \quad (2)$$

D. Quality Factor Measurement

The envelope of the ring-down signal contains information about the quality factor Q of the resonant sensor. To digitize Q , threshold-crossing detection can be applied [7]: as shown in

Fig. 4a, in successive ring-downs, the comparator is operated at two different threshold levels V_{th1} and V_{th2} , and detects the time instants at which the envelope drops below these levels, which are quantized by counting the corresponding number of threshold crossings N_1 and N_2 . An approximation of Q can then be obtained from the formula [7]

$$Q = \frac{\pi(N_2 - N_1)}{\ln\left(\frac{V_{th1} - V_{th0}}{V_{th2} - V_{th0}}\right)} \quad (3)$$

E. Proposed Combined Measurement

In previous integrated implementations of the ring-down technique [8, 9], multiple successive excitation and ring-down phases were applied to determine f_{res} and Q . The readout circuit presented in this paper, in contrast, determines f_{res} and Q from one ring-down signal, which reduces measurement time and energy consumption.

As illustrated in Fig. 4b, a single comparator with a dynamically-adjusted threshold is used. The threshold level starts at V_{th0} , which corresponds to the steady-state level of ring-down signal. After a few cycles, it is switched to a higher threshold V_{th1} and then switched back in steps to V_{th0} . These threshold-switching points happen when the envelope of the ring-down signal drops below the relevant threshold level. Then, the number of the threshold crossings (N_1, N_2) can be counted, from which Q can be calculated using equation (3).

The counting continues until a pre-defined number of N_{total} ring-down cycles has been reached. Thus, as before, f_{res} can be obtained using equation (2), by counting the number of references-clock cycles M within these N_{total} cycles.

A time-out mechanism is used to decide the threshold-switching points. A time-out detector counts the number of reference-clock cycles that have passed since the last level crossing, as illustrated in Fig. 5. If this number exceeds a time-out level, the envelope is assumed to have dropped below the threshold. This time-out level is set roughly equal to one

cycle of the ring-down signal by employing the same timer to measure the duration of the first cycle. After a time-out, the threshold is switched to V_{th0} first, to ensure that a next level crossing can be found by the comparator. Once found, the threshold is switched to the next threshold level. The frequency of the reference clock is about 40 times higher than f_{res} to achieve a sufficiently accurate time-out detection.

F. Excitation Signal and Frequency Tracking

Ring-down measurement requires that the resonator absorbs enough energy in its resonance mode during the excitation phase. In principle, the resonator can be excited by an impulse. However, for the resonators used in this work, the required voltage amplitude would be beyond the capability of standard CMOS technology and would drive the resonator into its nonlinear region. Moreover, if the resonator has multiple dominant resonance modes, impulse excitation will cause the resonator to vibrate at all modes simultaneously, complicating accurate ring-down measurement. Therefore, instead, we use periodic excitation at a fixed frequency close to the target resonance mode. Thus, an excitation signal with smaller voltage amplitude can be used to excite resonator sensor, and excitation of undesired resonance modes can be prevented. For implementation simplicity, square-wave excitation is used.

The initial amplitude of the ring-down signal depends on the frequency of the excitation signal. By sweeping the frequency of the excitation signal near the resonance, a maximum ring-down current can be found when the driving frequency is exactly equal to f_{res} , at which the impedance of the resonator is minimum and equals value R_m . The initial amplitude of the ring-down current I_{RLC} decreases as the difference between excitation frequency and the resonance frequency Δf increases, as expressed by the following equation:

$$I_{RLC} = \frac{V_{exc}}{R_m \cdot \sqrt{1 + \left(2 \frac{f_{res} - f_{exc}}{f_{res}} Q\right)^2}} \quad (4)$$

where V_{exc} and f_{exc} are the amplitude and frequency of a sinusoidal excitation signal, respectively. (For square-wave excitation, this corresponds to the first harmonic of the excitation signal.) Hence, in order to get a sufficiently large ring-down current, the excitation frequency should be close enough to f_{res} , relative to the bandwidth f_{res}/Q , which is about 1.5 kHz for our sensor.

Because f_{res} shifts in response to changes in the physical quantity being measured, a fixed excitation frequency can typically not be used. Instead, after a ring-down measurement, the excitation frequency is updated based on the measured f_{res} . For a slowly-changing f_{res} , this will keep the excitation frequency within the bandwidth of the resonator. After startup, or in case of larger abrupt changes in f_{res} , the excitation frequency is swept across the frequency range of interest in steps small enough to allow the resonance to be found. With this approach, typically only a single excitation will be needed per measurement.

III. CIRCUIT REALIZATION

Fig. 6 shows a block diagram of the proposed readout circuit, which consists of an inverter-based TIA with reconfigurable feedback network, a comparator, a sensor driver, a period time-out detector and a capacitive DAC (CDAC). Control signals and the excitation clock are generated off-chip using an FPGA for flexibility.

A. TIA

The TIA is used to convert the ring-down current to a ring-down voltage for the comparator. It should have a small enough input impedance to ensure that most of the motional current

flows into the TIA instead of into the parasitic capacitor C_p ; a high enough trans-impedance to produce an output voltage with sufficient initial amplitude; and good enough noise performance for the targeted resolution, as will be discussed in Section IV-B. These requirements lead to a minimum trans-conductance for the operational transconductance amplifier (OTA) of the TIA.

To optimize the energy efficiency, an inverter-based OTA is used, as shown in Fig. 7 [11]. An auto-zeroing phase is adopted to set the bias point of the OTA's transistors using a floating current source. The difference between the gate voltages of these transistor and the input bias voltage is stored on capacitors C_{AZ1} and C_{AZ2} . Thus, the sensitivity of the bias current and the DC output voltage to process, supply-voltage and temperature variations is significantly less than in a conventional inverter [11]. The input transistors are biased in weak-inversion to optimize current efficiency. The output swing is limited to 1 V to prevent distortion of the ring-down envelope from causing significant errors in the Q measurement.

The TIA's feedback network has been made reconfigurable to allow the circuit to measure different resonant sensors. The gain value of TIA can be programmed from 160 k Ω to 5.12 M Ω . The TIA consumes about 12 μ A.

B. Comparator

Propagation delay of the comparator results in an error in f_{res} , since the input signal of the comparator has an exponentially decaying envelope, and the propagation delay is typically amplitude-dependent. Hence the delay at the start of the ring-down will be smaller than that at the end, and will not cancel out in the time measurement. Therefore, we design the comparator to have a delay that is sufficiently small compared to the resonance period. To do so in an energy-efficient way, an inverter-based auto-zeroing comparator is utilized.

Fig. 8 shows the circuit diagram of the comparator, which consists of two inverter stages. The input stage is dynamically biased by means of an auto-zeroing scheme similar to that of OTA in the TIA (but without cascode transistors). During the auto-zeroing phase, the two input transistors M_{n1} and M_{p1} are diode-connected and biased at the same bias current by a floating current source, while their gate-source voltages, including offset and $1/f$ noise, are stored on the capacitors C_{AZC1} and C_{AZC2} . The floating current source is kept in the circuit during the comparison phase and assists in class-AB biasing of the output stage, thus reducing its static current consumption while maintaining a high slew rate. The comparator consumes about 8 μA supply current during operating. The difference in propagation delay between a 20 mV peak-to-peak input voltage and a 400 mV peak-to-peak input voltage is below 18 ns.

