

Special Issue Reprint

MEMS/NEMS Devices and Applications, 2nd Edition

Edited by
Ching-Liang Dai, Yao-Chuan Tsai and Zhi-Xuan Dai

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Guest Editors

Ching-Liang Dai

Yao-Chuan Tsai

Zhi-Xuan Dai



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Guest Editors

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About the Editors

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Preface

Recently, nanoelectromechanical system (NEMS) and microelectromechanical system (MEMS) technologies have been employed to develop various microdevices and microstructures. Many sensors and actuators have been manufactured and commercialized using technologies such as pressure sensors, accelerometers, gyroscopes, tactile sensors, thermal sensors, flow sensors, optical sensors, image sensors, microphones, magnetic sensors, chemical sensors, gas sensors, biosensors, microchannels, ink jet heads, optical switches, RF switches, micromirror, motors, relays, resonators, filters, and energy harvesters. NEMS/MEMS devices have been widely applied in various fields.

Ching-Liang Dai, Yao-Chuan Tsai, and Zhi-Xuan Dai

Guest Editors

Editorial for the Special Issue on MEMS/NEMS Devices and Applications, 2nd Edition

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1. Introduction

Microelectromechanical systems (MEMSs) and nanoelectromechanical systems (NEMSs) are revolutionary technologies that merge mechanical and electronic components on microscopic and nanoscopic scales [1–3]. MEMSs operate at the microscale, while NEMSs represent the next evolution, with components functioning at the nanoscale. Together, these technologies form the backbone of numerous modern devices, driving innovation across industries ranging from healthcare to consumer electronics. MEMSs and NEMSs integrate mechanical elements, sensors, actuators, and electronics on a single silicon chip [4–6]. MEMS devices leverage microfabrication techniques derived from the semiconductor industry, including photolithography, etching, and deposition. Components such as micro-scale gears, levers, and membranes operate in conjunction with electronic circuits, enabling complex functionalities. NEMSs, on the other hand, build upon MEMS principles but operate at the nanoscale, utilizing materials like graphene, carbon nanotubes, and nanowires [7–9]. The smaller size of NEMSs offers advantages such as faster response times, reduced energy consumption, and the ability to detect extremely subtle forces or changes in the environment.

As shown in Figure 1, MEMS devices are applied in numerous fields. In the automotive industry, they serve as accelerometers for airbags and gyroscopes for stability control systems [10]. In healthcare, MEMSs enable innovations such as lab-on-a-chip devices for diagnostics, drug delivery systems, and biosensors for glucose monitoring [11,12]. MEMS-based microfluidic chips are also used for deoxyribonucleic acid (DNA) analysis and point-of-care testing. In consumer electronics, MEMS devices are integral to smartphones, tablets, and gaming consoles [13]. Applications include motion sensing for screen orientation, vibration motors, microphones, and optical image stabilization in cameras. In aerospace and defense, MEMS gyroscopes and accelerometers are critical in navigation systems for drones, aircraft, and missiles [14]. They provide precise motion and orientation data for stable flight and accurate targeting. In industrial automation, MEMS sensors monitor force, displacement, and vibration in industrial machinery [15]. These devices are essential for predictive maintenance and improving operational efficiency in manufacturing processes. In environmental monitoring, MEMS devices are used in air and water quality sensors to detect pollutants and measure environmental parameters such as humidity, temperature, and gas concentration [16,17]. These are vital for environmental protection and smart city initiatives. In optical communications, MEMS-based optical switches and mirrors play a key role in telecommunication networks [18–20]. They enable efficient routing of

light signals in fiber optic communications, reducing power consumption and enhancing speed. In energy harvesting, MEMS devices, such as thermoelectric energy harvesters, capture mechanical heat and convert it into electrical energy [21,22]. These are used in internet of things (IoT) devices for self-powered sensors. In inkjet printing, MEMS nozzles in inkjet printers enable precise ejection of ink droplets, ensuring high-resolution printing [23]. This application is also extended to 3D printing technologies. In wearable devices, MEMS accelerometers, gyroscopes, and pressure sensors are core components of wearable fitness trackers, smartwatches, and augmented reality glasses [24]. These devices track motion, monitor vital signs, and enhance user interaction. NEMS devices excel in applications requiring extreme sensitivity and precision. In medicine, NEMS devices are utilized to detect biomolecules, such as proteins, DNA, and disease markers, with exceptional sensitivity. These biosensors enable early diagnosis of diseases like cancer or infectious diseases by identifying trace quantities of specific biomarkers [25]. In scientific research, NEMS resonators are used for studying quantum phenomena. Additionally, NEMS are pivotal in next-generation data storage and computing technologies due to their ultra-low power requirements and high-speed operation. As advancements in fabrication and integration continue, these systems will remain critical to the development of smarter, smaller, and more efficient technologies across diverse fields.

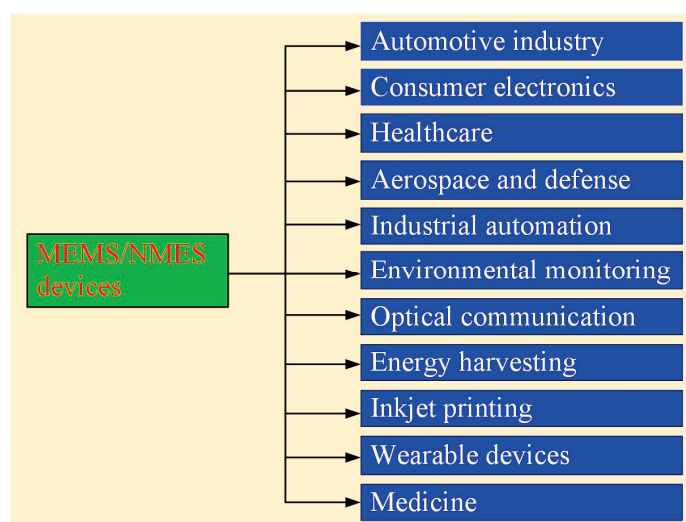


Figure 1. Applications of MEMS/NEMS devices.

This Special Issue has compiled numerous outstanding studies on NEMS/MEMS devices and their applications. These include a wide range of sensors [26–30], actuators [31–34], energy harvesters [35], and antennas [36].

2. MEMS/NEMS Sensors in This Special Issue

This Special Issue includes many studies on MEMS/NEMS sensors. Their key features and highlights are introduced below. Wang et al. [26] present the development and evaluation of a piezoelectric pressure transducer utilizing polyvinylidene fluoride (PVDF) for investigating unsteady phenomena in supersonic shock wave/boundary layer interactions. Unlike traditional sensors such as MEMS-based devices or pressure-sensitive paint, this research introduces a low-cost, flexible, and lightweight solution that enhances both sensitivity and adaptability. The PVDF sensor array, designed specifically for capturing low-frequency pressure fluctuations, features a unique structure fabricated via a double-sided screen-printing technique. The array comprises 110 μm -thick PVDF films with printed silver electrodes, arranged in a 3×6 configuration. The researchers employed

finite element simulations to optimize the design, ensuring minimal signal crosstalk and stable performance up to 300 kHz. Experimental calibration and validation were conducted using a custom acoustic setup and wind tunnel tests at Mach 2, confirming the sensor's capability to measure dynamic pressure changes effectively. Compared to conventional Kulite sensors, the PVDF array demonstrated competitive sensitivity while being significantly more flexible and scalable. Additionally, the modular assembly of the sensing and circuit components addressed challenges like signal attenuation and crosstalk. This innovative approach highlights the potential of PVDF sensors in capturing complex aerodynamic behaviors, providing a promising alternative for advanced research in high-speed flows and low-frequency unsteady phenomena.

Lv et al. [27] study the design and optimization of a MEMS resonant pressure sensor to achieve a wide measurement range and high sensitivity, addressing the challenges in traditional sensor designs. A novel vertical resonator structure is introduced, with the resonant beam aligned perpendicularly to the sensitive diaphragm. This configuration improves diaphragm rigidity without increasing its thickness, enabling a broader range while maintaining sensitivity. To optimize the structural parameters, the study employs a combination of back propagation neural networks and the non-dominated sorting genetic algorithm II. These methods ensure the sensor's performance by resolving modal interference issues and enhancing linearity. The core materials include silicon-based components for the diaphragm and resonator, while the fabrication process integrates micro-scale engineering techniques. The optimized sensor achieves a sensitivity of 4.23 Hz/kPa over a range of 1–10 MPa, with a linear influence factor of 38.07—significantly outperforming conventional designs. This work represents high-performance MEMS pressure sensors, with potential applications in aerospace, marine exploration, and other industries requiring precise pressure measurements.

Zhou et al. [28] present the design of a quad mass gyroscope (QMG) integrating a compliant mechanical amplification structure to significantly improve sensitivity and performance. The research introduces a rhombus-shaped V-spring system for mechanical motion amplification, amplifying the Coriolis force before it is processed electronically, thereby reducing the noise interference and enhancing the signal clarity. The gyroscope is fabricated using silicon-on-insulator (SOI) and gold–silicon eutectic bonding techniques, featuring a compact 4×5 mm structure. The QMG's design incorporates four symmetrically arranged proof masses connected by U-shaped and V-shaped springs, which facilitate mechanical amplification and decouple drive and sense modes. Finite element analyses validated the structural integrity and performance, with the gyroscope achieving a resonant frequency of 11.67 kHz and a mechanical amplification factor of 3.65 times. Testing demonstrated a scale factor nonlinearity of 54.69 ppm over a $\pm 400^\circ/\text{s}$ range and a bias stability of $44.53^\circ/\text{h}$. This work advances gyroscope technology by integrating mechanical amplification at the front end, offering enhanced sensitivity and robustness against external vibrations, making it suitable for high-performance inertial sensing applications.

Xu et al. [29] explore the development of a microscale gas chromatography (μGC) system based on a novel multi-sensing progressive cellular architecture. The research aims to optimize the detection and analysis of volatile organic compounds (VOCs) through a compact, efficient, and highly integrated design. The innovative aspect of the study lies in its use of multithreaded control software for managing the complex interplay of various hardware components in real-time, ensuring enhanced performance and flexibility. The μGC system comprises three micro-gas chromatography cells, each tailored to specific ranges of analytic volatility. These cells incorporate preconcentrators, separation columns, and three complementary detectors, leveraging diverse sensing principles, including capacitive and photoionization detection. The device uses silicon-based microfabrication with

precise heater and sensor integration for temperature and flow control. Fabrication includes advanced photolithography and etching techniques for component miniaturization and integration. The control software, implemented on a Raspberry Pi platform, enables the concurrent operation of pumps, valves, heaters, and sensors via a multithreading architecture. This approach facilitates precise synchronization and efficient resource utilization, achieving minimal timing errors and robust system reliability. Experimentation demonstrates the system's capability to analyze 18 VOCs with high sensitivity and reproducibility, supported by a flexible graphical user interface for real-time monitoring and configuration. This work highlights the potential of μ GC systems in environmental monitoring, industrial applications, and portable chemical analyses, establishing the foundation for future innovations in integrated microscale systems.

Song et al. [30] focus on designing and optimizing a milling force measurement tool system embedded with thin-film strain sensors. The research addresses the challenge of accurately measuring milling forces by employing an innovative double-end-supported thin-film strain sensor. This novel design enhances sensitivity and strain measurement reliability through the integration of a resistance grid beam on the transition layer. The sensors comprise multiple layers: a stainless-steel substrate, a TiN transition layer for improved binding, a Si_3N_4 insulating layer, and a $\text{Ni}_{80}\text{Cr}_{20}$ resistance grid. Advanced microfabrication techniques, including magnetron sputtering, photolithography, ion beam etching, and wet etching, were utilized to create the sensors. The structural optimization process introduced features like H-type and arc-type substrates to minimize stress concentrations and improve mechanical performance. Experimentally, the system demonstrated a resistive sensitivity coefficient increased by 20%, achieving 51.2% compared to conventional designs. The results were validated using various characterization tools, including confocal and atomic force microscopy, revealing high precision in the sensor's microstructure. This research represents a significant advancement in sensor technology, enabling precise real-time force measurement in milling processes, with potential applications in manufacturing and machining industries.

3. MEMS/NEMS Actuators in This Special Issue

This Special Issue also features many studies on MEMS/NEMS actuators. The features and highlights of these actuators are described below. Shuaibu et al. [31] introduce a laterally actuated silicon-to-silicon direct current MEMS switch designed for power switching in high-voltage and harsh environments. The research highlights the innovation of employing a chevron-shaped electrothermal actuator that uses an aluminum heater layer for actuation, combined with a silicon dioxide insulation layer to ensure electrical isolation between the heater and transmission line. This design overcomes the limitations seen in traditional MEMS switches, such as high contact resistance and material degradation. The switch is fabricated using the PiezoMUMPs process [31], involving multi-layered structures with precise alignment of silicon, aluminum, and silicon dioxide. The actuator achieves an in-plane displacement of $2.52\text{ }\mu\text{m}$ to close the contact gap at an actuation voltage of 1.2 V and a current of 205 mA, with a switching speed below 5 ms. Rigorous simulations and testing confirmed the switch's capability to withstand a breakdown voltage of up to 350 V, with low contact resistance measured at $150\text{ }\Omega$. Innovative techniques, such as controlling the etching angle and optimizing the thermal actuator's design, mitigate challenges such as out-of-plane deformation and contact misalignment. Additionally, the integration of narrow silicon beams reduces thermal gradients, enhancing stability and performance. This work significantly advances the field by providing a robust and reliable MEMS switch solution for high-voltage applications.

Wang et al. [32] explore strain-induced frequency splitting in parity-time (PT) symmetric coupled silicon resonators. It innovatively leverages the unique properties of PT symmetry, which involves balancing gain and loss within coupled resonators, to enhance sensitivity to strain near the exceptional point. This study integrates theoretical modeling, finite element simulations, and experimental validations. The resonators, constructed using a SOI process, employ precise feedback circuits to achieve the required negative damping for PT symmetry. Key components include silicon beams of 25 μm thickness, 400 μm substrates, and a gap of 2 μm . The fabrication process involves deep reactive-ion etching and gold layer patterning. Experimental results confirm the system's heightened strain sensitivity, achieving significant frequency splitting under applied strain. This work broadens PT symmetry applications and sets the foundation for advanced sensing technologies.

Chimerad et al. [33] present an innovative bio-inspired soft actuator designed to autonomously remove copper ions from aqueous environments. This miniaturized device integrates fuel-free, self-propelling capabilities utilizing the Marangoni effect, mimicking the propulsion mechanisms of semi-aquatic beetles. The actuator's unique design and functionality represent a significant advancement over traditional stationary hydrogels used for heavy metal remediation, addressing mobility challenges in uneven contamination scenarios. The device's structure consists of a hydrogel fabricated from acrylamide derivatives and cross-linked with polyethylene glycol diacrylate. A gear-shaped mold, created using computer-aided design and 3D printing, facilitates its rotational movement and increases its surface area for absorption. The manufacturing process incorporates ultraviolet polymerization and freeze-drying, yielding a porous material capable of dynamic environmental responses. Through surface tension gradients induced by hydrophilic and hydrophobic surface transitions, the actuator achieves autonomous locomotion, exposing fresh hydrogel surfaces for continuous copper ion absorption. Experimental results reveal its capacity to operate effectively in copper-ion concentrations ranging from 500 to 3000 ppm, with absorption efficiency validated through scanning electron microscope, Fourier-transform infrared spectroscopy, and colorimetric analysis. The actuator demonstrated scalability and adaptability, autonomously covering large water surface areas without external energy sources. This study marks a breakthrough in active, eco-friendly water purification technologies, paving the way for versatile applications in environmental remediation.

Shan et al. [34] focuses on developing and analyzing a vertical comb-drive actuator for MEMS micromirrors, aimed at achieving enhanced stability and precise control during optical scanning. It introduces a self-alignment fabrication technique that ensures a height difference between movable and fixed comb structures, reducing alignment errors and improving reliability. This actuator leverages the quasi-static driving principle to address instability issues such as lateral bending, displacement, and rotational failures of comb fingers. The micromirror's structural components include a SOI substrate and torsion beams, with a 0.8 mm mirror diameter and 30 μm device layer thickness. The height difference between comb structures is achieved through precise dry etching and photolithography processes. The actuator demonstrates stable operation below 60 V, and reaches a static mechanical deflection angle of 2.25° at 60 V, aligning closely with the theoretical models. Critical failure modes were observed at higher voltages, validating the proposed instability models. This research offers a robust foundation for designing reliable MEMS micromirrors with potential applications in optical communications and autonomous systems.

4. MEMS/NEMS Energy Harvesters and Antennas in This Special Issue

This Special Issue features a study on a MEMS energy harvester, with its key features described below. Chen et al. [35] explore the fabrication and performance evaluation of a photovoltaic microgenerator created using the complementary metal oxide semiconductor

(CMOS)-MEMS process, aiming to achieve efficient energy conversion within a compact structure. The research emphasizes innovative modifications, such as a mesh-patterned p–n junction design and the post-fabrication removal of the silicon dioxide layer, which enhance light absorption and energy conversion efficiency. The microgenerator is constructed using silicon-based materials, featuring patterned p-type and n-type doping layers that optimize charge separation and minimize leakage currents. The fabrication process employs the standardized 0.18 μm CMOS process, followed by a buffered oxide etching step to expose the p–n junctions, enabling direct illumination and improved power generation capabilities. Simulations and experimental validation demonstrate the device's effectiveness, achieving an energy-conversion efficiency of 12.5% under an irradiance of 1000 W/m^2 . It records an open-circuit voltage of 0.53 V, a short-circuit current of 233 μA , and a maximum output power of 99 μW within a compact area of 0.79 mm^2 . This study provides a significant step forward in the development of scalable, efficient photovoltaic microgenerators, suitable for powering IoT devices, sensors, and other compact electronics.

Li et al. [36] investigate a magnetoelectric thin-film antenna, focusing on optimizing its structure to enhance bandwidth and radiation performance. This study addresses the challenges of narrow operating bandwidths and low antenna gains inherent in traditional magnetoelectric antennas. By analyzing the interplay between structural dimensions and electromagnetic properties, the research introduces a ring-shaped design with variable inner radii, leveraging stress concentration effects to improve performance. The core of the antenna comprises piezoelectric AlN and magnetostrictive FeGaB layers, fabricated using a thin-film bulk-acoustic resonator approach. Through COMSOL [36] multiphysics simulations, the team analyzed resonance behavior, stress distribution, and radiation patterns, revealing that an inner radius of 50 μm maximized bandwidth by 104% and enhanced radiation efficiency. The study also validated its findings against experimental data from previous designs, ensuring reliability. This research introduces a scalable optimization model that correlates stress distribution with radiation capability, advancing magnetoelectric antenna applications in compact communication systems operating at lower frequencies.

Conflicts of Interest: The authors declare no conflict of interest.

References

1. Judy, J.W. Microelectromechanical systems (MEMS): Fabrication, design and applications. *Smart Mater. Struct.* **2001**, *10*, 1115. [CrossRef]
2. Ekinici, K.L. Electromechanical transducers at the nanoscale: Actuation and sensing of motion in nanoelectromechanical systems (NEMS). *Small* **2005**, *1*, 786–797. [CrossRef] [PubMed]
3. Huang, Y.; Vasan, A.S.S.; Doraiswami, R.; Osterman, M.; Pecht, M. MEMS Reliability Review. *IEEE Trans. Device Mater. Reliab.* **2012**, *12*, 482–493. [CrossRef]
4. Blencowe, M.P. Nanoelectromechanical systems. *Contemp. Phys.* **2005**, *46*, 249–264. [CrossRef]
5. Qu, H. CMOS MEMS Fabrication Technologies and Devices. *Micromachines* **2016**, *7*, 14. [CrossRef] [PubMed]
6. Ko, W.H. Trends and frontiers of MEMS. *Sens. Actuators A* **2007**, *136*, 62–67. [CrossRef]
7. Torkashvand, Z.; Shayeganfar, F.; Ramazani, A. Nanomaterials based micro/nanoelectromechanical system (MEMS and NEMS) devices. *Micromachines* **2024**, *15*, 175. [CrossRef]
8. Jiang, S.L.; Shi, T.L.; Zhan, X.B.; Xi, S.; Long, H.; Gong, B.; Li, J.J.; Cheng, S.Y.; Huang, Y.Y.; Tang, Z.R. Scalable fabrication of carbon-based MEMS/NEMS and their applications: A review. *J. Micromech. Microeng.* **2015**, *25*, 113001. [CrossRef]
9. Zang, X.N.; Zhou, Q.; Chang, J.Y.; Liu, Y.M.; Lin, L.W. Graphene and carbon nanotube (CNT) in MEMS/NEMS applications. *Microelectron. Eng.* **2015**, *132*, 192–206. [CrossRef]
10. Acar, C.; Schofield, A.R.; Trusov, A.A.; Costlow, L.E.; Shkel, A.M. Environmentally robust MEMS vibratory gyroscopes for automotive applications. *IEEE Sens. J.* **2009**, *9*, 1895–1906. [CrossRef]
11. Shikida, M.; Hasegawa, Y.; Farisi, M.S.A.; Matsushima, M.; Kawabe, T. Advancements in MEMS technology for medical applications: Microneedles and miniaturized sensors. *Jpn. J. Appl. Phys.* **2022**, *61*, SA0803. [CrossRef]

12. Yang, M.Z.; Dai, C.L.; Hung, C.B. Fabrication of a glucose sensor with oscillator circuit using CMOS-MEMS technique. *Microelectron. Eng.* **2012**, *97*, 353–356. [CrossRef]
13. Capuano, V.; Xu, L.; Estrada Benavides, J. Smartphone MEMS accelerometer and gyroscope measurement errors: Laboratory testing and analysis of the effects on positioning performance. *Sensors* **2023**, *23*, 7609. [CrossRef] [PubMed]
14. Yuan, W.Z. Development and application of high-end aerospace MEMS. *Front. Mech. Eng.* **2017**, *12*, 567–573. [CrossRef]
15. Liang, Q.K.; Zhang, D.; Coppola, G.; Wang, Y.N.; Wei, S.; Ge, Y.J. Multi-dimensional MEMS/Micro sensor for force and moment sensing: A review. *IEEE Sens. J.* **2014**, *14*, 2643–2657. [CrossRef]
16. Yang, M.Z.; Dai, C.L. Ethanol microsensors with a readout circuit manufactured using the CMOS-MEMS technique. *Sensors* **2015**, *15*, 1623–1634. [CrossRef]
17. Shen, W.C.; Shih, P.J.; Tsai, Y.C.; Hsu, C.C.; Dai, C.L. Low-concentration ammonia gas sensors manufactured using the CMOS-MEMS technique. *Micromachines* **2020**, *11*, 92. [CrossRef]
18. Kao, P.H.; Dai, C.L.; Hsu, C.C.; Wu, C.C. Manufacture of micromirror arrays using a CMOS-MEMS technique. *Sensors* **2009**, *9*, 6219–6231. [CrossRef] [PubMed]
19. Ma, J. Advanced MEMS-based technologies and displays. *Displays* **2015**, *37*, 2–7. [CrossRef]
20. Cheng, Y.C.; Dai, C.L.; Lee, C.Y.; Chen, P.H.; Chang, P.Z. A circular micromirror array fabricated by a maskless post-CMOS process. *Microsyst. Technol.* **2005**, *11*, 444–451. [CrossRef]
21. Lee, W.L.; Shih, P.J.; Hsu, C.C.; Dai, C.L. Fabrication and characterization of flexible thermoelectric generators using micromachining and electroplating techniques. *Micromachines* **2019**, *10*, 660. [CrossRef]
22. Chen, Y.W.; Wu, C.C.; Hsu, C.C.; Dai, C.L. Fabrication and testing of thermoelectric CMOS-MEMS microgenerators with CNCs Film. *Appl. Sci.* **2018**, *8*, 1047. [CrossRef]
23. Kim, Y.J.; Lee, S.M.; Kim, S.; Hwang, J.; Kim, Y.J. A hybrid electrohydrodynamic drop-on-demand printing system using a piezoelectric MEMS nozzle. *J. Micromech. Microeng.* **2012**, *22*, 045013. [CrossRef]
24. Sigcha, L.; Pavón, I.; Arezes, P.; Costa, N.; De Arcas, G.; López, J.M. Occupational risk prevention through smartwatches: Precision and uncertainty effects of the built-in accelerometer. *Sensors* **2018**, *18*, 3805. [CrossRef] [PubMed]
25. Elseddawy, A.M.; Phillips, A.H.; Bayoumi, A.S. Carbon nanotube-based nanoelectromechanical resonator as mass biosensor. *Chin. Phys. B* **2020**, *29*, 078501. [CrossRef]
26. Wang, B.; Corsi, C.; Weiland, T.; Wang, Z.; Grund, T.; Pohl, O.; Bienia, J.M.; Weiss, J.; Ngo, H.D. Screen-printed PVDF piezoelectric pressure transducer for unsteadiness study of oblique shock wave boundary layer interaction. *Micromachines* **2024**, *15*, 1423. [CrossRef] [PubMed]
27. Lv, M.; Li, P.; Miao, J.; Qiao, Q.; Liang, R.; Li, G.; Zhuang, X. Design and optimization of MEMS resonant pressure sensors with wide range and high sensitivity based on BP and NSGA-II. *Micromachines* **2024**, *15*, 509. [CrossRef] [PubMed]
28. Zhou, J.; Xu, D.; Li, X.; Chen, F. Design of a micro-electro mechanical system quad mass gyroscope with compliant mechanical amplification. *Micromachines* **2024**, *15*, 124. [CrossRef] [PubMed]
29. Xu, Q.; Zhao, X.; Qin, Y.; Gianchandani, Y.B. Control software design for a multisensing multicellular microscale gas chromatography system. *Micromachines* **2024**, *15*, 95. [CrossRef]
30. Song, X.; Wu, W.; Zhao, Y.; Cheng, Y.; Liu, L. Structural design and optimization of the milling force measurement tool system embedded with thin-film strain sensors. *Micromachines* **2023**, *14*, 2133. [CrossRef] [PubMed]
31. Shuaibu, A.H.; Rabih, A.A.S.; Blaquièrre, Y.; Nabki, F. Laterally actuated Si-to-Si DC MEMS switch for power switching applications. *Micromachines* **2024**, *15*, 1295. [CrossRef] [PubMed]
32. Wang, L.; Zhang, S.; Yuan, Q. Strain-induced frequency splitting in PT symmetric coupled silicon resonators. *Micromachines* **2024**, *15*, 1278. [CrossRef]
33. Chimerad, M.; Borjian, P.; Pathak, P.; Fasano, J.; Cho, H.J. A miniaturized, fuel-free, self-propelled, bio-Inspired soft actuator for copper ion removal. *Micromachines* **2024**, *15*, 1208. [CrossRef] [PubMed]
34. Shan, Y.; Qian, L.; Wang, J.; Wang, K.; Zhou, P.; Li, W.; Shen, W. Driving principle and stability analysis of vertical comb-drive actuator for scanning micromirrors. *Micromachines* **2024**, *15*, 226. [CrossRef]
35. Chen, C.Y.; Dai, Z.-X. Fabrication and characterization of photovoltaic microgenerators using the complementary metal oxide semiconductor process. *Micromachines* **2023**, *14*, 2038. [CrossRef]
36. Li, X.; Zhao, P.; Wang, G.; Li, N.; Zhang, Y. Structural optimization design of magnetoelectric thin-film antenna for bandwidth and radiation enhancement. *Micromachines* **2024**, *15*, 810. [CrossRef]

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Article

Screen-Printed PVDF Piezoelectric Pressure Transducer for Unsteadiness Study of Oblique Shock Wave Boundary Layer Interaction

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Abstract: Shock wave boundary/layer interactions (SWBLIs) are critical in high-speed aerodynamic flows, particularly within supersonic regimes, where unsteady dynamics can induce structural fatigue and degrade vehicle performance. Conventional measurement techniques, such as pressure-sensitive paint (PSP), face limitations in frequency response, calibration complexity, and intrusive instrumentation. Similarly, MEMS-based sensors, like Kulite[®] sensors, present challenges in terms of intrusiveness, cost, and integration complexity. This study presents a flexible, lightweight polyvinylidene fluoride (PVDF) piezoelectric sensor array designed for high-resolution wall-pressure measurements in SWBLI research. The primary objective is to optimize low-frequency pressure fluctuation detection, addressing SWBLI's need for accurate, real-time measurements of low-frequency unsteadiness. Fabricated using a double-sided screen-printing technique, this sensor array is low-cost, flexible, and provides stable, high-sensitivity data. Finite Element Method (FEM) simulations indicate that the sensor structure also has potential for high-frequency responses, behaving as a high-pass filter with minimal signal attenuation up to 300 kHz, although the current study's experimental testing is focused on low-frequency calibration and validation. A custom low-frequency sound pressure setup was used to calibrate the PVDF sensor array, ensuring uniform pressure distribution across sensor elements. Wind tunnel tests at Mach 2 verified the PVDF sensor's ability to capture pressure fluctuations and unsteady behaviors consistent with those recorded by Kulite sensors. The findings suggest that PVDF sensors are promising alternatives for capturing low-frequency disturbances and intricate flow structures in advanced aerodynamic research, with high-frequency performance to be further explored in future work.

Keywords: supersonic; shock wave/boundary layer interaction (SWBLI); PVDF; piezoelectric pressure transducer; FEM simulation; dynamic response in frequency domain; crosstalk effect; wall pressure; unsteadiness

1. Introduction

A shock wave/boundary layer interaction (SWBLI) occurs in high-speed aerodynamic flows, particularly in supersonic and hypersonic regimes. This interaction is influenced by factors such as shock wave strength, the angle of incidence on the boundary layer, Reynolds number, and boundary layer characteristics. These factors lead to abrupt changes in flow properties, resulting in shock wave formation and regions with pressure and temperature

gradients. Near solid surfaces, a thin boundary layer forms, characterized by gradual changes in flow properties from the surface to freestream conditions [1–3]. When the shock wave impinges on the boundary layer, interactions occur, resulting in significant flow modifications, as shown in Figure 1. These modifications can cause the boundary layer to either separate from the surface or undergo substantial thickening or thinning [2,3]. Changes in the boundary layer induce unsteady behaviors, such as low-frequency oscillations, flow separations, turbulence generation, and pressure fluctuations [2–5]. These unsteady phenomena can lead to vibrations, buffeting, and even structural fatigue in aircraft and other high-speed vehicles [6,7]. Thus, understanding and predicting the unsteadiness caused by shock wave/boundary layer interactions is essential for designing aerodynamically stable and structurally robust systems.

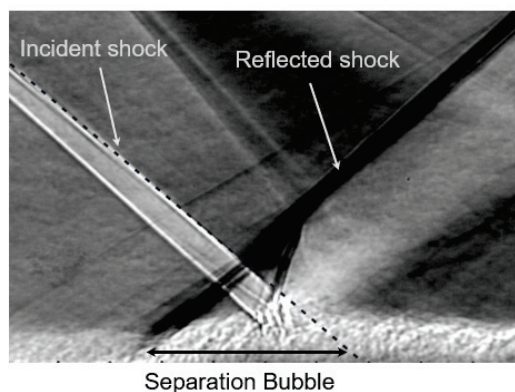


Figure 1. Schlieren image of the flow field of the shock wave/boundary layer interaction [3].

For many years, researchers have employed both experimental measurements and computational flow simulations to observe and analyze interactions, visualize flow fields, and obtain qualitative information. Optical methods like schlieren imaging capture refractive index changes due to density gradients but are limited by calibration and small density differences [8]. Particle Image Velocimetry (PIV) visualizes flow fields by tracking tracer particles, although its accuracy is affected by factors such as particle size, lighting, and background noise [9–12].

Wall-pressure measurements are an essential technique in both ground and flight tests for understanding SWBLIs. These measurements help identify high- and low-pressure regions, such as separation bubbles, validate simulations, and capture the frequency characteristics of unsteadiness [13–17]. Wall-pressure measurements directly assess pressure distribution, offering a more accurate depiction of unsteadiness associated with SWBLIs. However, the use of pressure-sensitive paint (PSP) or Kulite® sensors (Kulite Semiconductor Products, Inc., One Willow Tree Road, Leonia, NJ, USA) is limited by factors like installation, structure, and cost, making them unsuitable for large-area sensing and integration [13,18].

Polyvinylidene Fluoride (PVDF), a thermoplastic fluoropolymer, has been valued for decades due to its flexibility and stretchability, enabling PVDF sensors to adapt to various shapes and surfaces [19–24]. These sensors are highly sensitive and structurally simple, making them effective for detecting mechanical stimuli such as pressure, strain, and vibration [25–27].

In the study of SWBLIs, low-frequency unsteadiness often dominates critical flow behaviors, including flow separation and recirculation bubble formation, making accurate low-frequency pressure measurements essential. To address this need, this study develops a flexible polyvinylidene fluoride (PVDF) piezoelectric sensor array optimized specifically for low-frequency pressure fluctuation detection in SWBLI environments. Our design and processing methods differ significantly from traditional MEMS sensors and existing approaches [28,29]. The PVDF sensor array, fabricated using a double-sided screen-printing technique, provides a low-cost, lightweight, and flexible solution that can adapt to varying surface shapes while delivering high-sensitivity measurements.

Finite Element Method (FEM) simulations indicate that the PVDF sensor array has potential for a stable high-frequency response up to 300 kHz, demonstrating high-pass filter characteristics with minimal signal attenuation at higher frequencies. Although the current experimental testing focuses on low-frequency calibration and validation to meet the primary needs of SWBLI research, the sensor's design and simulation results provide a foundation for exploring its high-frequency response in future studies. This approach allows us to address immediate research needs in low-frequency unsteadiness while setting a trajectory for expanded applications.

In this study, we present the design, fabrication, and performance evaluation of the PVDF sensor array in a supersonic flow environment (Mach 2). The calibration and validation tests were performed to assess the sensor's low-frequency detection capabilities and benchmark its performance against the widely used Kulite® MEMS sensors. The results highlight the PVDF sensor array's potential as an effective alternative for unsteady wall-pressure measurements in SWBLI studies, with future work planned to extend testing across a broader frequency range.

2. Methodology: Processing and Characterization

2.1. Material Properties and Design

The flexible sensor array was fabricated using screen printing, as shown in Figure 2. A 110 μm -thick PVDF foil (PolyK®) (PolyK Technologies, LLC, 2124 Old Gatesburg Road, State College, PA, USA) served as the piezoelectric sensing material (Table 1). The electrode nodes, each 4 mm in diameter, were printed onto the PVDF foil in a 3×6 configuration. Nodes were spaced 6 mm apart within each row and 10 mm apart in each column (Figure 2).

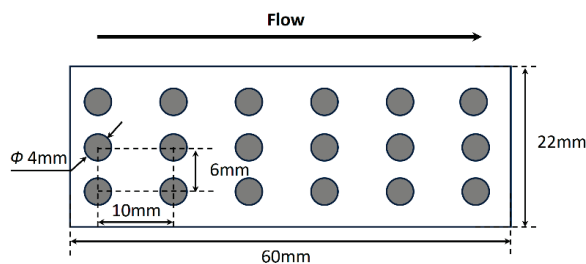


Figure 2. Schematic of the design of sensor elements.

Table 1. Characteristic parameters of PVDF piezoelectric foil [30,31].

Item	Value	Unit
Piezoelectric Strain Constand d_{33}	28	pC/N
Piezoelectric Voltage Constant g_{33}	0.3	V·m/N
Relative Dielectric Constant ϵ_r	13.5	
Maximum Operating Voltage	>100	V/ μm
Tensile Strength σ_{MD}	$\sim 0.4 \times 10^9$	N/m ²
Young's Modulus γ_{MD}	$\sim 2.8 \times 10^9$	N/m ²
Poisson's Ratio	0.3	
Density	1.78×10^3	kg/m ³
Frequency Range	$10^{-3} \sim 10^{10}$	Hz
Pressure Range	$0 \sim 10^9$	Pa
Operating Temperature Range	$-40 \sim 80$	°C
Thickness	110	μm

2.2. FEM Simulation in Frequency Domain

The FEM simulation of the PVDF piezoelectric sensor elements was carried out by using COMSOL® Multiphysics 6.2. The model consists of a PVDF piezoelectric film (110 μm) and silver ink electrodes (10 μm) screen-printed in mirrored positions on its top and bottom surfaces. The silver electrodes on the top surface of the PVDF film are

defined as the anode, while the other on the bottom is defined as the cathode. Due to the limitations of computational capacities and simulation time, the model was simplified, and only two adjacent sensor element pairs (with a center-to-center spacing of 6 mm) were simulated. A 1 kPa pressure load was applied vertically on the left silver anode. Figure 3a,b shows the deformation/displacement and potential distributions of the model induced by a 1 kHz pressure load, by defining a Profile Line on the surface of the silver anodes (as shown in Figure 4), Figure 5 illustrates the impact of distance ΔW between adjacent sensor elements on the output signal crosstalk under identical boundary conditions. Furthermore, Figure 6a,b represents the deformation/displacement and the electric potential of the silver anodes (including the adjacent electrode element) in the model at varying frequencies.

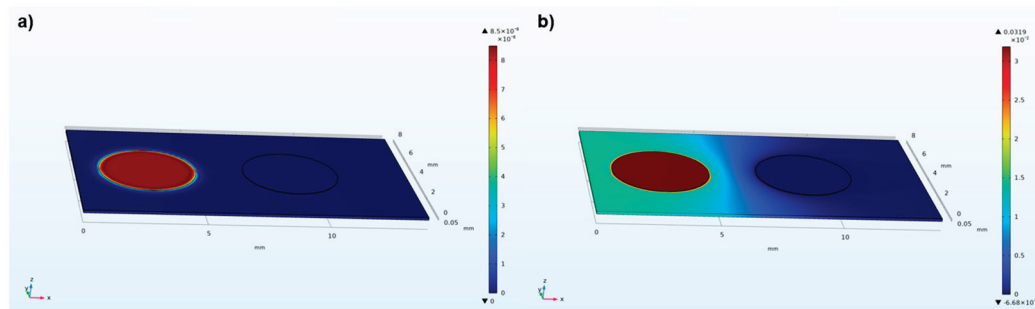


Figure 3. (a) Displacement (mm) of PVDF film caused by 1 kPa pressure vertically to the left sensor electrode at 1 kHz, and (b) the electric potential caused by the piezoelectric effect of the PVDF film.

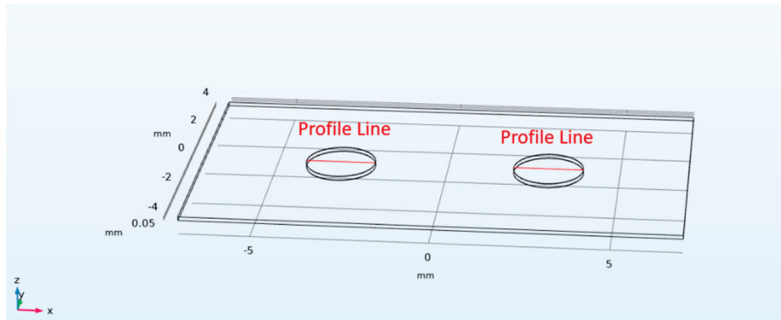


Figure 4. Defined profile lines on the top of the Ag electrodes printed on PVDF film.

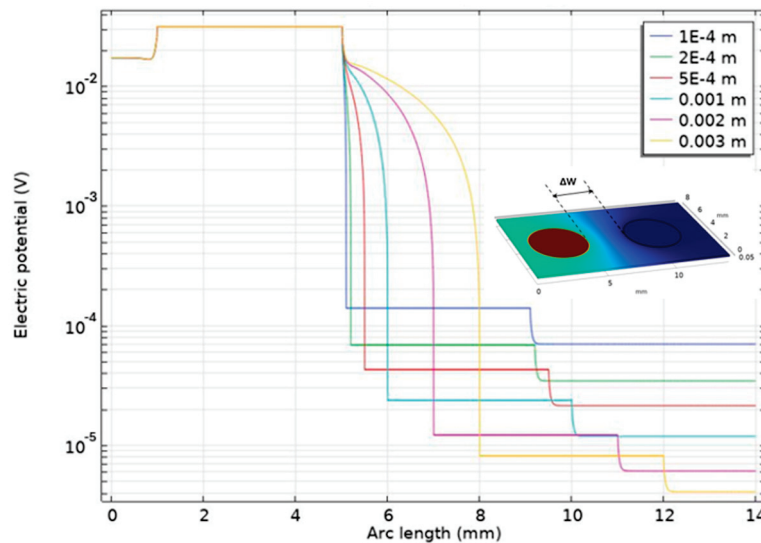


Figure 5. Simulation of signal crosstalk between two adjacent sensor elements at varying distances ΔW .

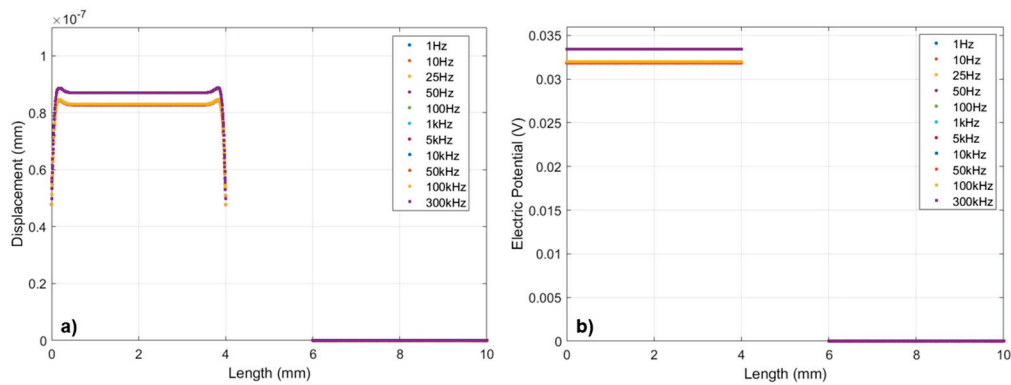


Figure 6. (a) The displacement (mm) on the profile lines of FEM simulation in frequency domain analysis, and (b) the electric potential caused by the displacement in frequency domain analysis.

Figure 5 shows the impact of sensor distance ΔW , on the output signal crosstalk under a defined load of 1 kPa and the frequency of 1 kHz at the sensor element on the left side. The output potential of the loaded sensor element on the left is 0.031 V. As the sensor distance increases, the crosstalk effect on the unloaded sensor element on the right gradually diminishes. With a 0.1 mm distance, the output potential of the right sensor element is $1.5 \cdot 10^{-4}$ V, which is 0.48% of the output potential of the left sensor element. At 2 mm of distance (the design dimension mentioned above), as indicated by the purple line in Figure 5, the output potential of the right sensor element is $1.3 \cdot 10^{-5}$ V, approximately 0.042% of the output potential of the left sensor element. Therefore, at the sensor design dimension used in this work, signal crosstalk can be considered negligible.

Figure 6 shows the curves of deformation/displacement and the potential of the silver anodes as a function of load pressure frequency in the finite element frequency domain analysis. Due to our target frequency range up to 200 kHz, the simulation was defined in the frequency range up to 300 kHz.

In Figure 6a, the displacement profile illustrates deformations of the PVDF film under varying frequencies of applied pressure. As frequency increases, the displacement becomes more complex, indicating that the dynamic mechanical response of the film is frequency-dependent. Figure 6b shows the electric potential distribution caused by the piezoelectric effect. As expected, an increase in frequency alters the electric potential, as the strain-induced charge generation by the piezoelectric effect changes with deformation/displacement.

Figure 7 shows the relationship between frequency and both displacement and voltage. The curves indicate that both displacement and electric potential evolved with increasing frequency, and there appears to be a threshold where the responses increased obviously. In the frequency range up to 1 kHz, the mechanical deformation/displacement of the model remained constant at $8.27 \cdot 10^{-8}$ mm, with a piezoelectric response voltage of 0.031 V. Between 1 kHz and 50 kHz, both displacement and potential response show a slight increase with frequency. From 50 kHz to 300 kHz, the displacement and potential responses exhibit a linear upward trend, reaching $8.69 \cdot 10^{-8}$ mm (0.033 V) at 300 kHz, about +5% compared to the response in the low frequency range. This behavior suggests that the PVDF sensor has resonant frequency in the test range [32,33].

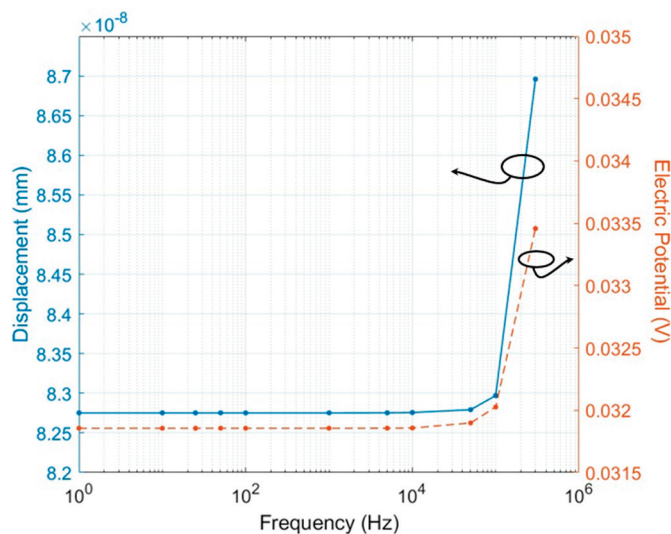


Figure 7. Spectra representation of displacement (mm) and electric potential (V) in frequency domain up to 300 kHz.

The analysis of Figure 7 supports the hypothesis that, at higher frequencies, the deformation/displacement of the PVDF film increases in the frequency domain. This finding is due to the dynamic characteristics of the PVDF material, where its mechanical and piezoelectric properties interact with the frequency of the applied pressure load. As frequency increases, the inertia of the film can lead to more pronounced oscillatory behavior, increasing the deformation until a certain resonant frequency is reached. Furthermore, Figure 6 suggests that the electrical response of the PVDF sensor can behave similarly to a high-pass filter, where, at higher frequencies, the parasitic capacitance and resistance of the sensor become more significant. As frequency increases, the impedance of the capacitive elements decreases, allowing more current (or voltage in this case) to pass through, effectively filtering out low-frequency signals. In summary, the PVDF piezoelectric film's displacement increases with frequency due to its resonant mechanical behavior. Additionally, at high frequencies, the parasitic capacitance and resistance form a high-pass filter, with the cutoff frequency dependent on the material's intrinsic properties and the sensor's electrical design. Notably, the results in Figure 6a,b also indicate that, for the metal electrodes, which have a radius of 2 mm, printed on a 110 μm thick PVDF film, with an edge-to-edge distance of 2 mm between adjacent electrodes, their resulting sensor crosstalk effect in frequency range up to 300 kHz is negligible. These results highlight the sensor's ability to capture a wide range of unsteady flow phenomena, from low-frequency pressure oscillations to high-frequency turbulence-induced fluctuations.