C. On-chip Sensor Driver

As shown in Fig. 6, the square-wave excitation signal is generated using an on-chip driver, which is clocked by an off-chip excitation clock. The circuit diagram of the driver is shown in Fig. 9. When ϕ_{EX} is high, V_{exc} is pulled to ground by M_{n8} . When ϕ_{EX} is low, the output V_{exc} is pulled up by source follower M_{n9} . The input of this source follower is derived from a reference voltage V_{ref} using a diode-connected transistor M_{n6} . Thus, the amplitude of V_{exc} is defined by V_{ref} . Transistors M_{p3} and M_{n7} power down the source follower and its biasing when ϕ_{EX} is high. When active, the sensor driver consumes an average supply current of 1 μA .

D. Implementation and Timing Details

As shown in Fig. 10, several auto-zeroing steps are applied to correct for the offset of the front-end and other errors in every ring-down measurement, so as to maximize the number of zero-crossings that can be detected.

Before the excitation phase, the OTA and comparator are auto-zeroed, as described above. In Fig. 10, single offset-storage capacitors are shown at their inputs for simplicity, but the actual implementation involves pairs of capacitors, as shown in Fig. 7 and 8. During phase ϕ_{AZ} , both the OTA and the comparator are configured in unity-gain feedback, so that their input offset is stored on capacitors C_{AZ} and C_{AZC} , respectively. Then, during ϕ_{AZC} , the comparator is kept in unity-gain feedback, while the OTA is connected to its feedback network and the resonator. By doing this, any residual output offset of the TIA will be stored on capacitor C_c , which capacitively couples the TIA's output to the input of the comparator.

The binary-weighted capacitors of the 3-bit CDAC are also capacitively coupled to the comparator's input and are driven using logic signals $D_{1,2,3}$. During auto-zeroing, these signals are zero. During the ring-down phase, the CDAC's inputs are switched, changing the voltage on C_c via charge-redistribution, and thus changing the comparator's threshold as desired.

After auto-zeroing, the sensor is excited during phase ϕ_{exc} . Then, the feed-through parasitic capacitor C_p is discharged during phase ϕ_{cp} , in order to prevent the charge on C_p from shifting the output of the TIA during the ring-down. As shown Fig. 11, when C_p is large compared to the sensor's impedance, any charge that has been left on C_p at the end of excitation phase will cause an undesired transient at the output of TIA, which will cause errors in the Q measurement. This is solved by introducing a short phase ϕ_{cp} to discharge C_p after the excitation phase to make the time for recovery shorter. In order to realize this, the TIA is held in unity-gain during ϕ_{cp} .

IV. NOISE ANALYSIS

The resolution of the resonance frequency measurement is determined by two effects: quantization errors and timing jitter associated with the thermal noise sources in the circuit.

A. Quantization Errors

In order to quantize the duration of the N_{total} ring-down cycles, a digital counter, realized on FPGA, counts the number of cycles M of a reference clock. This is associated with two quantization errors, which are uncorrelated, at the first cycle and the last cycle, leading to an RMS quantization error (σ_q) of

$$\sigma_q \left(\frac{\Delta f}{f} \right) = \frac{f_{res}}{\sqrt{6} N_{total} f_{clk}} \quad (5)$$

where f_{clk} is the reference clock frequency. To arrive at a resolution better than 30 ppm, with $400 \text{ kHz} < f_{res} < 600 \text{ kHz}$ and $300 < N_{total} < 600$, f_{clk} should be at least 25 MHz.

B. Timing Jitter

There are two dominant thermal-noise sources in the ring-down-based readout circuit: noise of the TIA and noise of the comparator. The combined noise at the input of comparator can be estimated by adding the TIA's output noise and the equivalent input noise of the comparator. This combined noise will result in timing jitter in the zero-crossing detection. The expression for the RMS timing jitter Δt in terms of the initial ring-down signal amplitude A_0 , the combined rms voltage noise v_n and the ratio of N_{total} and Q is obtained as

$$\Delta t = \frac{v_n}{2\pi f_{res} A_0 e^{-\frac{\pi N_{total}}{Q}}} \quad (6)$$

in which the denominator represents the slope of at the zero-crossings of the ring-down signal at the end of the measurement. By assuming $N_{total} \approx Q$, the timing jitter simplifies to

$$\Delta t = \frac{v_n}{2\pi f_{res} A_0 e^{-\pi}} \quad (7)$$

The uncertainty in the resonance-frequency measurement as a result of this timing jitter can be expressed as

$$\sigma_n \left(\frac{\Delta f}{f} \right) = \frac{\Delta t \cdot f_{res}}{N_{total}} = \frac{v_n}{2\pi A_0 N_{total} e^{-\pi}} \quad (8)$$

Using this equation, the targeted resolution can be translated into a requirement on the signal-to-noise ratio A_0/v_n at the start of the ring-down, which for $N_{total} = Q = 300$ is about 430 and determines the current levels in the inverter-based OTA and comparator.

V. MEASUREMENT RESULTS

The readout circuit has been realized in 0.35 μm standard CMOS technology and tested with three different resonators with $f_{res} = 469 \text{ kHz}$, 537 kHz , 592 kHz and $Q = 521, 386, 376$ (under ambient conditions). The TIA's reconfigurable feedback network was used to adapt its gain to these different resonators. Fig. 12 shows a micrograph of the chip, with an active area of 0.24 mm^2 , and of the three resonant sensors. The readout operates from supply voltages from 1.8 V to 3.3 V. The experimental results reported here were obtained at a supply voltage of 1.8 V, at which the current consumption is $34 \mu\text{A}$, including the current consumption of the on-chip driver that provides energy to the resonator and of the period time-out detector.

The resolution of the measured f_{res} as a function of the number of ring-down cycles N_{total} is shown in Fig. 13. The resolution at a particular value N_{total} is calculated as the standard deviation of 1000 successive measurements. When N_{total} is smaller than 200, the resolution is limited by the quantization noise associated with the 25 MHz reference clock of the counter, in good agreement with equation (5). For larger values of N_{total} , jitter due to the thermal noise of the TIA and the comparator limits the resolution. For a measurement time of 1.3 ms, including all phases of the ring-down measurement, an optimum better than 30 ppm can be reached for all three resonators when N_{total} is roughly equal to Q . The resolution of Q is limited by quantization noise and is better than 1%.

To demonstrate the combined measurement of f_{res} and Q , the three resonant sensors have been exposed to different pressure levels in a vacuum chamber. Pressure is swept from 200 mbar to 1000 mbar, which leads to significant changes in Q due to changing air damping, and a smaller variation in f_{res} , as shown in Fig. 14. The results obtained using our readout circuit are in good agreement with results from a bench-top impedance analyzer (Agilent 4294A). The mismatch in f_{res} between the two data sets is smaller than 200 Hz and that in Q is smaller than 15. The Q values obtained using our readout circuit are systematically slightly lower than those obtained with the impedance analyzer. This is likely due to the finite input resistance of the TIA, which slightly increases the losses in the resonator.

In order to illustrate the frequency tracking technique, one resonant sensor and a reference PT100 thermometer have been exposed in a climate chamber to temperature changes, since these will cause a significant variation of f_{res} . Temperature is varied step-wise between 10°C and 40°C while the humidity level is kept constant. As shown in Fig. 15, a 30°C temperature variation causes a frequency shift of about 8 kHz, which far exceeds the bandwidth of the resonator of 1.37 kHz and can therefore not be measured using a fixed excitation frequency. The ring-down measurement circuit, employing the frequency-tracking technique, nevertheless tracks this large shift and maintains a good resolution (31ppm).

For 9 samples of the readout chip, the measured sample-to-sample variation in f_{res} and Q (for the same resonator under constant ambient conditions), is less than ± 90 Hz and ± 2.5 , respectively. The measured shift of f_{res} and Q due to power supply variation in the range from 1.8 V to 3.3 V is below 100 Hz and 3, respectively. (In these measurements, averaging was applied to properly distinguish between systematic effects and noise).

In Table I, the performance of the readout circuit is summarized and compared with several previously published readout circuits for similar resonant sensors. Oscillator-based readout circuits [2, 3] achieve similar or better resolution, however, at much higher power

consumption and much longer measurement time, without providing information about Q . Compared with earlier ring-down based circuits [9], this readout circuit consumes $1.7\times$ less power and provides a $1.6\times$ better resolution in a $5\times$ shorter measurement time .

VI. CONCLUSION

Ring-down-based readout of resonant sensors is a promising readout approach for use in energy-constrained applications. Compared with traditional oscillator-based readout, it can determine f_{res} at much lower energy consumption. Moreover, it is insensitive to the parasitic feedthrough capacitance and allows simultaneous Q measurement. The proposed readout circuit, optimized for SiN micro-machined resonant sensors, is implemented in $0.35\ \mu\text{m}$ standard CMOS technology and has an active area of $0.24\ \text{mm}^2$. It employs a dynamically-switching inverter-based level-crossing detector to determine f_{res} and Q from a single ring-down transient signal. The frequency tracking technique, rather than the use of a fixed excitation frequency, allows the detection of large shifts in f_{res} . A programmable feedback network is applied to enable the readout of different resonant sensors. The results obtained with three different resonant sensors are in good agreement with conventional impedance analysis. The circuit achieves a frequency resolution of 30 ppm while consuming less than 80 nJ/meas from a 1.8V supply, which is $7.8\times$ less than the state-of-the-art.

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Figure captions:

Fig. 1: Block diagram of (a) an oscillator-based readout circuit; (b) the proposed ring-down-based readout circuit.

Fig. 2: Measured typical impedance characteristic of the piezo-electrically actuated SiN resonators targeted in this work.

Fig. 3: Basic architecture of the ring-down measurement circuit.

Fig. 4: Timing diagram of f_{res} and Q measurements (a) using multiple successive ring-downs; (b) combined in a single ring-down.

Fig. 5: Dynamic threshold adjustment for the combined f_{res} and Q measurement.

Fig. 6: Block diagram of the ring-down readout.

Fig. 7: Circuit diagram of the OTA.

Fig. 8: Circuit diagram of the comparator.

Fig. 9: Circuit diagram of sensor driver.

Fig. 10: Implementation details and timing diagram.

Fig. 11: Effect of parasitic capacitance C_p on the TIA's output.

Fig. 12: Micrographs of the readout chip and prototype resonators.

Fig. 13: Measured resolution of f_{res} as a function N_{total} .

Fig. 14: Measured frequency and quality-factor shift associated with different pressure levels.

Fig. 15: Measured resonance frequency tracking during temperature transients.

Table captions:

Table I: Performance summary and benchmarking.

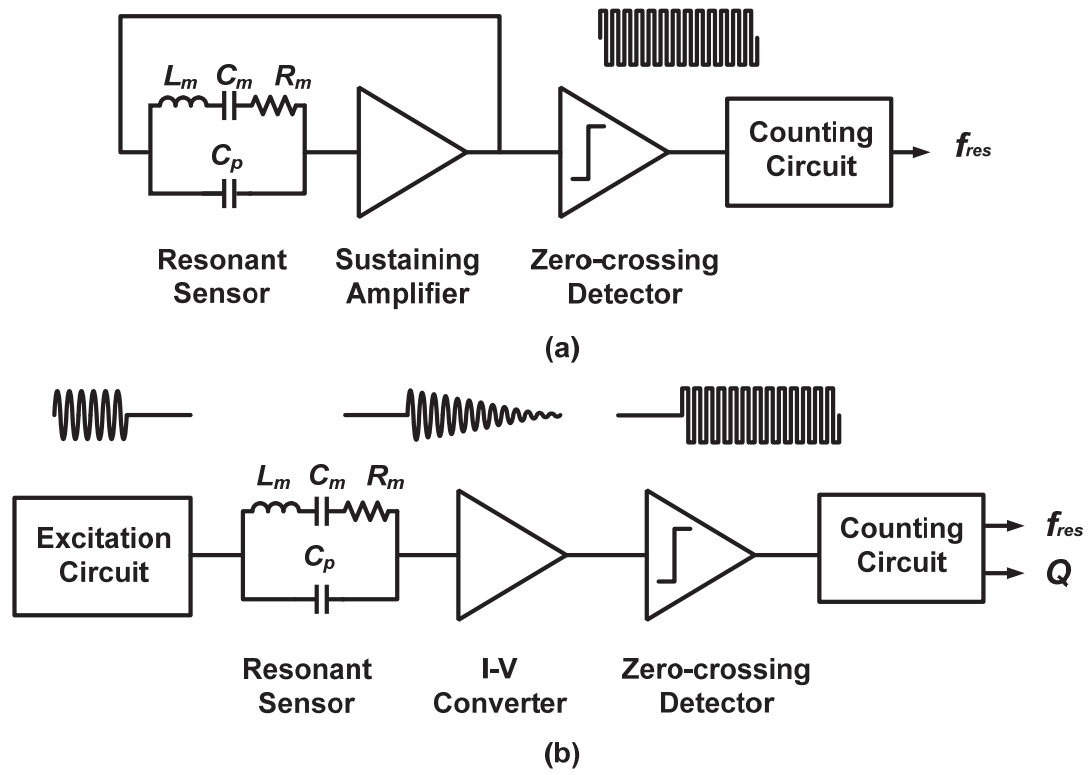


Fig. 1. Block diagram of (a) an oscillator-based readout circuit; (b) the proposed ring-down-based readout circuit.