2.3. Processing and Its Electrical Circuit

Although PVDF sensors offer benefits such as lightweight, flexibility, good ductility, and ease of processing, the high density of sensor elements introduces several challenges, including increased signal line area, crosstalk between signal lines, and potential signal attenuation due to the signal line length. These factors limit the overall sensitivity and stability of the PVDF sensor elements. To address this issue, we propose an innovative assembly method where the sensing and circuit parts are processed separately and then assembled, as shown in Figure 8. For the sensing part, the sensor cathode nodes were printed on the top side of the PVDF film using silver conductive ink Bectron[®] CP 6667 (ELANTAS Europe, Grossmannstr. 105, Hamburg, Germany). After curing at 50 $^{\circ}\text{C}$ in an oven for 30 min, followed by 1 h of room temperature curing and cooling, the sensor anode nodes were printed on the backside of the PVDF film. The printed sensor anodes also underwent a 50 $^{\circ}\text{C}$ curing process, followed by room temperature curing and cooling, completing the fabrication of the sensing layer as shown in Figure 8a.

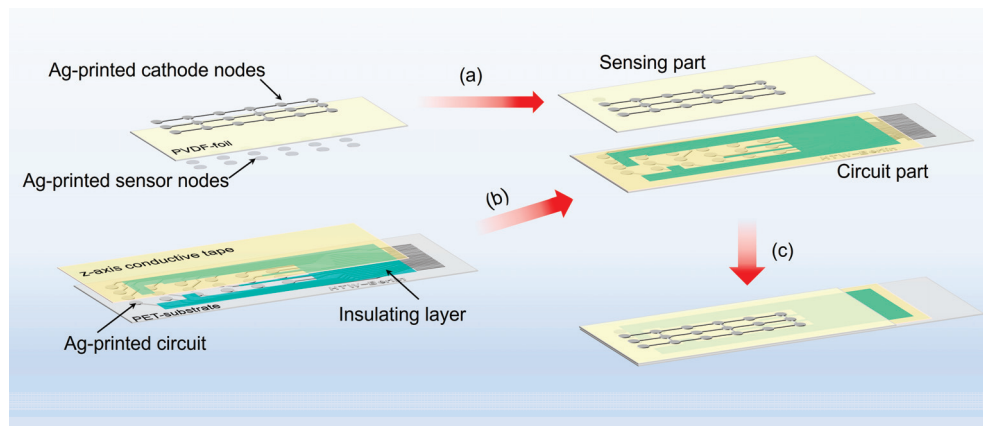


Figure 8. Illustration of 3D-integrated thin-film flexible PVDF sensor array: (a–c) show the process steps.

For the circuit part, the signal circuit was printed using silver conductive ink on a 110 μm -thick polyethylene terephthalate (PET) film, which is non-piezoelectric and serves as an insulating substrate, followed by the same thermal curing process. To protect signal transmission and isolate the conductive circuits from the sensing area, insulating ink DP 8442 (the green layer in Figure 8b) was printed over the circuits after the previous layer was cured. The curing process of the insulating ink requires 30 min of UV exposure at 5 mJ/cm^2 . Subsequently, 3MTM 9703 z-axis conductive tape (St. Paul, MN, USA) was applied on top of the circuits and their insulating layer to ensure a reliable electrical connection with the sensing part, as shown in Figure 8b. Finally, the sensing and circuit parts were bonded together to create a 3D-integrated thin-film flexible PVDF sensor array, as shown in Figure 8c.

Figure 9 shows an image of the processed sensor array with 4 mm diameter sensor elements. The cathode of the sensor array was connected to the GND pad of the electric circuit on the PET substrate, ensuring that all cathode nodes shared the same electrical potential of 0 V.

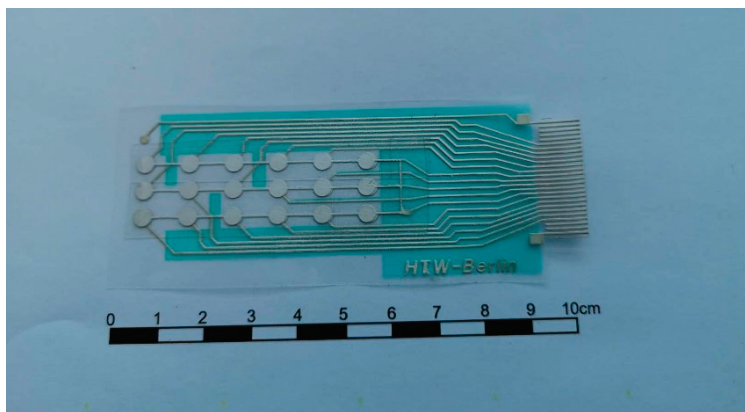


Figure 9. Images of PVDF sensor array with 4 mm sensor elements.

The charge amplifier was constructed using a TLC-272 from Texas Instruments, Dallas, TX, USA, with a feedback resistance $R_f = 100 \text{ M}\Omega$ and a feedback capacitance $C_f = 100 \text{ pF}$ connected to the inverting input terminal. An input resistance $R_{in} = 56 \text{ }\Omega$, providing overcurrent protection, was placed between the piezoelectric sensor and the inverting input terminal, as depicted in Figure 10a. This resistance also served as electrostatic discharge (ESD) protection, as indicated in the TLC-272 datasheet [34].

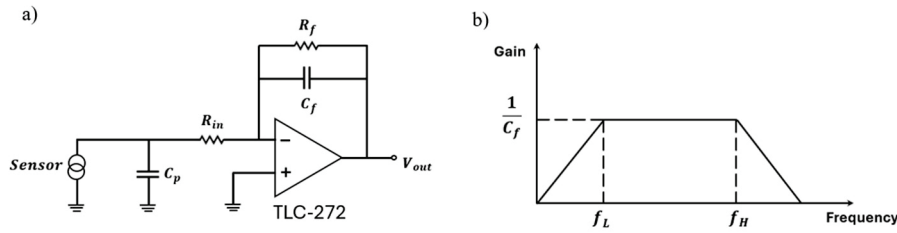


Figure 10. (a) The charge amplifier circuit for the PVDF piezoelectric sensor array, and (b) the band-pass behavior of the charge amplifier circuit [31,34].

The charge amplifier circuit functions as a band-pass filter, with a low-frequency cutoff at

$$f_L = \frac{1}{2\pi R_f C_f}, \quad (1)$$

and a high-frequency cutoff at

$$f_H = \frac{1}{2\pi R_{in} C_p}. \quad (2)$$

Here, C_p includes the sensor's internal capacitance, cable capacitance, and any other parasitic capacitances in parallel with the sensor.

In this setup, C_p is approximately 10 pF, resulting in $f_L = 15.9$ Hz and $f_H = 2.84$ GHz, as presented in Figure 10b. The data acquisition (DAQ) system NI-USB 6353 from National Instruments™, Austin, TX, USA was used to obtain power spectral densities (PSDs) up to 100 kHz at a sample rate of $f_s = 200$ kHz with $N = 200,000$ samples collected.

3. Calibration

Ensuring the reliable performance of PVDF piezoelectric sensors, especially at low pressures, requires precise calibration. The acoustic pressure calibration setup is essential for calibrating PVDF thin-film piezoelectric sensor arrays under low-pressure conditions. The key feature of this setup is that it ensures the PVDF sensor array receives a uniform and consistent pressure distribution at the specified sound frequency within the sound pressure detection cavity. The acoustic pressure test setup generated controlled acoustic pressure waves, which were applied to the PVDF sensor array to calibrate and evaluate its performance in comparison with a Kulite sensor under identical conditions, as shown in Figure 11.

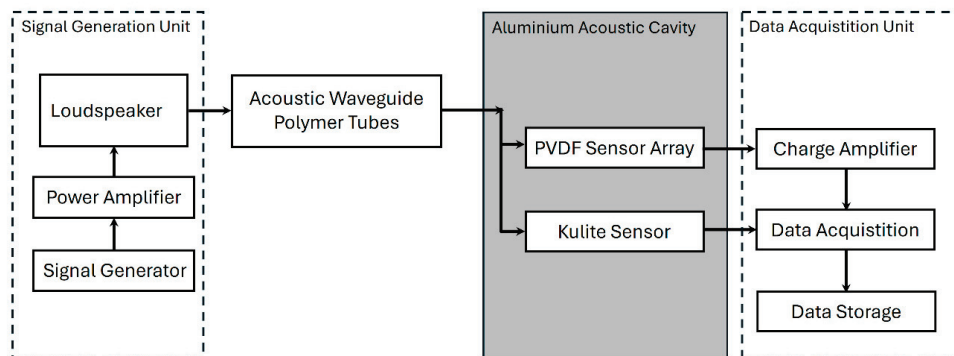


Figure 11. Block diagram of calibration setup.

In the signal generation unit, a 25 Hz acoustic signal was amplified and transmitted through a loudspeaker, then directed via five plastic acoustic wave conduction tubes positioned on its top cover into an aluminum sound pressure detection cavity, as shown in Figure 11. The plastic acoustic wave conduction tubes have a diameter of 2 cm and a length of 50 cm. Of the five tubes (colored blue in Figure 12), one was directly connected

to a central through-hole at the top of the aluminum acoustic cavity, while the other four were connected to equidistant through-holes on the cavity's sides. These through-holes were located 2 cm from the top surface of the acoustic cavity. The PVDF sensor array was secured at the bottom of the acoustic cavity and sealed with O-rings to ensure airtightness and prevent pressure leakage. A Kulite sensor was also mounted on the side of the cavity, as shown in Figure 12.

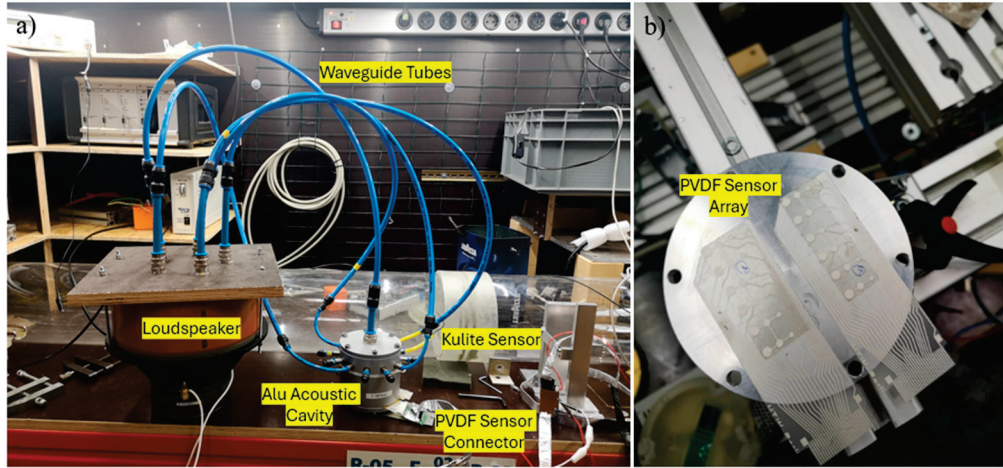


Figure 12. Setup of the sound pressure calibration test equipment (a) and the PVDF sensor array test samples (b).

At room temperature, with the speed of sound in air $c = 343$ m/s, the wavelength λ of the sound in the tubes is calculated as

$$\lambda = \frac{c}{f} = \frac{343 \text{ m/s}}{25 \text{ Hz}} = 13.72 \text{ m}. \quad (3)$$

According to [35], the resonance frequency f_n for the propagation of mechanical waves in a closed-ended tube is calculated as

$$f_n = \frac{n \cdot c}{2L}, \quad (4)$$

where n is the resonance mode, c is the speed of sound in air, and $L = 0.5$ m is the length of the tube.

For the first mode $n = 1$, the cutoff frequency f_{cutoff} is calculated as follows:

$$f_{cutoff} = \frac{1 \cdot 343 \text{ m/s}}{2 \cdot 0.5 \text{ m}} = 343 \text{ Hz} \quad (5)$$

This cutoff frequency is much higher than the driven frequency of 25 Hz, indicating that the sound propagates in the fundamental mode without any cutoff or higher-order transmission modes. Therefore, the acoustic pressure distributions and velocity fields in the fundamental mode are uniform across the cross-section. Over the length of the tubes, the sound wave may undergo phase change, but the long wavelength ensures that this effect is minimal. The phase shift $\Delta\phi$ over the length L is given by the following:

$$\Delta\phi = \frac{2\pi L}{\lambda} = \frac{2\pi \cdot 0.5 \text{ m}}{13.72 \text{ m}} \approx 0.229 \text{ radians}. \quad (6)$$

This mathematical analysis confirms that the sound wave will primarily propagate in the fundamental mode with negligible influence from the tube's geometry at the specified frequency and dimensions.

The physical bending of the plastic tubes and the mismatch in acoustic signal propagation (e.g., from a narrow waveguide to a wider one) could result in an uneven final sound pressure distribution. To address this issue, we utilized FEM simulations using COMSOL® Multiphysics, as shown in Figure 13. To reduce the number of mesh elements and the computational load, we implemented symmetrical modeling (Figure 13a) and conducted sound pressure simulations on only one half of the mirror-symmetrical structure (Figure 13b).

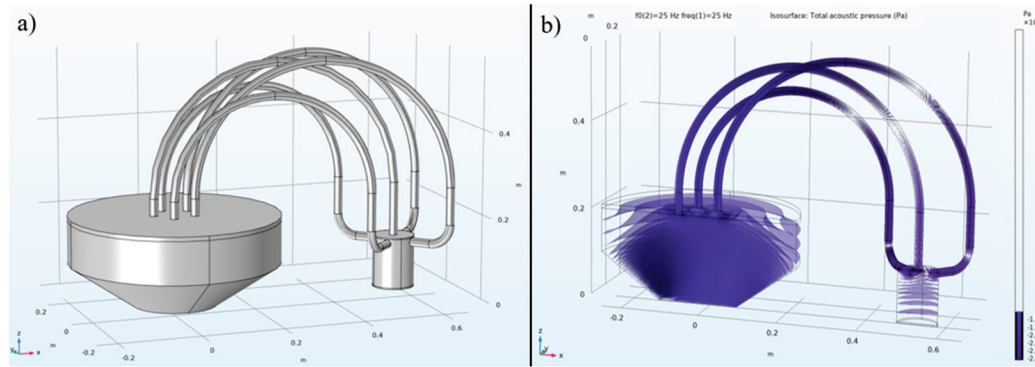


Figure 13. Geometric modeling of the sound pressure calibration test setup (a), and the pressure contour distribution (b).

To further investigate the distribution of the acoustic pressure field on the bottom surface of the aluminum acoustic cavity, we defined 10 cm profile lines on the cavity's bottom surface (as shown by the red lines in Figure 14a) and conducted simulations with acoustic signals at various frequencies. The results showing the total sound pressure level (dB) are presented in Figure 14b). The results indicate that, within the low-frequency range (≤ 500 Hz), the pressure field distribution on the bottom surface of the acoustic cavity is uniform. This result implies that each sensor element in the PVDF sensor array receives consistent sound pressure during calibration. However, as the frequency of the acoustic signal increases, the sound pressure is affected by transmission modes and mechanical bending of the tubes, leading to a non-uniform pressure distribution across the PVDF sensor array, as shown in the total sound pressure level (dB) of the 10 kHz sound signal in Figure 14b. In our 25 Hz low-frequency sound pressure calibration platform, the sound pressure measured by the PVDF sensors at each sensor element remained consistent.

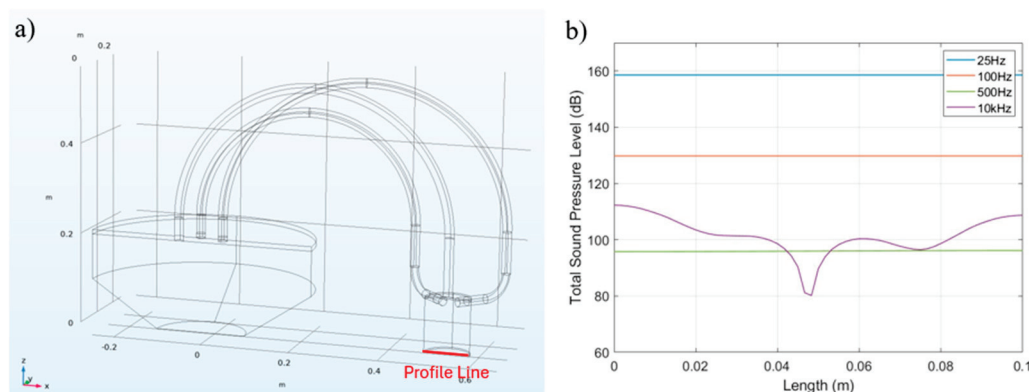


Figure 14. Schematic presentation of the defined profile line in acoustic simulation (a), and (b) the total sound pressure level (dB) on the profile line at frequency sweep.

During the calibration characterization, we utilized both homogeneous and heterogeneous sensor arrays. As shown in Figure 15a, the heterogeneous sensor array consisted of three groups of sensor elements with different diameters arranged along the y-axis into

three columns labeled “L” (left), “C” (center), and “R” (right). In this setup, the sensor elements in the L column were circular pads with a diameter of 4 mm, those in the C column had a diameter of 3 mm, and those in the R column had a diameter of 2 mm. In Figure 15b, the sensor elements were also organized into three columns; however, in this array, all nodes were uniform, with a diameter of 4 mm. The sensor elements were further divided along the x-axis into six parallel rows, labeled 1 through 6. The piezoelectric signals from all sensors were collected at each sensor element, transmitted through an interface to the charge amplifier module, and then sent to the DAQ for data conversion and processing.

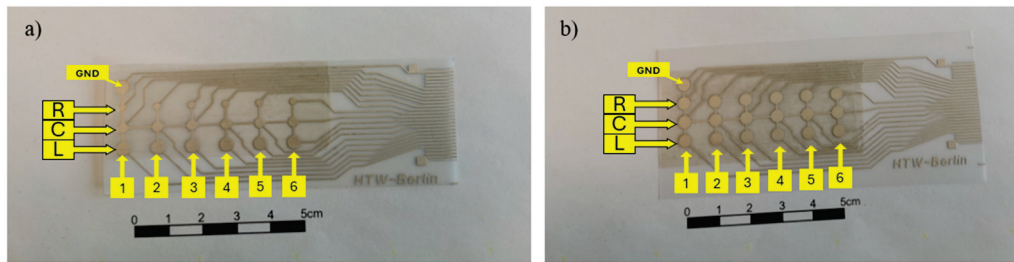


Figure 15. (a) Heterogeneous and (b) homogeneous sensor arrays.

During the signal processing stage, the amplified time-domain voltage signals from the sensors were filtered through a custom-designed IIR Butterworth band-pass filter with cutoff frequencies at 10 Hz and 40 Hz. The sampling rate is 20 kHz. Figure 16 presents the time-domain responses of the 18 sensor elements in the heterogeneous sensor array after filtering. In this figure, 1C represents the first-row sensor elements in the center column, 1L the first-row sensor element in the left column, and 1R the first-row sensor element in the right column. The figure demonstrates that, due to the consistent pressure distribution of the sound wavefront on the sensor surface, there is no phase shift between the time-domain signals. The variations in response are attributed to the different sizes of the sensor elements.

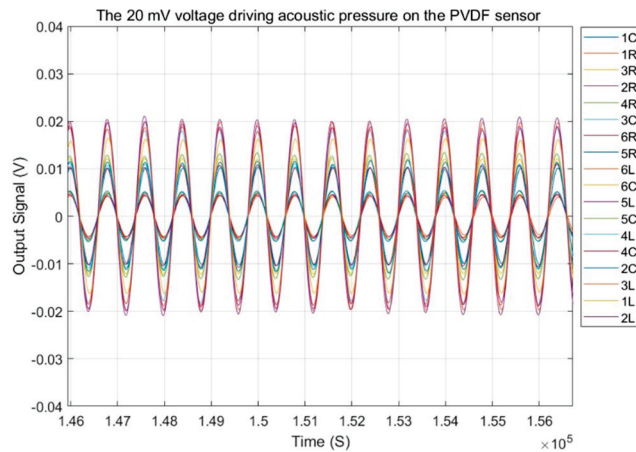


Figure 16. Sensor response of the heterogeneous PVDF sensor array in time domain.

Figure 17 illustrates the output responses of the heterogeneous sensor array. In Figure 17a, the amplitude response of sensor elements of varying sizes is shown as the sound pressure from the acoustic wave generator increases linearly. As the input from the acoustic wave generator increases, so does the sound pressure amplitude (P) on the sensor surface. The x-axis (P) represents the input acoustic pressure in pascals (Pa), while the y-axis represents the voltage response output by the charge amplification circuit in volts (V). This figure highlights how the voltage response changes with input pressure for three sensor sizes ($d_1 = 2$ mm, $d_2 = 3$ mm, $d_3 = 4$ mm). As input pressure increases, all sensors exhibit a corresponding rise in output voltage, with larger sensors generating higher

voltage responses. The largest sensor (4 mm) shows the highest output, producing a signal approximately four times greater than that of the smallest sensor (2 mm) at $P = 2117$ Pa (input acoustic pressure of 300 mV power input), indicating a proportional relationship between size and response. Figure 17b highlights the sensitivity of 18 sensor elements in the array under a driving voltage of $P = 240$ Pa (input acoustic pressure of 20 mV power input). The 4 mm sensors (L1 to L6) exhibit higher sensitivity values in the range of $14 \mu\text{V} \pm 6\%$ /Pa compared to their average value. The 3 mm sensors (C1 to C6) have their sensitivities in the range of $7.7 \pm 9\% \mu\text{V}/\text{Pa}$, while the 2 mm sensors (R1 to R6) show lower sensitivities in the range of $3.2 \pm 10\% \mu\text{V}/\text{Pa}$.

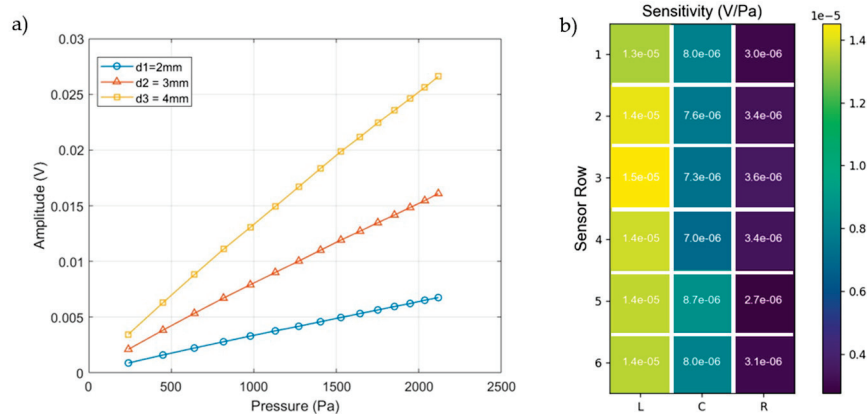


Figure 17. (a) Amplitude responses of sensors of different sizes as the driven voltage from the acoustic wave generator increases, and (b) sensitivities of each sensor element in the heterogeneous array at a constant pressure of $P = 240$ Pa.

Figure 18a compares the sensitivity of the 4 mm PVDF sensor elements to the Kulite sensor across different pressure levels. The Kulite sensor exhibits consistent sensitivity throughout the range, with a value of $27.241 \mu\text{V}/\text{Pa}$. In contrast, the 4 mm PVDF sensor, marked by black dots, shows sensitivities between $10.96 \mu\text{V}/\text{Pa}$ and $13.55 \mu\text{V}/\text{Pa}$, with a slight decline as pressure increases. Its average sensitivity is $12.1 \mu\text{V}/\text{Pa}$, with an error margin of $\pm 10\%$. Unlike the PVDF sensor, the Kulite sensor maintained higher and more stable sensitivity. The PVDF sensor's performance variation could be due to its piezoelectric material's response to ambient temperature or environmental factors during assembly.

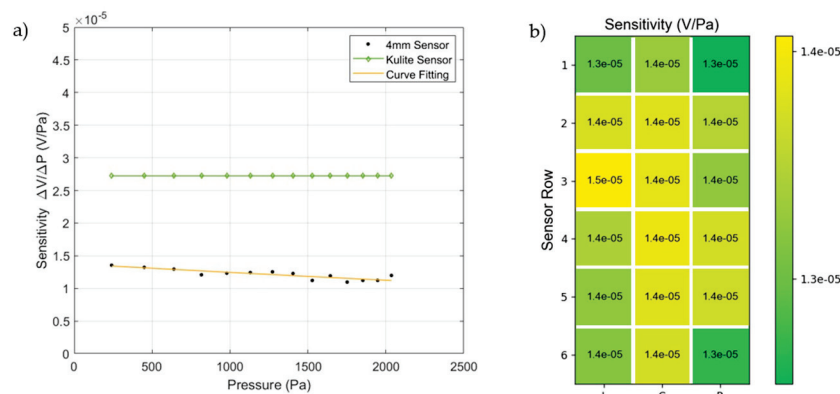


Figure 18. (a) The sensitivity of PVDF sensors (4 mm) and the Kulite sensor vs. increasing pressure, and (b) the sensor sensitivity of each sensor element of the test homogeneous PVDF sensor array at pressure $P = 240$ Pa.

Figure 18b displays the sensitivity distribution across different positions in a 4 mm PVDF sensor array at 240 Pa acoustic pressure. Sensitivity values range from $12.82 \mu\text{V}/\text{Pa}$

to $14.52 \mu\text{V}/\text{Pa}$, with the highest sensitivity at position 3L and the lowest at position 1R. The average sensitivity is $13.86 \mu\text{V}/\text{Pa}$, with an error margin of $\pm 8\%$. This error may be due to environmental factors in the laboratory and equipment errors during sensor assembly. Despite applying high pressure to exhaust the gas microbubbles before calibration, microbubbles may remain trapped at the sensor's edges, affecting the sensitivity of sensors located there.

4. SWBLI Study in Mach 2 Wind Tunnel

4.1. Experimental and Measurement Setup

The PVDF sensors were positioned beneath the separation bubble to capture pressure fluctuations in the interaction region. Kulite sensors were also installed for comparison, offering a benchmark for the PVDF sensor performance.

The experiments were conducted in the supersonic wind tunnel at the Technical University of Berlin. The supersonic wind tunnel was driven up to 16,000 rpm by a DC-motor with 400 kW power, as Figure 14 shows. The centrifugal compressor of the wind tunnel drew the air from outside into the tunnel, and the air was dried as it flowed through the silica gel drying chamber.

The air flow reached the Laval nozzle, which accelerates the air up to Mach number 2 at the entrance to the test section. A wedge shock generator with an angle of $\Phi = 10^\circ$ was installed at the top of the test section creating an oblique SWBLI. As shown in Figure 19, a plug was employed as a platform at the bottom of the test section for installing sensors to capture physical parameters in the test section. The PVDF sensor array was glued flat on the plug. As shown in Figure 20a,b, its signal conductive wires passed through the reserved slot on the plug board and connected with the electric circuit part from the backside of the plug. It is worth noting that the embedded plug must be installed into the wind tunnel with high tightness to avoid the acoustic noise from the connection gaps around the plug edges. The configuration of the wind tunnel is shown in Table 2.

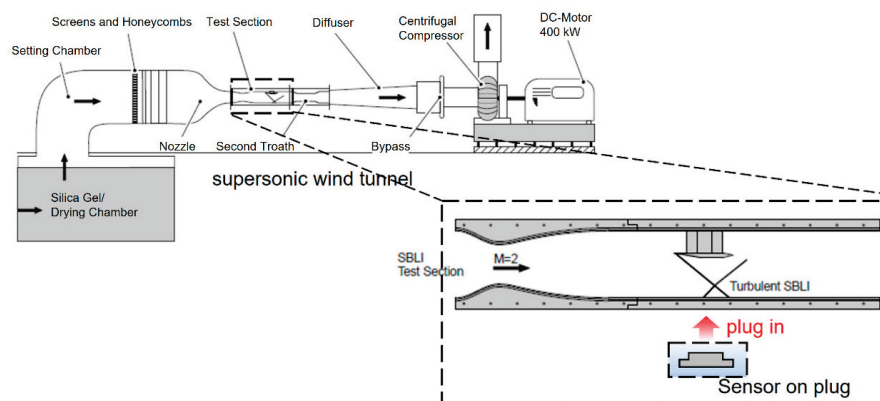


Figure 19. Schematic representation of the supersonic wind tunnel present at the Technical University of Berlin with a focus on the turbulent SBLI test section [28].

Table 2. Configuration of the wind tunnel for the SWBLI [30].

Item	Value	Unit
Mach Number M	2	
Freestream Unit Reynolds Number ReU	1.5×10^6	m^{-1}
Boundary Layer Velocity Thickness δ_0	7.2	mm
Shock Generator Angle Φ	10	degree
Test Section Length L_t	500	mm
Cross-Section Dimension	150×150	mm $\times$ mm

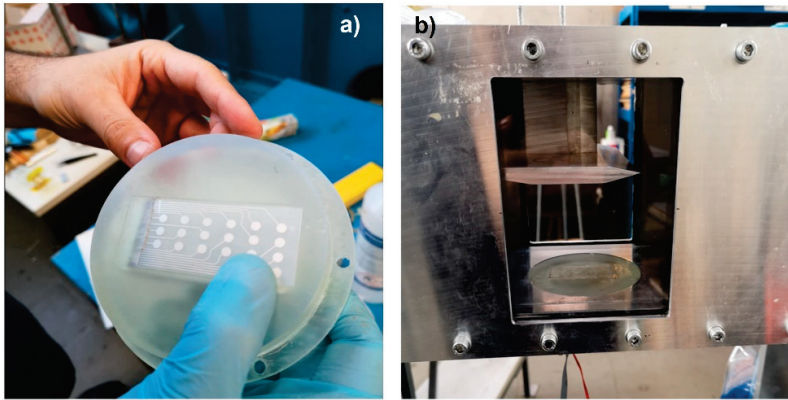


Figure 20. Installation of pressure sensor on plug (a), and (b) the plug was embedded in the test section in the wind tunnel.

When the Mach 2 flow inside the wind tunnel test section meets the shock generator, a shockwave is formed. The resulting oblique shockwave impinges on a fully turbulent boundary layer with a thickness of $\delta_0 = 7.2$ mm on the surface of the wind tunnel floor, which was also observed experimentally by Rohlf et al. [28,31]. As depicted in Figure 21a, this interaction creates a strong adverse pressure gradient, causing the boundary layer to separate. This separation results in a bubble of reversed flow with a characteristic length L_{sep} . The curvature of this bubble induces converging compression waves that merge into separation shock, interacting with the incident shock to form the characteristic λ -foot structure [36,37]. Another crucial parameter for subsequent analysis is the interaction length, L_{int} . This parameter is defined as the distance between the extrapolated impingement point, x_{imp} (based on inviscid shock theory) and the interaction onset, x_0 , where the wall pressure begins to increase [1,4,6]. To investigate unsteadiness of this kind of flow, the piezoelectric sensor is placed beneath the separation bubble, and the sensor is defined by the row marked ① to ⑥ in Figure 21b.

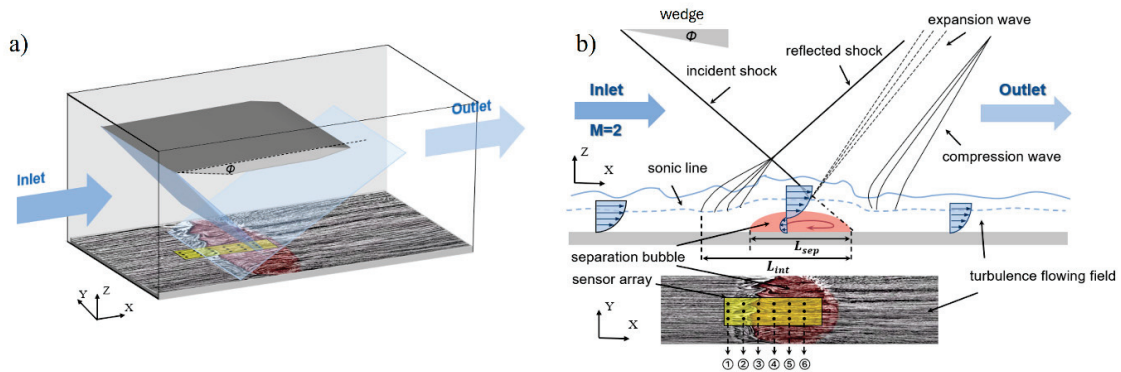


Figure 21. Schematic of (a) shock wave/boundary layer interaction with Mach 2.0 flow intake in 3d-presentation, and (b) the sensor mounting in the measurement.

In the field of shock wave/boundary layer interaction research, Kulite pressure transducers have gained wide recognition as a prominent technique for measuring wall pressure [18,28,31]. In this study, the Kulite sensor XCQ-062 (Kulite Semiconductor Products, Inc., One Willow Tree Road, Leonia, NJ, USA) [38] was employed as a reference and subjected to the same supersonic conditions. Unlike the PVDF thin-film pressure sensor, Kulite sensors are expensive and must be flush-mounted in the test plug, which entails drilling a hole at each detection position before installation. This requirement arises from the cylindrical packaging of the Kulite sensor, which is 1.7 mm in diameter and 9.5 mm in length. The intricacies associated with this installation process are regarded as one of the primary drawbacks of utilizing Kulite sensors in shock wave/boundary layer interaction study.

4.2. Results

Wind tunnel tests at Mach 2 yielded real-time voltage responses from the sensors, which were processed using a high-order FIR filter. After the spectral analysis process, the resulting PSD data provided insights into the frequency characteristics of the unsteady flow. The PVDF sensors detected pressure fluctuations across different positions within the separation and reattachment regions, capturing the energy distribution within the flow field (Figure 21) [15,28,31,39,40].

Through wind tunnel tests at a wind speed of Mach 2, we obtained the voltage response of the sensors (S1 to S6) in time domain, as shown in Figure 22. To further filter the low-frequency noise, the signal was processed through a designed high-order FIR (finite impulse response) high-pass filter using a Kaiser window function. The Kaiser window function adjustment parameter was $\beta = 5$. The FIR filter has an order of 500 and a cutoff frequency of 70 Hz.

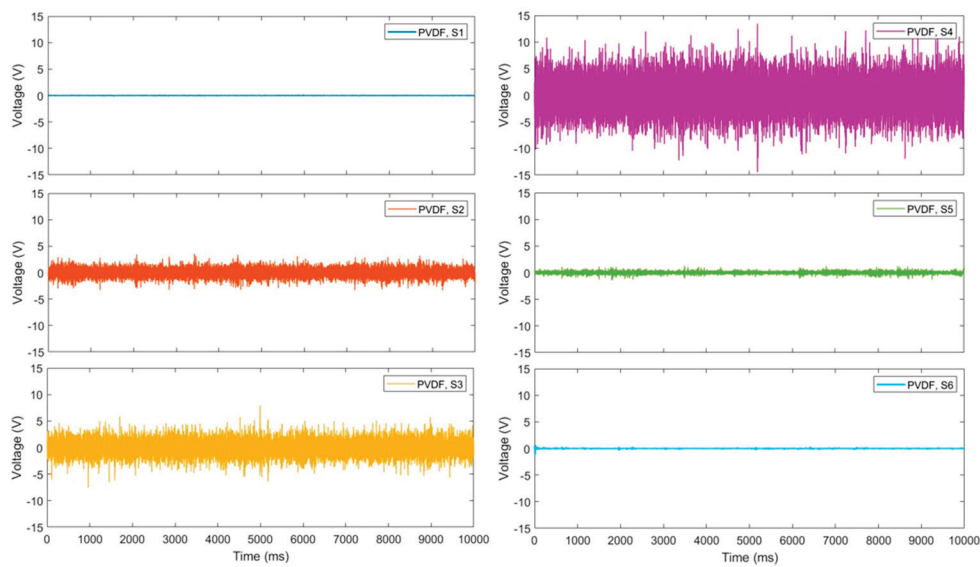


Figure 22. PVDF sensor responses at 6 positions in time domain.

By utilizing the pwelch function in Matlab® R2022b [41], these voltage response data were converted into Power Spectral Density (PSD). In the pwelch function, the Hanning window was used with a window length of 1/20th of the signal length and 50% overlap. Further, to more effectively demonstrate the energy distribution of the signal at different frequencies, particularly the low-frequency unsteadiness, pre-multiplied Power Spectral Density ($f \cdot PSD$) was used. Figure 23 shows the $f \cdot PSD$ values from sensors along the x-direction in the middle row of the PVDF sensor array, namely, S1 to S6. The plot highlights the variation in frequency responses at different positions within the separation and reattachment regions. The PVDF sensors provide insights into the pressure fluctuations and energy distribution within the flow field.

Figure 24 shows the $f \cdot PSD$ values from sensors along the x-direction in the middle row of the PVDF sensor array. The frequency has been normalized by the Strouhal number S_t which is defined as $S_t = fL_{int}/u_{\infty}$, where f is the frequency, the free stream velocity is $u_{\infty} = 512$ m/s, and the interaction length obtained is $L_{int} = 54.5$ mm with the $\theta = 10^\circ$ edge shock generator [28,31]. Referring to the parameters from a paper based on the same wind tunnel experiment [28,31,42], the normalized position parameter of the sensors, $X^* = (x_i - x_0)/L_{int}$, was calculated based on the interaction length L_{int} .

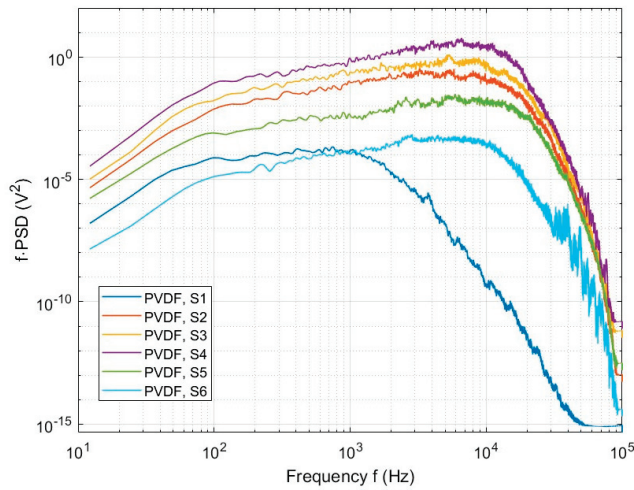


Figure 23. Value $f \cdot PSD$ of 6 PVDF sensors S1 to S6 in frequency domain.

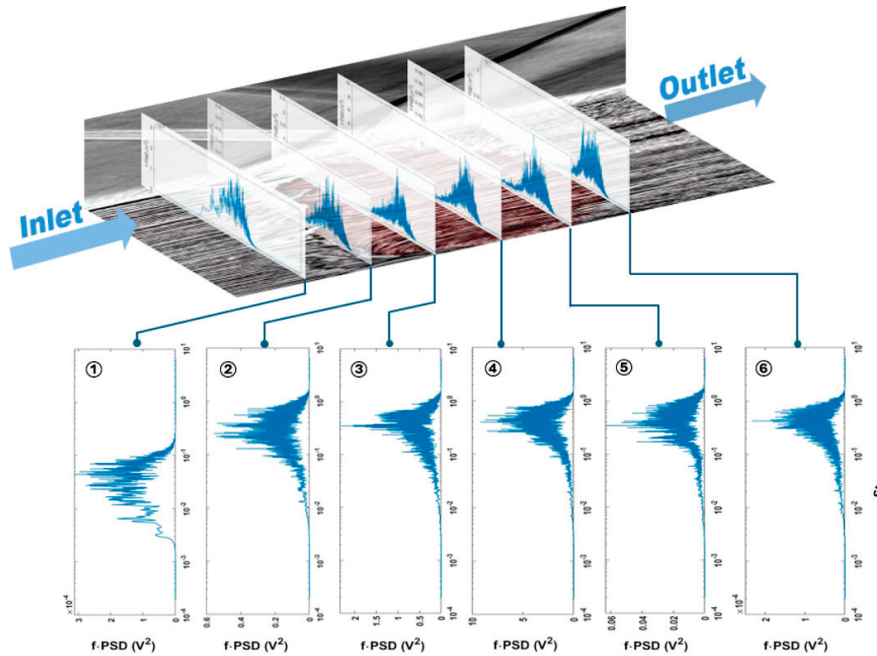


Figure 24. Schematic of the flow field and respective $f \cdot PSD$ spectra along the streamwise direction for the PVDF sensor array.

The provided plots offer significant insights into the energy distribution of the flow across various frequencies and locations within the shock wave/boundary layer interaction (SBLI). Consistent with the established literature, distinct frequency ranges are observed, each categorized into specific zones characterized by unique temporal scales [43–46].

These zones are as follows:

Low Frequency Zone (LFZ): This zone is captured by sensor S1, and it is situated at the onset of the interaction region, near the foot of the reflected shock at approximately X^* . It is marked by very low-frequency fluctuations, featuring a broad peak centered around $0.04 < St < 0.06$, which indicates large-scale pressure fluctuations due to the low frequency motion of the reflected shockwave.

Intermediate Frequency Zone (IFZ): Captured by sensors S2 to S6 and spanning the interaction region of $0.2 < X^* < 1$, inside this region, the frequency response has a broader range compared to the low frequency zone and exhibits medium-frequency characteristics at $St \approx 0.5$, which is in between the low frequencies' fluctuations of the reflected shock foot and the high frequencies of the small-scale turbulence of the supersonic boundary layer.

Usually, the IFZ region is associated with the shedding of large-scale coherent structures in the shear layer that develops over the shock-induced separation bubble.

For comparison, we integrated six Kulite sensors in the sensor plug, parallel to the S1 to S6 PVDF sensors. Both the PVDF sensors and the Kulite sensors were tested under the same condition simultaneously.

As shown in Figure 25, the power spectral density $f \cdot PSD$ show that the PVDF sensors exhibit a strong consistency with the Kulite sensors located at the same positions along the x-direction of the air flow. At location ①, the PVDF sensor and the Kulite sensor have very low frequency fluctuations and a broad hump centered in the Strouhal number range $0.04 < S_t < 0.06$. The peaks of the Kulite response at $S_t = 0.0033$ indicates the main hum noise at 50 Hz. From the location ② to ⑥, both the PVDF sensors and the Kulite sensors exhibit significant narrow humps in the range $0.4 < S_t < 0.5$.

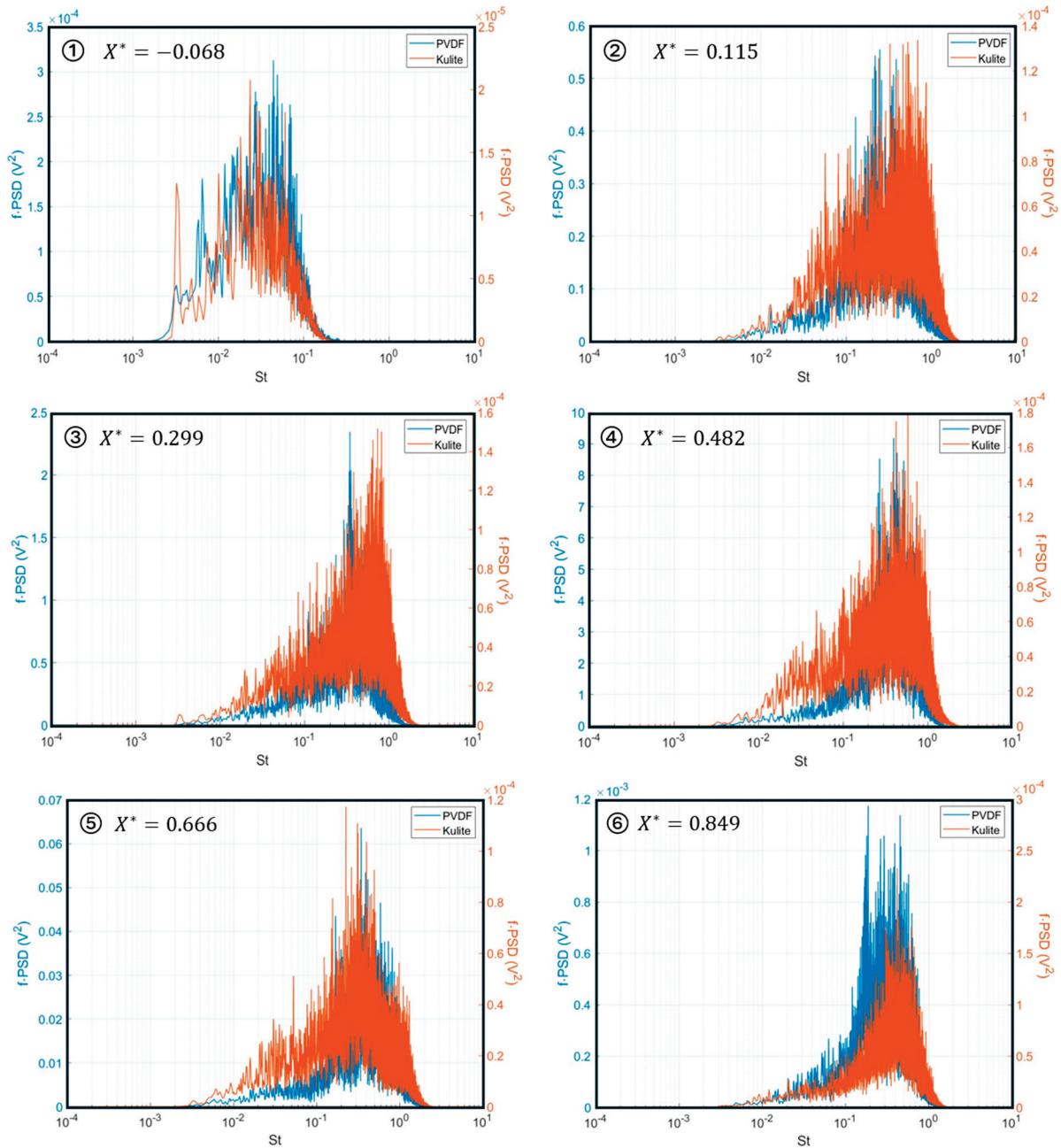


Figure 25. Comparison of values $f \cdot PSD$ vs. Strouhal number S_t at six investigation positions ① to ⑥ with Kulite sensors under the same conditions in the wind tunnel.

Since the PVDF sensors and Kulite sensors were measured under the same conditions during the wind tunnel experiment, there is no need to consider the time lag between the data. We chose the Pearson's Correlation Coefficient r to measure the correlation between the normalized power spectral densities of both the PVDF sensor and the Kulite sensor [47]:

$$r = \frac{\sum(x_i - \bar{x})(y_i - \bar{y})}{\sqrt{\sum(x_i - \bar{x})^2 \sum(y_i - \bar{y})^2}}, \quad (7)$$

where x_i and y_i are the observed values of the two data sets, and $\bar{x}$ and $\bar{y}$ are the means of the two data sets.

Comparison with the Kulite sensors showed that the PVDF sensors demonstrated strong consistency in detecting unsteadiness, particularly in the low-frequency range. Pearson's correlation coefficient analysis indicated moderate positive correlation ($0 < r < 1$) between the two sensor types, validating the PVDF sensor array's effectiveness. As in Figure 26, the Pearson correlation coefficient fluctuates between different X^* values but consistently remains between 0.42 and 0.52. This finding indicates that the correlation between sensor variables at different positions or time points is relatively stable. Notably, in sensor region ①, the correlation coefficient is the highest, reaching 0.5. In sensor region ③, the correlation is the weakest, at 0.39. From region ④ to ⑥, the correlation shows an increasing trend. The possible reasons for these differences may be attributed to the varying design size of the sensing transducers, sensitivity, and spectral response resulting from the different sensing mechanisms of piezoelectric (PVDF sensor) and piezoresistive (Kulite sensor) sensors as well. Additionally, the positioning of the sensors within the wind tunnel, measurement environment errors, human error, or even a combination of multiple factors could contribute to these variations.