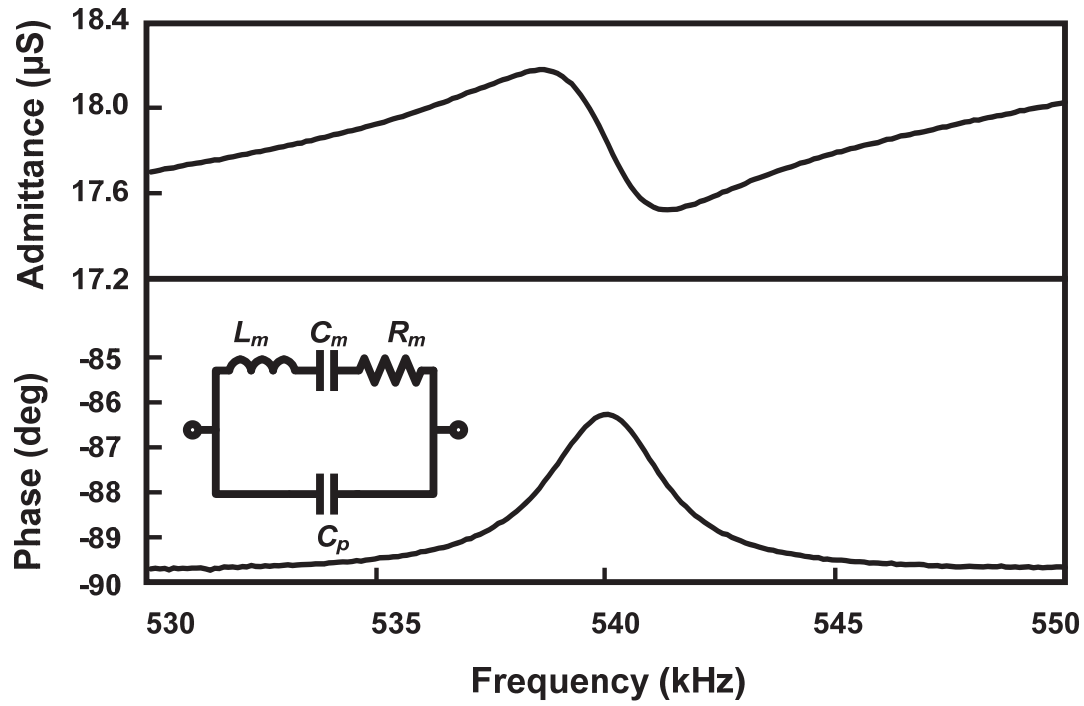


Fig. 2. Measured typical impedance characteristic of the piezo-electrically actuated SiN resonators targeted in this work.

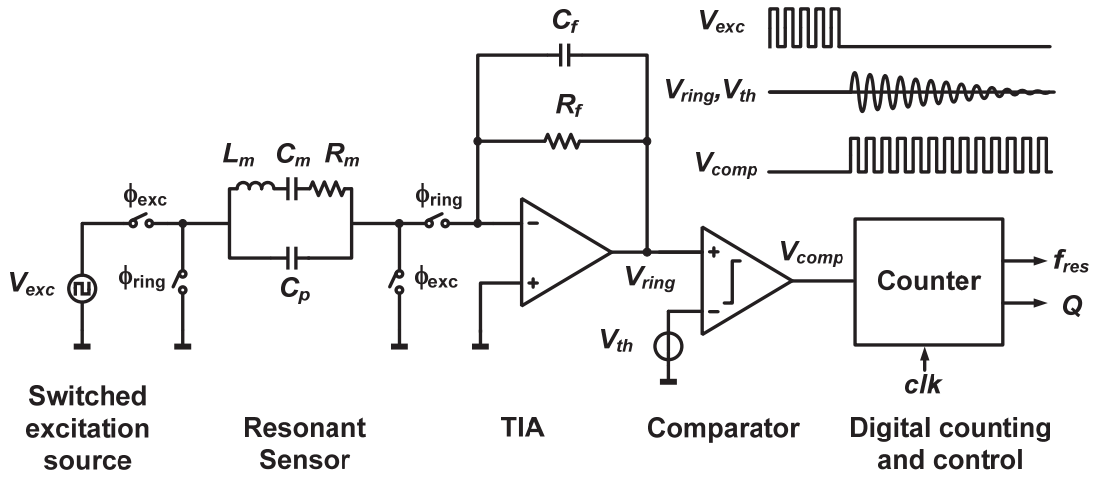


Fig. 3. Basic architecture of the ring-down measurement circuit.

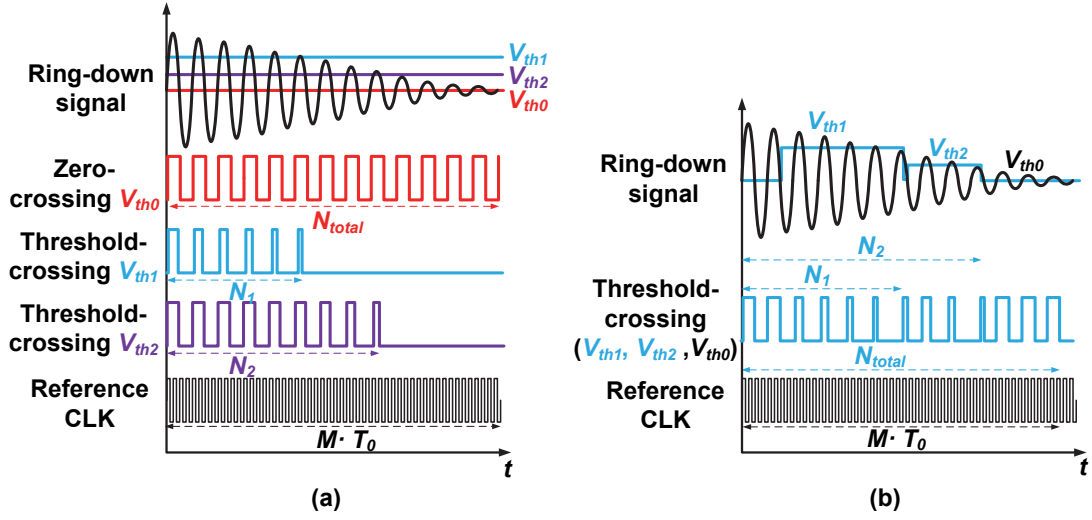


Fig. 4. Timing diagram of f_{res} and Q measurements (a) using multiple successive ring-downs; (b) combined in a single ring-down.

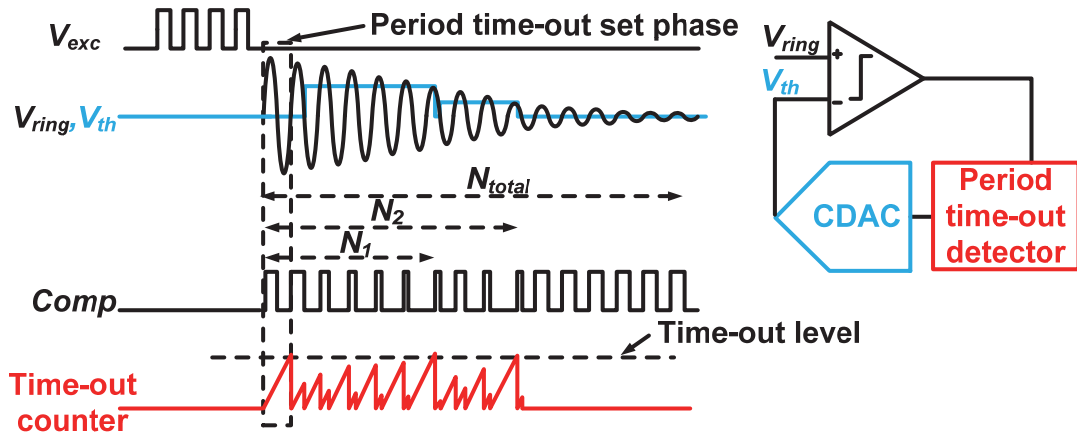


Fig. 5. Dynamic threshold adjustment for the combined f_{res} and Q measurement.

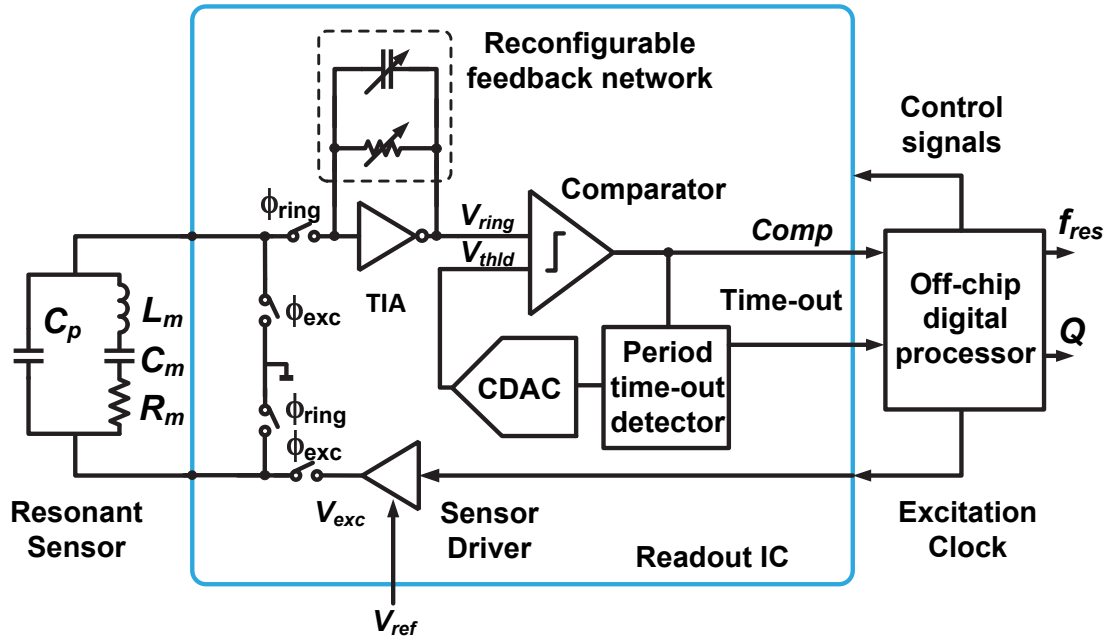


Fig. 6. Block diagram of the ring-down readout.

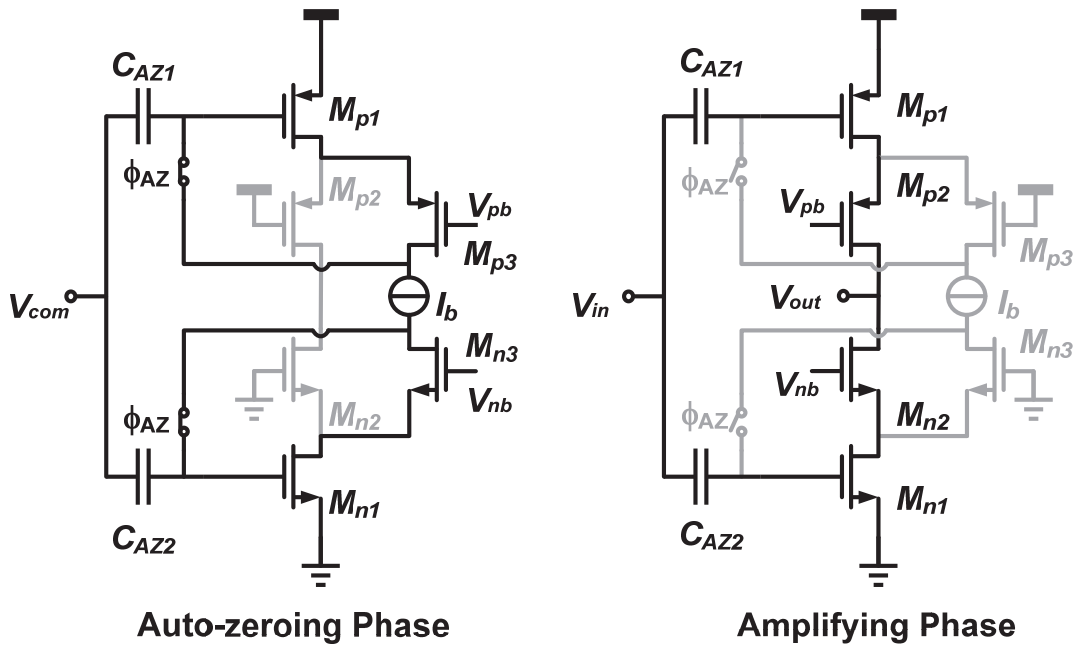


Fig. 7. Circuit diagram of the OTA.

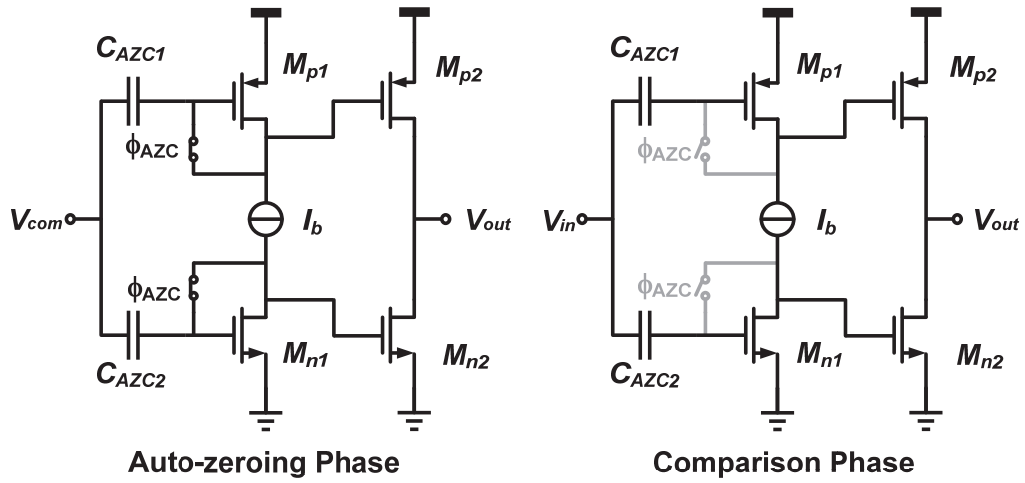


Fig. 8. Circuit diagram of the comparator.