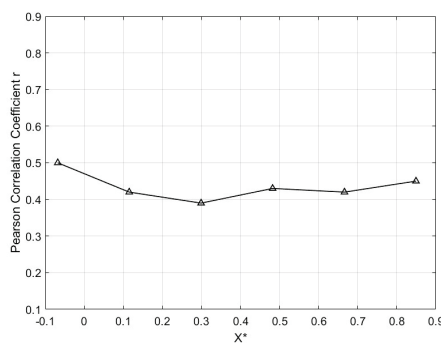


Figure 26. Illustrates the similarity of the pressure signal values received by the two types of sensors at the same location and time.

5. Conclusions

Finite Element Method (FEM) simulations were conducted to investigate the PVDF sensor array's frequency response characteristics, particularly its potential for high-frequency operations up to 300 kHz. The simulations revealed that the sensor exhibits high-pass filter behavior due to the interaction of its capacitive and resistive properties with circuit parasitics. This behavior suggests that the sensor can effectively minimize low-frequency noise while maintaining stability in the high-frequency range, a promising feature for applications that require sensitivity to dynamic pressure fluctuations, such as SWBLI studies.

The current experimental work, however, concentrated on validating the PVDF sensor array's low-frequency response, which aligns with the primary needs of SWBLI research. Low-frequency unsteadiness is a key factor in SWBLIs, driving phenomena like flow separation and recirculation, making accurate detections within this range essential. By first establishing a reliable low-frequency performance, this study addresses the immediate requirements for SWBLI research while setting a foundation for future high-frequency testing.

In wind tunnel experiments conducted at Mach 2, the PVDF sensor array demonstrated the effective detection of unsteady low-frequency pressure fluctuations. The results were consistent with data from Kulite sensors, confirming the PVDF array's sensitivity and reliability as a viable alternative for SWBLI applications. These findings underscore the PVDF sensor array's suitability for low-frequency unsteadiness studies, with its high-frequency potential—indicated by an FEM simulation—reserved for subsequent experimental validation.

This study's main contributions include the development and initial validation of a PVDF sensor array tailored for low-frequency wall-pressure measurements in SWBLI environments. The FEM simulations suggest that the sensor could maintain stable responses up to 300 kHz, providing a theoretical basis for high-frequency applications. While current experiments focused on low-frequency calibration, future work will extend testing to cover the high-frequency range. This further testing will allow for a comprehensive evaluation of the PVDF sensor array's operational range and dynamic response under varied flow conditions, enhancing its potential as a versatile tool in broader aerodynamic and aerospace applications, including high-frequency pressure measurements.

Future work will involve controlled testing of the PVDF sensor array under varying temperature and humidity conditions to assess its long-term stability and sensitivity. These tests will help evaluate the robustness of the PVDF material and validate its performance in diverse environments commonly encountered in aerodynamic research.

As an alternative piezoelectric sensing material, PVDF polymers, particularly in their nanofiber form, exhibit remarkable sensitivity to various mechanical stimuli, including pressure, strain, vibration, and motion [48]. This exceptional versatility has enabled their widespread application across diverse sensing technologies, further highlighting the potential of PVDF-based sensors for innovative advancements in the field.

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References

1. Harvey, J.; Babinsky, H. *Shock Wave-Boundary-Layer Interactions*; Cambridge University Press: Navi Mumbai, India, 2011; pp. 25–42. [CrossRef]
2. Hu, W.; Hickel, S.; van Oudheusden, B.W. Low-frequency unsteadiness mechanisms in shock wave/turbulent boundary layer interactions over a backward-facing step. *J. Fluid Mech.* **2021**, *915*, A107. [CrossRef]

3. Bernardini, M.; Posta, G.D.; Salvatore, F.; Martelli, E. Unsteadiness characterization of shock wave/turbulent boundary-layer interaction at moderate Reynolds number. *J. Fluid Mech.* **2023**, *954*, A43. [CrossRef]
4. Gaitonde, D.V. Progress in shock wave/boundary layer interactions. *Prog. Aerosp. Sci.* **2015**, *72*, 80–99. [CrossRef]
5. Mullenix, N.J.; Gaitonde, D.V. Analysis of unsteady behavior in shock/turbulent boundary layer interactions with Large-Eddy simulations. In Proceedings of the 51st AIAA Aerospace Sciences Meeting including the New Horizons Forum and Aerospace Exposition, Grapevine, TX, USA, 7–10 January 2013. [CrossRef]
6. Pasquariello, V.; Hickel, S.; Adams, N.A. Unsteady effects of strong shock-wave/boundary-layer interaction at high Reynolds number. *J. Fluid Mech.* **2017**, *823*, 617–657. [CrossRef]
7. Masini, L.; Timme, S.; Peace, A.J. Analysis of a civil aircraft wing transonic shock buffet experiment. *J. Mech.* **2019**, *884*, A1–A41. [CrossRef]
8. Lin, H.; Lu, X.G. Visualization of the shock wave/boundary layer interaction using polarization imaging. *J. Vis.* **2020**, *23*, 839–850. [CrossRef]
9. Atkins, M.D. Velocity field measurement using Particle Image Velocimetry (PIV). In *Thermo-Fluidic Measurement Techniques*; Butterworth-Heinemann: Oxford, UK, 2016; pp. 116–125. [CrossRef]
10. Humble, R.A.; Scarano, F.; van Oudheusden, B.W. Particle image velocimetry measurements of a shock wave/turbulent boundary layer interaction. *Exp. Fluids* **2007**, *43*, 173–183. [CrossRef]
11. Humble, R.A.; Elsinga, G.E.; Scarano, F.; Van Oudheusden, B.W. Three-dimensional instantaneous structure of a shock wave/turbulent boundary layer interaction. *J. Fluid Mech.* **2009**, *622*, 33. [CrossRef]
12. Ganapathisubramani, B.; Clemens, N.; Dolling, D. Effects of upstream boundary layer on the unsteadiness of shock-induced separation. *J. Fluid Mech.* **2007**, *585*, 369–394. [CrossRef]
13. Merienne, M.C.; Molton, P.; Bur, R.; Le Sant, Y. Pressure-Sensitive paint application to an oscillating shock wave in transonic flow. *AIAA J.* **2015**, *53*, 3208–3220. [CrossRef]
14. Standfield, S.; Kimmel, R.; Adamczak, D. HIFiRE-1 flight data analysis: Turbulent Shock-Boundary-Layer interaction experiment during ascent. In Proceedings of the 42nd AIAA Fluid Dynamics Conference and Exhibit, New Orleans, LA, USA, 25–28 June 2012. [CrossRef]
15. Clemens, N.T.; Narayanaswamy, V. Low-Frequency unsteadiness of shock wave/Turbulent boundary layer interactions. *Annu. Rev. Fluid Mech.* **2014**, *46*, 469–492. [CrossRef]
16. Huang, W.; Wu, H.; Yang, Y.G.; Yan, L.; Li, S.B. Recent advances in the shock wave/boundary layer interaction and its control in internal and external flows. *Acta Astronaut.* **2020**, *174*, 103–122. [CrossRef]
17. Hegde, A.; Changanthi, N.; Hubner, J.P.; Olcmen, S.M. Spectral study of wall pressure measurements in an impinging shock-wave boundary layer interaction. In Proceedings of the AIAA Scitech 2021 Forum, Virtual, 11–15 January 2021; p. 1558. [CrossRef]
18. Erengil, M.; Dolling, D. Unsteady wave structure near separation in a Mach 5 compression ramp interaction. *AIAA J.* **1991**, *29*, 728–735. [CrossRef]
19. Varadan, V.V.; Roh, Y.R.; Varadan, V.K. *Measurement of All the Elastic and Dielectric Constants of Poled PVDF Films*; IEEE: Piscataway, NJ, USA, 1989; Volume 2, pp. 727–730. [CrossRef]
20. Tadaki, D.; Ma, T.; Yamamiya, S.; Matsumoto, S.; Imai, Y.; Hirano-Iwata, A.; Niwano, M. Piezoelectric PVDF-based sensors with high pressure sensitivity induced by chemical modification of electrode surfaces. *Sens. Actuators A Phys.* **2020**, *316*, 112424. [CrossRef]
21. Lynch, C.S. Polyvinylidene fluoride (PVDF) elastic, piezoelectric, pyroelectric, and dielectric coefficients and their non-linearities. *Ferroelectrics* **1993**, *150*, 331–342. [CrossRef]
22. Zhu, M.; Luo, M.; Abdalla, I.; Yu, J.; Li, Z.; Ding, B. Highly shape adaptive fiber based electronic skin for sensitive joint motion monitoring and tactile sensing. *Nano Energy* **2020**, *69*, 104429. [CrossRef]
23. Lu, K.; Huang, W.; Guo, J.X.; Gong, T.X.; Wei, X.B.; Lu, B.W.; Liu, S.Y.; Yu, B. Ultra-sensitive strain sensor based on flexible poly(vinylidene fluoride) piezoelectric film. *Nanoscale Res. Lett.* **2018**, *13*, 83. [CrossRef]
24. Guo, Z.; Hu, Y.; Hu, X.R.; Shang, F.; Xiang, Y. In-situ polarization enhanced high-sensitivity acoustic locating and recognition sensor based on flexible PVDF-TrFE film array. In Proceedings of the IEEE 2020 3rd International Conference on Advanced Electronic Materials, Computers and Software Engineering (AEMCSE), Shenzhen, China, 24–26 April 2020. [CrossRef]
25. Zhou, Y.M.; Wang, H.B.; Qi, K.; You, X.L.; Shao, W.L.; Wang, L.D.; Ding, B.; Cui, S.Z. Highly sensitive, self-powered and wearable electronic skin based on pressure-sensitive nanofiber woven fabric sensor. *Sci. Rep.* **2017**, *7*, 12949. [CrossRef]
26. Luo, Z.; Chen, S.; Wang, L.L.; Jiang, K.; Shen, G.Z. An ultra-sensitive and rapid response speed graphene pressure sensors for electronic skin and health monitoring. *Nano Energy* **2016**, *23*, 7–14. [CrossRef]
27. Lee, J.H.; Yoon, H.J.; Kim, T.Y.; Gupta, M.K.; Lee, J.H.; Seung, W.; Ryu, H.; Kim, S.W. Micropatterned P(VDF-TrFE) film-based piezoelectric nanogenerators for highly sensitive self-powered pressure sensors. *Adv. Funct. Mater.* **2015**, *25*, 3203–3209. [CrossRef]
28. Rohlf, L.; Stab, I.; Weiss, J. Experimental Investigations of Incident Shockwave Boundary Layer Interactions in a Continuously Operating Supersonic Wind Tunnel. In Proceedings of the AIAA Aviation 2022 Forum, Chicago, IL, USA, 27 June–1 July 2022; American Institute of Aeronautics and Astronautics: Reston, VA, USA, 2022. [CrossRef]
29. Dussauge, J.P.; Dupont, P.; Debieve, J.F. Unsteadiness in shock wave boundary layer interactions with separation. *Aerosp. Sci. Technol.* **2006**, *10*, 85–91. [CrossRef]
30. Datasheet of PVDF Film. Available online: <https://piezopvdf.com/piezo-pvdf-film-kit/> (accessed on 10 July 2023).

31. Corsi, C.; Rohlf, L.; Weiss, J.; Wang, B.; Kahf, M.; Obloch, P.; Ngo, H.D. Development of PVDF piezo-film sensor array for unsteady wall-pressure measurements in a turbulent SBLI. In Proceedings of the AIAA Aviation 2022 Forum, Chicago, IL, USA, 27 June–1 July 2022. [CrossRef]
32. Kumar, Y.; Trivedi, A.; Shukla, S.K. Investigating the Influence of Frequency on Piezo-dynamics of Polyvinylidene Fluoride (PVDF) Films Embedded in Confined Geomaterials. *J. Vib. Eng. Technol.* **2024**. [CrossRef]
33. Saxena, P.; Shukla, P. A comprehensive review on fundamental properties and applications of poly(vinylidene fluoride) (PVDF). *Adv. Compos. Hybrid Mater.* **2021**, *4*, 8–26. [CrossRef]
34. Datasheet of TLC272. Available online: https://www.ti.com/product/TLC272?utm_source=google&utm_medium=cpc&utm_campaign=asc-null-null-GPN_EN-cpc-pf-google-eu&utm_content=TLC272&ds_k=TLC272+Datasheet&DCM=yes&gclid=EAIaIQobChMIorrz-2DgAMVjLnVCh1qqAbNEAAYASAAEgKu7PD_BwE&gclidsrc=aw.ds (accessed on 10 September 2024).
35. Tavoularis, S. *Measurement in Fluid Mechanics*, 1st ed.; Cambridge University Press: Navi Mumbai, India, 2005; pp. 201–205. ISBN 978-0-521-81518-5.
36. Doerffer, P.; Dallmann, U. Reynolds number effect on separation structures at normal shock wave/boundary-layer interaction. *AIAA J.* **1989**, *27*, 1206–1212. [CrossRef]
37. Doerffer, P. An experimental investigation of the Mach number effect upon a normal shock wave—Turbulent boundary layer interaction on a curved wall. *Acta Mech.* **1989**, *76*, 35–51. [CrossRef]
38. Datasheet. Available online: <https://kulite.com/products/product-advisor/product-catalog/ultraminiature-pressure-transducer-xcq-062/> (accessed on 10 July 2023).
39. Hurst, A.M.; Olsen, T.R.; Goodman, S.; Van De Weert, J.; Shang, T.H. An experimental frequency response characterization of MEMS piezoresistive pressure transducers. In Proceedings of the ASME Turbo Expo 2014, Turbine Technical Conference and Exposition, Düsseldorf, Germany, 16–20 June 2014; Volume 6. GT2014-27159. [CrossRef]
40. Hu, N. Sensor-size-related attenuation correction of wall pressure spectra measurements. *Phys. Fluids* **2022**, *34*, 067119. [CrossRef]
41. Welch's Power Spectral Density Estimate in Matlab. Available online: <https://de.mathworks.com/help/signal/ref/pwelch.html> (accessed on 10 July 2023).
42. Montgomery, D.C.; Runger, G.C. *Applied Statistics and Probability for Engineers*, 7th ed.; John Wiley & Sons: Hoboken, NJ, USA, 2018; pp. 299–303. ISBN 978-1-119-40036-3.
43. Threadgill, J.A.S.; Bruce, P.J.K. Unsteady Flow Features Across Different Shock/Boundary-Layer Interaction Configurations. *AIAA J.* **2020**, *58*, 3063–3075. [CrossRef]
44. Dussauge, J.-P.; Piponniau, S. Shock/boundary-layer interactions: Possible sources of unsteadiness. *J. Fluids Struct.* **2008**, *24*, 1166–1175. [CrossRef]
45. Weiss, J.; Little, J.C.; Threadgill, J.A.; Gross, A. Low-Frequency Unsteadiness in Pressure-Induced Separation Bubbles. In Proceedings of the AIAA Scitech 2021 Forum, Virtual, 11–15 January 2021; American Institute of Aeronautics and Astronautics: Reston, VA, USA, 2021; pp. 1–19. [CrossRef]
46. Agostini, L.; Larchevêque, L.; Dupont, P. Mechanism of shock unsteadiness in separated shock/boundary-layer interactions. *Phys. Fluids* **2015**, *27*, 126103. [CrossRef]
47. Kiya, M.; Sasaki, K. Structure of a Turbulent Separation Bubble. *J. Fluid Mech.* **1983**, *137*, 83–113. [CrossRef]
48. Bhatta, T.; Maharjan, P.; Cho, H.; Park, C.; Yoon, S.H.; Sharma, S.; Salauddin, M.; Toyabur Rahman, M.; Sohel Rana, S.M.; Park, J.Y. High-performance triboelectric nanogenerator based on MXene functionalized polyvinylidene fluoride composite nanofibers. *Nano Energy* **2021**, *81*, 105670. [CrossRef]

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Article

Laterally Actuated Si-to-Si DC MEMS Switch for Power Switching Applications

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Abstract: Electrothermal actuators are highly advantageous for microelectromechanical systems (MEMS) due to their capability to generate significant force and large displacements. Despite these benefits, their application in reconfigurable conduction line switches is limited, particularly when employing commercial processes. In DC MEMS switches, electrothermal actuators require electrical insulation between the biasing voltage and the transmission line to prevent interference and maintain the integrity of the switch. This work presents a chevron-type electrothermal actuator utilizing a stack of SiO₂/ Al thin films on a silicon (Si) structural layer beam to create a DC MEMS switch. The design leverages a thin film Al heater to drive the actuator while the SiO₂ layer provides electrical insulation, suppressing crosstalk with the Si layer. The electrical contact resistance of a Si-to-Si interface was evaluated by applying a controlled current and measuring the resultant voltage. A low contact resistance of 150 Ω was achieved when an initial contact gap of 2.52 μm was closed using an actuator with an actuation voltage of 1.2 V and a current of 205 mA, with a switching speed of less than 5 ms. Factors such as the contact force, the temperature, and the residual device layer etching angle significantly impact the Si-to-Si contact resistance and the switch's longevity. The switch withstands a breakdown voltage up to 350 V at its terminal contacts. Thus, it will be robust to self-actuation caused by unwanted voltage contributions, making it suitable for high-voltage and harsh environment applications.

Keywords: electrothermal actuator; microelectromechanical systems (MEMS); PiezoMUMPs; DC MEMS switch; Si-to-Si contact resistance

1. Introduction

In recent years, reconfigurable, integrated, and miniaturized switches have been in high demand in order to reduce the complexity of power switching and fault protection systems in the transport industry [1–3]. Microelectromechanical systems (MEMS) switches/relays combine the benefits of solid-state switches, such as their compact size, low energy consumption, low cost, and compatibility with other integrated circuits on the same substrate [4]. Most importantly, they mirror the desired characteristics of conventional electromechanical relays through direct physical contact and low leakage currents [5]. Accordingly, MEMS switches can feature low insertion losses, high linearity, high isolation, and wide operation electrical bandwidth [6,7]. This can be advantageous for implementing configurable power networks and other payload systems [8–10]. These make switches based on MEMS technology an interesting and active area of research. However, despite the progress of DC MEMS switches, the endurance to withstand large currents and breakdown voltages has been a bottleneck [11,12], causing some unsatisfactory reliability issues. These issues include contact stiction, wear and micro-welding, material transfer between the contact, high contact resistance, and dielectric breakdown of the air gap at the off-state [13]. Many of these issues are found in electrostatically actuated switches [13,14]. In this regard,

MEMS switches based on thermal actuation stand to be of great interest to many applications such as RF MEMS switches [15,16], optical switches [17,18], and tensile loading [19], due to promising features such as high contact force, robustness, wide travel range, in-plane design flexibility, and low fabrication complexity [20–22]. As the actuation mechanism is built on a resistive heating element, it requires a DC current to maintain its active state. This can also be customized to achieve a bistable switch to eliminate the current and power consumption during operation [23]. The thermal actuator's robustness in environments where high-voltage spikes can be observed is another advantage making this actuation mechanism suitable for applications in the transportation sector, which require high-voltage resilience [24].

Several works on electrothermal MEMS switches for power switching applications using commercial foundry processes provided by MEMSCAP (Crolles, France) were published, mostly using the Metal Multi-User MEMS Process (MetalMUMPS) [8,23,25]. However, that process lacks a silicon (Si) structural layer, which brings challenges with regard to the robustness of the designs for harsh environments. In [26], a study was reported on electrostatically actuated Si-to-Si contact microswitches that were custom-fabricated using an ultraclean encapsulation process and operated within a temperature range of $-60\text{ }^{\circ}\text{C}$ – $300\text{ }^{\circ}\text{C}$, but the contact resistance was in the hundreds of kilohms. In [27], the MEMSCAP Silicon-on-Insulator Multi Users Micromachining process (SOIMUMPS) was utilized to design a structure with a hybrid actuation mechanism. The design used an electrostatic latching voltage of 21 V to keep the switch in an active state. However, it was not possible to isolate the RF signal line from the actuator signal. In [28], a test platform that utilized a V-shaped thermal actuator driven by a thin-film aluminum (Al) heater stacked onto a SiO_2 layer was reported. The actuator was robust against out-of-plane deformation due to its high aspect ratio, with a Si structural layer thickness of $50\text{ }\mu\text{m}$. However, it required substantial energy consumption equivalent to 200 mA at 3 V to close the contact. Recently, in [29], an electrostatically actuated MEMS switch for microwave band application using the MEMSCAP Piezoelectric Multi-User MEMS Process (PiezoMUMPs) was reported. The authors observed the non-uniformity of the contact with a scallop diameter of $0.33\text{ }\mu\text{m}$ on the sidewall interface. As such, non-ohmic contact was observed and demonstrated. This observation implies that the contact interface can be improved by mechanical polishing.

Previously, in [30], we demonstrated an electro-thermal chevron-type (i.e., V-shape) MEMS actuator fabricated using the PiezoMUMPs process and aimed at power switching applications that require high breakdown voltage, taking advantage of the high-robustness silicon structural layer provided by the process [31]. However, the Si-to-Si contact interface has shown inconsistency in ohmic contact resistance measurement. A close observation of the interface demonstrated scallops that could cause out-of-plane misalignments of the conducting lines upon contact. Other issues observed included the influence of temperature and the possible appearance of a native oxide across the sidewall interface, besides the restricted lifespan of the switch due to the high current density sustained by the heater.

Accordingly, this work further investigates the design, fabrication, and characterization of a V-shape MEMS actuator with an improved force, using the PiezoMUMPs process. The actuator is integrated with a shuttle to implement a direct contact (DC) switch aimed at power switching applications and high-voltage resilience. Additionally, this work offers potential engineering approaches to enhance the understanding of the impact of silicon sidewall morphology and its negative effects on the formation of reliable ohmic contacts. The study also investigates the effect of the contact temperature on the contact resistance by implementing a near zero-temperature actuation test, i.e., by manual contact operation through a micro-needle.

This paper is structured as follows: Section 2 presents the proposed design along with the simulation results. Section 3 provides a brief overview of the prototype fabrication process. Section 4 details the experimental results of the fabricated design, including the DC characterization of the actuator, the I – V curve of the Si-to-Si contact interface, the etching angle profile, and the switching speed of the actuator. Section 5 discusses the influence

of the sidewall contour on the contact interface of the MEMS switch. Finally, Section 6 concludes the paper.

2. Design and Simulation

2.1. Design Configuration

The schematic shown in Figure 1 illustrates the proposed concept of an in-plane electrothermally actuated Si-to-Si DC MEMS switch. In this work, variations in the widths of different layers are determined by fabrication constraints, which require specific enclosure distances between the layers to ensure proper functionality and manufacturability. The main part of the design is a chevron type electrothermal actuator driven by a heater built from the Padmetal layer made of aluminum (Al) and chromium (Cr), stacked onto a silicon Si structural layer via a thin silicon dioxide SiO_2 layer. The SiO_2 layer, positioned between the Al and Si layers, provides electrical isolation between the switch transmission line and the driving heater current.

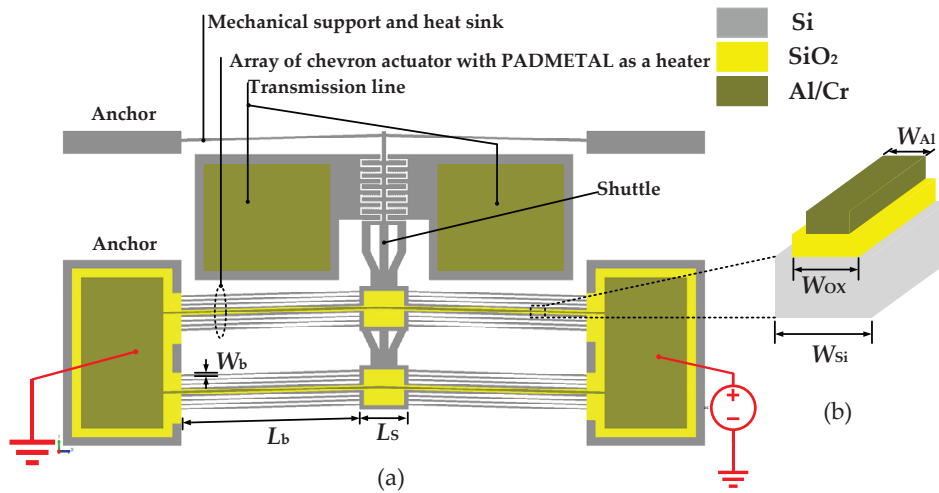


Figure 1. Schematic illustration of the thermally actuated DC MEMS switch: (a) top view and (b) cross-section of the Al heater on top of the Si layer.

Typically, the chevron actuator features a symmetric geometry that is composed of an array of beams slanted at a tilted angle, β , in such a way that the generated heat enables lateral motion due to joule heating and thermal expansion.

To reduce the out-of-plane deformation, only one pair of the chevron beams is covered with the SiO_2 / Al-Cr layer to act as a heater, allowing the remaining chevron beams to have a reduced width of $6 \mu\text{m}$, as depicted in Figure 1. This is advantageous, as it reduces the stiffness and serves as a heat sink to reduce the contact temperature. In addition, these narrower beams reduce the high impact of the lower aspect ratio of the heater beam (i.e., the Si thickness, $t_{Si} = 10 \mu\text{m}$, is significantly smaller than the Si width, $W_{Si} = 21 \mu\text{m}$) due to fabrication process constraints. Figure 1b illustrates the patterning of the silicon chevron beams with SiO_2 / Al-Cr layers, having a total width of $21 \mu\text{m}$. In this configuration, the SiO_2 layer encloses the side edges of the Al-Cr layer by $3 \mu\text{m}$. The design includes an air gap of $3 \mu\text{m}$ between the switch contacts, which the actuator must traverse to bring the shuttle comb structure into contact with the transmission line structures, thereby closing the switch. In addition, the multilayered structure is vulnerable to intrinsic stress gradients due to the wide coefficient of linear thermal expansion (CTE) mismatches between the silicon ($\alpha_{Si} = 2.6 \times 10^{-6}/\text{K}$) and Al ($\alpha_{Si} = 23.1 \times 10^{-6}/\text{K}$) layers. As a result, the beam's moment of inertia is smaller for out-of-plane motion compared to in-plane motion, leading to an undesired tendency for the actuator to favor out-of-plane motion during operation [12]. This needs to be carefully assessed in the simulation.

2.2. Simulation

The CoventorWare11.1 FEM software package with an electro-thermomechanical solver was utilized to study the displacement, temperature, and current of the heater over an applied voltage range. In the solid mechanics physics, the two anchors of the V-shaped beam of the actuator are set to be fixed boundaries, where the voltage is applied across the anchor points. The room temperature is set at the bottom of the anchors and the rest of the structure is configured to exhibit conductive heat transfer [32]. The designed dimensional parameters in Figure 1 are listed in Table 1.

Table 1. Design dimensions of the thermal chevron actuator.

Geometry Parameter	Symbol	Value
Length of the chevron actuator	L_b	500 μm
Length of the shuttle	L_s	110 μm
Width of the chevron actuator	W_b	6 μm
Width of the chevron actuator	W_{Si}	21 μm
Width of the Padoxide	W_{OX}	11 μm
Width of Padmetal	W_{Al}	5 μm
Angle of chevron beams	β	1.4°

The displacement of the actuator at the shuttle tip for a voltage ranging from 0 to 1.2 V applied across the actuator terminals (i.e., heater) is shown in Figure 2a. This 1.2 V is the typical voltage required to provide in-plane displacement of 2.52 μm to close the initial contact gap. At this displacement, the out-of-plane displacement (misalignment with respect to the fixed parts) was recorded to be 100 nm, as illustrated in Figure 2b. It is deemed acceptable. The maximum generated temperature on the actuator was 525.2 K, as shown in Figure 2c, whereas the recorded temperature on the contact of the switch was 450 K. Note that the actuator can operate only at temperatures that are below the melting point of the heater's aluminum material, which is 960 K [33].

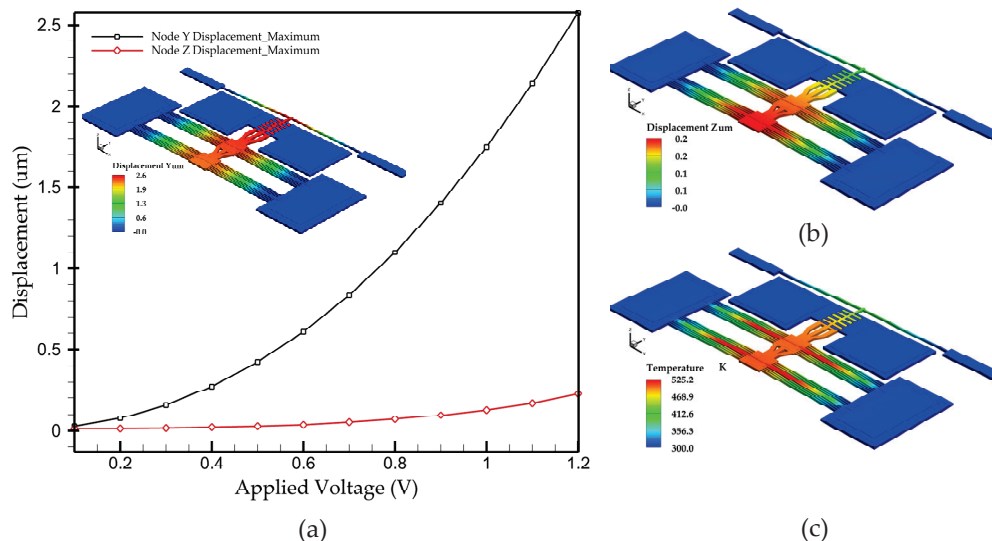


Figure 2. (a) Simulation results of the displacement vs. actuation voltage, (b) out-of-plane displacement profile, and (c) temperature profile of the thermal actuator.

3. Fabrication and Characterization

The switch devices were fabricated following the PiezoMUMPs process technology provided by Science Corporation [34]. The process with a total of five masks begins with a 400 μm -thick handle substrate, an n-type (100) silicon orientation wafer with a $1 \pm 0.05 \mu\text{m}$ buried oxide (BOX) insulation layer, and a 10 μm -thick phosphorus-doped silicon device

layer. Figure 3a,d summarize the sequence of steps involved in the fabrication process of the thermally actuated MEMS switch devices. A thermally grown 200 nm-thick SiO_2 layer is patterned to provide electrical isolation, as shown in Figure 3b. The microheater, composed of a 1 μm -thick Al layer and a 20 nm-thick Cr adhesion layer, is deposited on top of the device structural layer using e-beam evaporation and patterned through a lift-off technique, as shown in Figure 3c. The suspended silicon structural layer in Figure 3d is patterned through front side reactive ion etching (RIE) to remove the thermal silicon dioxide layer and deep reactive ion etching (DRIE) to remove the silicon structural layer. This is followed by coating the front side with a protective polyimide layer and the backside RIE on the silicon dioxide layer. The silicon handle wafer is then removed through DRIE. After the etch is completed, the photoresist is removed, and a wet oxide etch process is then used to remove the SiO_2 layer in regions defined by the trench mask. The frontside polyimide layer is finally removed using an oxygen dry etch process. The released device cross-section is shown in Figure 3d.

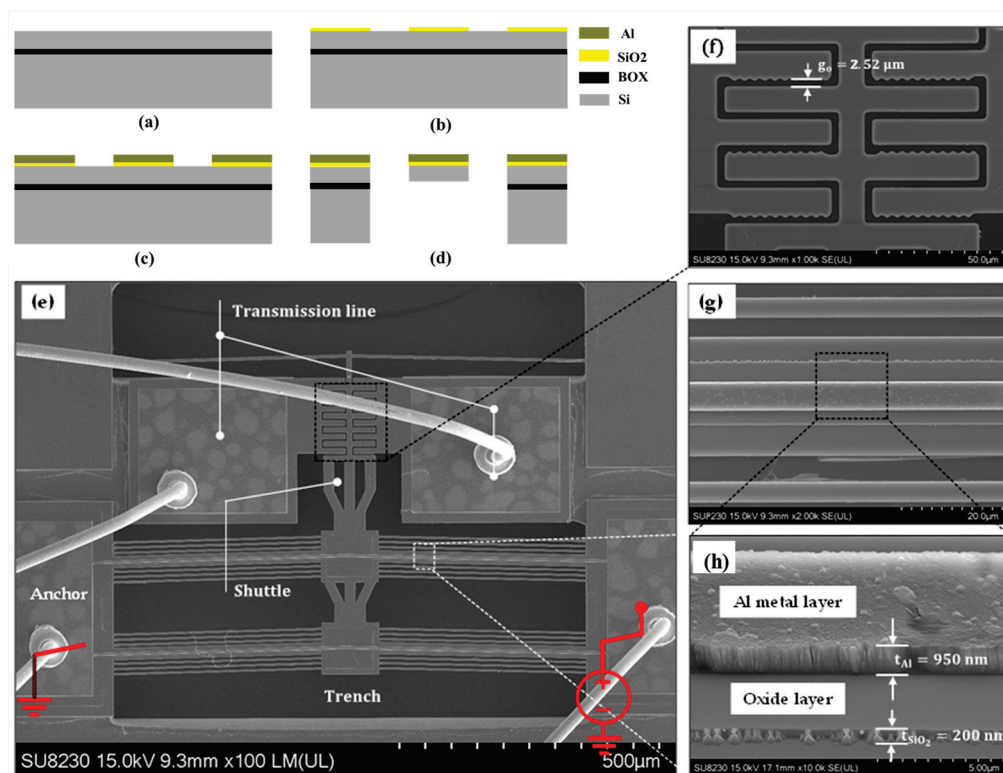


Figure 3. Structure of the fabricated thermally actuated Si-to-Si DC MEMS switch: (a–d) simplified fabrication process flow, (e) SEM of the switch, (f) zoomed-in image of the switching contact area, and (g,h) zoomed-in image of the chevron beam detailing the Al heater and silicon oxide insulating layer.

Prior to experimental testing, the fabricated devices were visually inspected using a scanning electron microscope (SEM) model SU6 300, as shown in Figure 3e–h. The measured thickness of the Al layer is ~950 nm compared to the designed value of 1 μm , whereas the designed contact gap of 3 μm was measured to be 2.52 μm . These small discrepancies could be attributed to variations in the deposition process parameters. Figure 3h shows a higher grain boundaries defect for the Al layer compared to the oxide layer.

4. Measurements and Results

This section discusses the switch electrical characterization and behavior of the ohmic contact. Initially, the on/off characteristics are measured, followed by an evaluation of the

contact resistance during quasi-static actuation. Additionally, the transient response of the chevron actuator upon establishing the ohmic contact is analyzed.

4.1. Displacement Measurement

Figure 4 illustrates the schematic diagram for the DC characterization setup used to measure displacement and contact resistance. The setup includes a Tektronix PWS4205 programmable DC power supply (0–20 V, 5 A), a source meter measurement unit, and a Keyence VHX7000 optical microscope. The device under study is mounted on a ceramic DIP package. The device is wire-bonded using gold wires. Subsequently, the package is connected into a prototype PCB board for measurement.

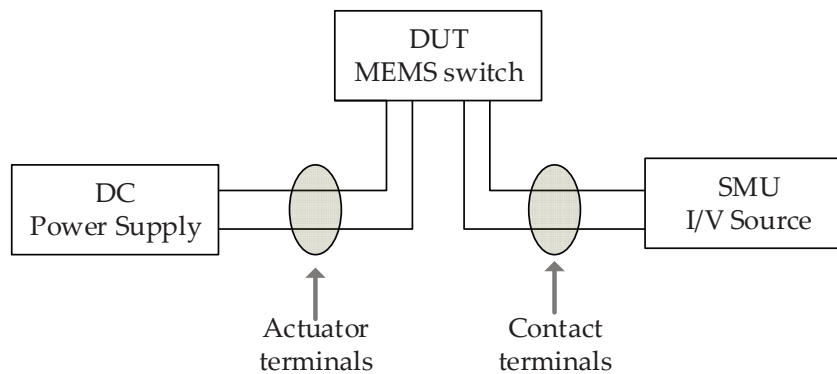


Figure 4. Schematic of the DC measurement setup for MEMS switch characterization.

To verify the functionality of the fabricated devices, the static response of the chevron thermal actuator driven by the Al heater was examined. The displacement of the thermal actuator's contact tip was measured by capturing its motion at each increment of the biasing voltage using an optical microscope until the gap was entirely closed at an average measured displacement of $2.52\ \mu\text{m}$. The actuation voltage and current (I – V) necessary to close the switch gap were found to be 205 mA and 1.2 V, respectively, as shown in Figure 5. The observed difference between the measured and simulated I – V curves in Figure 5 is mainly due to the simulation model assuming ideal electrical conditions. However, real-world factors, including environmental conditions and fabrication tolerances, such as variations in the thickness and width of the aluminum layer across the interconnect geometry, introduce deviations that are not captured by the simulation.

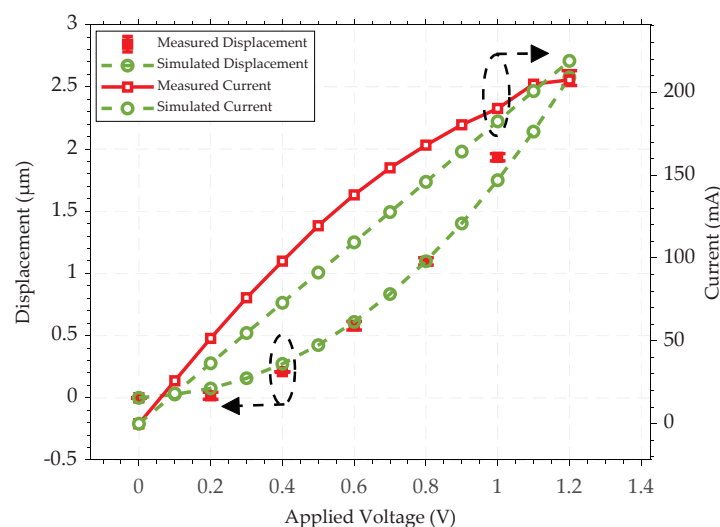


Figure 5. Switch displacement and actuator current as a function of applied heater voltage.

4.2. Contact Resistance Measurement

The contact resistance of the MEMS switch was measured using a source meter unit (SMU) after the switching interfaces were closed and mechanically settled. The test setup utilized to determine the contact resistance is depicted in Figure 4. Upon closing the MEMS switch, the initial contact resistance was observed to be around $400\ \Omega$. The current–voltage (I – V) characteristics, shown in Figure 6, were obtained with a test limiting a current of 5 mA passing through the switch. Within the tested limited current, the voltage drop across the contact was found to vary due to the variation in the contact resistance, where the initial measured $400\ \Omega$ resistance corresponds to a ~ 0.25 mA testing current and a 0.1 V drop across the contact. At the maximum testing current of 5 mA, the measured contact resistance dropped to only $240\ \Omega$; therefore, the maximum measured voltage drop was 1.2 V, as shown in Figure 6. This progressive decrease in the contact resistance with the increase in the testing current results in erratic behavior and inconsistencies in the ohmic contact. The initial establishment of the contact resistance suggests that instability may occur when the contact force is below the threshold value, coupled with the high temperature across the shuttle resulting from resistive heating. Notably, this variation in resistance with voltage may be due to the resulting self-heating of the switch causing additional force to be generated by the actuator. This highlights the importance of optimizing both the mechanical force and thermal management to ensure uniform switch operation.

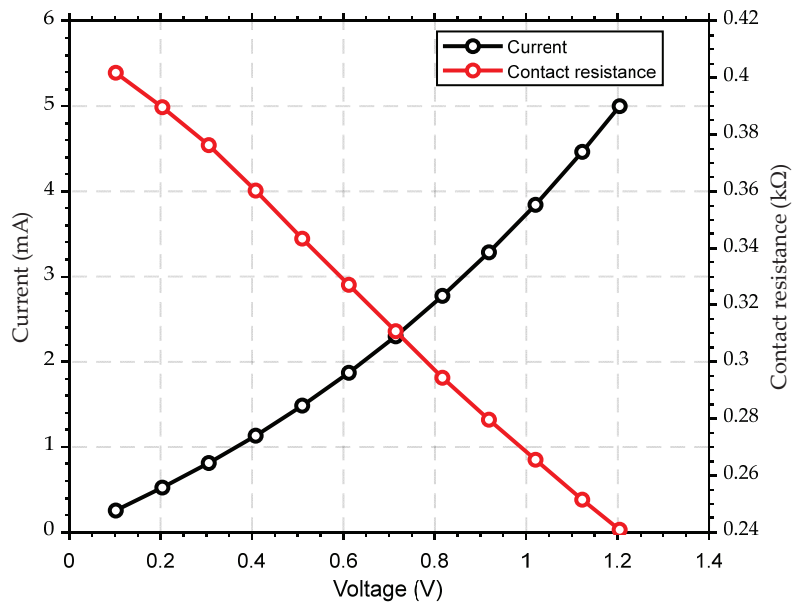


Figure 6. I – V plot with the extracted electrical contact resistance of the Si-to-Si contact.

4.3. Switching Time

The dynamic transient response was measured using a square wave signal toggling between 0 and 1 V at a rate of 1 Hz and applied to the Al heater. The switching speed of the proposed MEMS switch is explored in ambient air with a simple test circuit depicted in Figure 7a. V_{Drive} generates a driving signal for the heater and activates the switch. The switching contact is used as a sensor to detect the response time of the actuator when excited to the desired displacement. Thus, this creates a current path from the supply voltage, V_S , through the floating contact to the off-the-shelf resistor, R_{Load} . Both V_{Drive} and the voltage across the resistor R_{Load} are monitored using an oscilloscope. The actual setup used to measure the response time of the switch is shown in Figure 7b. It was observed that the reaction time for a rising edge τ_{rise} was ~ 5.7 ms, whereas the falling edge, τ_{fall} , was found to be ~ 1.5 ms, as measured in Figure 8. The rise time is longer than the fall time, and this is attributed to the Joule heating time constant and the cooling time constant being different. Note that the variability of the voltage across R_{Load} is due to the variation in

the contact resistance, as mentioned above. Moreover, potential reasons for the deviation include fabrication imperfections, mainly over-etching and the etching angle during the DRIE process. The high aspect ratio also increases the stiffness and out-plane displacement.

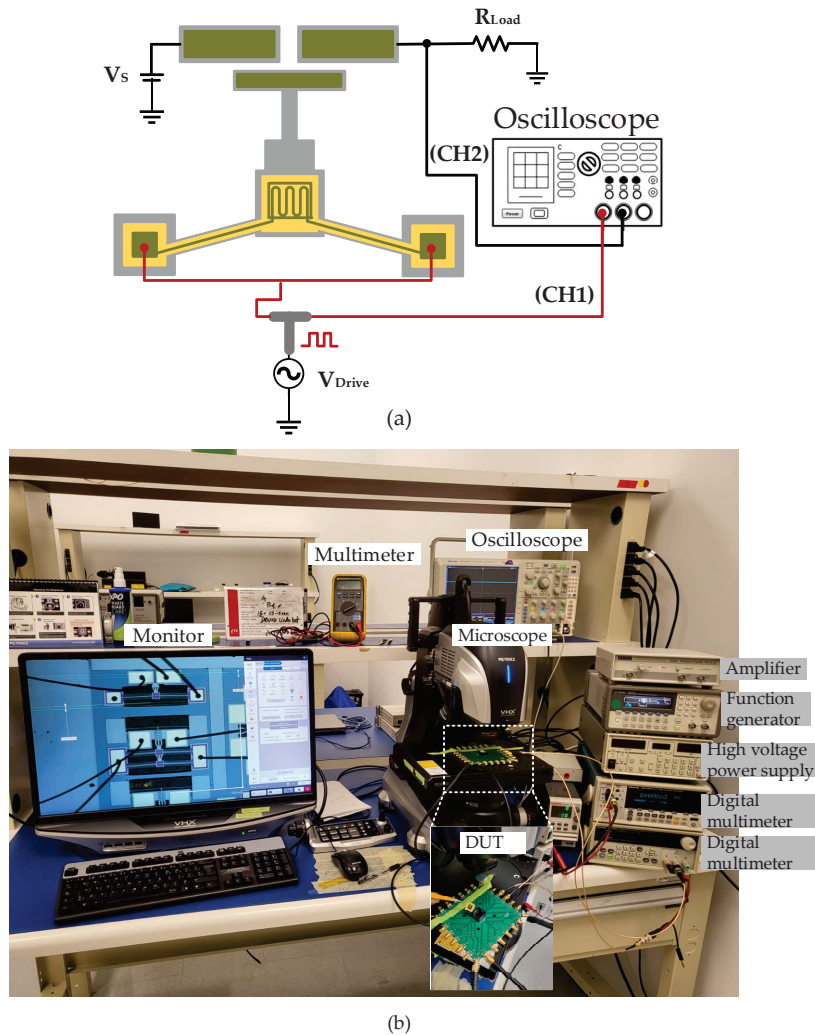


Figure 7. Switching speed test measurement setup: (a) schematic and (b) photograph.

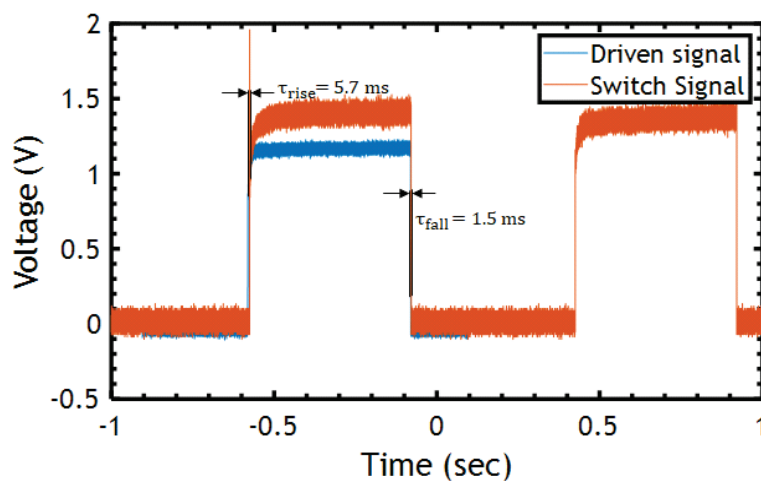


Figure 8. Measured switching speed of the actuator at a maximum range of $2.52 \mu\text{m}$.

4.4. Breakdown Voltage Test

Using a high-voltage source manufactured by Stanford Research Systems (model PS310), the switch was kept in an open state and a high voltage was applied to one of the transmission terminals while the other was grounded, as given in the setup depicted in Figure 9a. The voltage required to break down the structure was visually monitored on the microscope and found to be 350 V, and it is limited by the $2.52\ \mu\text{m}$ air gap between the shuttle and the switch transmission line, which results in arcing, as shown in Figure 9b. The stiffness of the chevron actuator also plays a significant role by preventing self-actuation due to electrostatic forces. This can be enhanced by increasing the gap size at the cost of the larger actuation displacement required to close the switch and, consequently, more actuation power.

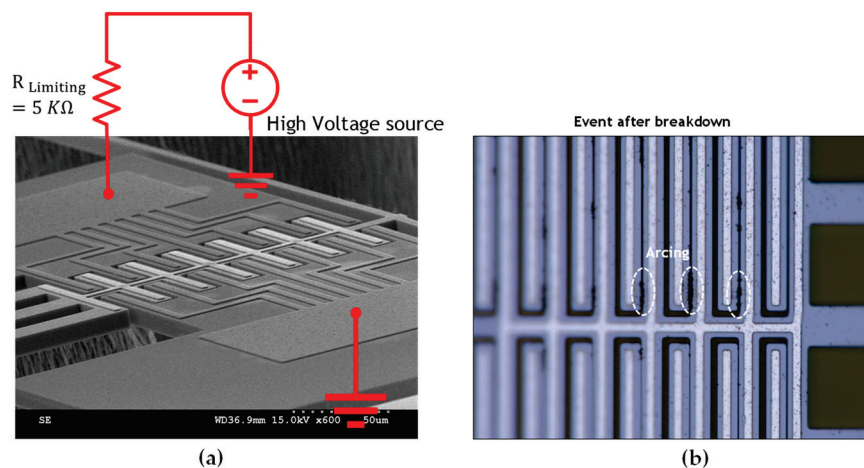


Figure 9. Breakdown voltage measurement: (a) test schematic, (b) microscope image after breakdown event.

5. Discussion

The results presented in the previous section demonstrated that driving the chevron thermal actuator through Joule heating, with a stacked Al layer as the resistive heating element for DC MEMS switching applications, requires an insulating layer between the resistive heating and the transmission line. However, this approach can potentially cause unwanted out-of-plane displacement due to the stress of both insulating and metal layers, which is a trade-off for achieving the desired in-plane motion crucial to this research work. In this study, the effect of the out-of-plane deformation was mitigated by the parallel connection of several narrow Si beams with the stacked ($\text{SiO}_2/\text{Al-Cr}$) chevron beams. The simulated out-of-plane displacement was found to be 100 nm. Additionally, reducing the width of the narrow beams to $6\ \mu\text{m}$ significantly mitigates the temperature gradient across the actuator's shuttle, as compared to the local temperature of the actuator beams. The lateral stiffness of the actuator is also mitigated due to the presence of the narrow beams. Furthermore, the stress across the switching contact is mitigated, as no Al layer is deposited on the Si-to-Si contact. Notably, the initial contact resistance, as depicted in Figure 5, showed an improvement over that in our previously reported study [30], where all the pairs of the chevron beams were patterned with a $\text{SiO}_2/\text{Al-Cr}$ stack. The thermal actuator closes a measured switch gap of $2.5\ \mu\text{m}$, with resistive heating at 33 mA/2.4 V, as depicted in Figure 10. Despite this, infinite electrical contact resistance (ECR) was observed at this initial setting. Consequently, increasing the heater bias voltage by an additional 0.6 V resulted in achieving a contact resistance of approximately $100\ \text{M}\Omega$. Furthermore, as the resistive heating current of the thermal actuator was increased to 42 mA/3.9 V, the contact resistance continued to decrease, ultimately stabilizing at an average value of $80\ \text{k}\Omega$. This reveals that higher levels of resistive heating significantly improve the contact resistance performance of the thermal actuator. It is noteworthy that the doping

surface of the 10 μm -thick Si layer is approximately 1–2 μm deep only as mentioned in the process handbook, and the actuator characterization revealed an out-of-plane displacement of ~ 0.5 μm [30]. However, operating with an overdrive value of 42 mA/3.9 V, following the main actuation voltage sufficient to close the switch, caused degradation in the Al heater. Similar characteristics were observed in the current study, where an overdrive power exceeding 205 mA/1.2 V resulted in damage to the Al heater with irreversible misalignment. This degradation is attributed to the formation of “hillocks” in the direction of current flow and the development of voids at the grain boundaries, as seen in Figure 11. Over time, this degradation led to reduced current flow until heater failure.

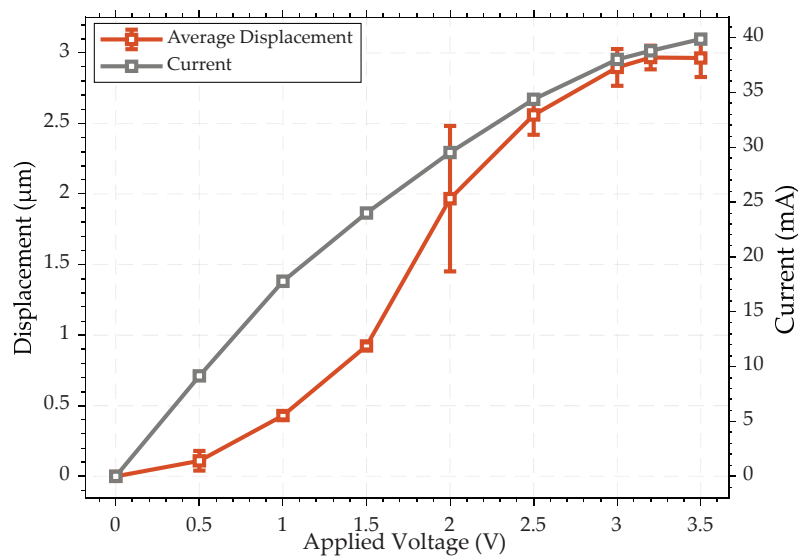


Figure 10. Displacement and current as a function of applied voltage for the previous actuator reported in [30].

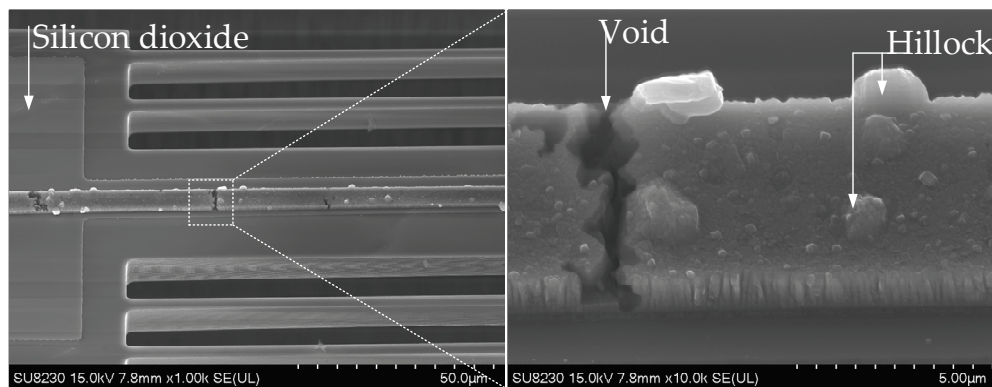


Figure 11. Degradation of aluminum heater over time.

In addition to observing these alterations to the aluminum heater, other factors can impact the ability to maintain a consistent and stable contact resistance. As described in the next subsections, these factors include the sidewall contour of the MEMS switch, the contact force exerted at the interface, and the rise in the contact interface temperature resulting from localized heating by the Al heater. The sidewall contour can influence the mechanical alignment and contact quality, while the contact force affects the physical engagement between contact surfaces, leading to a reduced contact area. Additionally, the localized heating can lead to thermal expansion and variations in material properties, further complicating the stability of the contact resistance. The management techniques of these combined effects must be carefully elaborated to ensure the reliable and consistent performance of the MEMS switch, which is beyond the scope of this work.

5.1. Sidewall Topography Characterization

The sidewall surface topography of the contacting surfaces was examined through SEM imaging analysis of the fabricated MEMS switches. These studies revealed that the scalloping outlines and asperity sizes of the sidewall surfaces varied between different fabrication runs, which could potentially limit the repeatability of the ohmic contact. Figure 12a presents an SEM image depicting the cross-sectional profile of the residual etching angle and the sidewall contours of the contact interface of the MEMS switches. The etch angles at the contact interface were measured to be approximately 1.92° . This etching angle resulted in a 608 nm air gap between the centers of the switching contacts upon initial contact. To further evaluate and validate the accuracy of the procedure used to extract the etching angle profile shown in Figure 12a, the MEMS die was mechanically flipped to observe the differences between the top edge-to-edge distance and the bottom edge-to-edge distance of the silicon structure. Figure 12b–d depicted a 690 nm residual etching after closure, which aligns well with the contact analysis. However, these factors could lead to inconsistencies in achieving well-aligned doped Si-to-Si contacts. The scalloping of the silicon sidewall is unavoidable due to the repeated alternation between DRIE etching and passivation cycles.

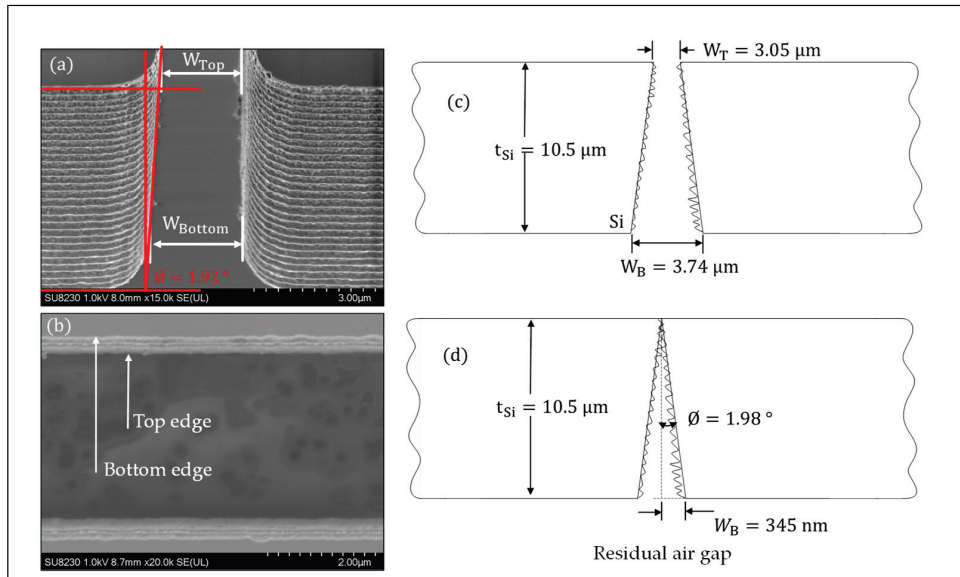


Figure 12. Switch contact interface: (a) cross-sectional view, (b) bottom view of the device, (c) sketched analysis of contact interface before activation, and (d) after activation.

5.2. Manual Activation of the MEMS Switches with a Probe Needle

The main objective of this experiment was to confirm the impact of high contact force and elevated temperature across the contact tip, resulting from Joule heating, on the value and the repeatability of the ohmic resistance of the MEMS switch. The on-state resistance was characterized using a Source Meter Unit (SMU) to obtain the I - V plot in Figure 13. The switches were manually triggered using a probe needle until a reasonable resistance was detected despite the less-controlled force applied with the needle. Once the switch closed, a voltage sweep was applied with a testing current (I) limited to 1, 5, 10, and 15 mA. The I - V plot in Figure 13 shows that the calculated resistance value (R_{ON}) ranges from $150\ \Omega$ to $410\ \Omega$, with a 5 min interval between each testing current. This method allowed for the evaluation of the switch's performance by measuring the contact resistance at room temperature.

As was observed in the previous graphs, despite the low Si-to-Si contact resistance obtained, for testing currents higher than 5 mA, the I - V relation was found to follow a less linear pattern. This could be due to the temperature generated by the higher testing currents. To better investigate this behavior, an additional experiment was conducted on a

bare silicon beam with metal pads at its two ends, as shown in Figure 14a. The resistance of the beam was measured to be $116\ \Omega$. This experiment is essential to understanding the behavior of the barrier between a uniformly surface-doped Si and a top metal (Al) pad, which are the major parts of our proposed Si-to-Si contact switch. The reliability and durability of the MEMS switches will significantly be influenced by the integrity of this junction, which depends on the type of metal used, the doping type and concentration of the Si, and the annealing process after metal deposition [35].

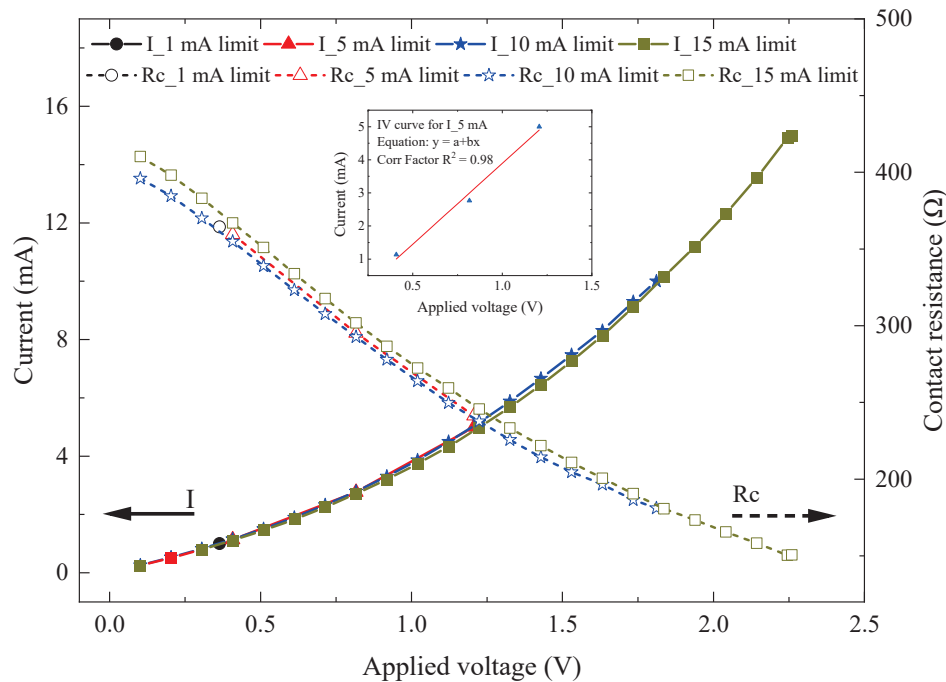


Figure 13. I - V plot of the switching contact and calculated contact on-resistance (R_{ON}), with an inset picture showing the linear I - V curve for the 5 mA limit current.

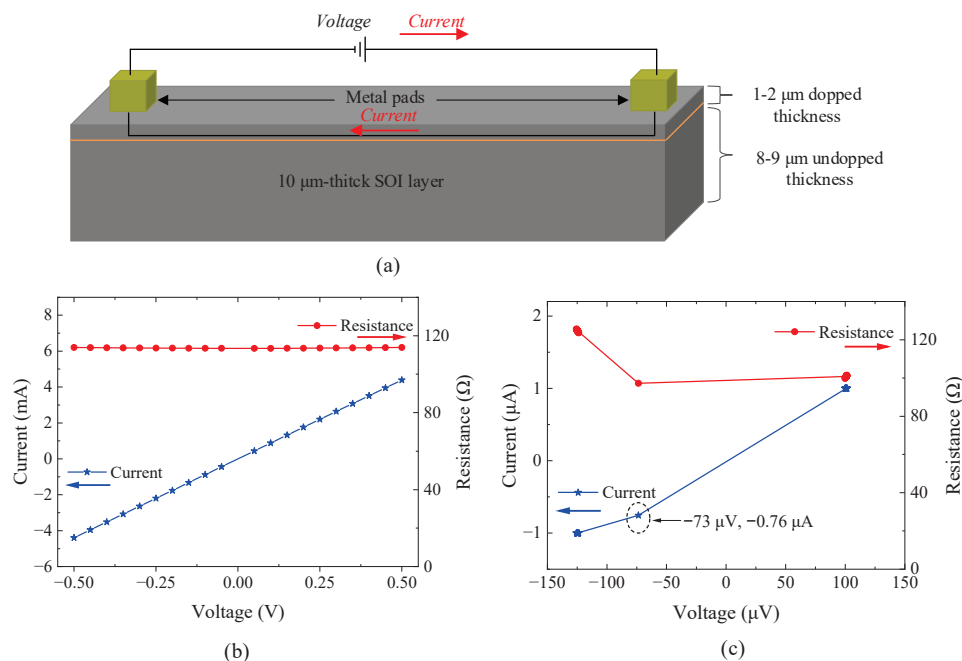


Figure 14. Silicon-Al contact: (a) schematic diagram of the bare silicon beam with an Al pad metal, (b) I - V plot in the range of -0.5 V to 0.5 V , and (c) I - V plot in the range of $-150\ \mu\text{V}$ to $150\ \mu\text{V}$.

As illustrated by the I - V characteristic curve shown in Figure 14b, a voltage ranging from -0.5 to 0.5 V was applied on the Al pads to allow current through the Si beam and observe any possible Schottky barrier between the Si layer and the Al layer. As seen in Figure 14b, the current exhibited a linear relationship with the voltage, resulting in a constant resistance value. This linear response is expected, as surface doping of the Si layer for the PiezoMUMPs process (formerly SOIMUMPs) is very high (10^{20} cm $^{-3}$) [36]. With such a significant surface doping concentration, carriers possess sufficient energy to efficiently tunnel through the inherent barrier between metals and semiconductors [37] and make the barrier behave like a ohmic one. However, applying a very low voltage in the range of -150 to 150 μ V resulted in less ohmic contact where a nearby Schottky barrier was observed, as seen in Figure 14c. Nonetheless, with special annealing treatments and the careful selection of the metal, the height barrier responsible for the non-ohmic behavior could be reduced to near zero, and more ohmic behavior could be obtained [38].

Given the linear I - V curve observed in Figure 14b, the non-linear behavior seen in previous graphs is expected to stem largely from the silicon-to-silicon (Si-to-Si) contact when the switch is closed. This can be attributed to several factors. First, even though the silicon layer is heavily surface-doped, as shown in Figures 12 and 14a, ensuring good and continuous contact between the 1 – 2 μ m-deep doped regions of the two contacting edges is challenging. Etching processes can disturb the doping concentration near the contact edges. Moreover, the doping concentration is maximal at the top surface and gradually decreases toward the bottom surface of the Si layer. Therefore, as continuous force is applied to the switch to achieve good contact in the doped regions, the surface contact area of the less-doped regions also increases, potentially causing the non-linear response.

Second, residual etching materials used to pattern the silicon contact could form a thin sidewall coating, generating a barrier between the Si-to-Si contact when the switch is closed. Third, there is a possibility of the natural formation of a thin oxide layer on the edges of the Si contact, which could also act as a barrier [38]. The results and the foregoing analysis provide insights into necessary design optimizations for enhancing the contact behavior and, consequently, the longevity and repeatability of MEMS switches, particularly in applications requiring precise and stable electrical contact.

5.3. Comparison with State-of-the-Art MEMS Switches

To position this work within the context of state-of-the-art silicon-to-silicon (Si-to-Si) contact MEMS switches, Table 2 compares the proposed thermally actuated DC MEMS switch with previously reported devices, focusing specifically on Si-to-Si contact resistance. Among these, the switch presented in this work demonstrates a significant reduction in contact resistance, achieving values in the range of 150 Ω to 450 Ω , along with a satisfactory breakdown voltage. Additionally, despite the absence of heater optimization, which resulted in a shortened switch lifetime, the design maintains a reasonable electrical isolation between the heater and the signal line. Note that the high aspect ratio introduces a trade-off in the form of out-of-plane displacement.

Table 2. State-of-the-art MEMS switches compared to the proposed Si-to-Si MEMS switch.

Ref.	Actuation	Material	Motion	Actuation Voltage	R_{ON}	$V_{Breakdown}$
[26]	Electrostatic	Si-to-Si	Lateral		~ 35 k Ω	-
[29]	Electrostatic	Si-to-Si	In-plane	~ 18 V	No data	-
[39]	Electrostatic	Tungsten (W)	Out-of-plane	11.3 V	~ 7.5 k Ω	-
[40]	Electrostatic	Platinum (Pt)	In-plane	20 V	4.4 -to- 4.6 k Ω	-
This work	Electrothermal	Si-to-Si	In-plane	1.2 V	0.15 k Ω	350 V

6. Conclusions and Perspective

This study presented the design, fabrication, and characterization of an electro-thermally actuated Si-to-Si DC MEMS switch for power switching applications. The experimental results showed that upon closure, the switch exhibited an initial contact

resistance of approximately 400 Ω and an average response time of 5.7 ms. The electrical contact resistance displayed non-linear behavior for currents ranging from 1 to 15 mA due to self-heating within the switch, while lower currents had negligible effects. These findings highlight the necessity of optimizing both mechanical force and thermal management to achieve consistent and reliable switch operation.

Detailed examination revealed that the etching angles of the sidewall contact rendered the normal contact force insufficient to adequately compress and polish the scallops and asperities of the sidewall, affecting contact resistance reliability and repeatability. Increasing the compressive load led to out-of-plane misalignment, affecting the contact resistance, due to doping confinement to the surface of the SOI device layer.

The impact of contact temperature on contact resistance was further investigated through a near-zero-temperature actuation test using manual contact via a micro-needle actuation. A linear I – V response was observed for currents below 5 mA, whereas higher currents induced non-linear responses due to Joule heating.

Importantly, the switching contacts sustained a maximum breakdown voltage of 350 V, owing to the actuator's thermoelectric nature. The switch's ability to withstand up to 350 V, coupled with low contact resistance, underscores its potential for demanding applications such as power systems and fault protection networks. Compared to other state-of-the-art MEMS switches, the proposed design offers good performance and a competitive breakdown voltage, despite challenges related to the heater lifetime.

Future work will focus on reducing the effects of Joule heating at the transmission line and optimizing the heater design to extend the switch's lifespan.

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References

1. Pal, J.; Zhu, Y.; Lu, J.; Dao, D.; Khan, F. High Power and Reliable SPST/SP3T RF MEMS Switches for Wireless Applications. *IEEE Electron Device Lett.* **2016**, *37*, 1219–1222. [CrossRef]
2. Keimel, C.; Claydon, G.; Li, B.; Park, J.N.; Valdes, M.E. Microelectromechanical-Systems-Based Switches for Power Applications. *IEEE Trans. Ind. Appl.* **2012**, *48*, 1163–1169. [CrossRef]
3. Pothier, A.; Hitier, S.; El khatib, M.; Blondy, P.; Orlianges, J.C.; Champeaux, C.; Catherinot, A.; Vendier, O.; Cazaux, J.L. MEMS DC contact micro relays on ceramic substrate for space communication switching network. In Proceedings of the 2005 European Microwave Conference, Paris, France, 4–6 October 2005; Volume 1, p. 4. [CrossRef]
4. Nguyen, V.H.; Aimaier, N.; Nobert, G.; Pham, T.; Constantin, N.; Blaquiére, Y.; Cowan, G. A Reconfigurable Power System-in-Package Module using GaN HEMTs and IC Bare Dies on LTCC Substrate: Design—Implementation—Experiment and Future Directions. In Proceedings of the 2022 20th IEEE Interregional NEWCAS Conference (NEWCAS), Quebec City, QC, Canada, 19–22 June 2022; IEEE: Piscataway, NJ, USA, 2022. [CrossRef]
5. Maciel, J.; Majumder, S.; Lampen, J.; Guthy, C. Rugged and reliable ohmic MEMS switches. In Proceedings of the 2012 IEEE/MTT-S International Microwave Symposium Digest, Montreal, QC, Canada, 17–22 June 2012; pp. 1–3. [CrossRef]
6. Rebeiz, G.M. *RF MEMS*; John Wiley & Sons, Inc.: Hoboken, NJ, USA, 2003. [CrossRef]
7. Song, Y.-H.; Han, C.-H.; Kim, M.-W.; Lee, J.O.; Yoon, J.-B. An Electrostatically Actuated Stacked-Electrode MEMS Relay with a Levering and Torsional Spring for Power Applications. *J. Microelectromech. Syst.* **2012**, *21*, 1209–1217. [CrossRef]
8. Bakri-Kassem, M.; Mansour, R.R. High Power Latching RF MEMS Switches. *IEEE Trans. Microw. Theory Tech.* **2015**, *63*, 222–232. [CrossRef]

9. Pal, J.; Zhu, Y.; Lu, J.; Dao, D.V.; Khan, F. RF MEMS switches for smart antennas. *Microsyst. Technol.* **2015**, *21*, 487–495. [CrossRef]
10. Stefanini, R.; Chatras, M.; Blondy, P.; Rebeiz, G.M. Miniature MEMS Switches for RF Applications. *J. Microelectromech. Syst.* **2011**, *20*, 1324–1335. [CrossRef]
11. Uvarov, I.; Kupriyanov, A. Stiction-protected MEMS switch with low actuation voltage. *Microsyst. Technol.* **2019**, *25*, 3243–3251. [CrossRef]
12. Czaplewski, D.A.; Nordquist, C.D.; Dyck, C.W.; Patrizi, G.A.; Kraus, G.M.; Cowan, W.D. Lifetime limitations of ohmic, contacting RF MEMS switches with Au, Pt and Ir contact materials due to accumulation of ‘friction polymer’ on the contacts. *J. Micromech. Microeng.* **2012**, *22*, 105005. [CrossRef]
13. Newman, H.S.; Ebel, J.L.; Judy, D.; Maciel, J. Lifetime Measurements on a High-Reliability RF-MEMS Contact Switch. *IEEE Microw. Wirel. Compon. Lett.* **2008**, *18*, 100–102. [CrossRef]
14. Ma, B.; You, Z.; Ruan, Y.; Chang, S.; Zhang, G. Electrostatically actuated MEMS relay arrays for high-power applications. *Microsyst. Technol.* **2016**, *22*, 911–920. [CrossRef]
15. Zhang, Y.; Gong, Z.; Liu, H.; Liu, Z. Power Handling Capability Enhanced RF MEMS Switch Using Modified-Width Cantilevers Structure. In Proceedings of the 2021 IEEE 16th International Conference on Nano/Micro Engineered and Molecular Systems (NEMS), Xiamen, China, 25–29 April 2021; pp. 695–698. [CrossRef]
16. Iannacci, J. *RF-MEMS Technology for High-Performance Passives: The Challenge of 5G Mobile Applications*; IOP Publishing: Bristol, UK, 2017.
17. Sharma, S.; Nabavi, S.; Rabih, A.A.S.; Ménard, M.; Nabki, F. Hybrid MEMS Actuator With 3 Degrees-of-Freedom for Efficient Planar Optical Switching. *J. Microelectromech. Syst.* **2023**, *32*, 593–603. [CrossRef]
18. Rabih, A.A.S.; Nabavi, S.; Ménard, M.; Nabki, F. A 3 Degrees-of-Freedom Electrothermal Micro-Positioner for Optical Chip-to-Chip Alignment. *J. Microelectromech. Syst.* **2024**, *33*, 260–273. [CrossRef]
19. Rajagopalan, J. Microelectromechanical Systems (MEMS)-Based Testing of Materials. In *Handbook of Mechanics of Materials*; Schmauder, S., Chen, C.-S., Chawla, K.K., Chawla, N., Chen, W., Kagawa, Y., Eds.; Springer: Singapore, 2019; pp. 1955–1979.
20. Agrawal, V. A latching MEMS relay for DC and RF applications. In Proceedings of the 50th IEEE Holm Conference on Electrical Contacts and the 22nd International Conference on Electrical Contacts, 2004, Seattle, WA, USA, 23 September 2004; pp. 222–225. [CrossRef]
21. Shojaei-Asanjan, D.; Bakri-Kassem, M.; Mansour, R.R. Analysis of Thermally Actuated RF-MEMS Switches for Power Limiter Applications. *J. Microelectromech. Syst.* **2019**, *28*, 107–113. [CrossRef]
22. Qiu, J.; Lang, J.; Schmidt, M. An Electrothermally-Actuated Bistable MEMS Relay for Power Applications. Ph.D. Thesis, Massachusetts Institute of Technology, Cambridge, MA, USA, 2003.
23. Daneshmand, M.; Fouladi, S.; Mansour, R.R.; Lisi, M.; Stajcer, T. Thermally-actuated latching RF MEMS switch. In Proceedings of the 2009 IEEE MTT-S International Microwave Symposium Digest, Boston, MA, USA, 7–12 June 2009; pp. 1217–1220. [CrossRef]
24. Mahameed, R.; Rebeiz, G.M. A High-Power Temperature-Stable Electrostatic RF MEMS Capacitive Switch Based on a Thermal Buckle-Beam Design. *J. Microelectromech. Syst.* **2010**, *19*, 816–826. [CrossRef]
25. Wang, L.; Jin, Y. A push-pull double-contact MEMS relay fabricated by MetalMUMPs process. *Microsyst. Technol.* **2017**, *23*, 2257–2262. [CrossRef]
26. Soon, B.W.; Qian, Y.; Ng, E.J.; Hong, V.A.; Yang, Y.; Ahn, C.H.; Kenny, T.W.; Lee, C. Investigation of a Vacuum Encapsulated Si-to-Si Contact Microswitch Operated From -60°C to 400°C . *J. Microelectromech. Syst.* **2015**, *24*, 1906–1915. [CrossRef]
27. Zhu, Y.; Pal, J. Low-Voltage and High-Reliability RF MEMS Switch with Combined Electrothermal and Electrostatic Actuation. *Micromachines* **2021**, *12*, 1237. [CrossRef]
28. Meszmer, P.; Hahn, S.; Hiller, K.; Wunderle, B. Highly Miniaturized MEMS-Based Test Platforms for Thermo-Mechanical and Reliability Characterization of Nano-Functional Elements – Technology and Functional Test Results. *Phys. Status Solidi (A)* **2019**, *216*, 1800936. [CrossRef]
29. Bengashier, M.; Grech, I.; Casha, O.; Portelli, B.; Farrugia, R. Mechanical Contact Type RF-MEMS Switches for Microwave Band Applications. In Proceedings of the 2023 Symposium on Design, Test, Integration & Packaging of MEMS/MOEMS (DTIP), Valetta, Malta, 28–31 May 2023; pp. 1–4. [CrossRef]
30. Shuaibu, A.H.; Nabki, F.; Blaquièrre, Y. A MEMS Electrothermal Actuator Designed for a DC Switch Aimed at Power Switching Applications and High Voltage Resilience. In Proceedings of the 2022 20th IEEE Interregional NEWCAS Conference (NEWCAS), Quebec City, QC, Canada, 19–22 June 2022; pp. 317–321. [CrossRef]
31. Cowen, A.; Hames, G.; Glukh, K.; Hardy, B. *PiezoMUMPs™ Design Handbook*; MEMSCAP Inc.: Durham, NC, USA, 2014; p. 28.
32. Coutu, R.A.; LaFleur, R.S.; Walton, J.P.K.; Starman, L.A. Thermal Management Using MEMS Bimorph Cantilever Beams. *Exp. Mech.* **2016**, *56*, 1293–1303. [CrossRef]
33. Available online: <https://aluminium-guide.com/melting-point-aluminium/> (accessed on 24 June 2020).
34. Science Corporation. Available online: <https://science.xyz/docs/d/mems-piezo/index> (accessed on 3 October 2023).
35. Benouattas, N.; Tamaarat, B.; Bouabellou, A.; Halimi, R.; Mosser, A. Electrical properties of Cr/Si(p) structures. *Solid-State Electron.* **1999**, *43*, 439–446. [CrossRef]
36. Ta, Q. Local Synthesis and Direct Integration of Carbon Nanotubes into Microsystems for Sensor Applications. Ph.D. Thesis, Buskerud and Vestfold University College, Drammen, Norway, 2014.
37. Available online: <https://cleanroom.byu.edu/ohmic-schottky> (accessed on 9 February 2014).

38. Card, H.C. Aluminum—Silicon Schottky barriers and ohmic contacts in integrated circuits. *IEEE Trans. Electron Devices* **1976**, *23*, 538–544. [CrossRef]
39. Osoba, B.; Almeida, S.F.; Sikder, U.; Ye, Z.A.; Hu, X.; Esatu, T.K.; Liu, T.-J.K. Study of MEM Relay Contact Design and Body-Bias Effects on on-State Resistance Stability. *J. Microelectromech. Syst.* **2020**, *29*, 1531–1536. [CrossRef]
40. Grogg, D.; Drechsler, U.; Knoll, A.; Duerig, U.; Pu, Y.; Hagleitner, C.; Despont, M. Curved in-plane electromechanical relay for low power logic applications. *J. Micromech. Microeng.* **2013**, *23*, 025024. [CrossRef]

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Article

Strain-Induced Frequency Splitting in PT Symmetric Coupled Silicon Resonators

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Abstract: When two resonators of coupled silicon resonators are identical and the gain on one side is equal to the loss on the other side, a parity-time (PT) symmetric-coupled silicon resonator is formed. As non-Hermitian systems, the PT-symmetric systems have exhibited many special properties and interesting phenomena. This paper proposes the strain-induced frequency splitting in PT symmetry-coupled silicon resonators. The frequency splitting of the PT system caused by strain perturbations is derived and simulated. Theory and simulation both indicate that the PT system is more sensitive to strain perturbation near the exceptional point (EP) point. Then, a feedback circuit is designed to achieve the negative damping required for PT symmetry. Based on a simple silicon-on-insulator (SOI) process, the silicon resonator chip is successfully fabricated. After that, the PT-symmetric-coupled silicon resonators are successfully constructed, and the frequency splitting phenomenon caused by strain is observed experimentally.

Keywords: PT symmetry; coupled resonators; silicon resonators; strain-induced; frequency splitting

1. Introduction

In 1998, Professor C. M. Bender of University of Washington proposed a parity-time (PT)-symmetric Hamiltonian. This PT-symmetric Hamiltonian does not have Hermitian properties, but still has a real eigenvalue spectrum [1]. Here, P and T represent parity transformation and time transformation, respectively. According to Professor Bender's theory, the condition for Hamiltonian H to have PT symmetry is that the inner product of the PT transformation operator and the H operator is 0, denoted as $[PT, H] = 0$. It can also be understood as the condition that a system's Hamiltonian remains unchanged after undergoing parity transformation (coordinate x transformation to $-x$) and time transformation (time t transformation to $-t$). After Professor Bender proposed the concept of PT symmetry, many researchers quickly joined the study on this non-Hermitian PT symmetry system. So far, the theoretical framework of PT-symmetric non-Hermitian quantum systems has been established [2,3].

Compared to Hermitian symmetry in mathematical form, PT symmetry is closer to physical symmetry. Therefore, in addition to research in quantum fields, PT symmetry theory has also been validated in different types of classical physical systems. In 2007, R. El Ganainy et al. from the University of Central Florida first proposed an optical PT-symmetric structure and formulated a coupled-mode theory (CMT) appropriate for PT-symmetric optical elements [4]. In 2008, Technion-Israel Institute of Technology's S. Klaiman et al. first observed the phase transition process of PT-symmetric optical structures from exact PT symmetry to broken PT symmetry and identified the PT symmetry exceptional point (EP) [5]. In 2011, J. Schindler from Wesleyan University first proposed a PT-symmetric electrical system based on a coupled LCR resonant circuit. The theoretical model for the electrical PT system was established and experimentally verified [6]. In 2014, X. Zhu et al. from the University of California introduced the concept of acoustic PT symmetry. Combining this acoustic PT-symmetric medium with transformation acoustics, two-dimensional

symmetric acoustic cloaks that were unidirectionally invisible in a prescribed direction were designed [7]. In 2016, R. Fleury et al. from the University of Texas systematically expounded the theoretical model and implementation method of acoustic PT-symmetric system [8].

Gradually, some special properties and interesting phenomena of PT symmetry systems have been discovered [9–14]. Z. Lin et al. from Wesleyan University found that PT-symmetric optical periodic structures exhibit unidirectional invisibility properties. Near the PT breaking point, if the light wave enters from the left side of the PT structure, it will exit from the right side completely unaffected; and if the light wave enters from the right side, its reflected wave will be enhanced [9]. Wesleyan University's H. Ramezani et al. discovered the Talbot effect in PT-symmetric optical systems. When the input field pattern respects specific periodicities, the complex PT-symmetric photonic lattices can lead to a new class of self-imaging Talbot effects [10]. University of Central Florida's H. Hodaei et al. developed a micro-resonant ring laser using the PT symmetry principle. By harnessing notions from PT symmetry, stable single-longitudinal mode operation can be readily achieved in a system of coupled micro-ring lasers. The selective breaking of PT symmetry can be used to systematically enhance the maximum attainable output power in the desired mode [12]. S. Assawaworrarit et al. from Stanford University proposed theoretically and demonstrated experimentally that a parity-time symmetric circuit incorporating a nonlinear gain saturation element provides robust wireless power transfer. The transfer efficiency can remain near unity over a distance variation of approximately one meter, without the need for any tuning [14]. Wayne State University's P. Chen et al. proposed the PTX symmetric theory and significantly improved the readout signal strength of LC sensors using this theory [15]. B. Zhou et al. from Southeast University constructed a PT-symmetric LC sensing system working in the broken PT regime. Experiments showed that its wireless readout distance was nearly four times longer than traditional methods [16]. Recently, a type of artificial material called metamaterial has become a research hotspot due to its unusual electromagnetic properties [17,18]. P. R. Charumathi et al. revealed that the PT metamaterials exhibit several unusual light-matter interactions that could be exploited for realizing novel applications [19].

As a special point in PT phase transition, the EP point exhibits some outstanding characteristics. Z. Liu et al. from Tsinghua University showed that it is possible to realize ultrasensitive optomechanical PT-symmetric microcavities whose sensitivity is at least two orders of magnitude better than single-cavity optomechanical transducers [20]; Tongji University's C. Zeng et al. reported the study of high-order EP in PT-symmetric electrical systems, and found that the sensitivity of the system was enhanced by nearly 10 times when the perturbation frequency was less than 0.4 kHz [21]. H. Hodaei et al. from the University of Central Florida found that the higher the order of EP in PT-symmetric micro-resonant rings, the higher the sensitivity of their perturbation response [22]. Z. Dong et al. from the National University of Singapore used an EP-locked PT-symmetric circuit to increase the sensitivity of implantable passive wireless sensors by 3.2 times [23]. The University of Texas at Austin's Z. Xiao et al. designed PT-symmetric circuits with a sixth order EP, which can be used for readout of capacitive sensors and has the advantages of high sensitivity, high resolution, and low thermal noise [24].

Our team successfully constructed PT systems based on coupled LC resonators and coupled silicon resonators, and established corresponding theoretical models [16,25,26]. Based on our previous work, this paper studies the strain-induced frequency splitting in PT symmetry-coupled silicon resonators. Theoretical derivation, simulation analysis and experimental measurements will be conducted to study the strain perturbation on the PT system. Due to the particularity of the EP, our observation will start from the EP point of the PT system. The main content of this paper is organized as follows: Section 2 derives the theory of PT symmetry enhanced strain sensing; Section 3 designs the structure of the PT-symmetric-coupled silicon resonators and simulates its mechanical and circuit model;

Section 4 shows the fabrication process of the silicon resonators, and gives the experimental results; finally, Section 5 makes some conclusions.

2. Principle

Figure 1 shows the schematic of the PT-symmetric-coupled silicon resonators. Resonator 1 and Resonator 2 have the same mass m and elastic coefficient k , and the coupling coefficient between the two resonators is k_c . Different from the traditional coupled silicon resonators which includes two loss resonators [27,28], the proposed coupled silicon resonators here consist of one loss resonator and one gain resonator. When the positive damping of Resonator 2 and the negative damping of Resonator 1 are equal in magnitude, the two resonators form a PT-symmetric system.

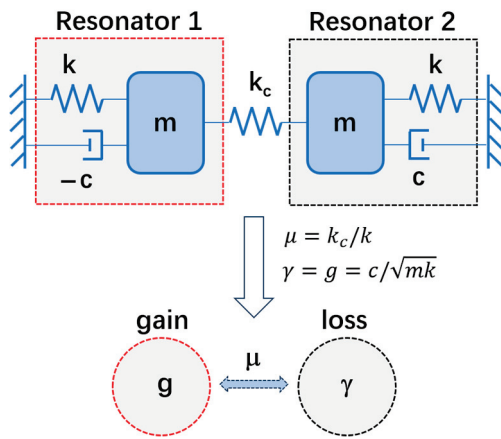


Figure 1. Schematic of the PT-symmetric-coupled silicon resonators.

According to Newton's law, the equations of the motions of the coupled silicon resonators are given as:

$$m \frac{d^2 x_1}{dt^2} - c \frac{dx_1}{dt} + kx_1 + k_c(x_1 - x_2) = 0 \quad (1a)$$

$$m \frac{d^2 x_2}{dt^2} + c \frac{dx_2}{dt} + kx_2 + k_c(x_2 - x_1) = 0 \quad (1b)$$

in which c is the damping coefficient, and x_n ($n = 1, 2$) denotes the vibration displacements of the two resonators, respectively. In Equation (1), swapping the indices of x (P reverse) and changing the sign of t (T reverse) leaves the equations unchanged, which conforms to the definition of PT symmetry. The gain of Resonator 1 and the loss of Resonator 2 can be defined as $g = \gamma = c/\sqrt{mk}$, the coupling strength between the two resonators can be defined as $\mu = k_c/k$, then the PT symmetry coupled silicon resonators can be simplified as a PT-symmetric dimer, as shown in Figure 1.

Taking $x_n(t) \propto x_n e^{i\omega t}$, Equation (1) can be rewritten as:

$$\begin{bmatrix} 1 + \mu - i\omega g - \omega^2 & -\mu \\ -\mu & 1 + \mu + i\omega \gamma - \omega^2 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} = 0 \quad (2)$$

where ω is scaled by $\omega_0 = \sqrt{k/m}$, which is the resonant frequency of the single silicon resonator.

Here, the coupled silicon resonators are set in a weakly coupled state ($\mu \ll 1$). When the system is in weak coupling, Equation (2) can be simplified to:

$$\begin{bmatrix} 1 + \frac{\mu - ig}{2} - \omega & -\frac{\mu}{2} \\ -\frac{\mu}{2} & 1 + \frac{\mu + i\gamma}{2} - \omega \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} = 0 \quad (3)$$

Solving Equation (3), the eigenfrequencies of the PT system can be obtained as:

$$\omega_{\pm} = 1 + \frac{\mu}{2} \pm \frac{1}{2} \sqrt{\mu^2 - \gamma^2} \quad (4)$$

In Equation (4), when $\mu > \gamma$, the system is in an exact PT-symmetric phase. The system has two real eigenfrequencies and the frequency splitting $\Delta\omega$ is:

$$\Delta\omega = \sqrt{\mu^2 - \gamma^2} \quad (5)$$

In Equation (4), when $\mu < \gamma$, the system is in a broken PT-symmetric phase. The eigenfrequencies have imaginary parts, and the system has only one real eigenfrequency; between these two regimes, when $\mu = \gamma$, the system is in an EP, where the transition of the two different phases of PT symmetry occurs.

According to the above analysis, the eigenfrequencies merge in the broken PT symmetry regime. In another word, the frequency splitting $\Delta\omega$ of the PT system remains zero at this regime. Therefore, the strain-induced frequency splitting only occurs in the exact PT symmetry regime starting from EP.

When strain is applied to the coupled silicon resonators, both resonators are simultaneously subjected to elastic coefficient perturbations δ . After perturbation, the parameters of the coupled resonator are rewritten as:

$$\mu' = \frac{\mu}{1 - \delta} \quad (6a)$$

$$\gamma' = \frac{\gamma}{\sqrt{1 - \delta}} \quad (6b)$$

$$\omega'_0 = \sqrt{1 - \delta} \omega_0 \quad (6c)$$

Substituting Equation (6) into Equation (3), and setting the initial state of the PT system at EP ($\mu = \gamma$), Equation (5) can be rewritten as:

$$\Delta\omega' = \sqrt{\mu'^2 - \gamma'^2} = \frac{\sqrt{\delta}}{1 - \delta} \mu \quad (7)$$

Normalizing $\Delta\omega'$ by ω_0 , the frequency splitting induced by strain can be obtained as:

$$\Delta\omega = \frac{\sqrt{\delta}}{\sqrt{1 - \delta}} \mu \quad (8)$$

3. Design and Simulation

3.1. Structural Design

Figure 2 shows the structural design of the PT-symmetric-coupled silicon resonators according to Figure 1. Resonator 1 and Resonator 2 are structurally symmetrical, while Resonator 1 is a gain resonator and Resonator 2 is a loss resonator. To create a gain effect on Resonator 1, the electrode V_{sen} detects the vibration of Resonator 1 and applies back to the V_{fb} electrode through a feedback circuit loop. The magnitude of gain can be adjusted by the feedback circuit. The loss of Resonator 2 is determined by its own parameters as well as the ambient air pressure, and its magnitude can be adjusted by the air pressure. The two resonators are coupled through electrical coupling. The coupling strength can be adjusted by the DC voltages V_{d1} and V_{d2} . AC voltage V_{a1} initiates the motion of the resonator. The DC voltage V_{d3} is used to eliminate the asymmetric effects caused by fabrication deviations.

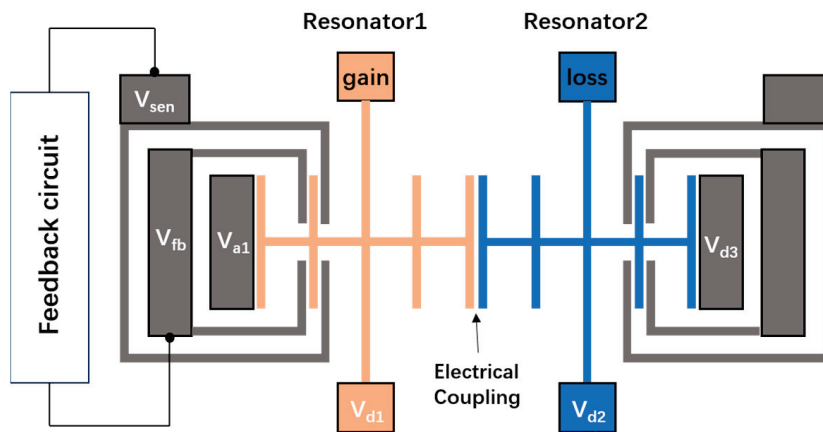


Figure 2. The structural design of the PT-symmetric-coupled silicon resonators.

3.2. Mechanical Model Simulation

The finite element simulation software COMSOL Multiphysics 6.2 was used to simulate and analyze the PT-symmetric-coupled silicon resonators, as shown in Figure 3a. The silicon resonators were fabricated based on silicon-on-insulator (SOI) process. The thickness of the silicon beam was 25 μm , the thickness of the substrate was 400 μm , and the gap between the silicon beam and the substrate was 2 μm . The size of silicon resonators is shown in Table 1.