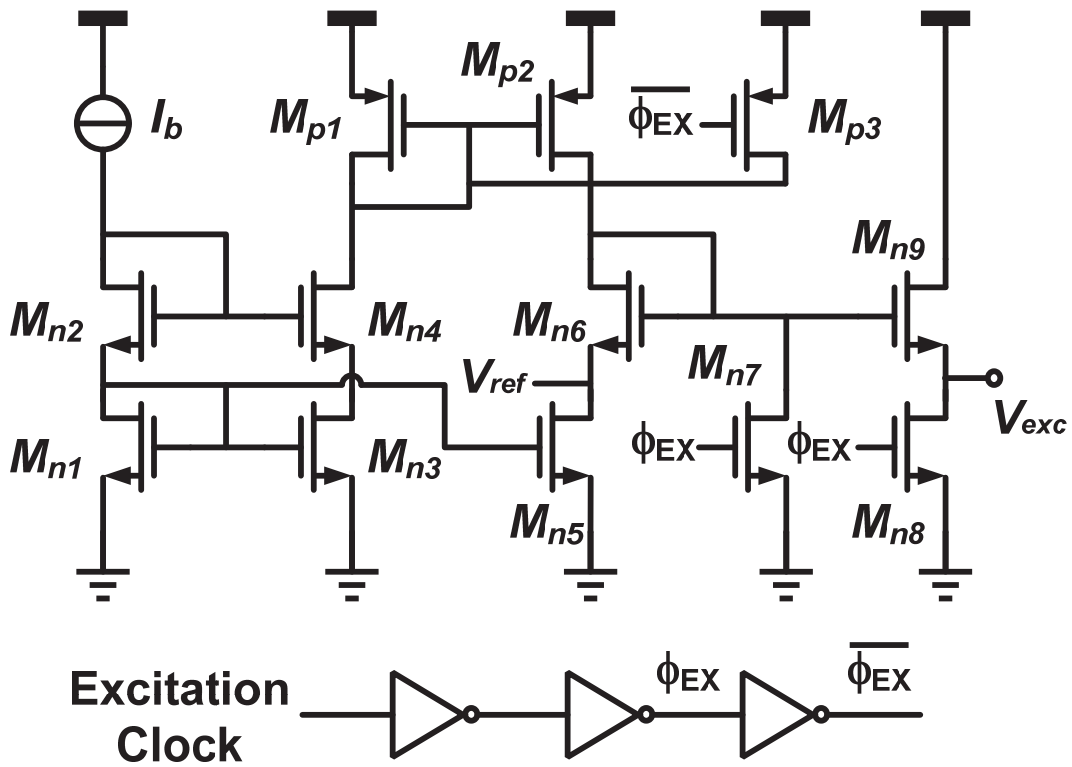


Fig.9. Circuit diagram of the sensor driver.

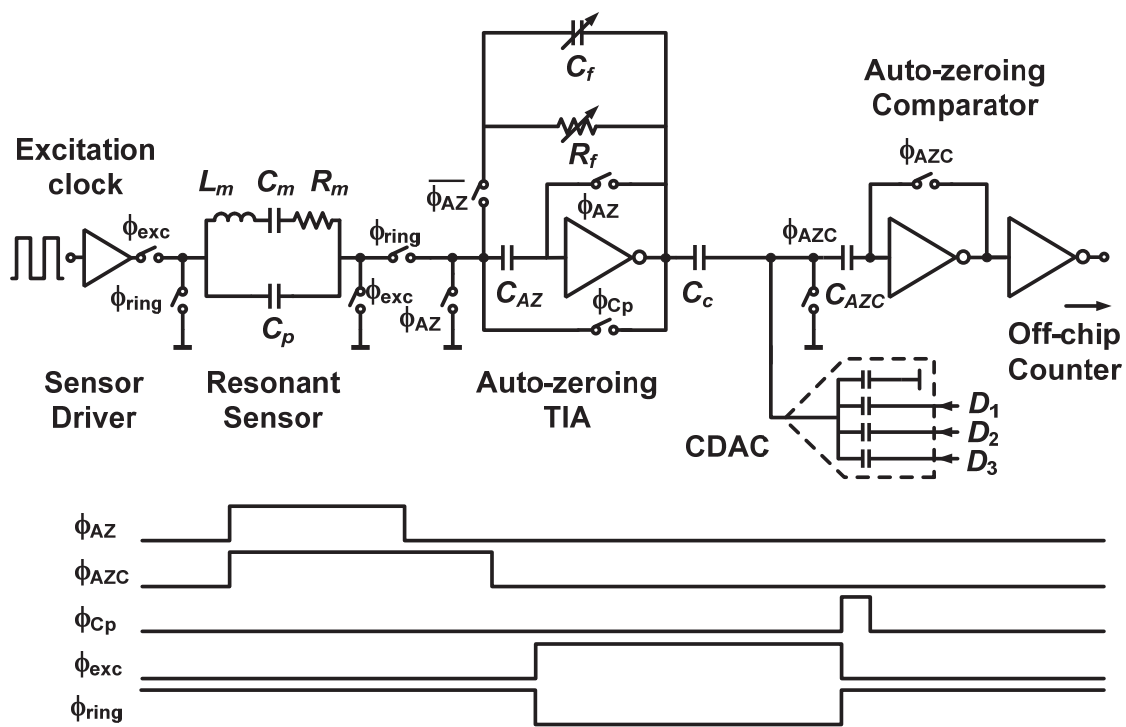


Fig.10. Implementation details and timing diagram.