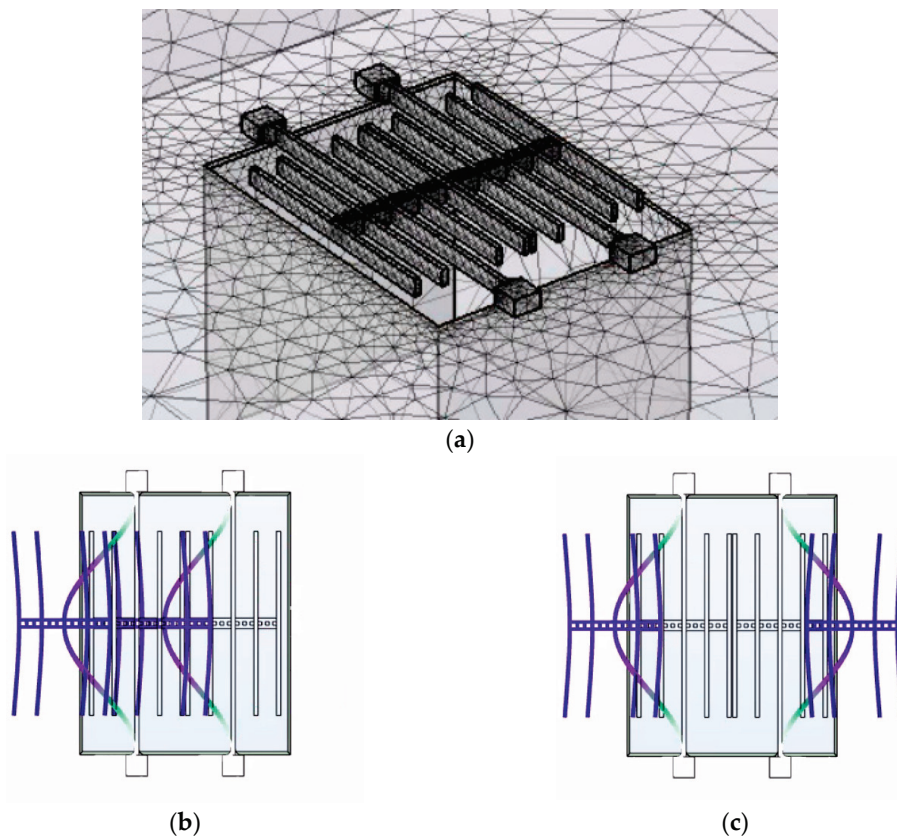


Figure 3. FEM simulation of the PT-symmetric-coupled silicon resonators. (a) simulation model; (b) the in-phase mode of vibration and (c) out-of-phase mode of vibration.

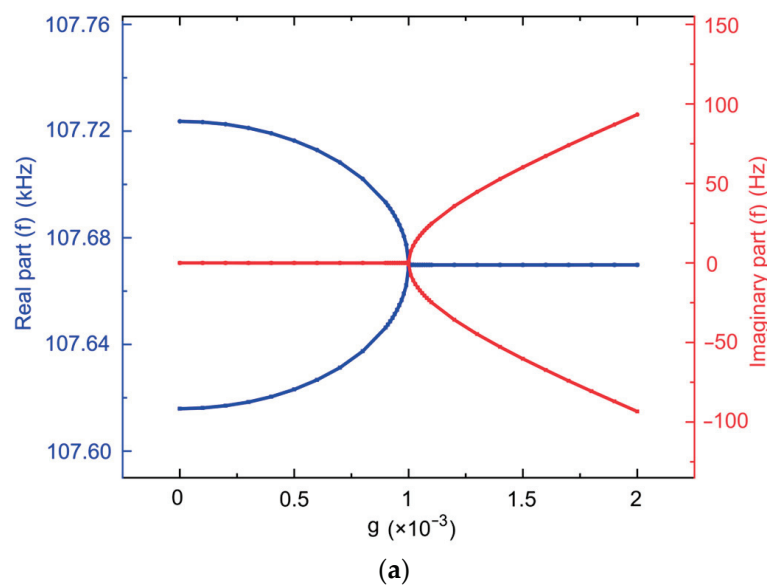
Table 1. The key parameters of the coupled silicon resonators.

Parameter	Value
Length of the resonators	500 μm
Width of the resonators	8 μm
Thickness of the resonators	25 μm
Length of the capacitor plates	260 μm
Width of the capacitor plates	8 μm
Electrical coupling gap	2.5 μm
Young's modulus	165 GPa
Poisson's ratio	0.27
Density	2350 kg/m^3

Figure 3b,c show the simulated two basic vibration modes of the coupled resonators, the in-phase mode (first mode) and the out-of-phase mode (second mode). The in-phase mode corresponds to the lower eigenfrequency of the coupled resonator system, while the out-of-phase mode corresponds to a higher eigenfrequency of the coupled resonator system.

The coupling coefficient μ of the coupled resonators was set to 0.001, while scanning the gain factor g . The variation curve of the eigenfrequencies of the PT-symmetric-coupled silicon resonators is depicted in Figure 4a. When g is larger than 0.001, the system is in the broken PT symmetry regime, where the real part of the eigenfrequencies merges and the imaginary part splits. When g is equal to 0.001, the system is at the phase-transition point EP. When g is less than 0.001, the system is in the exact PT symmetry regime, where the real part of the eigenfrequency splits and the imaginary part returns to zero.

The gain factor g of the coupled resonators was set to 0.001, while scanning the coupling coefficient μ . The variation curve of the eigenfrequencies of the system is shown in Figure 4b. It can be seen that when μ is small the system is in broken PT, and when μ is large, the system is in exact PT. Furthermore, as the coupling coefficient μ increases, the real part of the eigenfrequencies show a downward trend.

**Figure 4.** Cont.

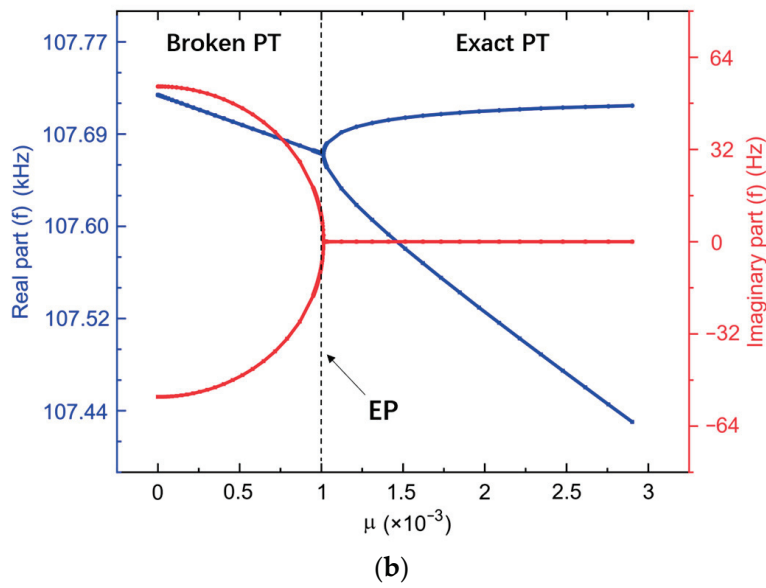


Figure 4. Simulation results of the PT-symmetric-coupled silicon resonators. (a) Keeping $\mu = 0.001$, the variation of the eigenfrequencies versus g ; (b) keeping $g = 0.001$, the variation of the eigenfrequencies versus μ .

After setting the initial state of the PT system at EP, strain perturbations were applied onto the coupled silicon resonators. The simulated frequency splitting of the PT system was obtained and shown in Figure 5. For comparison, the theoretical results are also given in Figure 5. The inset in Figure 5 shows how strain was applied onto the coupled silicon resonators. As can be seen, the simulation results are consistent with the calculation results. Both results show that the PT system has a larger response slope at small applied strains, and as the applied strain increases, the response slope decreases and tends to a constant value.

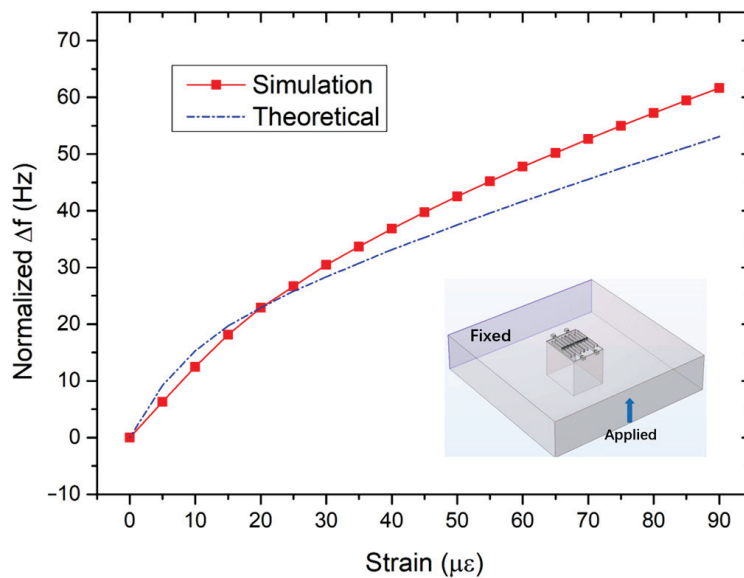


Figure 5. FEM simulated and theoretical calculated strain-induced frequency splitting in the PT system. The inset shows how strain is applied onto the coupled silicon resonators.

3.3. Implementation of Negative Damping

According to PT symmetry theory, Resonator 1 requires a negative damping, which is equal in magnitude to the positive damping of Resonator 2. Here, the negative damping was implemented by a feedback loop.

When applying a feedback force F_{fb} proportional to the velocity of Resonator 1, the equivalent damping coefficient c_{gain} generated by F_{fb} is:

$$F_{fb} = c_{gain}\dot{x} \quad (9)$$

Under the influence of this feedback force, the dynamic equation of Resonator 1 becomes:

$$m\ddot{x} + c\dot{x} + kx = F_{fb} \quad (10)$$

So, due to the feedback force, the equivalent damping coefficient c_{eff} of Resonator 1 is calculated as:

$$c_{eff} = c - c_{gain} \quad (11)$$

Therefore, when the damping coefficient c_{gain} caused by the feedback force F_{fb} is larger than the initial damping c of Resonator 1, it obtains an equivalent negative damping. Adjusting the feedback force applied on Resonator 1, when the negative damping c_{eff} of Resonator 1 and the positive damping c of Resonator 2 satisfy $c_{eff} = -c$, the coupled silicon resonators form PT symmetry.

The feedback force could be generated by the feedback circuit loop. The feedback force adjustment circuit is composed of a transconductance amplifier (TIA), bandpass filter (BPF), voltage-controlled amplifier (VCA), and phase shifter (PS). Firstly, the current signal of the resonator motion obtained from the V_{sen} electrode is converted into a voltage signal through TIA. Then, the voltage signal is filtered by a BPF. After that, the amplitude and phase of the voltage signal are adjusted using a VCA and PS. Finally, the feedback signal is applied back to the resonator through the V_{fb} electrode.

A circuit model of PT-symmetric-coupled silicon resonators was established for the design and verification of the feedback circuit. The circuit model with the feedback circuit is simulated in Multisim 14, as shown in Figure 6. Among them, $R1$, $L1$, and $C1$ are the equivalent circuit model of Resonator 1; $R2$, $L2$, and $C2$ are the equivalent circuit model of Resonator 2; and $C3$ is an equivalent capacitor representing the coupling between the two resonators. According to the mechanical parameters, the equivalent resistance, equivalent inductance, and equivalent capacitance values of a mechanical resonator can be calculated by conversion formulas [29]. Here, for verification of the feedback circuit, we set $R1 = R2 = 28 \text{ M}\Omega$, $C1 = C2 = 0.04 \text{ fF}$, and $L1 = L2 = 56,429 \text{ H}$.

By adjusting the parameters of each module in the feedback loop, the gain Resonator 1 is balanced with the loss Resonator 2, $g = \gamma = 0.01$, and the system achieves PT symmetry. Then, by adjusting the coupling capacitor $C3$, the signal waveforms of the PT system in different regimes can be displayed, as shown in Figure 7. When $C3 = 1.3 \text{ fF}$, the equivalent coupling coefficient $\mu = 0.03$, which is larger than the gain/loss coefficient of 0.01. The PT system enters the exact PT symmetry regime and displays two split resonance peaks; When $C3 = 13 \text{ fF}$, the equivalent coupling coefficient $\mu = 0.003$, which is less than the gain/loss coefficient of 0.01. The PT system enters the broken PT symmetry regime and shows one merged resonance peak. The simulation results of the circuit model verify the effectiveness of the feedback loop. Compared to mechanical model in Section 3.2, the advantage of the circuit model here is that it is convenient for the design and simulation of the feedback loop.

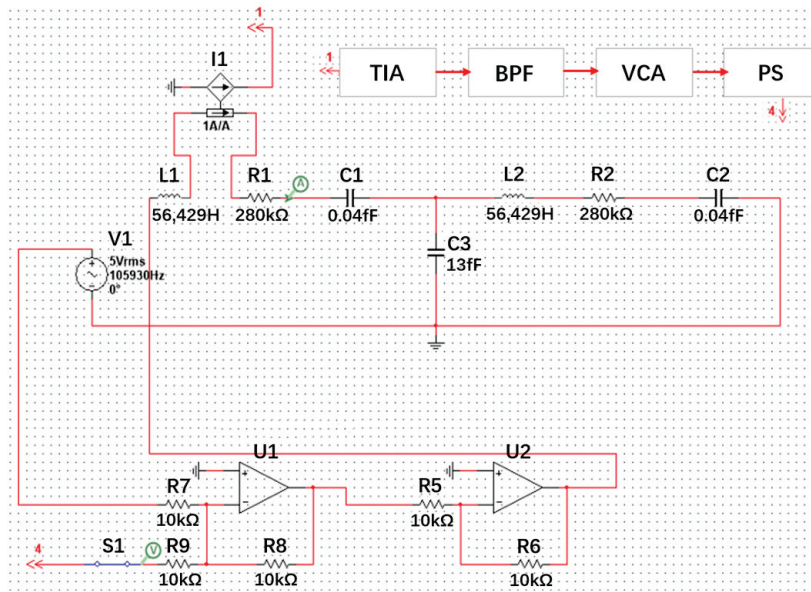


Figure 6. A circuit model of PT-symmetric-coupled silicon resonators for the design and verification of the feedback circuit.

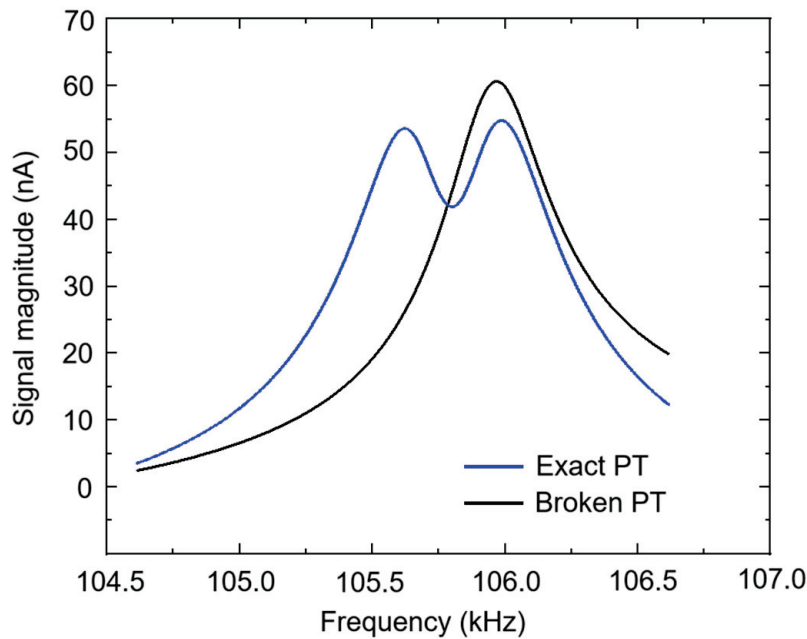


Figure 7. The signal waveforms of the PT system in different regimes, in which the blue line denotes the exact PT symmetry regime when $C3 = 1.3$ fF; the black line denotes the broken PT symmetry regime when $C3 = 13$ fF.

4. Measurements and Results

4.1. Fabrication

The coupled silicon resonators were fabricated based on a simple SOI process. The fabrication process is depicted in Figure 8: (Figure 8a) A SOI wafer with 25 μm device layer, 2 μm SiO_2 layer, and 400 μm substrate was prepared; (Figure 8b) then, a 50/500 nm Cr/Au layer was sputtered and patterned as the electrodes; (Figure 8c) after that, the device layer was patterned and etched by DRIE to form the structures of the resonators; (Figure 8d) finally, the SiO_2 layer was etched by HF gas, and the resonators were released.

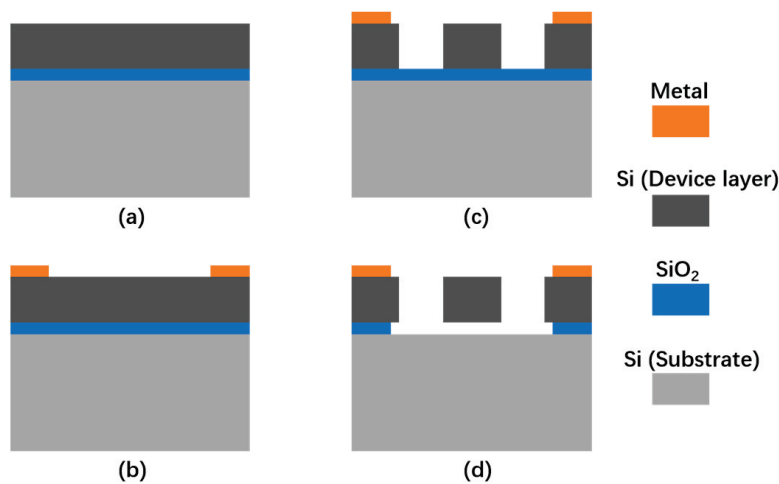


Figure 8. The fabrication process of the coupled silicon resonators. (a) A SOI wafer was prepared; (b) The electrodes were formed; (c) The structures of the resonators were formed; (d) The resonators were released.

The fabricated resonator chip is shown in Figure 9a. In order to ensure symmetry, the resonant structures and electrode structures on both sides are designed to be completely identical, even if some electrodes are not used during the testing process. To conduct the strain test, the fabricated resonator chip was adhered onto a 65 Mn elastic steel sheet with a thickness of 1 mm and a length of 9 cm, as shown in Figure 9b. Then, the resonator electrodes were wire-bonded to the surrounding PCB board for testing. Here the high-temperature-resistant epoxy resin with the product model 353 ND was adopted as the adhesion layer.

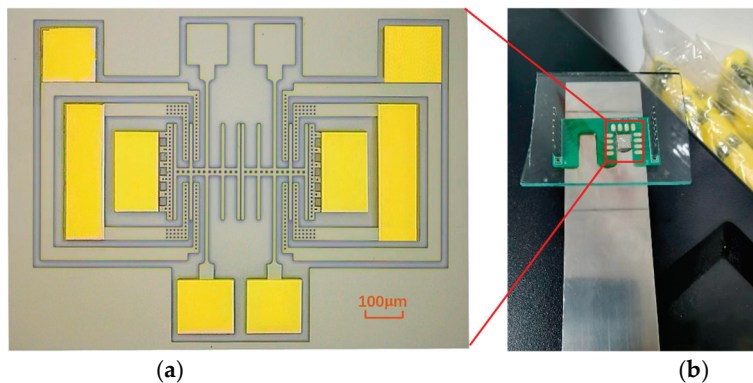


Figure 9. (a) The fabricated resonator chip. (b) The resonator chip adhered onto an elastic steel sheet for strain testing.

4.2. Measurements and Results

The testing platform for PT-symmetric-coupled silicon resonators is shown in Figure 10. The entire testing system was placed on a vibration isolation platform. In order to accurately adjust the strain of the elastic steel sheets, a servo control system was developed. The servo control system and elastic steel plate were placed in a steel chamber with adjustable air pressure. By connecting an external vacuum pump, the air pressure in the steel chamber can be adjusted between 1 and 10^5 Pa. The feedback circuits loop included a transconductance amplifier, bandpass filter, voltage-controlled amplifier, and phase shifter. The frequency responses of the resonators were read out using a Zurich Instruments HF2LI lock-in amplifier.

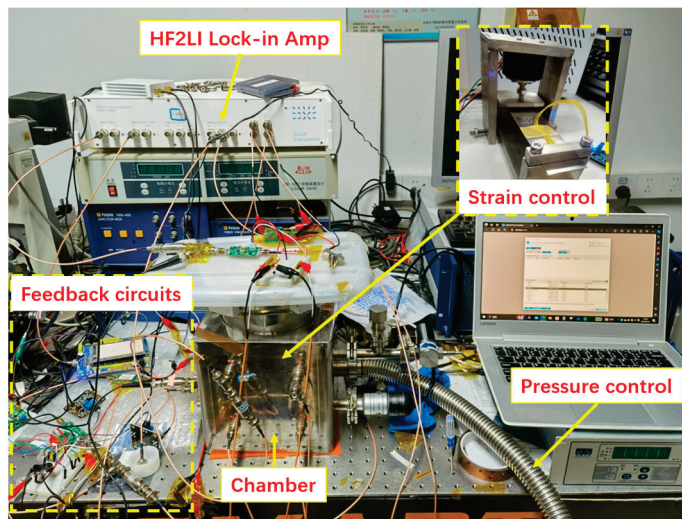


Figure 10. The testing platform for the PT-symmetric-coupled silicon resonators.

Firstly, the coupled silicon resonators were adjusted to be PT symmetry. The pressure of the vacuum chamber was set to 200 Pa, at which point the loss coefficient of the loss resonator was 7×10^{-4} . The gain and phase of the feedback circuit were carefully adjusted to make the gain coefficient of the gain resonator also 7×10^{-4} . When the gain and loss are equal, the system is in PT symmetry.

Then, the coupling coefficient was scanned to find and bias the system near the EP. By adjusting the coupling voltage between two resonators, the variations of eigenfrequencies versus the coupling coefficient was obtained in Figure 11. It can be seen that when the coupling coefficient is 7×10^{-4} , i.e., the corresponding coupling voltage is 9.8 V, the system is near the EP point.

After that, the strain-induced frequency splitting of the PT-symmetric-coupled silicon resonators was tested. The strain loading method was similar to our previous work [30], and the strain applied to the resonator chip ε_x can be calculated as [31]:

$$\varepsilon_x = \frac{3t_c\delta_y(L_c - d)}{2L_c^3} \quad (12)$$

where L_c is the length of the cantilever beam, t_c is its thickness, and d is the distance from the resonator chip to the anchor. Here, the servo motor is used to precisely control the deflection δ_y of the steel plate.

Figure 12 shows the relationship curve between frequency splitting and the applied strain of the PT system. For comparison, the simulation and the theoretical results were recalculated according to the experimental settings and are also shown in Figure 12. All curves showed that the frequency splitting increases with the increase of applied strain. The measured frequency splitting in the PT system achieved 34 Hz at $90 \mu\epsilon$. In Figure 12, the measured frequency splitting is smaller compared to the simulation or theoretical results. This may be because a portion of the applied strain is lost on the adhesive layer between the resonator chip and the steel plate [30].

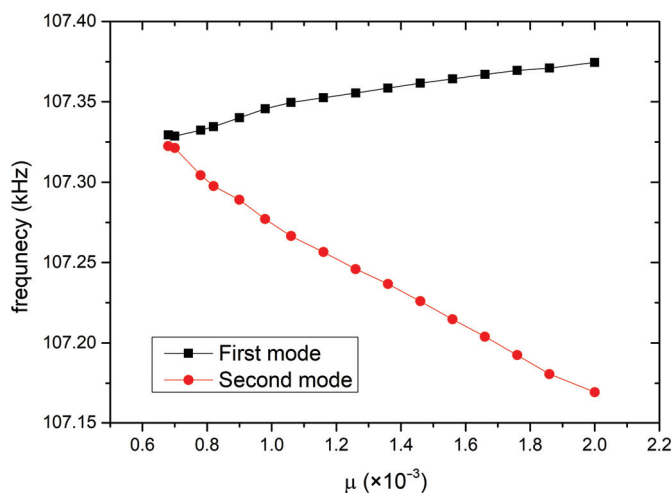


Figure 11. Variations of the eigenfrequencies of the PT system vs. the coupling coefficient.

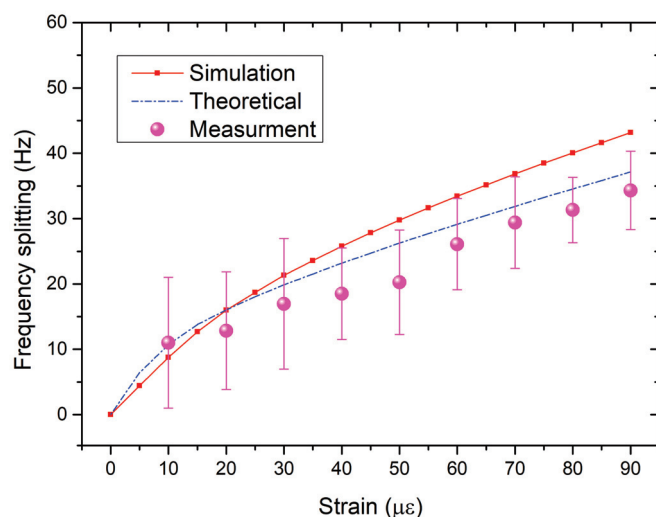


Figure 12. Measured strain-induced frequency splitting of the PT system. The simulation and theoretical results are also shown for comparison.

5. Conclusions

In summary, this paper proposes the strain-induced frequency splitting in PT symmetry-coupled silicon resonators. The frequency splitting of the PT system to perturbations caused by strain was derived. The mechanical model of the PT system was established in COMSOL software and the characteristic curves of its eigenfrequencies were obtained. Theory and simulation both indicate that the PT system is more sensitive to strain perturbation near the EP point. A feedback circuit was designed to achieve the negative damping required for PT symmetry. After design and simulation, the silicon resonator chip was successfully fabricated based on an SOI process. Finally, the PT-symmetric-coupled silicon resonators were successfully constructed, and the frequency splitting phenomenon caused by strain was observed experimentally. Our work enriches the PT symmetry theory and provides valuable references for potential applications.

Author Contributions: Conceptualization, L.W.; Methodology, S.Z.; Software, S.Z.; Validation, Q.Y.; Formal analysis, L.W. and S.Z.; Data curation, L.W.; Writing—original draft, L.W. and Q.Y.; Writing—review and editing, L.W.; All authors have read and agreed to the published version of the manuscript.

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References

1. Bender, C.M.; Boettcher, S. Real Spectra in Non-Hermitian Hamiltonians Having PT Symmetry. *Phys. Rev. Lett.* **1998**, *80*, 5243–5246. [CrossRef]
2. Bender, C.M. *PT Symmetry: In Quantum and Classical Physics*; World Scientific: London, UK, 2018.
3. El-Ganainy, R.; Makris, K.G.; Khajavikhan, M.; Musslimani, Z.H.; Rotter, S.; Christodoulides, D.N. Non-Hermitian physics and PT symmetry. *Nat. Phys.* **2018**, *14*, 11–19. [CrossRef]
4. El-Ganainy, R.; Makris, K.G.; Christodoulides, D.N.; Musslimani, Z.H. Theory of coupled optical PT-symmetric structures. *Opt. Lett.* **2007**, *32*, 2632–2634. [CrossRef]
5. Klaiman, S.; Günther, U.; Moiseyev, N. Visualization of Branch Points in PT-Symmetric Waveguides. *Phys. Rev. Lett.* **2008**, *101*, 080402. [CrossRef]
6. Schindler, J.; Li, A.; Zheng, M.C.; Ellis, F.M.; Kottos, T. Experimental study of active LRC circuits with PT symmetries. *Phys. Rev. A* **2011**, *84*, 040101. [CrossRef]
7. Zhu, X.; Ramezani, H.; Shi, C.; Zhu, J.; Zhang, X. PT-Symmetric Acoustics. *Phys. Rev. X* **2014**, *4*, 031042. [CrossRef]
8. Fleury, R.; Sounas, D.L.; Alù, A. Parity-Time Symmetry in Acoustics: Theory, Devices, and Potential Applications. *IEEE J. Sel. Top. Quantum Electron.* **2016**, *22*, 5000809. [CrossRef]
9. Lin, Z.; Ramezani, H.; Eichelkraut, T.; Kottos, T.; Cao, H.; Christodoulides, D.N. Unidirectional Invisibility Induced by PT-Symmetric Periodic Structures. *Phys. Rev. Lett.* **2011**, *106*, 213901. [CrossRef]
10. Ramezani, H.; Christodoulides, D.N.; Kovanis, V.; Vitebskiy, I.; Kottos, T. PT-Symmetric Talbot Effects. *Phys. Rev. Lett.* **2012**, *109*, 033902. [CrossRef]
11. Alexeeva, N.V.; Barashenkov, I.V.; Sukhorukov, A.A.; Kivshar, Y.S. Optical solitons in PT -symmetric nonlinear couplers with gain and loss. *Phys. Rev. A* **2012**, *85*, 063837. [CrossRef]
12. Hodaiei, H.; Miri, M.; Heinrich, M.; Christodoulides, D.N.; Khajavikhan, M. Parity-time-symmetric microring lasers. *Science* **2014**, *346*, 975–978. [CrossRef] [PubMed]
13. Zhou, H.; Lee, J.; Liu, S.; Zhen, B. Exceptional surfaces in PT -symmetric non-Hermitian photonic systems. *Optica* **2019**, *6*, 190–193. [CrossRef]
14. Assawaworrarit, S.; Yu, X.; Fan, S. Robust wireless power transfer using a nonlinear parity-time-symmetric circuit. *Nature* **2017**, *546*, 387–390. [CrossRef] [PubMed]
15. Chen, P.; Sakhdari, M.; Hajizadegan, M.; Cui, Q.; Cheng, M.M.; El-Ganainy, R.; Alù, A. Generalized parity-time symmetry condition for enhanced sensor telemetry. *Nat. Electron.* **2018**, *1*, 297–304. [CrossRef]
16. Zhou, B.; Deng, W.; Wang, L.; Dong, L.; Huang, Q. Enhancing the Remote Distance of LC Passive Wireless Sensors by Parity-Time Symmetry Breaking. *Phys. Rev. Appl.* **2020**, *13*, 064022. [CrossRef]
17. Cheng, S.; Li, W.; Zhang, H.; Akhtar, M.N.; Yi, Z.; Zeng, Q.; Ma, C.; Sun, T.; Wu, P.; Ahmad, S. High sensitivity five band tunable metamaterial absorption device based on block like Dirac semimetals. *Opt. Commun.* **2024**, *569*, 130816. [CrossRef]
18. Chen, Z.; Cheng, S.; Zhang, H.; Yi, Z.; Tang, B.; Chen, J.; Zhang, J.; Tang, C. Ultra wideband absorption absorber based on Dirac semimetallic and graphene metamaterials. *Phys. Lett. A* **2024**, *517*, 129675. [CrossRef]
19. Charumathi, P.R.; Senthilnathan, K. Unravelling PT Symmetry: Applications in Metamaterials. *Plasmonics* **2024**, 1–22.
20. Liu, Z.; Zhang, J.; Özdemir, Ş.K.; Peng, B.; Jing, H.; Lü, X.; Li, C.; Yang, L.; Nori, F.; Liu, Y. Metrology with PT -Symmetric Cavities: Enhanced Sensitivity near the PT -Phase Transition. *Phys. Rev. Lett.* **2016**, *117*, 110802. [CrossRef]
21. Zeng, C.; Sun, Y.; Li, G.; Li, Y.; Jiang, H.; Yang, Y.; Chen, H. Enhanced sensitivity at high-order exceptional points in a passive wireless sensing system. *Opt. Express* **2019**, *27*, 27562–27572. [CrossRef]
22. Hodaiei, H.; Hassan, A.U.; Wittek, S.; Garcia-Gracia, H.; El-Ganainy, R.; Christodoulides, D.N.; Khajavikhan, M. Enhanced sensitivity at higher-order exceptional points. *Nature* **2017**, *548*, 187–191. [CrossRef] [PubMed]
23. Dong, Z.; Li, Z.; Yang, F.; Qiu, C.; Ho, J.S. Sensitive readout of implantable microsensors using a wireless system locked to an exceptional point. *Nat. Electron.* **2019**, *2*, 335–342. [CrossRef]
24. Xiao, Z.; Li, H.; Kottos, T.; Alù, A. Enhanced Sensing and Nondegraded Thermal Noise Performance Based on PT -Symmetric Electronic Circuits with a Sixth-Order Exceptional Point. *Phys. Rev. Lett.* **2019**, *123*, 213901. [CrossRef] [PubMed]
25. Zhou, B.; Wang, L.; Dong, L.; Huang, Q. Observation of the perturbed eigenvalues of PT-symmetric LC resonator systems. *J. Phys. Commun.* **2021**, *5*, 045010. [CrossRef]
26. Zhang, M.; Dong, L.; Wang, L.; Huang, Q. Exceptional points enhance sensing in silicon micromechanical resonators. *Microsyst. Nanoeng.* **2024**, *10*, 12. [CrossRef]

27. Saleem, M.M.; Saghir, S.; Bukhari, S.A.R.; Hamza, A.; Shakoor, R.I.; Bazaz, S.A. A Low-g MEMS Accelerometer with High Sensitivity, Low Nonlinearity and Large Dynamic Range Based on Mode-Localization of 3-DoF Weakly Coupled Resonators. *Micromachines* **2021**, *12*, 310. [CrossRef]
28. Wang, K.; Xiong, X.; Wang, Z.; Cai, P.; Ma, L.; Zou, X. Utilizing the Intrinsic Mode of Weakly Coupled Resonators for Temperature Compensation. *Micromachines* **2022**, *13*, 1447. [CrossRef]
29. Lee, J.E. Silicon Micromechanical Resonators for Measurements of Mass and Charge. Ph.D. Thesis, University of Cambridge, Cambridge, UK, 2008.
30. Zhang, C.; Zhang, S.; Wang, L. A Sawtooth MEMS Capacitive Strain Sensor for Passive Telemetry in Bearings. *IEEE Sens. J.* **2021**, *21*, 22527–22535. [CrossRef]
31. Suster, M.; Guo, J.; Chaimanonart, N.; Ko, W.H.; Young, D.J. Low-noise CMOS integrated sensing electronics for capacitive MEMS strain sensors. In Proceedings of the IEEE 2004 Custom Integrated Circuits Conference, Orlando, FL, USA, 6 October 2004; pp. 693–696.

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Article

A Miniaturized, Fuel-Free, Self-Propelled, Bio-Inspired Soft Actuator for Copper Ion Removal

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Abstract: We present a novel miniaturized, gear-shaped, fuel-free actuator capable of autonomously propelling itself in an aquatic environment to absorb heavy metals, such as copper ions. While hydrogel-based absorbents are promising solutions for cationic pollutant remediation, their stationary nature limits their effectiveness in areas where contaminants are unevenly distributed. To address this, we developed a bio-inspired soft actuator that mimics natural propulsion mechanisms. The Marangoni effect, driven by its inherent chemical properties, demonstrated a self-propelled motion without requiring external fuel. The proof-of-concept actuator generated a plane motion lasting up to 2 h and swept over an area approximately 400 times bigger than its size. By harnessing the chemical and optical properties of the hydrogel, we efficiently removed and quantitatively analyzed copper ions through a colorimetric method. This innovative integration of self-propelled movement and efficient copper ion absorption underscores its potential for advancing miniaturized devices in environmental remediation, paving the way for more active and efficient pollutant removal systems in challenging aquatic environments.

Keywords: self-propelled; Marangoni effect; heavy metal ions; water pollution; fuel-free actuator

1. Introduction

Heavy metal ions have been identified as water and soil pollutants that can potentially endanger public health and cause long-term harm to diverse ecosystems [1]. These non-biodegradable materials possess extended biological half-lives, persisting in aquatic and terrestrial environments [2]. This persistence adversely affects humans and other living organisms [3]. There is a growing public health concern regarding heavy metal contamination, leading to stricter regulations governing the presence of heavy metals in water bodies [4].

Several techniques have been investigated for removing heavy metal ions from aqueous environments to secure safe water resources and human health [5,6]. These methods include ion exchange [7], membrane filtration [8], chemical precipitation [9], and electrochemical techniques [10]. Numerous industries have widely used several of these techniques. However, these procedures exhibit limitations, such as elevated maintenance expenses, substantial electrical power demands, and reduced thermostability [11,12].

Compared to the aforementioned methods, absorption has emerged as a promising candidate for heavy metal removal from the environment due to its unique advantages, including high metal uptake capacity, rapid adsorption kinetics, and potential metal ion selectivity [13,14]. Hydrogels, in particular, have been identified as highly effective absorbent materials with a wide range of applications in heavy metal remediation in wastewater [15–17]. They are soft materials created by cross-linking natural or synthetic polymers, forming a three-dimensional mesh structure [16]. Hydrogels possess inherent characteristics, including various functional groups tailored for specific analyte binding, biodegradability, and swellability, making them highly suitable for heavy metal removal [18,19].

Miniaturized self-propelled devices based on a loaded/unloaded power source have evolved as one of the effective instruments for a wide variety of medical and environmental science applications [20–25]. Many fuels and propulsion systems based on magnetic, acoustic, optical, and chemical actuation mechanisms have been used to drive such devices [26–28]. Among these, using bio-inspired surface-tension-based mechanisms to move the device on the water's surface has attracted a lot of interest due to their favorable scaling, low toxicity, and high efficiency [29–31].

Semi-aquatic arthropods, like rove beetles of the genus *Stenus* Latreille, are equipped with pygidial glands that emit a complex mix of piperidine and pyridine-derived alkaloids and several terpenes. This secretion composition enables them to skim rapidly and far over the aquatic environment without using their legs. Early observations by Billard and Bruyant noted that these species emit chemicals that act as surfactants, rapidly propelling them forward [32]. This locomotion is driven by Marangoni propulsion, a phenomenon where surface tension gradients created by these emissions allow the beetles to skim swiftly across water surfaces [33]. Self-propelled soft robots have been developed to mimic the locomotion of water striders and move freely on water surfaces [34].

To construct these surface-tension-driven devices, surfactants as fuel must be stored and released in a controlled manner. Numerous studies have focused on this mechanism and developed it to ensure the smooth motion of a swimming device propelled by the Marangoni effect [35–37]. While bio-inspired approaches have shown potential, developing self-propelled systems that operate without requiring a surfactant as a stimulus remains a significant challenge.

Despite the proven efficacy of hydrogels in heavy metal removal due to their high absorption capacity and chemical versatility, their static nature limits their application in unevenly distributed contamination scenarios. Additionally, while existing self-propelled devices demonstrate significant promise in medical and environmental applications, the reliance on external energy sources or complex fuel systems complicates their broader deployment. This study addresses these limitations by developing a novel fuel-free, self-propelled soft actuator inspired by the natural motility mechanisms observed in semi-aquatic arthropods, like the *Stenus* rove beetles. Leveraging Marangoni propulsion, driven by surface tension gradients similar to those used by these beetles, this actuator autonomously navigates water surfaces to target and remove copper ion contaminants effectively. By integrating actuation with remediation capabilities, this bio-inspired device not only overcomes the mobility constraints of traditional hydrogels but also eliminates the need for external power, offering a pioneering approach to environmental remediation.

2. Materials and Methods

2.1. Materials and Equipment

The following chemicals, materials, and equipment were used: copper (II) nitrate hydrate (heavy metal ions; Sigma Aldrich, St. Louis, MO, USA), 2-acrylamido-2-methylpropane sulfonic acid (AMPS), acrylic acid (AA) (monomers; Sigma Aldrich, St. Louis, MO, USA), polyethylene glycol diacrylate (PEGDA) (crosslinker; Sigma Aldrich, St. Louis, MO, USA), lithium phenyl 2,4,6-trimethyl-benzoyl phosphinite (photoinitiator; Sigma Aldrich, USA), and a laser-powered stereolithography (SLA) 3D printer (Formlabs, Somerville, MA, USA).

2.2. Working Principles

The novel fuel-free, self-propelled, bio-inspired actuator was designed to use the dynamic surface properties of hydrophilic–hydrophobic block copolymers and the Marangoni effect to achieve autonomous motion and effective copper ion removal.

The locomotion of rove beetles inspired the operational principle of the actuator. These beetles use the secretion of surfactants to propel themselves forward on the water surface by creating a gradient of surface tension known as the Marangoni effect, as shown in Figure 1a. Theoretically, according to the linearized equation of state for surface tension, changes in surface tension are proportional to changes in surfactant concentration [38].

Furthermore, balancing the tangential viscous stress in the fluid with the tangential force resulting from the gradient of surface tension can be expressed as:

$$\mu \frac{\partial u_s}{\partial z} = -\nabla_s \gamma \quad (1)$$

where μ , u_s , and γ represent dynamic viscosity, tangential velocity, and the fluid's surface tension. Hence, rove beetles applying the surfactant results in a surface tension gradient around the object. Considering the imbalance between the forces, this gradient induces the beetles to start moving and forms a velocity field around the arthropods.

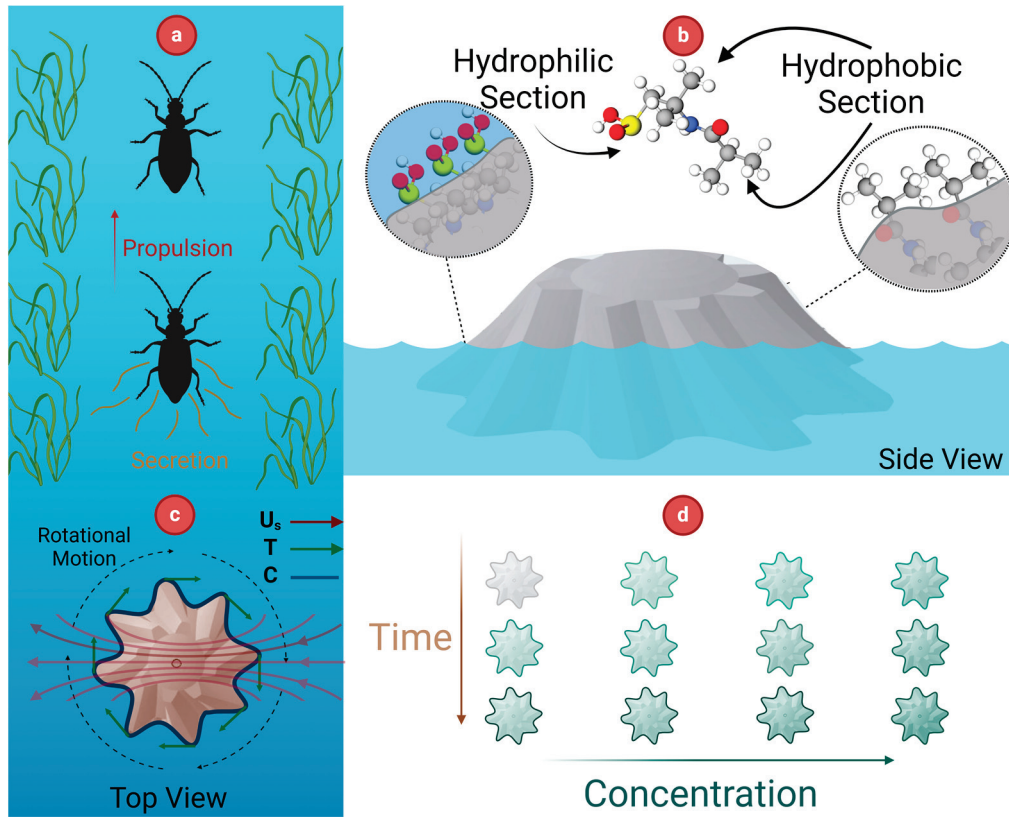


Figure 1. (a) Secretion of the surfactant by the beetle, (b) rearrangement of the surface configuration in response to the environment, (c) theoretical model behind the general plane motion of the gear-shaped actuator, and (d) absorption of copper ions indicated by color variations along the concentration and time scale.

Emulating rove beetles' biological mechanism, the actuator was fabricated from a chemical structure that undergoes quick surface rearrangement in response to the nature of environmental stimuli [39–42]. Under dry conditions, the surface of the actuator becomes hydrophobic, whereas it turns hydrophilic upon exposure to water (Figure 1b). This dynamic alteration in surface properties is associated with a change in the contact angle (θ), which, in turn, using Young's equation, Equation (2), facilitates the generation of the gradient of surface tension [43]. In this equation, γ_{sg} , γ_{sl} , and γ_{lv} stand for solid/liquid tension, solid/vapor tension, and liquid/vapor interfacial tension, respectively.

$$\gamma_{sg} - \gamma_{sl} = \gamma_{lv} \cos \theta \quad (2)$$

In our device, the integration of Equations (1) and (2) demonstrates that variations in the contact angle are correlated with changes in surface tension, resulting in generating a velocity field around the object. This mechanism parallels the locomotion observed in rove beetles, which is induced by surfactant secretion.

The fabricated actuator benefits from the gear-shaped design, enabling a smooth rotational motion of the object in an aquatic environment. The rotational motion of the actuator influenced by the Marangoni effect can be justified using the circulation concept. It quantifies the tendency of the fluid to induce rotational movement in the actuator.

$$F_{Prop} : \Gamma = \oint U_s \cdot dr = \oint U_s \cdot T ds \quad (3)$$

Circulation (Γ) can be defined as a line integral of the velocity field U_s around the actuator's perimeter, represented by a simple closed curve C , as presented in Equation (3). This can be expressed in terms of the unit tangent vector T and the differential arc length ds along the actuator's perimeter. In the design depicted in Figure 1c, the changes in the T vector and the velocity field are due to the asymmetrical geometry and dynamic reorientation of the actuator's structure, respectively. This results in a non-zero dot product of $U_s \cdot T$ during the actuator's interval motion. Consequently, as shown in Figure 1c and described by Equation (3), this leads to non-zero circulation and rotational motion of the actuator.

Furthermore, the gear-shaped design provides a higher surface area compared to the other geometries, which could increase the device's capacity for copper ion absorption, as demonstrated in Figure 1d.

2.3. Method of Fabrication

In this study, hydrogel was used to form the actuator's physical structure. As shown in Figure 2, AA (1 mL, 98%) and AMPS (1.4 mL, 10%) as monomers were copolymerized to form the device's configuration. Polyethylene glycol diacrylate (PEGDA) (0.04 mL) cross-linked the monomers. The solution contained distilled water as the solvent (2 mL) and lithium phenyl 2,4,6-trimethyl-benzoyl phosphinite (0.009 gr) as the photoinitiator. In contrast, the gear-shaped mold was designed using CAD software (SolidWorks, 2024 version) to create the gear configuration, and the CAD file was transferred to an SLA 3D printer to fabricate the gear mold, as depicted in Figure 2.

As shown in Figure 2a–e, a homogenous AMPS/AA solution was obtained by stirring the combined component solution for 1 h. The stirred solution was then poured into the gear-shaped mold, which was then exposed to UV light to cause the hydrogel to polymerize. The hydrogel was put through a freeze-drying procedure to create a porous structure. Next, the actuator was immersed in a copper-ion-enriched solution, with concentrations ranging from 500 to 3000 ppm, to show its functionality in heavy metal absorption. The actual image of the actuator, which absorbed copper ions, is in Figure 2f.

2.4. Material Characterization

The surface and chemical morphology of the AA/AMPS fabricated actuator were examined using a scanning electron microscope (SEM; Carl Zeiss, Jena, Germany). FTIR analysis was conducted to confirm the formation of functional groups in the copolymerized hydrogel. The Fourier-transform infrared spectroscopy (FTIR) spectrum was recorded at room temperature, covering the frequency range of 4000–500 cm^{-1} (Shimadzu, Kyoto, Japan).

2.5. Locomotion Characterization

The chemical principles underlying the actuator's locomotion mechanism were evaluated to validate the dynamic changes in the contact angle. To achieve this, contact angle measurements were conducted at room temperature using a contact angle meter (Holmarc, India). Additionally, the associated physical principles were analyzed to correlate the actuator's average velocity with the hydrogel's swelling capacity. The velocity was measured from recorded video, and the average velocity was calculated over a 5–10 min interval.