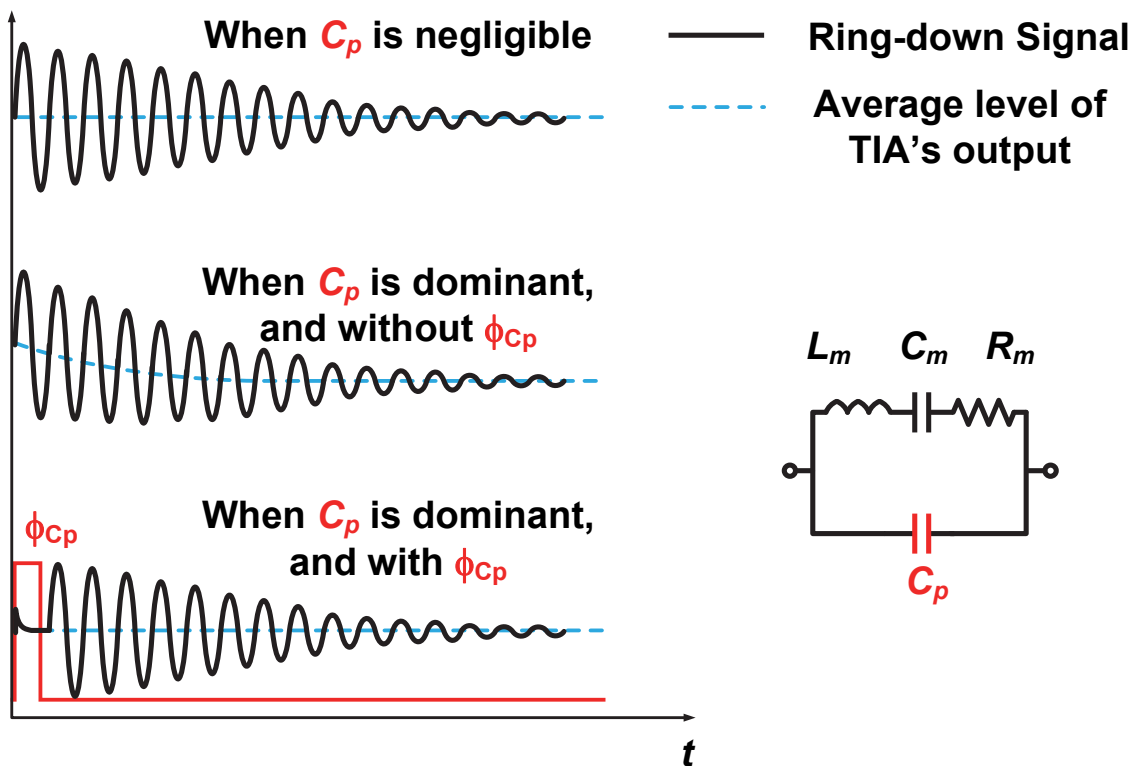


Fig.11. Effect of parasitic capacitance C_p on the TIA's output.

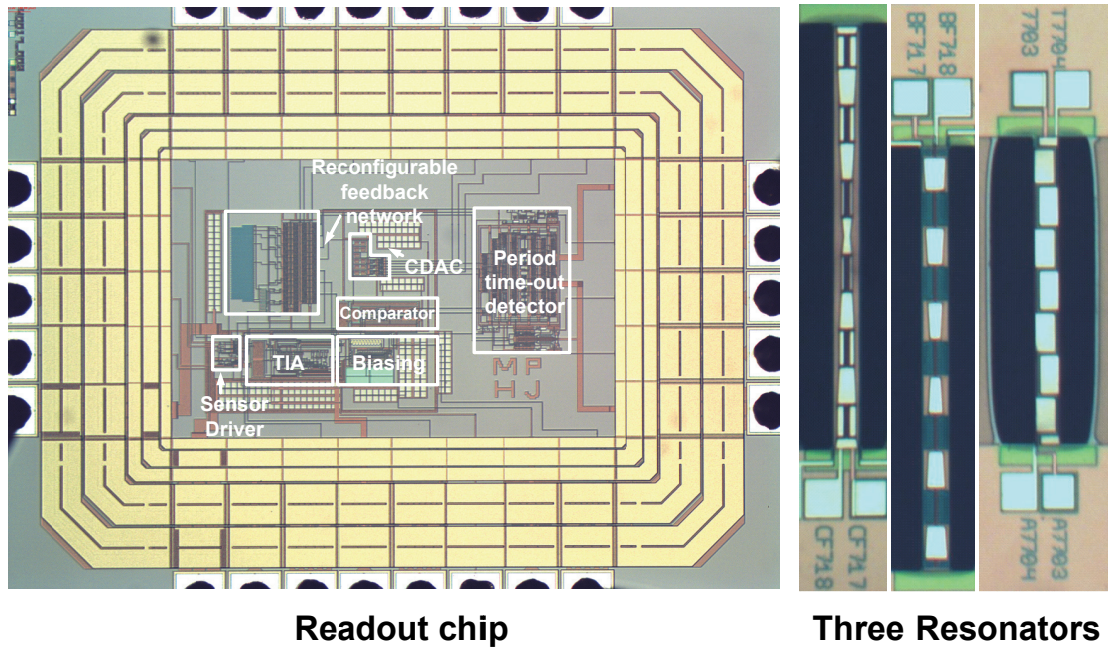


Fig.12. Micrographs of the readout chip and prototype resonators.

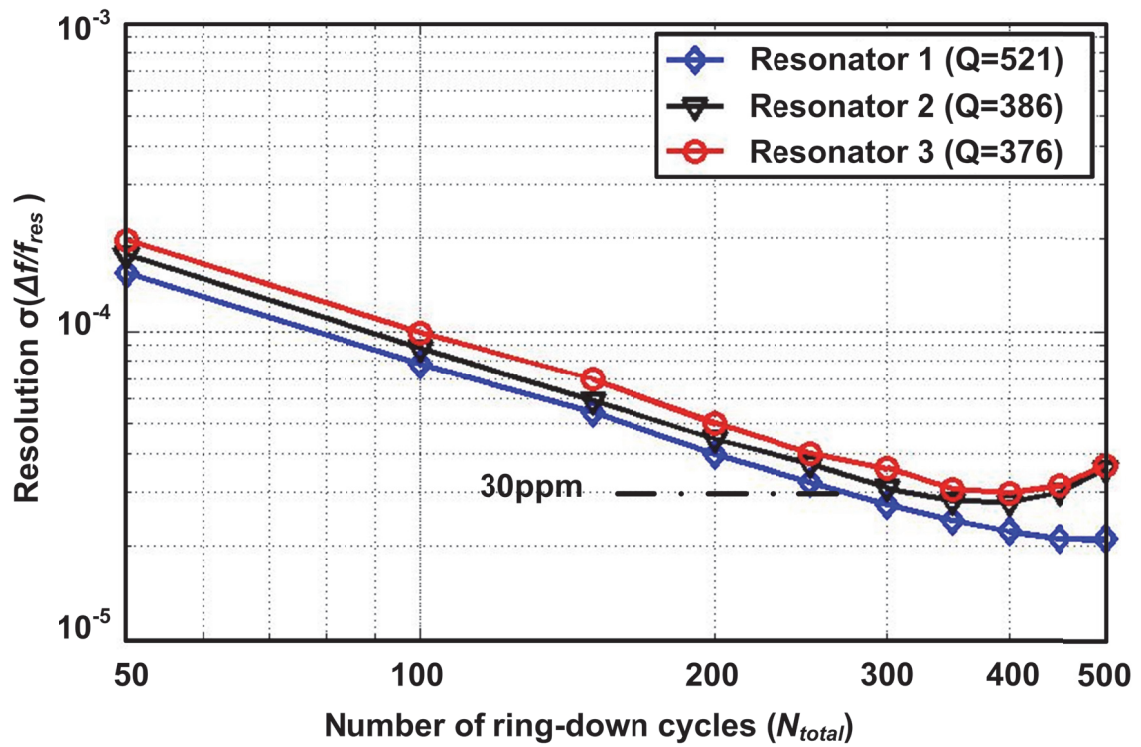


Fig.13. Measured resolution of f_{res} as a function N_{total} .