2.6. Characterization of Copper Ion Absorption

A series of copper (II) nitrate solutions were prepared at varying concentrations to demonstrate the actuator's efficacy in heavy metal absorption. These solutions were introduced into a container measuring 32 cm (W) $\times$ 48 cm (L) $\times$ 8 cm (H). The trajectory and capabilities of the device related to the absorption of copper ions were evaluated using color intensity parameters, including saturation and hue.

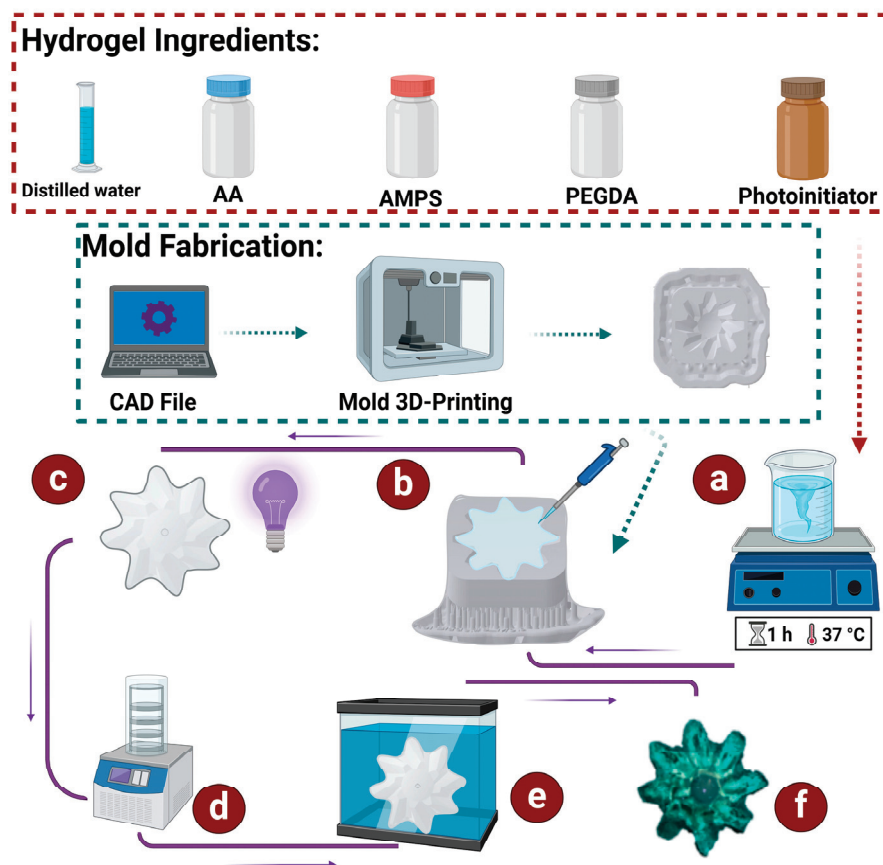


Figure 2. Steps to create and test a gear-shaped self-propelled actuator: (a) stirring hydrogel ingredients to form a homogenous solution, (b) dripping the prepared solution into a 3D-printed gear-shaped mold, (c) using UV irradiation to polymerize the hydrogel, (d) freeze-drying the hydrogel to form a porous structure, (e) immersing the hydrogel in a copper-ion-containing solution, and (f) the actual image of the actuator after copper absorption.

3. Results and Discussion

3.1. Morphological and Chemical Characterization

Figure 3 illustrates the surface morphology of the fibrous hydrogel that was fabricated, demonstrating the interconnected structure of the self-propelled hydrogel network. The provided SEM images of the actuator revealed a unique surface morphology, showcasing the intricate internal structure at different magnifications.

FTIR spectrum analysis confirmed the polymerization process among selected monomers and the existence of their chemical functional groups, as shown in Figure 4. Peaks at 1357 cm^{-1} and 1610 cm^{-1} are characteristic stretching vibrations of $S=O$ and $N-H$ associated with the sulfonic acid and amide groups in AMPS. Additionally, peaks at 3300 cm^{-1} indicate the elongation of the hydroxyl group in carboxylic acid, indicating AA's presence in the hydrogel.

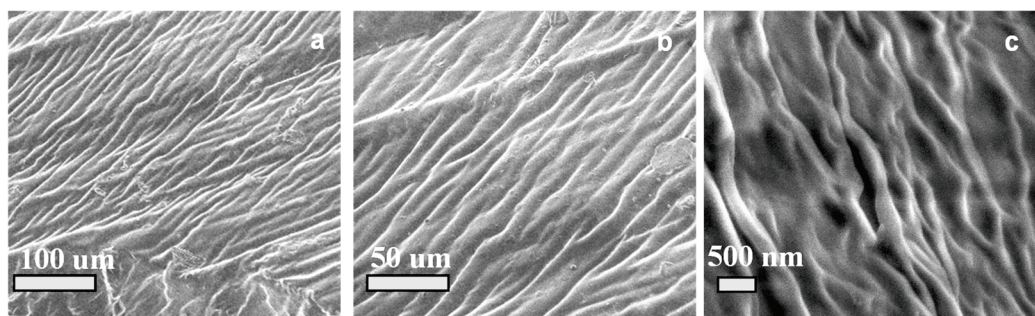


Figure 3. Morphology of the AA/AMPS-based actuator sample at different magnifications with a scale bar of (a) 100 μm , (b) 50 μm , and (c) 500 nm.

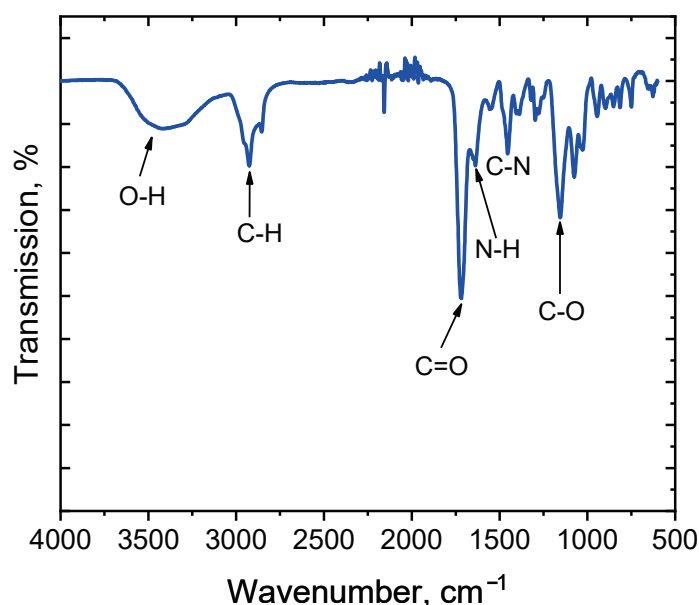


Figure 4. FTIR spectrum of the AA/AMPS-based actuator.

3.2. Chemical Aspect of Locomotion

Figure 5 illustrates the dynamic changes in the contact angle over a short period of time and with sequential droplet addition, indicating corresponding changes in surface tension, as described by Young's equation. Figure 5a demonstrates the variation in the contact angle versus time to show rapid contact angle changes on the outermost surface of the actuator. The results prove a general decreasing trend in the contact angle in 270 s, indicating a rapid initial decline that gradually stabilizes. Figure 5b shows the contact angles measured at the same spot with the sequential addition of water droplets to highlight a continuous decrease in the contact angle, which is crucial for sustaining the Marangoni propulsion mechanism. Initially, the contact angle was about 64 degrees, and with each new droplet, the contact angle progressively decreased and reached 47 degrees at the fourth one. This steady reduction in the contact angle correlated with a constant decrease in surface tension based on Young's equation. Hence, the flow in these actuators was generated by alterations in the contact angle, which subsequently changed the surface tension and created a surface tension gradient, resulting in Marangoni propulsion. All the contact angle measurements were performed using the DropSnake method and ImageJ software (Version 1.54k) (Figure 1a,b).

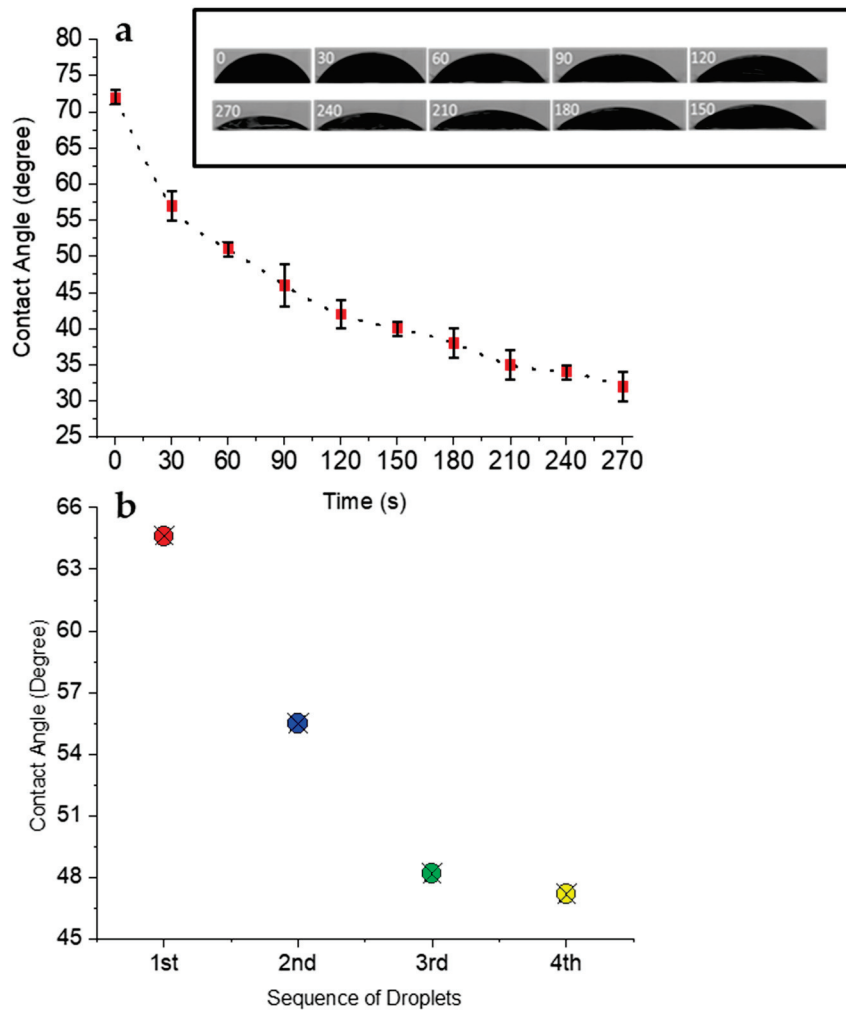


Figure 5. The changes in the contact angle vs. time (a) and vs. the addition of sequential droplets at the same spot (b).

3.3. The Physical Aspect of Locomotion

The water-absorbing capacity of the actuator was calculated by the Equation (4) as follows:

$$\text{Swelling ratio (\%)} = \frac{W_w - W_d}{W_d} \times 100 \quad (4)$$

where W_w and W_d are the weights of the swollen and the initial dried hydrogel, respectively.

The sample was immersed in an adequate copper-enriched solution at 10 min intervals to acquire different swelling ratios. The correlation between the average velocity and the hydrogel's swelling capacity is shown in Figure 6 to validate the actuator's locomotion from a physical perspective. As shown in Figure 6, the average velocity of the actuator decreased over time, while the actuator's swelling ratio increased. This inverse relationship indicates that as the hydrogel actuator absorbs water, it swells, causing a reduction in surface tension, as concluded from Figure 1. Consequently, the Marangoni effect, driven by the surface tension gradient, generates the propulsive force necessary for the actuator's motion.

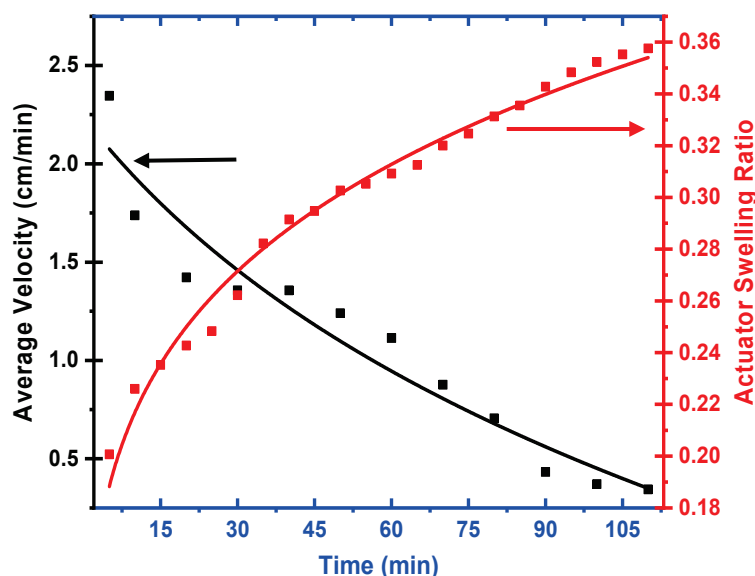


Figure 6. Actuator swelling ratio and average velocity during locomotion time.

3.4. Copper Ion Absorption Capacity

The Marangoni effect is crucial in propelling the actuator and enhancing its interaction with copper ions in solution. The hydrogel is composed of anionic functional groups derived from 2-acrylamido-2-methylpropane sulfonic acid (AMPS), featuring sulfonate groups that effectively attract and bind positively charged copper ions through electrostatic interactions. This interaction is dynamically enhanced by the actuator's continuous movement, driven by surface tension gradients. As the actuator moves, it exposes fresh hydrogel surfaces to the contaminated water, continuously renewing the contact boundary and preventing the saturation that typically reduces the effectiveness of stationary absorption systems. This mobility, coupled with the inherent chemical properties of the hydrogel designed for high metal uptake, enhances the overall absorption efficiency.

The trajectory of the device and its ability to absorb copper ions are shown in Figure 7, displaying its effectiveness over a 25 min period in an almost 1500 cm² area. Furthermore, displacement data were normalized relative to the size of the actuator to provide a scale-independent understanding of its mobility and coverage efficiency. During the trajectory, the actuator had general plane motion, which could be crucial for ensuring thorough exposure to and interaction with a copper-ion-enriched area. This experimental setup validates the actuator's effective absorption of copper ions over a broad area, confirming its potential as a scalable and efficient solution for heavy metal remediation.

3.5. Adsorption Study

As demonstrated in Figure 8, the AA/AMPS-based actuator was used for the colorimetric analysis of Cu^{2+} ions. As the actuator was immersed in the solution, the actuator's color started changing from the outermost surface area to the middle. The graph illustrates a parabolic and sigmoidal relationship between hue and saturation, respectively, with the concentration of copper (II) ions. As the Cu^{2+} ions increased from 0 to 3500 ppm, there was a noticeable transition in saturation from approximately 0.1 to over 0.5, paralleled by a change in hue from 40 to nearly 2000 degrees. This distinct pattern of color variation serves as an effective qualitative indicator of the concentration of Cu^{2+} ions.

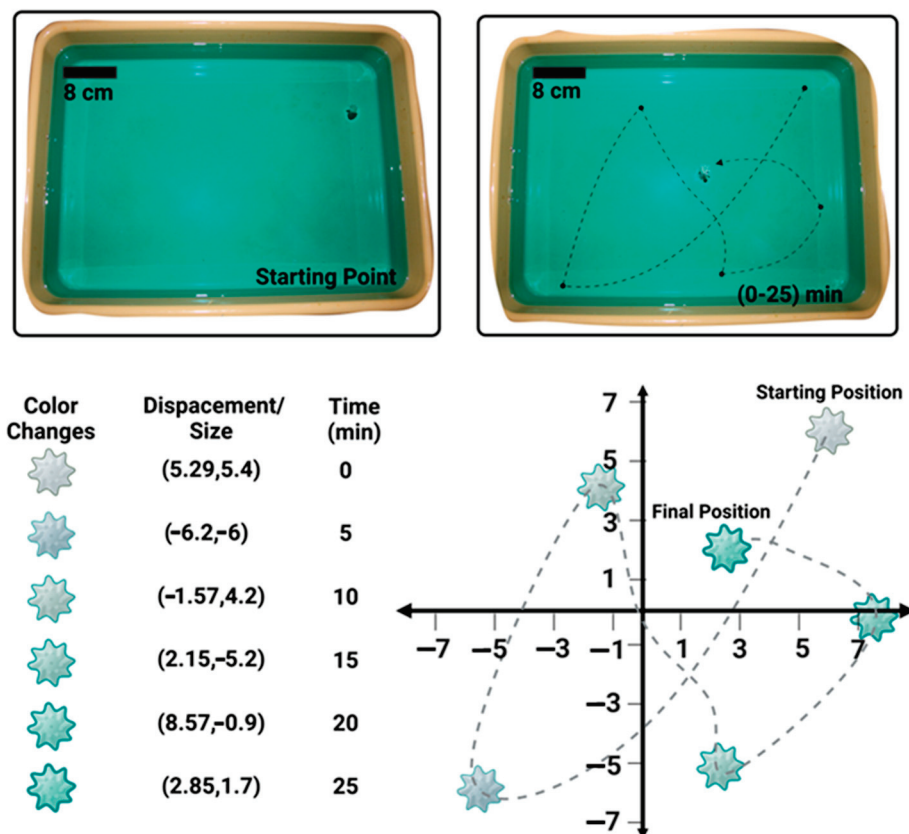


Figure 7. Trajectory of the fabricated actuator immersed in a copper nitrate solution plotted with the normalized scale (displacement/size).

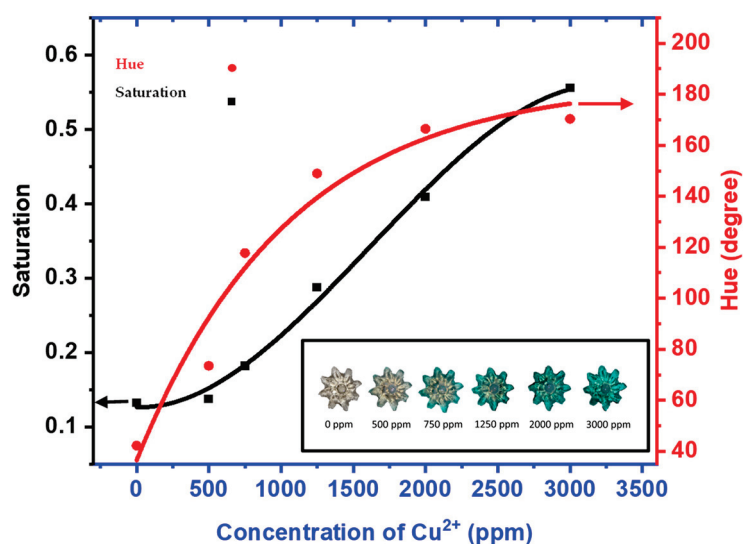


Figure 8. A colorimetric characterization graph of the actuator (inset) shows a color palette for different concentrations of copper ions (500–3000 ppm).

4. Conclusions

In conclusion, this study introduces a novel miniaturized, fuel-free, self-propelled bio-inspired soft actuator applicable to an aquatic environment. It demonstrates its usable features in removing heavy metal ions, particularly copper ions. Using the Marangoni effect for autonomous motion, the gear-shaped hydrogel actuator mimics the rove beetle's natural propulsion without surfactant loading and unloading.

The actuator is fabricated from a chemical structure that undergoes quick surface rearrangement of functional groups in response to the exposed environment, either air or water. Under dry conditions, the surface of the actuator becomes hydrophobic, whereas it turns hydrophilic upon autonomous exposure to water. This dynamic alteration in surface properties generates a surface tension gradient that enables propulsion.

This innovative approach eliminates the need for external fuel and overcomes the mobility limitations of traditional hydrogels. The actuator demonstrated efficient movement and copper ion absorption. Material characterization through SEM and FTIR confirmed the successful fabrication and functionalization of the hydrogel. The proposed soft actuator's dynamic contact angle progression enabled Marangoni propulsion, and its swelling behavior kept water absorption and motion in balance. This study provides a unique scalable strategy for removing heavy metals and contaminants from aquatic environments. This fuel-free, self-propelled actuator presents a promising solution for developing autonomous clean-up swarm devices.

Author Contributions: H.J.C. and M.C. conceived and designed the experiments; M.C., P.B., P.P., and J.F. performed the experiments and analyzed the data; M.C., P.B., and H.J.C. wrote the manuscript; and H.J.C. supervised the project. All authors have read and agreed to the published version of the manuscript.

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References

1. Wang, Z.; Luo, P.; Zha, X.; Xu, C.; Kang, S.; Zhou, M.; Nover, D.; Wang, Y. Overview assessment of risk evaluation and treatment technologies for heavy metal pollution of water and soil. *J. Clean. Prod.* **2022**, *379*, 134043. [CrossRef]
2. Borjian, P.; Chimera, M.; Pathak, P.; Childs, A.; Rajaraman, S.; Cho, H.J. Electrochemical Sensors for Lead Ion Detection Using Sodium Alginate Crosslinked with 2-Acrylamido-2-Methyl Propane Sulfonic Acid and Aluminum Microparticles. *IEEE Sens. Lett.* **2023**, *7*, 1–4. [CrossRef]
3. Bharti, R.; Sharma, R. Effect of heavy metals: An overview. *Mater. Today Proc.* **2022**, *51*, 880–885.
4. Briffa, J.; Sinagra, E.; Blundell, R. Heavy metal pollution in the environment and their toxicological effects on humans. *Heliyon* **2020**, *6*, e04691. [CrossRef]
5. Zamora-Ledezma, C.; Negrete-Bolagay, D.; Figueroa, F.; Zamora-Ledezma, E.; Ni, M.; Alexis, F.; Guerrero, V.H. Heavy metal water pollution: A fresh look about hazards, novel and conventional remediation methods. *Environ. Technol. Innov.* **2021**, *22*, 101504. [CrossRef]
6. Soni, R.; Suyal, D.C.; Morales-Oyervides, L.; Chauhan, J.S. *Current Status of Fresh Water Microbiology*; Springer: Berlin/Heidelberg, Germany, 2023.
7. Kurniawan, T.A.; Chan, G.Y.; Lo, W.-H.; Babel, S. Physico-chemical treatment techniques for wastewater laden with heavy metals. *Chem. Eng. J.* **2006**, *118*, 83–98. [CrossRef]
8. Abdullah, N.; Yusof, N.; Lau, W.; Jaafar, J.; Ismail, A. Recent trends of heavy metal removal from water/wastewater by membrane technologies. *J. Ind. Eng. Chem.* **2019**, *76*, 17–38. [CrossRef]
9. Barakat, M. New trends in removing heavy metals from industrial wastewater. *Arab. J. Chem.* **2011**, *4*, 361–377. [CrossRef]
10. Fu, F.; Wang, Q. Removal of heavy metal ions from wastewaters: A review. *J. Environ. Manag.* **2011**, *92*, 407–418. [CrossRef]
11. Qasem, N.A.; Mohammed, R.H.; Lawal, D.U. Removal of heavy metal ions from wastewater: A comprehensive and critical review. *Npj Clean Water* **2021**, *4*, 36. [CrossRef]
12. Kouhkord, A.; Hassani, F.; Amirmahani, M.; Golshani, A.; Naserifar, N.; Moghanlou, F.S.; Beris, A.T. Controllable Microfluidic System through Intelligent Framework: Data-Driven Modeling, Machine Learning Energy Analysis, Comparative Multiobjective Optimization, and Experimental Study. *Ind. Eng. Chem. Res.* **2024**, *63*, 13326–13344. [CrossRef]
13. Zhang, K.; Luo, X.; Yang, L.; Chang, Z.; Luo, S. Progress toward hydrogels in removing heavy metals from water: Problems and solutions—A review. *ACS EST Water* **2021**, *1*, 1098–1116. [CrossRef]
14. Chimera, M.; Barazesh, A.; Zandi, M.; Zarkesh, I.; Moghaddam, A.; Borjian, P.; Chimehrad, R.; Asghari, A.; Akbarnejad, Z.; Khonakdar, H.A. Tissue engineered scaffold fabrication methods for medical applications. *Int. J. Polym. Mater. Polym. Biomater.* **2023**, *72*, 1455–1479. [CrossRef]

15. Moghaddam, A.; Bahrami, M.; Mirzadeh, M.; Khatami, M.; Simorgh, S.; Chimehrad, M.; Kruppke, B.; Bagher, Z.; Mehrabani, D.; Khonakdar, H.A. Recent trends in bone tissue engineering: A review of materials, methods, and structures. *Biomed. Mater.* **2024**, *19*, 042007. [CrossRef]
16. Zhang, W.; Ou, J.; Wang, B.; Wang, H.; He, Q.; Song, J.; Zhang, H.; Tang, M.; Zhou, L.; Gao, Y. Efficient heavy metal removal from water by alginate-based porous nanocomposite hydrogels: The enhanced removal mechanism and influencing factor insight. *J. Hazard. Mater.* **2021**, *418*, 126358. [CrossRef] [PubMed]
17. Zhu, H.; Chen, S.; Luo, Y. Adsorption mechanisms of hydrogels for heavy metal and organic dyes removal: A short review. *J. Agric. Food Res.* **2023**, *12*, 100552. [CrossRef]
18. Zhang, Y.; Zhao, M.; Cheng, Q.; Wang, C.; Li, H.; Han, X.; Fan, Z.; Su, G.; Pan, D.; Li, Z. Research progress of adsorption and removal of heavy metals by chitosan and its derivatives: A review. *Chemosphere* **2021**, *279*, 130927. [CrossRef]
19. Pathak, P.; Borjian, P.; Chimera, M.; Cho, H.J. A ppt level PFOS (Perfluorooctanesulfonic Acid) sensor based on an eco-friendly chitosan biopolymer. In Proceedings of the 22nd International Conference on Solid-State Sensors, Actuators and Microsystems (Transducers), Kyoto, Japan, 25–29 June 2023.
20. Zhang, J.; Chen, Z.; Kankala, R.K.; Wang, S.-B.; Chen, A.-Z. Self-propelling micro-/nano-motors: Mechanisms, applications, and challenges in drug delivery. *Int. J. Pharm.* **2021**, *596*, 120275. [CrossRef]
21. Wu, Y.; Boymelgreen, A.; Yossifon, G. Micromotor-mediated label-free cargo manipulation. *Curr. Opin. Colloid Interface Sci.* **2022**, *61*, 101611. [CrossRef]
22. Choi, H.; Jeong, S.H.; Kim, T.Y.; Yi, J.; Hahn, S.K. Bioinspired urease-powered micromotor as an active oral drug delivery carrier in stomach. *Bioact. Mater.* **2022**, *9*, 54–62. [CrossRef]
23. Lin, R.; Yu, W.; Chen, X.; Gao, H. Self-Propelled Micro/Nanomotors for Tumor Targeting Delivery and Therapy. *Adv. Healthc. Mater.* **2021**, *10*, 2001212. [CrossRef] [PubMed]
24. Wang, K.; Ma, E.; Cui, H.; Hu, Z.; Wang, H. Bioinspired Self-Propelled Micromotors with Improved Transport Efficiency in the Subsurface Environment for Soil Decontamination. *Adv. Funct. Mater.* **2023**, *33*, 2307632. [CrossRef]
25. Hassani, F.; Sadegh Moghanlou, F.; Minaei, A.; Vajdi, M.; Golshani, A.; Kouhkord, A.; Dehghani, T. An efficient framework for controllable micromixer design through the fusion of data-driven modeling and machine learning insights: Numerical and experimental analysis. *Phys. Fluids* **2024**, *36*, 033606. [CrossRef]
26. Zeng, H.; Wang, Y.; Jiang, T.; Xia, H.; Gu, X.; Chen, H. Recent progress of biomimetic motions—From microscopic micro/nanomotors to macroscopic actuators and soft robotics. *RSC Adv.* **2021**, *11*, 27406–27419. [CrossRef] [PubMed]
27. Zarei, M.; Zarei, M. Self-propelled micro/nanomotors for sensing and environmental remediation. *Small* **2018**, *14*, 1800912. [CrossRef]
28. Hassani, F.; Kouhkord, A.; Golshani, A.; Amirmahani, M.; Moghanlou, F.S.; Naserifar, N.; Beris, A.T. Micro-electro-mechanical acoustofluidic mixing system: A response surface-metaheuristic machine learning fusion framework. *Expert Syst. Appl.* **2024**, *249*, 123638. [CrossRef]
29. Archer, R.; Kubodera, Y.; Ebbens, S.; Matsuo, M.; Shin-ichiro, M.N. Active mixing by self-propelled Janus sponge Marangoni motors with self-maintaining surface tension gradients. *ChemRxiv* **2024**. [CrossRef]
30. Luo, C.; Li, H.; Qiao, L.; Liu, X. Development of surface tension-driven microboats and microflotillas. *Microsyst. Technol.* **2012**, *18*, 1525–1541. [CrossRef]
31. Lin, C.-H.; Kinane, C.; Zhang, Z.; Pena-Francesch, A. Functional Chemical Motor Coatings for Modular Powering of Self-Propelled Particles. *ACS Appl. Mater. Interfaces* **2022**, *14*, 39332–39342. [CrossRef]
32. Billard, G.; Bruyant, C. Sur un mode particulier de locomotion de certains stenus. *CR Soc. Biol.* **1905**, *59*, 102.
33. Velarde, M.G.; Zeytounian, R.K. *Interfacial Phenomena and the Marangoni Effect*; Springer: Berlin/Heidelberg, Germany, 2002.
34. Lang, C.; Seifert, K.; Dettner, K. Skimming behaviour and spreading potential of Stenus species and Dianous coeruleus (Coleoptera: Staphylinidae). *Naturwissenschaften* **2012**, *99*, 937–947. [CrossRef] [PubMed]
35. Soh, S.; Bishop, K.J.; Grzybowski, B.A. Dynamic self-assembly in ensembles of camphor boats. *J. Phys. Chem. B* **2008**, *112*, 10848–10853. [CrossRef] [PubMed]
36. Liakos, I.L.; Salvagnini, P.; Scarpellini, A.; Carzino, R.; Beltran, C.; Mele, E.; Murino, V.; Athanassiou, A. Biomimetic locomotion on water of a porous natural polymeric composite. *Adv. Mater. Interfaces* **2016**, *3*, 1500854. [CrossRef]
37. Wu, H.; Chen, Y.; Xu, W.; Xin, C.; Wu, T.; Feng, W.; Yu, H.; Chen, C.; Jiang, S.; Zhang, Y.; et al. High-performance Marangoni hydrogel rotors with asymmetric porosity and drag reduction profile. *Nat. Commun.* **2023**, *14*, 20. [CrossRef]
38. Lauga, E.; Davis, A.M. Viscous marangoni propulsion. *J. Fluid Mech.* **2012**, *705*, 120–133. [CrossRef]
39. Pike, J.K.; Ho, T.; Wynne, K.J. Water-induced surface rearrangements of poly (dimethylsiloxane– urea– urethane) segmented block copolymers. *Chem. Mater.* **1996**, *8*, 856–860. [CrossRef]
40. Senshu, K.; Yamashita, S.; Mori, H.; Ito, M.; Hirao, A.; Nakahama, S. Time-resolved surface rearrangements of poly (2-hydroxyethyl methacrylate-block-isoprene) in response to environmental changes. *Langmuir* **1999**, *15*, 1754–1762. [CrossRef]
41. Lewis, K.B.; Ratner, B.D. Observation of surface rearrangement of polymers using ESCA. *J. Colloid Interface Sci.* **1993**, *159*, 77–85. [CrossRef]

42. Zhu, H.; Xu, B.; Wang, Y.; Pan, X.; Qu, Z.; Mei, Y. Self-powered locomotion of a hydrogel water strider. *Sci. Robot.* **2021**, *6*, eabe7925. [CrossRef]
43. Young, T. III. An essay on the cohesion of fluids. *Philos. Trans. R. Soc. Lond.* **1805**, *95*, 65–87. [CrossRef]

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Article

Structural Optimization Design of Magnetolectric Thin-Film Antenna for Bandwidth and Radiation Enhancement

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Abstract: The acoustically actuated nanomechanical magnetolectric (ME) antennas represent a promising new technology that can significantly reduce antenna size by 1–2 orders of magnitude compared to traditional antennas. However, current ME antennas face challenges such as low antenna gain and narrow operating bandwidth, limiting their engineering applications. In this paper, we enhance the bandwidth and radiation performance of ME antennas through structural optimization, leveraging theoretical analysis and numerical simulations. Our findings indicate that optimizing the inner diameter of the ring-shaped ME antenna can elevate the average stress of the magnetic layer, leading to improved radiation performance and bandwidth compared to circular ME antennas. We establish an optimization model for the radiation performance of the ME antenna and conduct shape optimization simulations using COMSOL Multiphysics. The results of the Multiphysics field optimization align with the stress concentration theory, demonstrating a strong correlation between the radiation performance and bandwidth of the ME antenna with the average stress of the magnetic film. The resonant frequency in the thickness vibration mode is determined to be 170 MHz. Furthermore, shape optimization can enhance the bandwidth by up to 104% compared to circular ME antenna structures of the same size.

Keywords: antenna miniaturization; magnetolectric antenna; structural optimization; bandwidth; radiation performance

1. Introduction

Antenna is a tool for receiving and transmitting electromagnetic (EM) waves, is an important part of modern communication systems, and is widely used in smart phones, smart watches, laptops and other devices. One of the most important challenges in the field of antenna research is the miniaturization of its size [1–3]. Traditional electric small antenna depends on EM oscillation. The size of the shortest antenna cannot be less than 1/10 of the EM wavelength; otherwise, the radiation and reception of EM waves become ineffective. This principle fundamentally limits the miniaturization of the antenna. Therefore, it becomes crucial to explore new mechanisms for EM wave reception and new mechanisms for the miniaturization of antennas.

In recent years, the concept of mechanical antennas, whose dimensions do not depend on the wavelength of EM waves, has been proposed and has attracted the attention of researchers [4,5]. ME antenna as a kind of mechanical antenna, based on the principle of the ME coupling effect, consists of an ME-composite-laminated structure. Yao et al. proposed for the first time the concept of using piezoelectric and magnetostrictive layers as antenna radiation, and used finite difference time-domain (FDTD) method to model sound waves and EM radiation [4,6]. Nan et al. firstly reported ME antennas and conducted experimental tests [7]. Xu et al. analyzed the converse ME effect and EM radiation of ME-laminated structures, and the results showed that the radiation of ME antennas is an

ideal magnetic dipole model [8]. In 2021, Sun et al. developed an ME antenna utilizing an SMR structure, achieving a center frequency of 1.575 GHz through innovative material and structural configurations. This design yielded a -5 dB bandwidth of 42 MHz, along with a fractional bandwidth of 2.6% [9]. In 2022, Li et al. used a Mason model to design a two-layer back-cavity-type ME antenna and used the COMSOL Multiphysics V6.0 to carry out ME performance simulation. The authors finally manufactured an ME antenna operating at 2.45 GHz with a gain of -15.59 dB [10]. Shi et al. proposed a multiphysical field coupling model considering nonlinear magnetostrictive models to model ME antennas [11]. However, the ME antenna still has issues with narrow bandwidth and low gain [1,12–14].

Nan et al. introduced the FBAR ME antenna, boasting a center frequency of 2.53 GHz and a -5 dB threshold based on S11, with a bandwidth of 75 MHz and a fractional bandwidth of 2.96% [7]. In 2019, Nikitin et al. conducted simulations on an ME antenna featuring a composition of ferromagnetic yttrium iron garnet and lead zirconate titanate film. The simulated results revealed a center frequency of 2.8 GHz, accompanied by an approximately 73.3 MHz -5 dB bandwidth and a fractional bandwidth of 2.6% [15]. Zaeimbashi et al. made strides in 2021 by unveiling three ME antennas designed in parallel, each boasting a center frequency of 2.51 GHz. These designs exhibited a -5 dB bandwidth of 60 MHz and a fractional bandwidth of 2.39% [14]. Yun et al. proposed a bandwidth-enhanced ME antenna made of Mo/AlN/FeGa sandwich layers composed of three different resonant regions; the antenna achieves a fractional bandwidth of 2.7% (-3 dB) while maintaining the advantage of small size [16]. In the same year, they manufactured and demonstrated an ME antenna driven by a high-overtone bulk-acoustic resonator (HBAR), which has a floating potential structure (FPA). Compared with directly grounded HBAR-ME antennas, FPA can significantly improve the gain and radiation efficiency of HBAR-ME antennas by over 10 dB [17]. In 2023, Shi et al. proposed a new acoustic-actuated antenna using embedded ME composites to improve the radiation of ME antennas through the enhanced strain transfer at the interfaces between different phases [18]. Jin et al. proposed a microbridge structure and array connection method for low-frequency thin-film ME antennas to address the problems of narrow working bandwidths and the weak radiation intensity of ME antennas, and verified them through experiments [19]. Luo et al. utilizes parallel and series antenna array topology to achieve a profound gain and radiation efficiency enhancement without degrading impedance mismatch and quality factor of ME resonators [20]. Dong et al. proposed an ME antenna array consisting of three units, which are constructed in a sandwich stack. The -3 dB operating bandwidth of 152.4–172.8 KHz is achieved, and the relative bandwidth is 12.5% [21].

In summary, current research on ME antennas based on thin-film bulk-acoustic resonators (FBARs) has mainly focused on operations in the GHz range, while relatively little attention has been paid to low-frequency variants below 200 MHz. Particularly lacking is the exploration of enhancing ME antenna bandwidth through structural design considerations. This paper aims to improve the bandwidth and radiation performance of ME antenna through structural optimization. We employ mathematical models and the Multiphysics field simulation software to optimize the internal radius of the circular ME antenna. Our aim is to enhance the radiation power of the antenna and determine the most optimal internal radius. The simulation results align with the predictions of stress concentration theory, demonstrating a consistent trend.

2. Theory and Simulation of Me Antenna

The 3D structure of a ring-shaped ME antenna is shown in Figure 1, where the illustration magnifies the core structure of an ME-laminated film. The composite magnetostrictive film is used as the core film stack of the antenna, and the piezoelectric layer below it generates acoustic excitation. Since the thickness of the electrode layer is very thin compared with the piezoelectric layer and magnetic layer, we regard the electrode layer as a finitely thin plane.

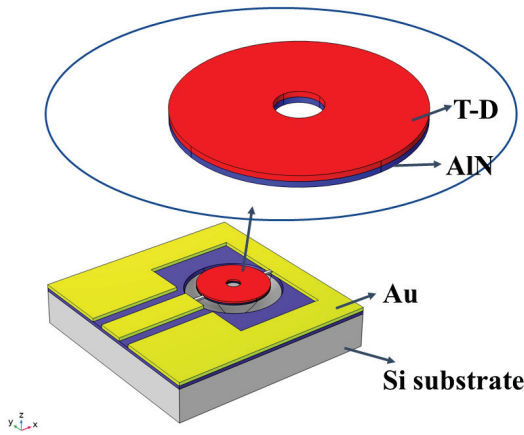


Figure 1. A 3D schematic diagram of ring-shaped ME antenna on released Si substrate.

In this paper, we demonstrate the ME antenna operating at HF frequencies based on the strong ME coupling EM and bulk-acoustic wave in the resonant ME heterostructures. To analyze the impact of the inner radius variations in the performance of the ME antenna, we employed COMSOL Multiphysics V6.0 to carry out a comprehensive Multiphysics field analysis on ME antennas with distinct inner radii. Initially, we conducted an S-parameter analysis to obtain the resonance frequency of ME antennas with varying inner radii. Subsequently, we analyzed the average stress in the magnetic layer at the resonance frequency. Finally, we utilized the piezoelectric and magnetic fields and EM-wave frequency-domain modules to simulate the entire ME antenna process and obtain its radiation pattern. The operational frequency of the ME antenna discussed in this article is 170 MHz. The piezoelectric layer employs AlN, while the magnetic layer is composed of FeGaB. The equivalent wave speed of the M-P laminated structure is given as [22]

$$v_{eq} = \sqrt{\frac{\frac{n_m}{s_{B11}} + \frac{n_p}{s_{E11}}}{n_m \rho_m + n_p \sigma_p}} \quad (1)$$

$$s_B = (c_H + e_H^T \mu_s^{-1} e_H)^{-1}$$

$$s_E = c_E^{-1}$$

where s_B and s_E are the mechanical compliance constants of the M phase and P phase, respectively; n_m and n_p are the volume ratios of the M phase and P phase, respectively. The simulation is validated against the existing simulation performed on FeGaB/AlN [7]. It can be observed that the proposed scheme is able to capture the S11 curve and the corresponding resonance frequency in the illustration in Figure 2. The two sets of results are in very close agreement, indicating that the simulation may be valid for simulating the electromechanical characteristics. The performance of the ME antenna is studied in the following subsection in which we present the stress–strain analysis, S11 curve, and far-field radiation of the ME antenna.

Additionally, since the magnetoelectric antenna was not fabricated this time, we validated the accuracy of our simulation method using both the simulation and experimental data from a previously designed 2.45 GHz magnetoelectric antenna, as referenced in [10]. The simulation method employed in this paper is identical to the one used in the previous study to simulate the electromechanical characteristics of the magnetoelectric antenna. The comparison between the simulated S11 data and the experimental data is shown in the Figure 3:

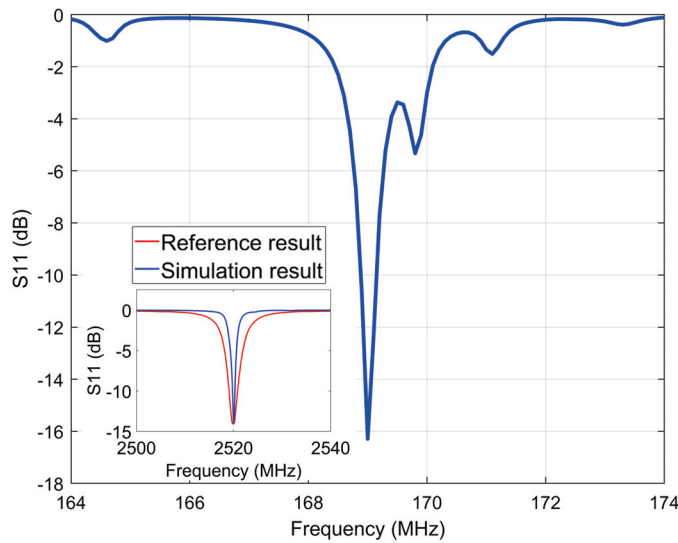


Figure 2. Simulated reflection coefficient S11 of the ME antenna.

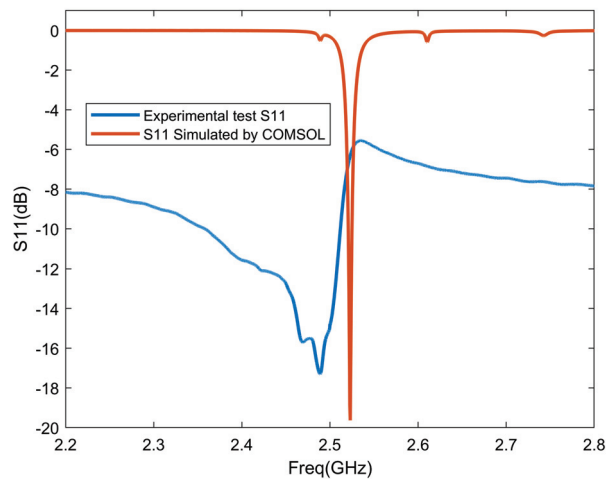


Figure 3. Comparison of simulated S11 and experimental S11 of magnetolectric antenna.

It can be observed that there is a slight difference between the resonance point of the simulated ME antenna and that of the actual tested antenna; however, both are essentially close to the designed resonance point. The primary reasons for this discrepancy include the precision of the processing equipment, environmental factors, and the inconsistent quality of the prepared film during fabrication. Additionally, there are some variations between the parameters of the materials used in the simulation and those of the actual processed materials. These factors contribute to the differences between the simulated S11 data and the actual test data of the ME antenna.

We consider a two-layer ME-laminated composite structure with $R = 484 \mu\text{m}$, $t_m = 20 \mu\text{m}$ (thickness of magnetic layer), and $t_p = 20 \mu\text{m}$ (thickness of piezoelectric layer). The piezoelectric material is AlN. The magnetostrictive (Terfenol-D) material parameters are obtained from [12]. The piezoelectric module in COMSOL Multiphysics was used for the electromechanical simulation of the ME antenna. Figure 2 shows the simulated reflection coefficient (S11) of the ME antenna, showing an estimated electromechanical resonance frequency of 169 MHz.

3. Result and Discussion

Firstly, we performed simulations for the S11 parameters of each structure with inner radii of 10 μm , 50 μm , 80 μm , 140 μm , 200 μm , 260 μm , and 320 μm , respectively. Secondly, we obtained the average stress values of the magnetic layer for various inner diameters, while considering the frequency variation. The simulation data are visualized in Figure 4.

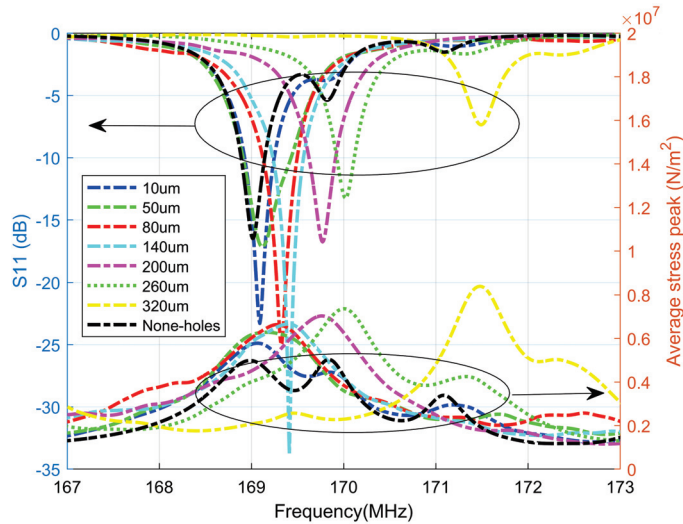


Figure 4. S11 and the average stress peak of the magnetosphere at different radii.

As shown in Figure 4, different radii will affect the resonance frequency and -10 dB bandwidth of the ME antenna. Except for the resonance frequency with a radius of 320 μm at 171.5 MHz, the resonance points of other ME antennas are ultimately between 169 and 170 MHz.

Meanwhile, we can see that the S11 resonance point corresponds to the maximum von Mises peak stress in the magnetic layer, verifying that the ME antenna is in electromechanical resonance at this frequency point. The bandwidth data are shown in Table 1.

Table 1. Bandwidth at different inner radii.

The Radius (μm)	-10 dB Bandwidth (MHz)	Fractional Bandwidth (FBW)
0	0.2533	0.1498%
10	0.3211	0.1899%
50	0.5212	0.3082%
80	0.3717	0.2195%
140	0.3324	0.1962%
200	0.2888	0.1702%
260	0.1896	0.1115%
320	0	0%

It can be found that, compared to circular ME antennas, as the inner radius increases, the bandwidth first increases and then decreases. When the radius is 50 μm , the bandwidth reaches its maximum value, which is 0.5212 MHz, with an increase of 104%. It is obvious that the stress concentration effect generated by the inner radius is effective for the bandwidth of the ME antenna.

Furthermore, we compared the existing papers related to bandwidth and gain indicators in the literature and listed in Table 2.

Table 2. Information of ME antenna bandwidth, gain and dimensions as reported in the literature.

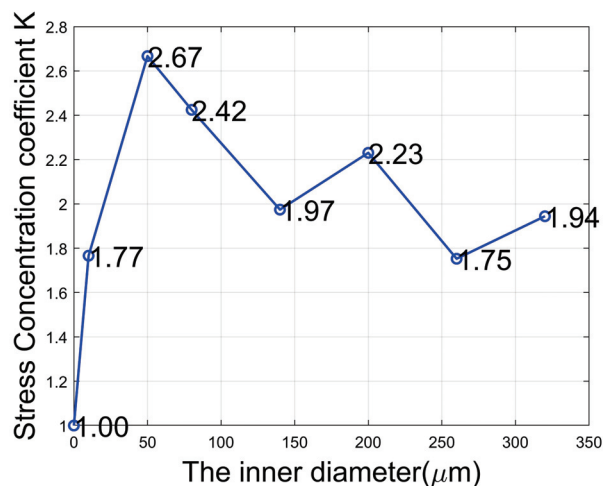
Paper	Bandwidth	Resonant Frequency (f0)	Dimension	Gain
Ref. [16]	−3 dB 21 MHz	800 MHz	0.49 mm ²	None
Ref. [9]	−3 dB 42 MHz	1.575 GHz	3 mm × 3 mm	−19.4 dBi
Ref. [23]	−3 dB 725 Hz + 1179 Hz	36 kHz	50 mm × 6 mm	None
Ref. [7]	None	2.53 GHz	0.7 mm × 0.8 mm	−18 dBi
This paper	−10 dB 0.5212 MHz	170 MHz	0.5 mm × 0.5 mm	−21.59 dBi

As can be seen from the table, compared with the literature that gives the size, bandwidth and gain data of ME antennas, the size of the ME antenna in this article is the smallest: 0.25 mm². Except for the third antenna, the rest are ME antennas based on thin-film bulk-acoustic resonators, and our antenna frequency is currently the lowest.

Furthermore, we use COMSOL software to observe the average stress magnetic layers with different radii. Here, we follow the concept of stress concentration and set a stress concentration coefficient K as follows:

$$K = \frac{\sigma_r}{\sigma_0} \quad (2)$$

where σ_r represents the average stress of the magnetic layer with an inner radius of r , and σ_0 is the average stress of the magnetic layer without an inner radius. We use COMSOL Multiphysics to extract the average stress of the magnetic layer under different inner radii. For the convenience of observation, we draw a line chart to display the data. From Figure 5, it can be seen that when the inner radius is 50 μm , the average stress of the magnetic layer is the highest, and compared to the circular ME antenna, the stress concentration coefficient is 2.67.

**Figure 5.** Stress concentration coefficient magnetic layer.

Given the critical role of the converse ME coefficient in the radiation performance of ME antennas, we conducted simulations to analyze the effects of varying the ring radius on this coefficient. The expression for the converse ME coefficient is given by: $\alpha_{CME} = |H_x/E_x|$. This coefficient is commonly used to characterize the ME coupling in a ME resonator, where H_x represents the average AC magnetic field induced in the magnetic layer, and E_z denotes the average electric field in the piezoelectric phase. Furthermore, we modeled the near-field radiation of the ME antenna by surrounding it with an air domain of radius $r = 500 \mu\text{m}$. Applying a 1mV AC voltage to the piezoelectric layer induces alternating stress in the layer. This stress is transferred to the magnetic layer, generating the piezomagnetic effect. The simulation incorporates the piezoelectric effect of the piezoelectric layer using the

AC/DC and solid mechanics modules within the structural mechanics module, while the magnetostrictive layer is modeled using the magnetic field and structural mechanics modules in the AC/DC module. We simulated the corresponding converse ME coefficients at various radii, and the results are shown in Figure 6:

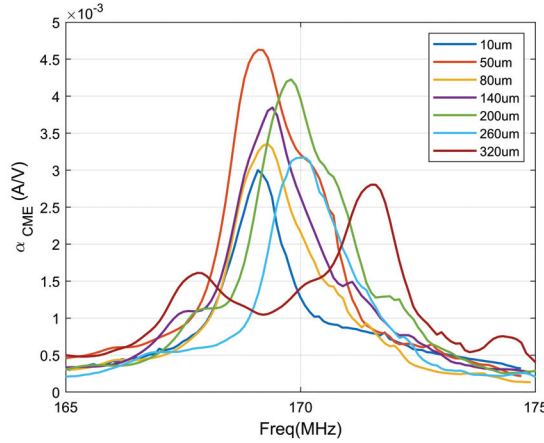


Figure 6. Variation of the converse ME coefficient and the inner diameter of the ring ME antenna.

We observed that the relationship between the inverse ME coefficient and the radius initially increases and then decreases as the radius increases. The inverse ME coefficient reaches its peak at 50 μm . This finding aligns with the law of the stress concentration coefficient, indicating that higher average stress in the magnetic layer results in stronger radiation from the ME antenna.

By using the one-dimensional ME-laminated-structure analytical model given in the Appendix A, the expression for the radiation power of a two-layer ME antenna is obtained as follows:

$$P_{rad} = \frac{A\omega^2 d^2 T_0^2 d_H^2}{16\eta_0} \quad (3)$$

From this expression, it can be reflected that the radiation power of an ME antenna is not only related to the average stress of the magnetic layer of the ME antenna, but also to the radiation area, the operating frequency and material parameters. For a ring-shaped ME antenna, the expression for its radiation area A is $A = \pi(R^2 - r^2)$, where R is the outer diameter of the ME antenna and r is the inner diameter. Based on this, we establish an optimization model targeting the radiation performance of ME antennas:

$$\left\{ \begin{array}{l} \text{Find : } r \in r_1, r_2, r_3 \dots, r_n \\ \text{Max : } P_{rad} = \frac{\pi(R^2 - r^2)\omega^2 d^2 T_0^2 d_H^2}{16\eta_0} \\ \text{s.t. } 0 < r < R \\ R = \text{const}, u_R = 0 \end{array} \right. \quad (4)$$

where ω is the circular frequency at which the ME antenna operates, d is the total thickness of the ME antenna, d_H is the piezomagnetic coefficient, and η_0 is the wave impedance in vacuum. Then, we use COMSOL for shape optimization, and the optimization results are shown in Figure 7.

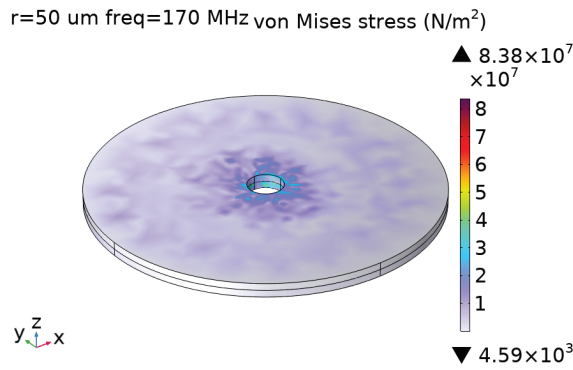


Figure 7. The results of shape optimization.

Lastly, we perform a direct radiation simulation of the ME antenna using COMSOL Multiphysics. We use four modules: electrostatic, solid mechanics, magnetic field, and EM waves to perform the simulation in the following steps:

- (1) The behavior of the ME antenna's near field is simulated using solid mechanics, electrostatics, and magnetic fields. Specifically, solid mechanics and electrostatics are applied to model the piezoelectric phase, while solid mechanics and magnetic fields are employed for the magnetostrictive phase.
- (2) The far field of the ME antenna is simulated using the EM wave module. In the vicinity of the ME antenna, a sphere is employed. Through the principle of surface equivalence, the equivalent magnetic current and equivalent current of the sphere are derived. These equivalent values are then utilized as the new radiation source for calculating the far-field radiation.

As shown in the inset of Figure 8, the radiation pattern of an ME antenna is similar to that of a dipole. Here, we only present the radiation pattern of the ME antenna with an inner radius of 50 μm ; its far-field mode maximum value is 1.79 $\mu\text{V}/\text{m}$ and the minimum value is 0.28 $\mu\text{V}/\text{m}$. The radiation patterns of other ME antennas are also similar to that of a dipole, with the difference being the maximum value. Here, we list the maximum values of far-field modes corresponding to different inner radius in Table 3. The far-field mode of the electric field is a physical quantity related to distance in the COMSOL software, which can reflect the antenna radiation capability. Under the spherical air domain simulation, ME antennas with different inner diameters have different maximum values of far-field modes, reflecting changes in the antenna's radiation ability as the inner diameter changes.

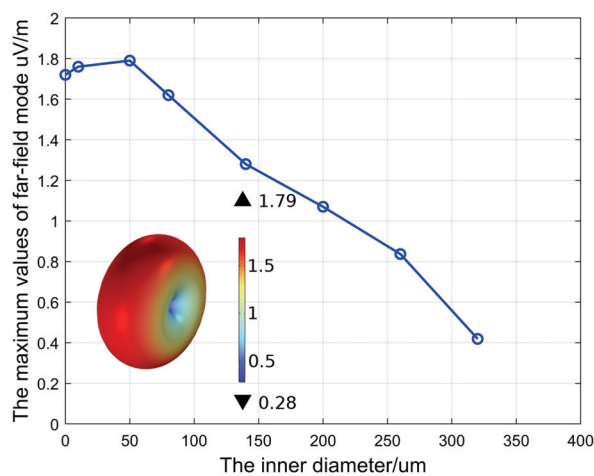


Figure 8. Maximum values of far-field modes under different inner radii.

Table 3. Maximum value of far-field modes under different inner radii.

The radius (μm)	0	10	50	80	140	200	260	320
Maximum value ($\mu\text{V/m}$)	1.72	1.76	1.79	1.62	1.28	1.07	0.84	0.44

From Figure 6, it can be clearly seen that as the inner diameter increases, the radiation capability first increases and then weakens. When the inner radius is 50 μm , the maximum far-field mode reaches its maximum value, which is 1.79 $\mu\text{V/m}$. Regarding the calculation method of the antenna gain in this paper, the directivity coefficient of the ME antenna obtained by COMSOL simulation is $D = 1.51$, and based on the radiation efficiency η given in Ref. [7] as $\eta = 0.28\%$, the gain of the antenna can be obtained: $\text{Gain} = -21.59 \text{ dBi}$.

4. Conclusions

In summary, the primary objective of this study is to enhance the bandwidth and radiation performance of ME antennas through structural optimization design. It has been determined that the ring-shaped ME antenna exhibits a comparatively higher average stress in its magnetic layer, which theoretically enables the generation of greater radiation power. To corroborate this finding, a comprehensive full-process simulation of the ME antenna was conducted, and the results obtained were consistent with the aforementioned conclusions.

This paper investigates the full-field coupling model of ME antennas and explores the potential of enhancing their bandwidth and radiation performance through structural optimization. These findings offer a promising avenue for designing ME antennas that are practically applicable in engineering applications.

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Abbreviations

The following abbreviations are used in this manuscript:

ME	magnetoelectric
EM	electromagnetic
FDTD	finite difference time-domain
HBAR	high-overtone bulk-acoustic resonator
FPA	floating potential structure

Appendix A

The radiation capability of the antenna can usually be described by the radiation Q factor. The Q factor represents the ratio of the average stored energy in the antenna structure to the radiation power per cycle:

$$Q = \omega \frac{W_T}{P_{rad}} \quad (\text{A1})$$

The potential energy in the ME antenna is equal to the sum of the potential energy in the piezoelectric and the piezomagnetic layers:

$$W_{PE} = \frac{1}{2} \iiint S \cdot T dv + \frac{1}{2} \iiint D \cdot E dv \quad (A2)$$

$$W_{PM} = \frac{1}{2} \iiint S \cdot T dv + \frac{1}{2} \iiint B \cdot H dv \quad (A3)$$

The dynamic strain is transferred from the piezoelectric layer to the magnetic layer, and then the dynamic strain excites a uniform dynamic magnetic flux within the magnetic layer. According to Faraday's law, this dynamic magnetic flux generates a dynamic electric field, which varies linearly along the thickness dimension until it reaches above the surface of the magnetic layer where an aperture electric field is formed. Denoting the aperture electric field by E , the average radiated power of the EM wave is thus calculated to be

$$P_{rad} = \frac{1}{2\eta_0} \iint |E_0|^2 ds \quad (A4)$$

Using Faraday's law $E = \omega h |B|$, where h is the thickness of the magnetic layer, Formula (8) can be written as

$$P_{rad} = \frac{\omega^2 h^2 d_H^2}{2\eta_0} \iint |T_H|^2 ds \quad (A5)$$

If the ME antenna has a two-layer structure, we can obtain a unique Q factor expression:

$$Q = \frac{A\omega^2 d^2 T_0^2 d_H^2}{16\eta_0} \quad (A6)$$

References

1. Mosallaei, H.; Sarabandi, K. Antenna miniaturization and bandwidth enhancement using a reactive impedance substrate. *IEEE Trans. Antennas Propag.* **2004**, *52*, 2403–2414. [CrossRef]
2. Skrivervik, A.K.; Zurcher, J.F.; Staub, O.; Mosig, J. PCS antenna design: The challenge of miniaturization. *IEEE Antennas Propag. Mag.* **2001**, *43*, 12–27. [CrossRef] [PubMed]
3. Gianvittorio, J.P.; Rahmat-Samii, Y. Fractal antennas: A novel antenna miniaturization technique, and applications. *IEEE Antennas Propag. Mag.* **2002**, *44*, 20–36. [CrossRef]
4. Yao, Z.; Wang, Y.E.; Keller, S.; Carman, G.P. Bulk acoustic wave-mediated multiferroic antennas: Architecture and performance bound. *IEEE Trans. Antennas Propag.* **2015**, *63*, 3335–3344. [CrossRef]
5. Domann, J.P.; Carman, G.P. Strain powered antennas. *J. Appl. Phys.* **2017**, *121*, 044905. [CrossRef]
6. Yao, Z.; Tok, R.U.; Itoh, T.; Wang, Y.E. A multiscale unconditionally stable time-domain (MUST) solver unifying electrodynamics and micromagnetics. *IEEE Trans. Microw. Theory Tech.* **2018**, *66*, 2683–2696. [CrossRef]
7. Nan, T.; Lin, H.; Gao, Y.; Matyushov, A.; Yu, G.; Chen, H.; Sun, N.; Wei, S.; Wang, Z.; Li, M.; et al. Acoustically actuated ultra-compact NEMS magnetoelectric antennas. *Nat. Commun.* **2017**, *8*, 296. [CrossRef]
8. Xu, G.; Xiao, S.; Li, Y.; Wang, B.Z. Modeling of electromagnetic radiation-induced from a magnetostrictive/piezoelectric laminated composite. *Phys. Lett. A* **2021**, *385*, 126959. [CrossRef]
9. Sun, N. RF Magnetoelectric Devices for Communication, Sensing, and Power Electronics. Ph.D. Thesis, Northeastern University, Boston, MA, USA, 2021.
10. Li, N.; Li, X.; Xu, B.; Zheng, B.; Zhao, P. Design and optimization of a micron-scale magnetoelectric antenna based on acoustic excitation. *Micromachines* **2022**, *13*, 1584. [CrossRef]
11. Shi, Y.; Xu, J.; Wang, Y.; Ye, J. Tuning the Performance of Acoustically Actuated Magnetoelectric Antenna via Magnetic and Mechanical Stimuli. *IEEE Trans. Antennas Propag.* **2023**, *71*, 3959–3967. [CrossRef]
12. Dong, C.; He, Y.; Li, M.; Tu, C.; Chu, Z.; Liang, X.; Chen, H.; Wei, Y.; Zaeimbashi, M.; Wang, X.; et al. A portable very low frequency (VLF) communication system based on acoustically actuated magnetoelectric antennas. *IEEE Antennas Wirel. Propag. Lett.* **2020**, *19*, 398–402. [CrossRef]
13. Zaeimbashi, M.; Lin, H.; Dong, C.; Liang, X.; Nasrollahpour, M.; Chen, H.; Sun, N.; Matyushov, A.; He, Y.; Wang, X.; et al. NanoNeuroRFID: A wireless implantable device based on magnetoelectric antennas. *IEEE J. Electromagn. RF Microwaves Med. Biol.* **2019**, *3*, 206–215. [CrossRef]

14. Zaeimbashi, M.; Nasrollahpour, M.; Khalifa, A.; Romano, A.; Liang, X.; Chen, H.; Sun, N.; Matyushov, A.; Lin, H.; Dong, C.; et al. Ultra-compact dual-band smart NEMS magnetoelectric antennas for simultaneous wireless energy harvesting and magnetic field sensing. *Nat. Commun.* **2021**, *12*, 3141. [CrossRef] [PubMed]
15. Nikitin, A.; Petrov, R.; Khavanova, M.; Tatarenko, A.; Bichurin, M. Modeling of magnetoelectric effect in multiferroic antenna. In Proceedings of the 2019 Photonics & Electromagnetics Research Symposium-Spring (PIERS-Spring), Rome, Italy, 17–20 June 2019; pp. 953–956.
16. Yun, X.; Lin, W.; Hu, R.; Liu, Y.; Wang, X.; Yu, G.; Zeng, Z.; Zhang, X.; Zhang, B. Bandwidth-enhanced magnetoelectric antenna based on composite bulk acoustic resonators. *Appl. Phys. Lett.* **2022**, *121*, 033501. [CrossRef]
17. Yun, X.; Lin, W.; Hu, R.; Liu, Y.; Wu, D.; Zhang, X.; Zeng, Z.; Zhang, B. Radiation-enhanced acoustically driven magnetoelectric antenna with floating potential architecture. *Appl. Phys. Lett.* **2022**, *121*, 203504. [CrossRef]
18. Shi, Y.; Xu, J.; Ye, J. Enhanced near-field radiation of acoustic-actuated antennas using embedded magnetoelectric composites. *Compos. Struct.* **2023**, *314*, 116975. [CrossRef]
19. Jin, J.; Jing, L.; Zuo, C.; Ma, Z.; Shi, Y.; Yang, X.; Chen, S. Array Study of VLF Thin-Film Magnetoelectric Antenna with a Microbridge Structure. *Micromachines* **2023**, *15*, 11. [CrossRef] [PubMed]
20. Luo, B.; Liang, X.; Chen, H.; Sun, N.; Lin, H.; Sun, N.X. Gain Enhancement and Ground Plane Immunity of Mechanically Driven Thin-Film Bulk Acoustic Resonator Magnetoelectric Antenna Arrays. *Adv. Funct. Mater.* **2024**, 2403244. [CrossRef]
21. Dong, B.; Yan, Z.; Zhang, Y.; Han, T.; Zhou, H.; Wang, Y. Analysis of near field mutual coupling in wideband magnetoelectric antennas array. *J. Appl. Phys.* **2023**, *134*, 114103. [CrossRef]
22. Blackburn, J.F.; Vopsaroiu, M.; Cain, M.G. Verified finite element simulation of multiferroic structures: Solutions for conducting and insulating systems. *J. Appl. Phys.* **2008**, *104*, 074104. [CrossRef]
23. Xiao, N.; Wang, Y.; Chen, L.; Wang, G.; Wen, Y.; Li, P. Low-frequency dual-driven magnetoelectric antennas with enhanced transmission efficiency and broad bandwidth. *IEEE Antennas Wirel. Propag. Lett.* **2022**, *22*, 34–38. [CrossRef]

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Article

Design and Optimization of MEMS Resonant Pressure Sensors with Wide Range and High Sensitivity Based on BP and NSGA-II

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Abstract: With the continuous progress of aerospace, military technology, and marine development, the MEMS resonance pressure sensor puts forward the requirements of not only a wide range but also high sensitivity. However, traditional resonators are hardly compatible with both. In response, we propose a new sensor structure. By arranging the resonant beam and the sensitive diaphragm vertically in space, the new structure improves the rigidity of the diaphragm without changing the thickness of the diaphragm and achieves the purpose of increasing the range without affecting the sensitivity. To find the optimal structural parameters for the sensor sensitivity and range, and to prevent the effects of modal disturbances, we propose a multi-objective optimization design scheme based on the BP and NSGA-II algorithms. The optimization of the structure parameters not only improved the sensitivity but also increased the interference frequency to solve the issue of mode interference. The optimized structure achieves a sensitivity and range of 4.23 Hz/kPa and 1–10 MPa, respectively. Its linear influence factor is 38.07, significantly higher than that of most resonant pressure sensors. The structural and algorithmic optimizations proposed in this paper provide a new method for designing resonant pressure sensors compatible with a wide range and high sensitivity.

Keywords: resonant pressure sensor; MEMS; BP; NSGA-II; algorithmic optimization

1. Introduction

As the global industry develops and advances, higher requirements are placed on pressure sensors for measurement accuracy, sensitivity, and range. For example, submarines need to accurately monitor depth positions in the 0–1000 m range. This requires highly accurate and sensitive pressure sensors with a range of 10 MPa. Resonant MEMS pressure sensors are the most accurate and stable pressure sensors available. They calculate the pressure by detecting the eigenfrequency change in the resonator, without analog-to-digital conversion, which makes signal acquisition and processing convenient. Device accuracy is mainly affected by the mechanical properties of a single-crystal silicon structure, strong anti-interference ability, and stable performance. Thus, it is very suitable for ocean exploration and other areas that require the accurate measurement of pressure in complex environments. Currently, researchers have conducted extensive studies on the range and sensitivity of resonant pressure sensors [1–3]. In the research on a wide range, Xiang C et al. [4] deployed silicon islands on the device layer to enhance the equivalent stiffness and structural stability of the pressure-sensitive diaphragm, achieving a measurement range of 0.2–7 MPa, sensitivity of 2.26 Hz/kPa, and a linear influence factor of 15.386. The linear influence factor means sensitivity $\times$ range, which is the product of the sensitivity and range. Yu J et al. [5,6] designed electromagnetic-excited sensors and electrostatically excited sensors, obtaining high measurement ranges of 0.11–30 MPa and 0.11–50 MPa, respectively, by increasing the thickness of the sensitive diaphragm. However, the sensitivity of these two sensors was only 0.43 Hz/kPa and 0.066 Hz/kPa, with linear influence factors of 12.83 and 3.3, respectively. Yan P et al. [7] proposed a resonator made

of SOI and SOG materials, with a sensitivity of 11.89 Hz/kPa within a range of 0.1–1 MPa, and a linear influence factor of 10.7. Lu Y et al. [8] introduced a resonant pressure sensor with a measurement range of 1 MPa, featuring a sensitivity of 13.1 Hz/kPa and a linear influence factor of 13.1. These pressure sensors increase the diaphragm stiffness by either increasing the diaphragm thickness or reducing the diaphragm area to minimize diaphragm deformation at high pressures, ensuring measurement accuracy under high-pressure conditions. However, thicker diaphragms may reduce the conversion ratio, thereby sacrificing sensitivity to varying degrees. To achieve a higher sensitivity, Han X et al. [9] proposed a high-precision differential resonator. The lower thickness of the sensitive diaphragm (75 μm) resulted in a high sensitivity of 35.5 Hz/kPa, but the low stiffness of the diaphragm made the structure fragile. Significant deformation under high pressure led to a decreased linearity and accuracy, resulting in a range of only 0–0.2 MPa and a linear influence factor of 7.1. Y. Li et al. [10] proposed a high-sensitivity MEMS resonant differential pressure sensor based on bulk silicon, with a sensitivity of up to 143.17 Hz/kPa, but with a range of only 0.11 MPa and a linear sensing factor of 15.75. Additionally, Shi X et al. [11] and Lu Y et al. [12] proposed resonator structures with linear influence factors of 11.12 and 14.14, respectively, achieving a high sensitivity. Currently, traditional resonant pressure sensors are unable to meet the performance requirements of both a wide range and high sensitivity due to structural limitations. Pressure sensors compatible with both range and sensitivity not only enhance the measurement accuracy, expand the application scope, and improve the system performance but also reduce sensor application costs. Therefore, investigating resonators compatible with both range and sensitivity is of significant importance.

The working principle of traditional resonant pressure sensors, as shown in Figure 1, utilizes a secondary stress transfer island–diaphragm structure [10]. When external pressure acts on the sensitive diaphragm, the stress generated by the deformation of the diaphragm is transmitted through the silicon islands to the resonant beam, causing a change in the resonant beam frequency. However, the sensor can only reliably convert pressure changes into frequency changes when the deflection deformation of the sensitive diaphragm is small. Under high pressure, excessive deformation of the sensitive diaphragm leads to a decrease in linearity and accuracy [13], affecting the sensor's range.

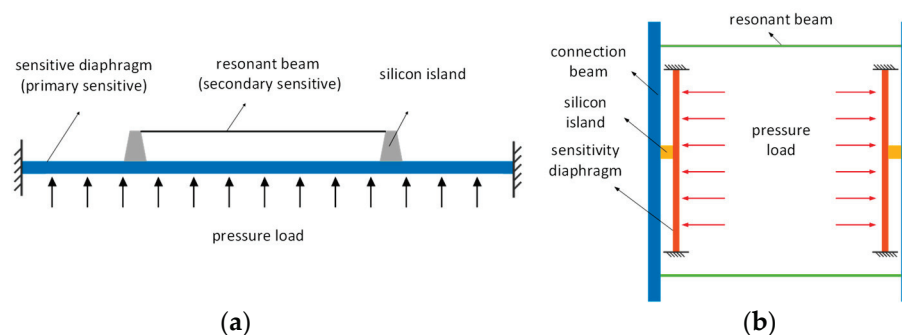


Figure 1. (a) Schematic diagram illustrating the working principle of the traditional beam–diaphragm structure; (b) schematic design of the new structure.

In response to the current shortcomings of resonator structures, this paper proposes a vertical resonator structure for resonant pressure sensors, and the structural principle is shown in Figure 1b. The new structure eliminates the influence of excessive diaphragm deformation on the linearity, providing feasibility for designing sensors compatible with a high sensitivity and wide range. Additionally, parasitic modes exist in resonant pressure sensors, and the proximity of parasitic mode frequencies to the operating mode frequency can lead to a decreased measurement accuracy. To eliminate the influence of mode interference on sensor performance in the new structure and improve two key indicators, namely the structure sensitivity and range, a multi-objective optimization design scheme

for resonant pressure sensors based on Back Propagation Neural Network (BP) and Non-dominated Sorting Genetic Algorithm-II (NSGA-II) algorithms is proposed. The optimized structure avoids the influence of parasitic modes and achieves device performance with a high sensitivity and wide range. The new structure provides a conceptual approach for designing resonant pressure sensors with a wide range and high sensitivity. Furthermore, the multi-objective optimization scheme for sensitivity and range proposed for resonant pressure sensors can be applied to the multi-parameter optimization of other sensor structures.

2. Principle and Design of High-Sensitivity and Wide Range Resonator Structure

The principle of the vertical resonator structure, with the resonant beam perpendicular to the sensitive diaphragm, is illustrated in Figure 1b. When external pressure acts on the sensitive diaphragm, the diaphragm transmits the pressure to the resonant beam through the connecting beam, causing a change in the resonant beam frequency. Compared to the original structure, in the new structure, when the diaphragm is subjected to external forces, it not only receives support force from the four edges of the diaphragm but also obtains support force from the resonant beam. Additionally, the sensor adopts a secondary stress transfer island–diaphragm structure to suppress the energy coupling between the resonator and the sensitive diaphragm, reducing the impact of resonance on sensor accuracy, and thus improving the Q factor and accuracy. The resonant beam in the new structure suppresses the deformation of the diaphragm, improving the equivalent stiffness of the pressure-sensitive diaphragm and the overall stability of the structure under high pressure. This effectively addresses the problem of linear attenuation and accuracy reduction caused by the large deflection deformation of the sensitive diaphragm under high pressure. Based on the principle shown in Figure 1b, we designed the resonant pressure sensor shown in Figure 2.

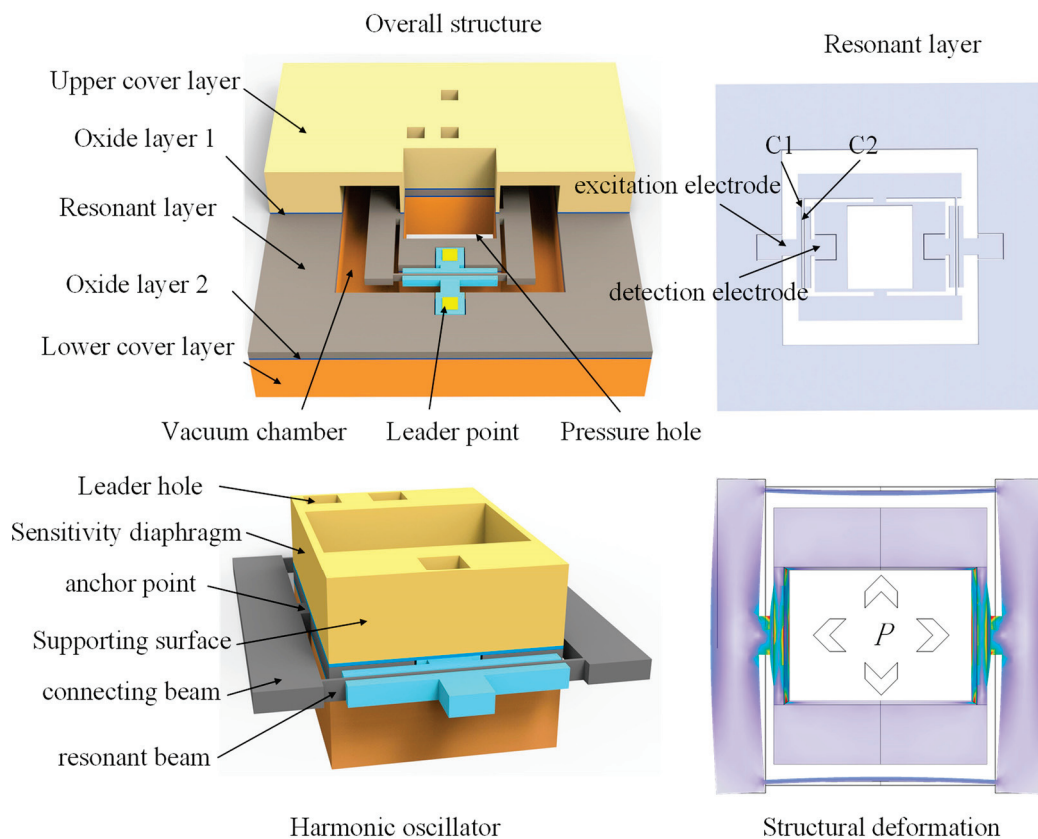


Figure 2. Schematic diagram of the vertical distribution resonator.

The structure of the resonant pressure sensor is depicted in Figure 2. The resonator consists of an upper cover layer, a resonant layer, and a lower cover layer. The square hole in the middle of the structure serves as the pressure hole, allowing fluid to apply pressure to the surroundings through the square hole. The pressure hole consists of two thick walls (support surfaces) and two thin walls (sensitive diaphragms). The support surfaces prevent the overall deformation of the structure and fix the sensitive diaphragms, which are responsible for sensing external pressure. The stress on the diaphragm is transmitted to the resonant beam through silicon islands and connecting beams, causing a corresponding frequency change in the resonant beam. Excitation and detection electrodes are located on both sides of the resonant beam, forming capacitors C1 and C2, respectively, with the resonant beam. The sensor operates by exciting C1 and collecting the capacitance frequency signal through C2. The signal is transmitted to external circuits through wire holes, and external pressure is calculated using the frequency–stress relationship.

3. Mathematical Modeling and Parametric Analysis of Resonators

The stress and deformation of the new structure can be equivalent to the structure shown in Figure 3. When pressure acts on the sensitive diaphragm, the structure will undergo deformation from Figure 3a to Figure 3b. According to the principle of force equilibrium in the same direction, the pressure P acting on the sensitive diaphragm should be equal to the support force of the diaphragm. The support force on the diaphragm can be divided into two sources: one is the support force around the diaphragm, and the other is the support force provided by the resonator beam. The force acting on the diaphragm satisfies Equation (1).

$$P = P_1 + 2P_2 \quad (1)$$

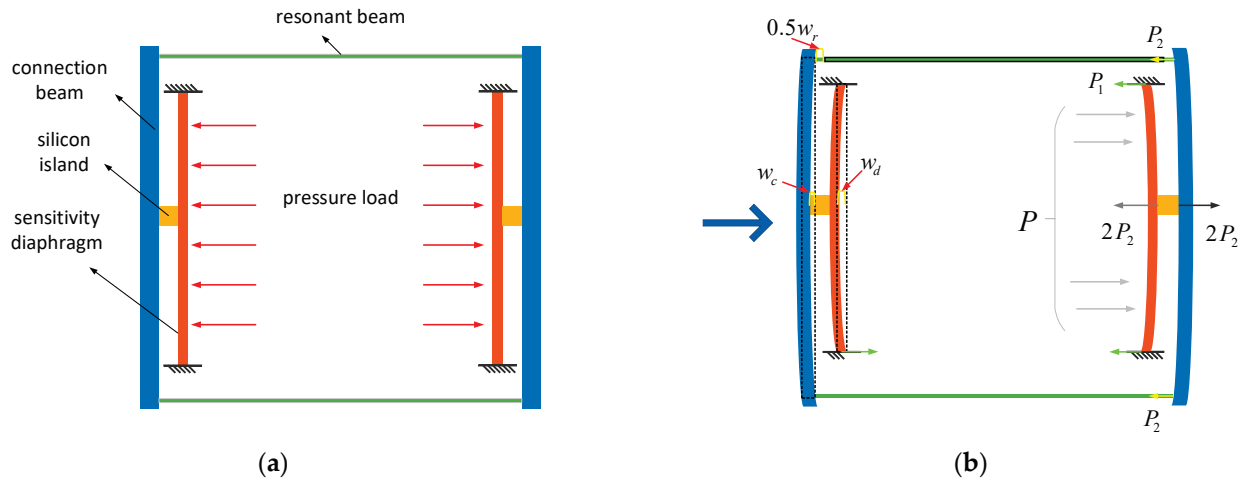


Figure 3. (a) Structural schematic diagram. (b) Stress and deformation analysis.

In Equation (1), P represents the external pressure on the diaphragm, P_1 denotes the support force around the sensitive diaphragm, and P_2 signifies the support force provided by a single resonator beam.

Furthermore, according to the deformation analysis depicted in Figure 3b, the sensitive diaphragm, connecting beam, and resonator beam satisfy relationship 2 in terms of deformation:

$$w_d = w_c + 0.5w_r \quad (2)$$

where w_d is the deformation of the diaphragm, w_c is the deformation of the connecting beam, and w_r is the deformation of the resonant beam.

By combining the displacement equation and the stress equation, the structure satisfies relationship 3:

$$\begin{cases} P = P_1 + 2P_2 \\ w_d = w_c + 0.5w_r \end{cases} \quad (3)$$

To solve the unknowns in Equation (3), calculate the stretching deformation w_r of the resonator beam, and subsequently determine the axial stress in the resonator beam.

3.1. Analysis of the Connecting Beam

According to Newton's third law, the forces acting on the connecting beam and the resonator beam are equal in magnitude and opposite in direction. Therefore, the magnitude of the forces acting on both ends of the connecting beam is P_2 . The deformation of the connecting beam satisfies the conditions for solving the cantilever beam deformation, and the deflection deformation is expressed by Equation (4):

$$w_c = \frac{P_2 L_c^3}{12EI_c} \quad (4)$$

where E , L_c , and I_c are the modulus of elasticity of the silicon, the length of the connecting beam, and the moment of inertia of the connecting beam cross-section, respectively.

3.2. Analysis of the Diaphragm

The sensitive diaphragm is a rectangular diaphragm with four fixed sides, and its deformation under pressure P_1 conforms to the theory of small deflection deformation of diaphragms, as shown in the deformation of the sensitive diaphragm in Figure 3. For a sensitive diaphragm with width $2a$ and length $2b$, its deflection deformation can be expressed as Equation (5):

$$\frac{Eh_d^3}{12(1-\nu^2)} \left(\frac{\partial^4 w}{\partial x^4} + 2 \frac{\partial^4 w}{\partial x^2 \partial y^2} + \frac{\partial^4 w}{\partial y^4} \right) = \frac{P_1}{S} \quad (5)$$

where w_d is the deflection of the diaphragm, S is the sensitive diaphragm area, ν and h_d are the silicon Poisson's ratio and the thickness of the diaphragm, respectively. The bending stiffness of the diaphragm can be expressed as

$$D = \frac{Eh_d^3}{12(1-\nu^2)} \quad (6)$$

The boundary conditions for a rectangular diaphragm with fixed connections on all four sides are

$$\left. \begin{aligned} w_{x=-a,a} &= 0, \frac{\partial w}{\partial x} \Big|_{x=-a,a} = 0 \\ w_{y=-b,b} &= 0, \frac{\partial w}{\partial y} \Big|_{y=-b,b} = 0 \end{aligned} \right\} \quad (7)$$

The deflection deformation of rectangular diaphragms with fixed connections on all four sides can be represented by a trigonometric function as Equation (8):

$$w_d(x, y) = \sum_{m=1,3,5,\dots}^{\infty} \sum_{n=1,3,5,\dots}^{\infty} C_{mn} \left(1 + \cos \frac{m\pi x}{a} \right) \left(1 + \cos \frac{n\pi y}{b} \right) \quad (8)$$

Substituting Equations (7) and (8) into Equation (5), the expression for the diaphragm deflection can be obtained as Equation (9):

$$w_d = \frac{4P_1 a^4 \left(\left(1 + \cos \frac{\pi x}{a} \right) \left(1 + \cos \frac{\pi y}{b} \right) \right)}{S\pi^4 D \left(3 + 2 \frac{a^2}{b^2} + 3 \frac{a^4}{b^4} \right)} \quad (9)$$

3.3. Resonant Beam Analysis

The stress–strain deformation of the resonant beam, as shown in Figure 3, can be represented by Equation (10):

$$w_r = \frac{P_2 l}{EA} \quad (10)$$

where A , w_r , and l are the cross-sectional area, elongation in length direction, and length of the resonant beam.

For a resonant beam without damping and driving effects, the formula for the resonant frequency is expressed as Equation (11):

$$EJ \frac{\partial^4 w(x, t)}{\partial x^4} - P_2 \frac{\partial^2 w(x, t)}{\partial x^2} + \rho_m \frac{\partial^2 w(x, t)}{\partial t^2} = 0 \quad (11)$$

where J is the moment of inertia of the resonant beam cross-section. ρ_m is the density of silicon.

The boundary conditions for the simply supported beam of length l at both ends are

$$w|_{x=0,l} = 0 \quad \left. \frac{\partial w}{\partial x} \right|_{x=0,l} = 0 \quad (12)$$

The solution of Equation (11) can be expressed as

$$w = w(x, t) = w(x) \cos \omega t \quad (13)$$

Substituting Equation (13) into Equation (11) and solving, under the action of axial force P_2 , the frequency equation for the first-order free vibration in the width direction of the beam with both ends fixed can be expressed as Equation (14):

$$w(P_2) = \left(\frac{E(4.73)^4 b_r^2}{12l^4 \rho_m} + \frac{(4.73)^4 (0.295) P_2}{12l^2 \rho_m h_r b_r} \right)^{\frac{1}{2}} \quad (14)$$

In Equation (14), l , h_r , and b_r represent the length, width, and height of the resonant beam, respectively.

The first-order resonant frequency of the resonant beam along the width direction under the action of axial force is

$$f = \frac{\omega_1(P_2)}{2\pi} = \frac{\left(\frac{E(4.73)^4 b_r^2}{12l^4 \rho_m} + \frac{(4.73)^4 (0.295) P_2}{12l^2 \rho_m h_r b_r} \right)^{\frac{1}{2}}}{2\pi} \quad (15)$$

3.4. Overall Structural Analysis

The stress and deformation of the structure under the pressure shown in Figure 3 can be summarized using Equation (3). By substituting Equations (4), (9), and (10) into the unknown variables of Equation (3), the axial force acting on the resonant beam satisfies the relationship expressed in Equation (16):

$$\frac{4P_1 a^4 \left((1 + \cos \frac{\pi x}{a}) (1 + \cos \frac{\pi y}{b}) \right)}{S\pi^4 D \left(3 + 2\frac{a^2}{b^2} + 3\frac{a^4}{b^4} \right)} = \frac{P_2 L_c^3}{12EI_c} + \frac{P_2 l}{2EA} \quad (16)$$

Using P_2 to represent P_1 , the equation can be expressed as

$$\frac{4(P - 2P_2) a^4 \left((1 + \cos \frac{\pi x}{a}) (1 + \cos \frac{\pi y}{b}) \right)}{S\pi^4 D \left(3 + 2\frac{a^2}{b^2} + 3\frac{a^4}{b^4} \right)} = \frac{P_2 L_c^3}{12EI_c} + \frac{P_2 l}{2EA} \quad (17)$$

The deflection at the midpoint of the diaphragm can be represented as

$$\frac{16(P - 2P_2)a^4}{S\pi^4 D \left(3 + 2\frac{a^2}{b^2} + 3\frac{a^4}{b^4}\right)} = \frac{P_2 L_c^3}{12EI_c} + \frac{P_2 l}{2EA} \quad (18)$$

The deflection coefficient at the midpoint of the diaphragm under a uniformly distributed load is given by

$$k = \frac{192a^4(1 - \nu^2)}{\pi^4 E h_d^3 \left(3 + 2\frac{a^2}{b^2} + 3\frac{a^4}{b^4}\right)} \quad (19)$$

The axial force P_2 in the resonant beam is given by

$$\frac{(P - 2P_2)k}{S} = \frac{P_2 L_c^3}{12EI_c} + \frac{P_2 l}{2EA} \quad (20)$$

$$P_2 = \frac{kP}{\frac{SL_c^3}{12EI_c} + \frac{Sl}{2EA} + 2k} \quad (21)$$

Substituting P_2 into Equation (15), the relationship between pressure P and frequency f is given by Equation (22):

$$f = \left(\frac{10.428Eb_r^2}{\pi^2 l^4 \rho} + \frac{SAL_c^3 + 6SII_c + 24EkAl_c}{3.9\pi^2 l^2 \rho h_r b_r kPEAl_c} \right)^{\frac{1}{2}} \quad (22)$$

Randomly selecting parameters such as the length, width, and height of the resonant beam, connecting beam, and sensitive diaphragm, we calculated the resonant frequency of the resonator under a pressure of 10 MPa using the proposed mathematical model. A comparison with data obtained from COMSOL 6.1 software revealed a maximum frequency difference of 2754 Hz and a maximum error of 1.96%, indicating the high accuracy of the proposed mathematical modeling. The calculated and comparison results are shown in Figure 4.

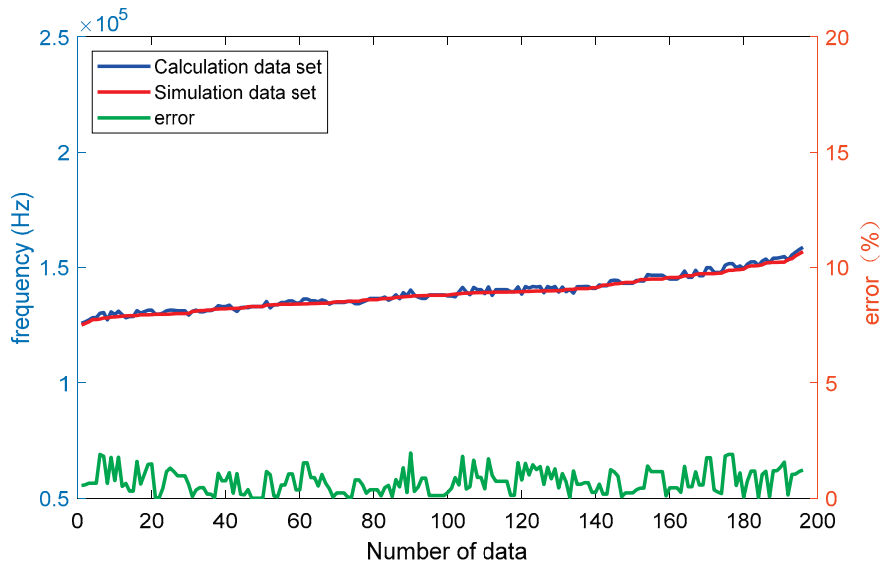


Figure 4. Comparison between mathematical model and simulation model.

4. Pressure Sensor Design and Optimization

To make resonant pressure sensors (absolute pressure sensors) not only capable of achieving a specific measurement range of 1–10 Mpa but also characterized by a high sensitivity, we systematically varied the individual parameters and analyzed their effects on the sensor sensitivity and interference frequency for the resonator beam, connecting beam, and diaphragm. Based on the analysis results, we selected the sensor structure with the highest sensitivity within the specified range.

4.1. Resonant Beam Structural Analysis and Design

According to Equation (15), the relationship between the length l , width h_r , and height b_r of the rectangular resonant beam and the initial frequency and sensitivity is shown in Figure 5. Here, when the length changes from 800 μm to 1500 μm , the resonant beam frequency decreases by 60% on average, and the sensitivity also decreases slightly. When the height changes from 40 μm to 50 μm , the resonant beam frequency remains basically unchanged, but the sensitivity decreases by 0.6 Hz/kPa on average, which is 20%. When the width is increased from 20 μm to 30 μm , the resonant beam frequency increases by 50% on average, while the sensitivity decreases by 50%. Reducing both the width and height can effectively improve the resonant beam sensitivity, but the close proximity of the width and height will lead to interference between the horizontal and vertical resonant modes, so there should be a certain difference between the height and width in the parameter selection. Compared with reducing the height, reducing the width of the resonant beam not only improves the higher sensitivity but also reduces the intrinsic frequency and increases the resonant beam amplitude. Therefore, a smaller width rather than height is chosen in the structure design. Based on the resonant beam parameters and resonant frequencies of the existing structures [10,11,14], a resonant beam structure with a length of 1200 μm , a width of 20 μm , and a height of 40 μm is selected with an intrinsic frequency of 120 kHz.

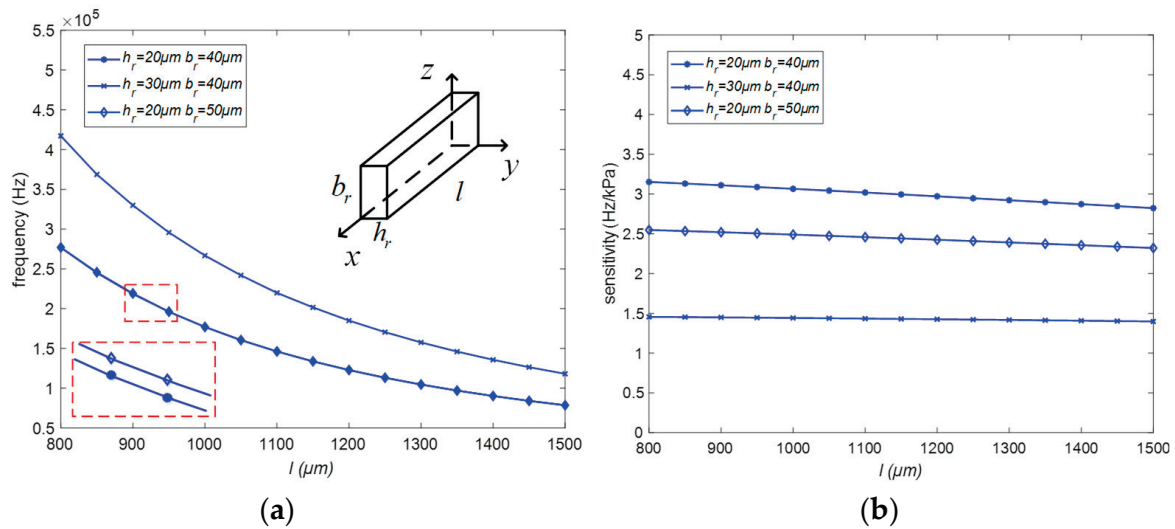


Figure 5. The influence of resonant beam parameters on the natural frequency and sensitivity: (a) Relationship between the parameters of the harmonic beam and the intrinsic frequency; (b) Trend of the influence of the variation of the parameters of the harmonic beam on the sensitivity.