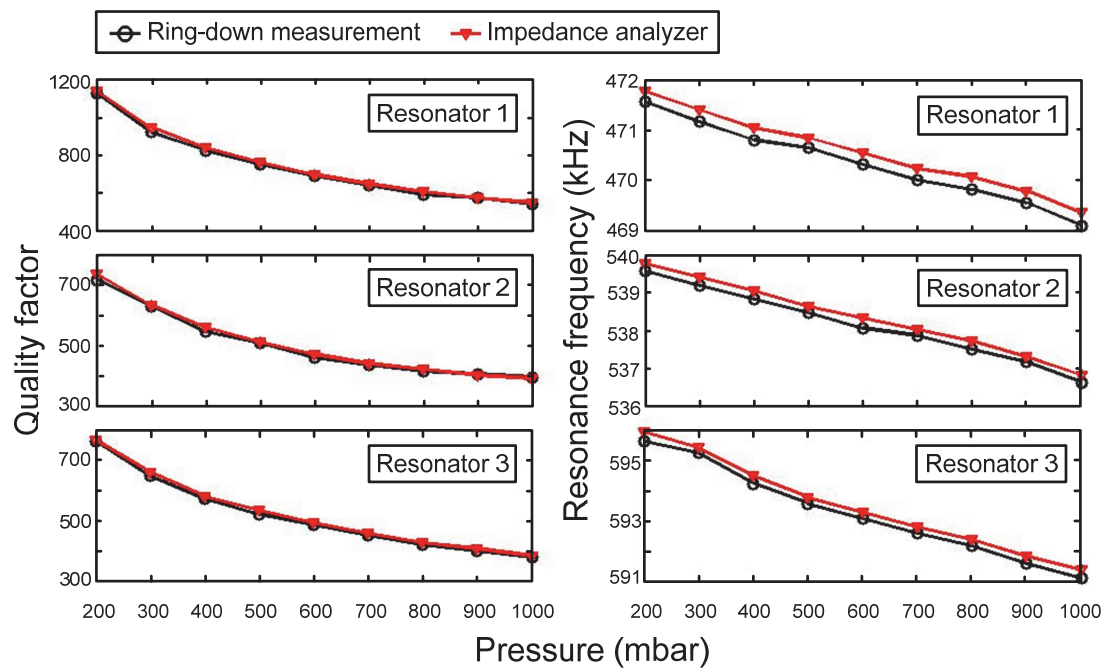


Fig.14. Measured frequency and quality-factor shift associated with different pressure levels.

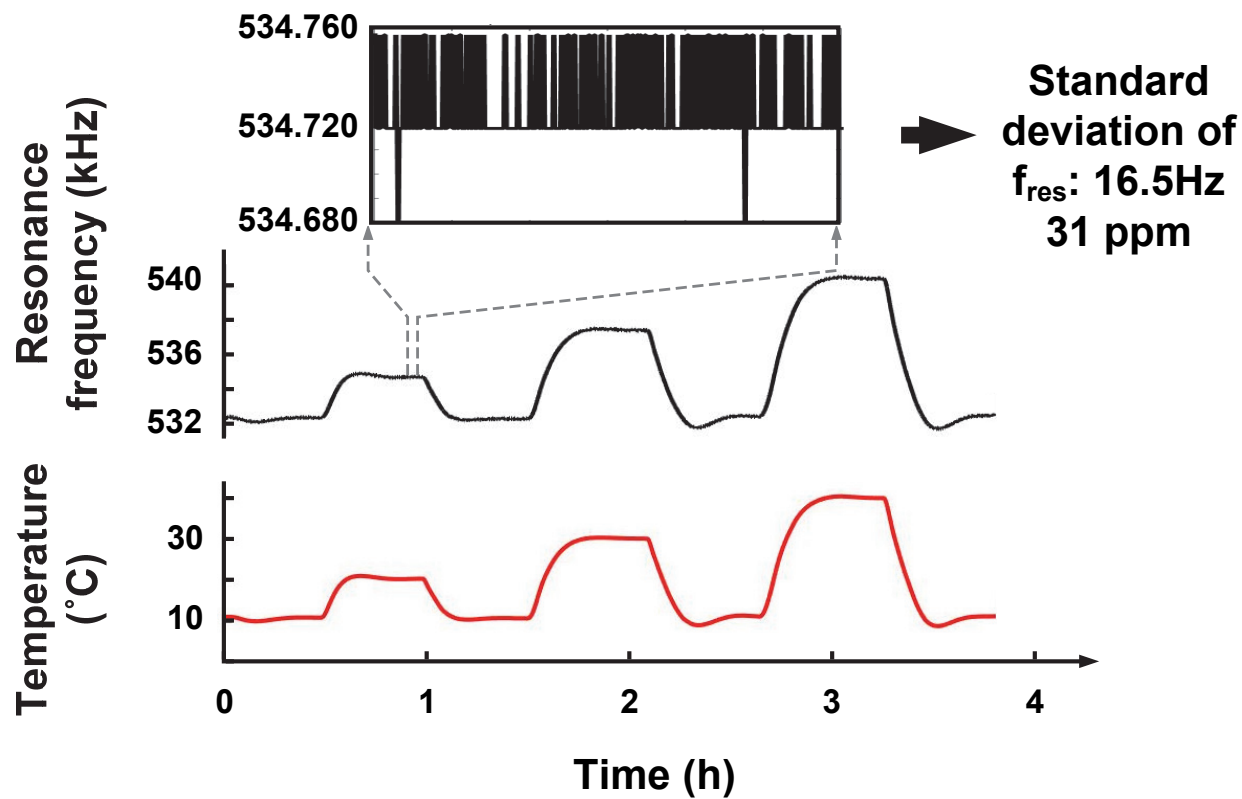


Fig.15. Measured resonance frequency tracking during temperature transients.

Table I. Performance summary and benchmarking.

Parameter	Petrescu [2]	Pettine [3]	Zeng [8]	Yan [9]	This work		
Technology	0.25 μ m CMOS	0.25 μ m CMOS	0.35 μ m CMOS	0.35 μ m CMOS	0.35 μ m CMOS		
Supports Q measurement	No	No	Yes	Yes	Yes		
Supply voltage (V)	3.3	3.3	3.3	3.3	1.8		
Supply current (μ A)	409	409	36	31.5	34		
Resonance freq. (kHz)	1980	2170	2010	535.8	469	536	592
Quality factor	280	450	667	386	521	386	376
Conversion time (ms)	5000	10000	3	6	1.3	1.2	1.1
Resolution (ppm)	27	2.4	250	50	21	27.7	29.8
Energy per meas. (μ J)	6748.5 ¹	13497 ¹	0.356 ^{2,3}	0.624 ^{2,3}	0.080 ²	0.073 ²	0.067 ²

¹ Excludes frequency counter; ² excludes counters; ³ excludes sensor driver