Preliminary structural models have been established, and dynamic analysis of the structure has been conducted using COMSOL. The first six mode shapes of the resonator structure are shown in Figure 6, while the first three mode shapes of the resonant beam are illustrated in Figure 7.

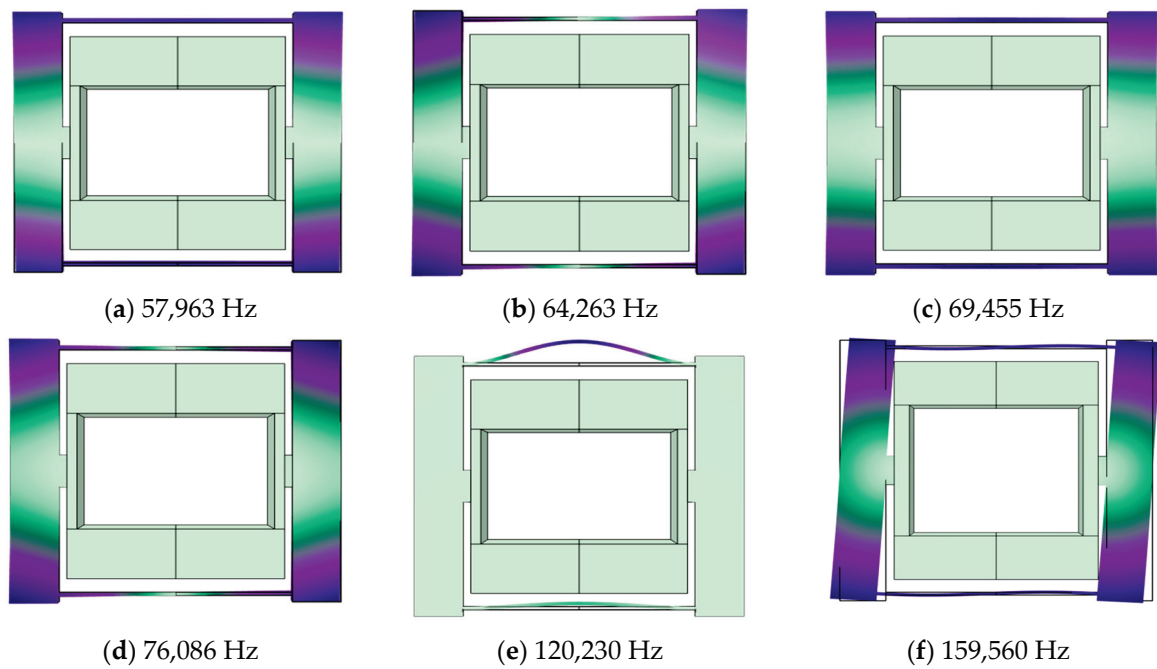


Figure 6. Dynamic analysis of the resonator: (a) the first mode shape; (b) the second mode shape; (c) the third mode shape; (d) the fourth mode shape; (e) the fifth mode shape; (f) the sixth mode shape, with the fifth mode as the operational mode and the sixth mode as the disturbance mode.

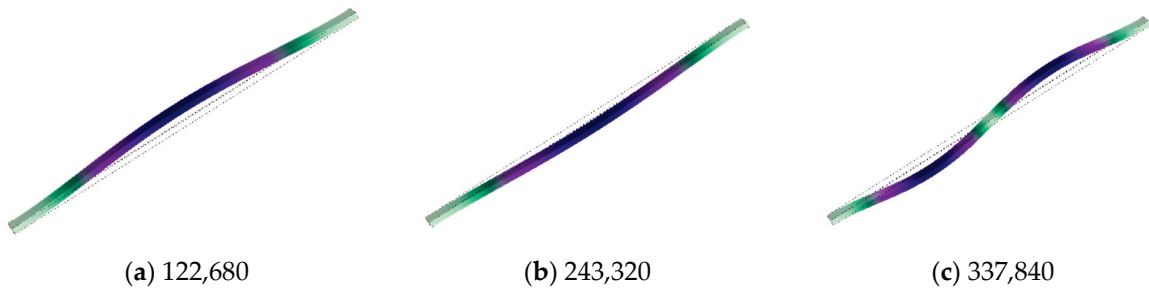


Figure 7. Vibration modes of the resonant beam: (a) the first mode shape; (b) the second mode shape; (c) the third mode shape.

Compared to higher-order modes, lower-order modes have advantages such as larger resonance amplitudes, easier excitation and detection, and stronger interference resistance. Therefore, in the selection of operating modes for resonators, lower-order modes are typically preferred. For this reason, in this study, the first-order horizontal resonance mode of the resonant beam is chosen as the operating mode, which is the fifth-order resonance mode in the overall structure. The frequencies of the first four modes of the structure are concentrated in the range of 50–80 kHz, which is significantly different from the operating mode at 120 kHz and can be disregarded. As the frequency of the resonant beam increases under external pressure, the frequency difference between it and the sixth-order mode gradually decreases until they intersect. To prevent the intersection of the operating mode and the sixth-order mode (interfering mode) within the measurement range, which would lead to an increase in the measurement error, finite element analysis is utilized to analyze the structural parameters affecting the interfering mode and the operating mode. Based on the analysis results, the structure is optimized to obtain the optimal sensitivity structure without mode interference within a specific range (1–10 MPa).

4.2. Connecting Beam Structure Analysis and Design

The relationship between the interference frequency of the sensor and the parameters of the connecting beam is shown in Figure 8 (left). The sensitivity and parameters of the connecting beam are depicted in Figure 8 (right), where w_c , L_c , and h_c represent the width, length, and height of the connecting beam, respectively. When the width w_c is increased from 200 μm to 500 μm , the average sensitivity of the structure increases by approximately 3.5 Hz/kPa, which is about a 230% increase. Meanwhile, the average interference frequency decreases by around 60 kHz, representing about a 30% decrease. Increasing the width significantly improves the sensitivity of the structure, but also reduces the interference frequency and compresses the range of frequency variation in the resonant beam operation.

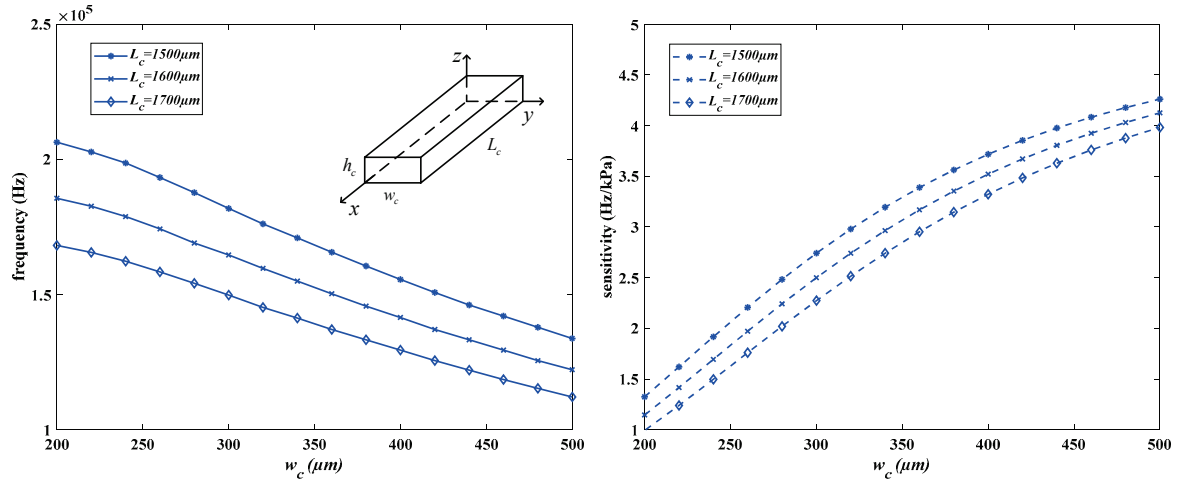


Figure 8. The impact of parameters on the disturbance frequency and sensitivity.

The connecting beam is located in the resonant layer along with the resonant beam, as shown in Figure 2, with a height of 40 μm . When the length L_c of the connecting beam increases from 1500 μm to 1700 μm , the average sensitivity of the structure decreases by 0.4 Hz/kPa, and the average interference frequency decreases by 40 kHz. Increasing the length leads to a decrease in both the sensitivity and interference modal frequency, thereby limiting the performance parameters of the structure. To achieve a higher sensitivity and interference frequency, the length should be minimized. The length of the connecting beam can be represented by the width of the diaphragm, as shown in Equation (23).

$$L_c = 2a + c \quad (23)$$

where L_c represents the length of the connecting beam, $2a$ is the width of the pressure-sensitive diaphragm, and c denotes the designed electrode placement distance, which is 200 μm .

4.3. Sensitive Diaphragm Structural Analysis and Design

The relationship between the diaphragm parameters and interference frequency and sensitivity is illustrated in Figure 9, where $2a$, $2b$, and h_d , respectively, represent the width, height, and thickness of the diaphragm. The width $2a$ increases from 600 μm to 1100 μm , resulting in an average decrease of 70 kHz in the interference frequency, representing a 30% reduction. The sensitivity of the structure with a height of 800 μm is maximized when the width is 900 μm , while the sensitivity of the structure with a height of 1000 μm is maximized when the width is 1100 μm . As the height $2b$ increases from 600 μm to 1100 μm , there is no significant change in the interference frequency, but the sensitivity increases on average by 1.5 Hz/kPa, representing a 150% increase. When the thickness h_d increases from 50 μm to 95 μm , the interference frequency increases on average by 30 kHz, a 20% increase, while the sensitivity decreases on average by 3 Hz/kPa, representing a 75% reduction.

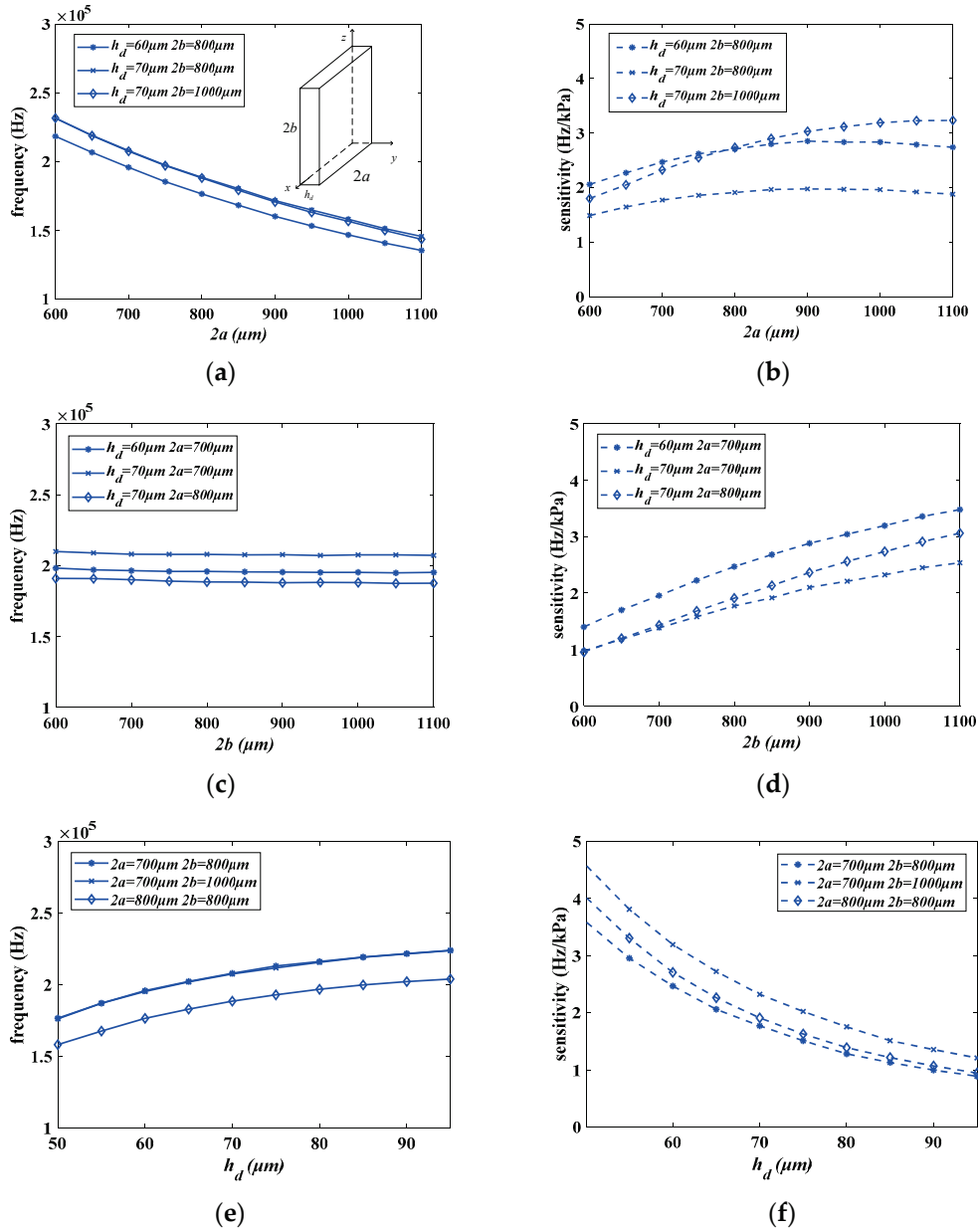


Figure 9. The relationship between diaphragm parameters and disturbance frequency and sensitivity. (a) Depicts the relationship between the width and disturbance frequency. (b) Illustrates the relationship between the width and sensitivity. (c,d) Show the influence of the height, while (e,f) demonstrate the effect of the thickness.

4.4. Resonator Structure Optimization

The pressure and frequency of the resonant structure exhibit a highly linear relationship; hence, the sensitivity s can be represented in terms of the operating frequency S at a full-scale pressure of 10 MPa, as shown in Equation (24):

$$S = f_0 + s \times r \quad (24)$$

where S is the full-scale frequency, f_0 is the natural frequency of the structure, the value is 120 kHz, s is the sensitivity of the structure, and r is the range; since resonator sensitivity is usually measured in Hz/kPa, the value here is 1000.

The design of the structure should aim for both a high range and high sensitivity. Therefore, within the determined range of 10 MPa, adjusting the parameters to achieve a higher sensitivity is necessary, meaning a higher full-scale operational frequency should be pursued. Additionally, to prevent interference mode effects in this structure, the interference frequency should be higher than the operational frequency at the full scale of the sensor.

Therefore, in the design of the structure, the goal should be to obtain the maximum full-scale resonant frequency while ensuring that the interference frequency is greater than the full-scale operational frequency.

In the structural parameters, the parameters of the resonant beam are related to sensitivity but not to interference frequency. The relationship between the length of the connecting beam and the width of the diaphragm is given by Equation (23), indicating a linear relationship between the two. The height of the connecting beam is the same as that of the resonant beam, which is 40 μm . These parameters can be selected through single-objective optimization.

For the width of the connecting beam and the width, height, and thickness of the sensitive diaphragm, there is usually an inverse relationship between the interference frequency and the parameters chosen for the full-scale resonance frequency. When parameters leading to higher full-scale resonance frequencies are selected, the interference frequency tends to be lower, whereas when parameters leading to lower full-scale resonance frequencies are chosen, the interference frequency tends to be higher. This contradictory relationship implies that single-objective optimization methods cannot meet the requirements for designing the optimal structure. Consequently, exhaustive search methods become one of the most common approaches for selecting structures from multiple parameters. However, exhaustive methods are not only inefficient and costly but also may fail to find the global optimum when dealing with large search spaces, leading to being trapped in local optima. Therefore, it becomes challenging to obtain the optimal structure of the resonator. To address this issue, a multi-objective optimization design scheme for resonant pressure sensors based on the BP algorithm and NSGA-II algorithm is proposed, achieving the simultaneous optimization of range and sensitivity.

4.4.1. BP- and NSGA II-Optimized Structures

The BP neural network [15] is widely utilized due to its simplicity, strong nonlinear mapping capability, and high operability. It serves as a black-box function to solve engineering optimization problems with complex input–output relationships, making it particularly suitable for fitting the unknown relationships between sensor structural parameters and performance. NSGA-I introduced a non-dominated sorting and elite strategy, while NSGA-II [16], based on NSGA-I, proposed congestion and crowding comparison operators, facilitating fast search and improving the convergence speed of optimization. It is often employed for handling multi-objective optimization problems.

In order to extract the parameter mapping relationship of the resonator and construct a high-precision BP network, 300 sets of data were randomly selected for the four parameters that simultaneously affect the interference frequency and operating frequency, namely the width of the connecting beam and the width, height, and thickness of the diaphragm. The corresponding operating frequency S and interference frequency F for each parameter set were calculated using the commercial software COMSOL. Among these, 260 sets of data were chosen as the training set and 40 sets as the validation set for training the BP neural network. Partial data are shown in Table 1.

Build a BP neural network with four inputs and two outputs, using diaphragm width $2a$, diaphragm height $2b$, diaphragm thickness h_d , and connection beam width w_c as input independent variables, and interference frequency F and operating frequency S as output dependent variables. The function relationship fitted by BP is used as the objective function for NSGA II iteration optimization to obtain the Pareto optimal solution set with two outputs. The optimization framework of the structure is illustrated in Figure 10. Calculate the interference frequency and operating frequency corresponding to the structural

parameters using finite element analysis and import them as training and testing sets into the BP training network. The weight model trained by BP is imported into NSGAII as the functional relationship equation, and the Pareto solution set after 1000 iterations of population iteration is taken as the optimal solution set. The solution most in line with the requirements is selected from the Pareto solution set as the optimal structure, and finite element analysis is conducted to validate the structure.

Table 1. BP network training set.

$h_d/\mu\text{m}$	$2b/\mu\text{m}$	$2a/\mu\text{m}$	$w_c/\mu\text{m}$	S/Hz	F/Hz
50	800	600	230	−145,745.2954	−206,431.3251
50	800	600	240	−147,728.3998	−203,625.0286
60	800	700	250	−144,926.1623	−195,884.3529
65	800	700	250	−140,845.8125	−202,368.2791
70	800	700	250	−137,959.511	−207,989.4299
75	800	700	250	−135,335.4896	−213,111.4485
70	750	700	250	−136,039.2568	−207,997.5868
70	800	700	250	−137,959.511	−207,989.4299
70	850	700	250	−139,393.7201	−207,660.0328
60	800	650	250	−142,946.4384	−206,745.8811
60	800	700	250	−144,926.1623	−195,884.3529
60	800	750	250	−146,436.1059	−185,449.8793
60	800	800	250	−147,310.5528	−176,572.7048
60	800	850	250	−148,220.33	−168,312.7138
...	...	...	...	...	...

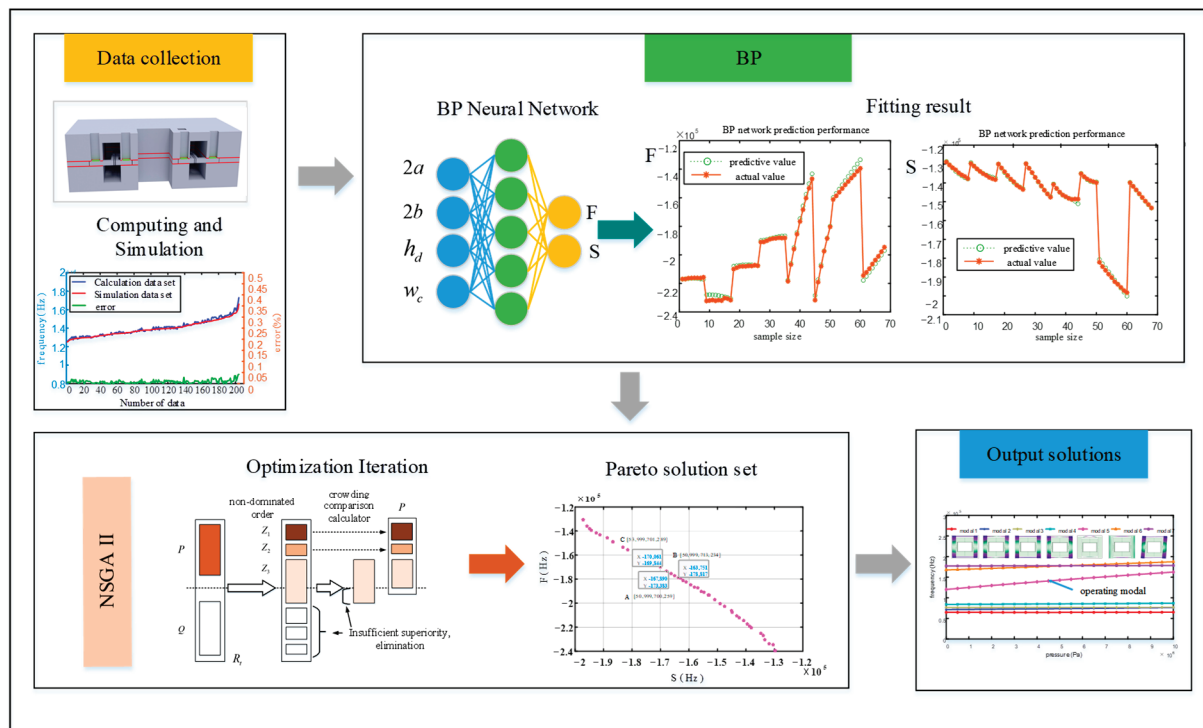


Figure 10. Optimization process.

The optimization model constructed employs a three-layer neural network structure for BP training, comprising an input layer, a hidden layer, and an output layer. Each layer is associated with corresponding weights and thresholds. The input matrix and output matrix are illustrated in Equations (25) and (26), respectively.

$$X = [2a \ 2b \ h_d \ w_c] \quad (25)$$

$$Y = [F S] \quad (26)$$

The three-layer neural network structure consists of four input parameters and two output parameters, with a hidden layer containing nine nodes. The specific formula is 27.

$$n = 2m + 1 \quad (27)$$

where n represents the number of hidden nodes, and m represents the number of input layer nodes.

In the BP training process, if there are significant differences in the range of the data, it can lead to large biases during weight updates, resulting in slow training. To improve the training speed of the BP network and mitigate the issues of gradient vanishing or exploding, all data are normalized using the normalization equation as shown in Equation (28). Subsequently, the data are de-normalized using Equation (29).

$$X_i = 2 \times \frac{(x_i - x_{\min})}{x_{\max} - x_{\min}} - 1 \quad (i = 1, 2, \dots, n) \quad (28)$$

$$Y_i = \frac{(y_i + 1) \times (y_{\max} - y_{\min})}{2} + y_{\min} \quad (i = 1, 2, \dots, n) \quad (29)$$

For the iteration of weights in the BP network model, adjustments are made using the error backpropagation algorithm. When the mean squared error E between the fitted output data and the actual values exceeds the expected value for the current iteration, the error is propagated backward from the output layer, and the weights and thresholds are updated until the error gradually decreases to achieve the desired accuracy. The mean squared error is calculated using the formula shown in Equation (30).

$$E = \frac{1}{n} \sum_{i=1}^n (p_i - r_i)^2 \quad (30)$$

where p_i is the predicted value of the BP fitting, and r_i is the true value of this training.

The NSGA-II algorithm employs a strategy of fast non-dominated sorting and crowding distance to achieve good convergence in multi-objective optimization problems. It can quickly converge to the Pareto front with a high convergence, solution set diversity, and uniformity. Its workflow is illustrated in Figure 11.

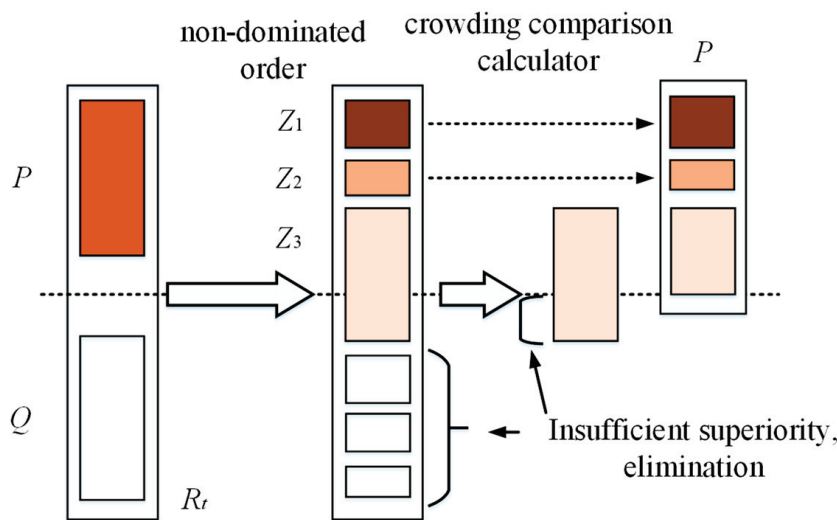


Figure 11. Elite strategy selection in NSGA-II algorithm.

An initial population Q_1 is generated by random selection, and a second-generation population Q_2 is generated by a genetic operator. Subsequently, during the population iteration process, the offspring population Q_t generated in generation t is merged together with the parent population P to form a new population of population size $2N$. Then, the population R_t is non-dominated sorted to find a series of non-dominated sets Z_1 , and the crowding degree of each individual is calculated. Since the individuals of both parent and child generations are contained in population R_t , the non-dominated set Z_1 after non-dominated sorting contains the best set of individuals in the whole population R_1 , so Z_1 is put into the new parent population P_{t+1} first. If at this point the size of the population is smaller than N , then it is necessary to continue to fill P_{t+1} with the next level of non-dominated set Z_2 until the size of the population exceeds N when the non-dominated set Z_n is added, then, individuals are extracted using the crowding comparison operator for each individual in Z_n to bring the size of the population P_{t+n} up to N . Genetic operators such as selection, crossover, and mutation are then used to generate new offspring populations Q_{t+1} in order to prevent the results from not converging due to the infinite size of some parameters in the optimization process, and at the same time to avoid the structure not being processed or processing difficulties due to some parameters being too small. According to the results of the above analysis, the boundary of the optimization parameters is divided, and the boundary conditions are shown in Equation (31):

$$\begin{cases} 700 \leq 2a \leq 1000 \\ 00 \leq 2b \leq 1000 \\ 50 \leq h_d \leq 95 \\ 200 \leq w_c \leq 400 \end{cases} \quad (31)$$

The weighted model trained by BP is imported into NSGA-II as the functional relationship, and the Pareto solution set obtained after 1000 iterations of the population is considered the optimal solution set. Each Pareto solution set contains the optimal points with their corresponding input parameters $2a$, $2b$, h_d , and w_c , as well as the optimized F and S . This process yields the optimal structure of the resonant pressure sensor and the highest sensitivity within the 10 MPa range.

4.4.2. Optimization Results

The trained BP neural network algorithm accurately predicts two output variables, the interference frequency and operating frequency, based on four input variables: the diaphragm length, width, and height, as well as the width of the connecting beam. The optimization solution set obtained through the NSGA-II algorithm is depicted in Figure 12, where the horizontal axis represents the resonant frequency S of the structure at the full scale of 10 MPa, and the vertical axis represents the interference mode frequency F . Points below point C satisfy the requirement that the operating frequency is less than the interference frequency. Among them, point A represents the structure with the highest operating frequency under the condition of the operating frequency being lower than the interference frequency. According to Equation (24), the highest operating frequency corresponds to the highest sensitivity; thus, point A exhibits the highest sensitivity. However, at higher initial frequencies and tens of thousands of quality factors [17], in order to avoid the decrease in sensor measurement accuracy caused by the interference frequency approaching the operating frequency, a large frequency difference needs to be maintained between the two. Zhang F et al. [18] reported an initial frequency of 37,431 Hz for a sensor, with a difference of 5000 Hz from the interference frequency, thus avoiding the influence of interference modes. Therefore, we select point B as the optimal solution, with a frequency difference of 15,558 Hz between the interference frequency and the operating frequency, effectively avoiding the interference mode. The parameters at this point include a diaphragm width $2a$ of 713 μm , a height $2b$ of 999 μm , a thickness h_d of 50 μm , and a connecting beam width w_c of 234 μm . The resonant frequency at full scale is 163,751 Hz, and the interference frequency is 178,817 Hz. The optimization results were validated using the commercial software

COMSOL, showing resonant and interference frequencies of 162,479 Hz and 178,493 Hz, respectively, which are in good agreement with the results of the NSGA-II multi-objective optimization. The specific structural parameters are listed in Table 2.

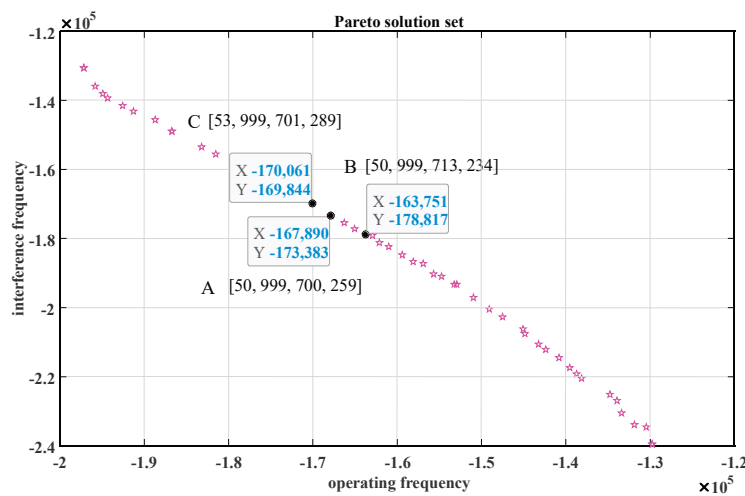


Figure 12. Pareto optimal solution set.

Table 2. Key parameters of optimized structure.

Name	Definition of Parameters	Range (μm)
2a	Width of diaphragm	713
2b	Height of diaphragm	999
h_d	Thickness of diaphragm	50
w_c	Width of connect beam	234
L_c	Length of connect beam	1626
h_c	Height of connect beam	40
l	Length of resonant beam	1200
b_r	Height of resonant beam	40
h_r	Width of resonant beam	20
S	Resonant frequency	162,479 Hz
F	Interference frequency	178,493 Hz

Within the full range of 10 MPa, the maximum internal stress of the resonator is 2.5×10^8 Pa. After optimization, within the 10 MPa range, the relationship between the operating mode (Mode 5) and other modes is depicted in Figure 13. The frequency difference between the operating mode and the nearest interference mode within the range is greater than 15 kHz, effectively eliminating the influence of the interference modes. The sensitivity of the sensor within the operating range of 1–10 MPa is calculated to be 4.23 Hz/kPa, with a linearity of 0.9984 and a linear influence factor of 38.07. Table 3 lists the resonant pressure sensors proposed in recent years. Among them, the structure proposed by Y Lu and PC Yan has a sensitivity greater than 10 Hz/kPa, but its range is only around 1 MPa, resulting in a linear influence factor of around 10. The structures proposed by Y Jie have wide measurement ranges of 30 MPa and 50 MPa, but their sensitivities are only 0.4 and 0.066, resulting in linear influence factors of 12 and 3, respectively. The sensor structure designed in this study demonstrates good compatibility between the range and sensitivity, with a linearity impact factor significantly higher than that of most traditional resonant pressure sensors reported to date.

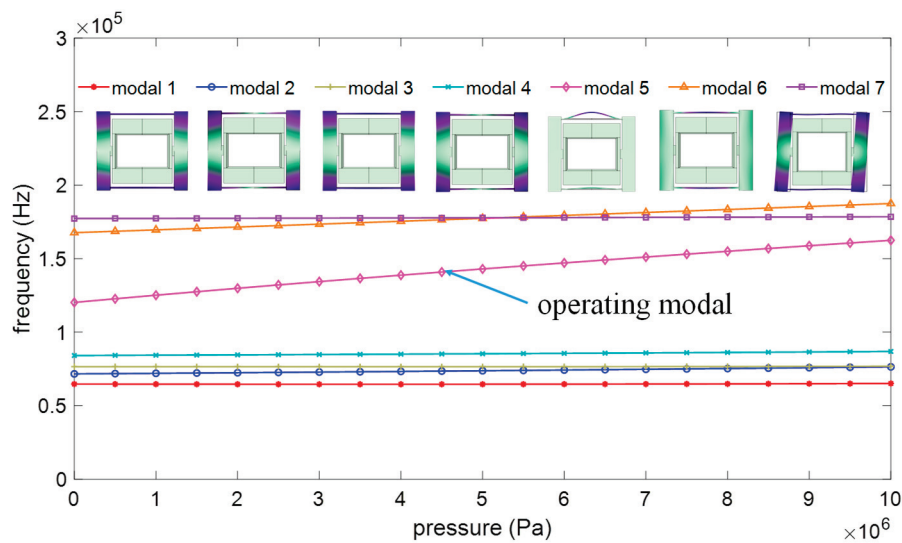


Figure 13. The relationship between resonant modes.

Table 3. Comparison of optimized structures.

Time	Author	Range (MPa)	Sensitivity (Hz/kPa)	Linear Influence Factor	Linearity
2019	Lu Y [8]	0.02–1	13.1	12.838	0.9999
2019	Yan P [7]	0.01–1	11.89	11.6522	0.9999
2020	Xiang C [19]	0.02–2	3.15	6.237	0.9999
2021	Xiang C [3]	0.02–7	2.26	18.0348	0.9998
2021	X Han [9]	0–0.2	35.5	7.01	0.9999
2022	Yu J [5]	0.11–30	0.428	12.84	0.9999
2022	Yu J [6]	0.11–50	0.066	3.3	0.9999
This structure		1–10	4.23	38.07	0.9984

5. Conclusions

Traditional resonant pressure sensors often struggle to achieve both a wide range and high sensitivity simultaneously, limiting their applications in scenarios requiring both. To address this issue, we propose a novel sensor structure. By vertically distributing the resonant beam and pressure-sensitive diaphragm, we enhance the overall stability of the resonant structure and the equivalent stiffness of the diaphragm under high pressure. This effectively resolves the problem of reduced accuracy due to excessive deformation of the sensitive diaphragm, achieving good compatibility between the range and sensitivity. To mitigate the impact of parasitic modes during sensor operation, we present a multi-objective optimization design approach for resonant pressure sensors based on BP and NSGA-II algorithms. This approach addresses modal interference while optimizing the structural sensitivity. Validation of the optimization results using commercial software demonstrates a frequency optimization error of 0.7768% and an interference frequency optimization error of 0.1812%, confirming the reliability of the optimization approach. The optimized resonant structure achieves a high sensitivity of 4.23 Hz/kPa and a linear range of 1–10 MPa. The design strategy of the new structure provides guidance for the design of resonant pressure sensors with a wide range and high sensitivity, effectively addressing the compatibility issue between the sensitivity and range. The multi-objective optimization algorithm proposed for resonant pressure sensors can also be applied to optimize multiple parameters in other MEMS sensor structures, thereby improving the structural performance.

Author Contributions: M.L. is the lead author who proposed and wrote this paper. P.L. guided the design of the structure of the paper. Q.Q. performed some of the data processing. R.L. and J.M. revised the format and checked the content of the article. G.L. corrected the English writing of the article. X.Z. guided the structural design and algorithmic modeling and corrected the manuscript before submission. All authors have read and agreed to the published version of the manuscript.

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References

1. Han, X.; Mao, Q.; Zhao, L.; Li, X.; Wang, L.; Yang, P.; Lu, D.; Wang, Y.; Yan, X.; Wang, S.; et al. Novel resonant pressure sensor based on piezoresistive detection and symmetrical in-plane mode vibration. *Microsyst. Nanoeng.* **2020**, *6*, 95. [CrossRef]
2. Zheng, Y.; Zhang, S.; Chen, D.; Wang, J.; Chen, J. A Micromachined Resonant Low-Pressure Sensor With High Quality Factor. *IEEE Sens. J.* **2021**, *21*, 19840–19846. [CrossRef]
3. Shi, X.; Lu, Y.; Xie, B.; Li, Y.; Wang, J.; Chen, D.; Chen, J. A Resonant Pressure Microsensor Based on Double-Ended Tuning Fork and Electrostatic Excitation/Piezoresistive Detection. *Sensors* **2018**, *18*, 2494. [CrossRef]
4. Xiang, C.; Lu, Y.; Cheng, C.; Wang, J.; Chen, D.; Chen, J. A resonant pressure microsensor with a wide pressure measurement range. *Micromachines* **2021**, *12*, 382. [CrossRef] [PubMed]
5. Yu, J.; Lu, Y.; Chen, D.; Wang, J.; Chen, J.; Xie, B. A resonant high-pressure sensor based on integrated resonator-diaphragm structure. *IEEE Sens. J.* **2021**, *22*, 3920–3927. [CrossRef]
6. Yu, J.; Lu, Y.; Xie, B.; Meng, Q.; Yu, Z.; Qin, J.; Chen, J.; Wang, J.; Chen, D. An Electrostatic Comb Excitation Resonant Pressure Sensor for High Pressure Applications. *IEEE Sens. J.* **2022**, *22*, 15759–15768. [CrossRef]
7. Yan, P.; Lu, Y.; Xiang, C.; Wang, J.; Chen, D.; Chen, J. A Temperature-Insensitive Resonant Pressure Micro Sensor Based on Silicon-on-Glass Vacuum Packaging. *Sensors* **2019**, *19*, 3866. [CrossRef] [PubMed]
8. Lu, Y.; Yan, P.; Xiang, C.; Chen, D.; Wang, J.; Xie, B.; Chen, J. A resonant pressure microsensor with the measurement range of 1 MPa based on sensitivities balanced dual resonators. *Sensors* **2019**, *19*, 2272. [CrossRef] [PubMed]
9. Han, X.; Zhao, L.; Wang, J.; Wang, L.; Huang, M.; Chen, C.; Yang, P.; Li, Z.; Zhu, N.; Wang, S.; et al. High-accuracy differential resonant pressure sensor with linear fitting method. *J. Micromech. Microeng.* **2021**, *31*, 045006. [CrossRef]
10. Li, Y.; Cheng, C.; Lu, Y.; Xie, B.; Chen, J.; Wang, J.; Chen, D. A High-Sensitivity Resonant Differential Pressure Microsensor Based on Bulk Micromachining. *IEEE Sens. J.* **2021**, *21*, 8927–8934. [CrossRef]
11. Shi, X.; Zhang, S.; Chen, D.; Wang, J.; Chen, J.; Xie, B.; Lu, Y.; Li, Y. A Resonant Pressure Sensor Based upon Electrostatically Comb Driven and Piezoresistively Sensed Lateral Resonators. *Micromachines* **2019**, *10*, 460. [CrossRef] [PubMed]
12. Lu, Y.; Yu, Z.; Xie, B.; Chen, D.; Wang, J. A pressure-temperature microsensor based on synergistical sensing of dual resonators. *Measurement* **2024**, *224*, 113946. [CrossRef]
13. Du, X.; Liu, Y.; Li, A.; Zhou, Z.; Sun, D.; Wang, L. Laterally Driven Resonant Pressure Sensor with Etched Silicon Dual Diaphragms and Combined Beams. *Sensors* **2016**, *16*, 158. [CrossRef]
14. Lu, Y.; Zhang, S.; Yan, P.; Li, Y.; Yu, J.; Chen, D.; Wang, J.; Xie, B.; Chen, J. Resonant Pressure Micro Sensors Based on Dual Double Ended Tuning Fork Resonators. *Micromachines* **2019**, *10*, 560. [CrossRef]
15. Liu, J.; Li, P.; Zhuang, X.; Sheng, Y.; Qiao, Q.; Lv, M.; Gao, Z.; Liao, J. Design and Optimization of Hemispherical Resonators Based on PSO-BP and NSGA-II. *Micromachines* **2023**, *14*, 1054. [CrossRef] [PubMed]
16. Deb, K.; Pratap, A.; Agarwal, S.; Meyarivan, T. A fast and elitist multi objective genetic algorithm: NSGA-II. *IEEE Trans. Evol. Comput.* **2002**, *6*, 182–197. [CrossRef]
17. Ren, S.; Yuan, W.; Qiao, D.; Deng, J.; Sun, X. A Micromachined Pressure Sensor with Integrated Resonator Operating at Atmospheric Pressure. *Sensors* **2013**, *13*, 17006–17024. [CrossRef]
18. Zhang, F.; Li, A.; Bu, Z.; Wang, L.; Sun, D.; Du, X.; Gu, D. Vibration modes interference in the MEMS resonant pressure sensor. *Int. J. Mod. Phys. B* **2017**, *31*, 1750223. [CrossRef]
19. Xiang, C.; Lu, Y.; Yan, P.; Chen, J.; Wang, J.; Chen, D. A resonant pressure microsensor with temperature compensation method based on differential outputs and a temperature sensor. *Micromachines* **2020**, *11*, 1022. [CrossRef]

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Article

Driving Principle and Stability Analysis of Vertical Comb-Drive Actuator for Scanning Micromirrors

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Abstract: We have developed a manufacturing process for micromirrors based on microelectromechanical systems (MEMS) technology. The process involves designing an electrostatic vertically comb-driven actuator and utilizing a self-alignment process to produce a height difference between the movable comb structure and the fixed comb structure of the micromirror. To improve the stability of the micromirror, we propose four instability models in micromirror operation with the quasi-static driving principle and structure of the micromirror considered, which can provide a basic guarantee for the performance of vertical comb actuators. This analysis pinpoints factors leading to instability, including the left and right gap of the movable comb, the torsion beams of the micromirror, and the comb-to-beams distance. Ultimately, the voltages at which device failure occurs can be determined. We successfully fabricated a one-dimensional micromirror featuring a 0.8 mm mirror diameter and a 30 μm device layer thickness. The height difference between the movable and fixed comb structures was 10 μm . The micromirror was able to achieve a static mechanical angle of 2.25° with 60 V@DC. Stable operation was observed at voltages below 60 V, in close agreement with the theoretical calculations and simulations. At the driving voltage of 80 V, we observed the longitudinal displacement movement of the comb fingers. Furthermore, at a voltage of 129 V, comb adhesion occurred, resulting in device failure. This failure voltage corresponds to the lateral torsional failure voltage.

Keywords: MEMS micromirrors; stability analysis; electrostatic actuators; self-alignment

1. Introduction

MEMS micromirrors are lightweight, small-size, and cost-effective, making them ideal for a wide range of applications in optical communications [1], autonomous driving [2], and augmented reality (AR)/virtual reality (VR) [3]. Comb-drive micromirrors in particular present notable benefits, including a susceptibility to easy oscillations, rapid response times, and a facilitated precise control of the mirror deflection angle [4–6]. They are highly appealing for optical applications, serving purposes in projection displays [7,8], spectrometers [9], optical switches [10], and other applications requiring precise optical control and adjustments [11]. These applications typically demand highly precise microfabrication technology to ensure optical performance and stability [12–14].

The driving force of the vertical comb-drive actuator is oriented vertically, effectively mitigating the occurrence of the pull-in effect commonly observed in parallel plate movements [15–17]. In an ideal setting, maintaining equal spacing between movable and fixed comb fingers results in mutually canceling lateral forces, establishing a steady state. However, the lateral forces may escalate at high voltages due to alignment accuracy and etching process error factors. In the presence of subtle disturbances, an imbalance in the lateral forces on the left and right sides of the vertical comb-drive structure can manifest, potentially leading to instability and contact between the movable and fixed comb fingers [18]. This contact triggers lateral adhesion, significantly impeding performance and resulting in device failure.

The research on the “pull-in” instability of MEMS devices began as early as the 1960s. Nathanson et al. reported an early analysis in which they used a mass–spring system to simulate the electrostatic deflection of parallel plate actuators, studying the impact of pull-in voltage on resonators [19]. With the widespread application of MEMS devices, the importance of the instability of device “pull-in” is increasingly growing. Hirano et al. conducted a modeling study on the lateral instability of in-plane comb-drive actuators driven by combs [20]. Bochobza-Degani et al. introduced the “pull-in” voltage and critical angle in out-of-plane torsional actuators using polynomial algebraic equations [21]. Pamidighantam et al. investigated the instability of electrostatically driven comb teeth based on the linear beam theory [22]. Zhou and Dowd analyzed the side instability of comb-drive actuators using a one-dimensional model with structural lateral translational stiffness [23]. Huang and Lu established a two-dimensional model to examine the influence of lateral and angular offsets on the critical voltage of in-plane comb-drive MEMS actuators [24]. Lee and Solgaard came up with three failure modes of the scanners, and analytical expressions of pull-in deflection were obtained by applying 2D analytical capacitance models to the derived pull-in equations [25].

In this work, a thorough analysis is conducted to refine the conditions leading to instability in the vertical comb-drive micromirror. Various scenarios are examined for a more comprehensive understanding of the underlying instability factors. This paper introduces the principle of quasi-static deflection for an electrostatic micromirror with a vertical comb-driven structure and provides a detailed explanation of the forces generated by the unequal-height comb-finger structure. Four potential instability models that may manifest during the operation of the micromirror are explored in connection with its structure. One model involves comb fingers with a relatively large length-to-width ratio, causing an uneven force distribution and resulting in the elastic deformation of the comb fingers themselves. The other three models involve comb fingers with high rigidity that concentrate the force on the twist beams, inducing a bending deformation of the fixed elastic beam at both ends. Consequently, the movable comb finger changes position and comes into contact with the fixed comb finger, causing the device to stop functioning. Failure voltages are determined based on these models. These models establish a robust foundation to support the stable operation of electrostatic micromirrors at high voltages, effectively compensating for the accuracy errors introduced in the fabrication process and ensuring that the mirror operates in a steady state. A test system is set up to validate the analysis of the vertical comb-driven micromirror device. This system ultimately determines the stable operating voltage for the device.

2. The Principle and Design of Electrostatically Actuated Micromirrors

2.1. Analysis of Drive Structure and Quasi-Static Deflection Principle

Designing a MEMS micromirror involves careful consideration of torsion beams, mirror size, and the structure of the micromirror comb [26]. For the vertical comb-drive micromirror, its primary operational mode is characterized by static deflection. Figure 1a illustrates an electrostatically actuated scanning micromirror with rectangular torsion beams. Four sets of comb-driven structures are connected to the sides of the circular mirror. It consists of two rows of undersized combs, one of which is attached to the substrate as the fixed comb, while the other is attached to the mirror as the movable comb.

According to the definition of torsion for noncircular cross-section beams in mechanics of materials, for a straight bar with a rectangular cross section undergoing free torsion, the calculation formula for the relative rotation angle θ between the ends of the beam with applied torque T is as follows:

$$\theta = \frac{Tl}{GI_t} \quad (1)$$

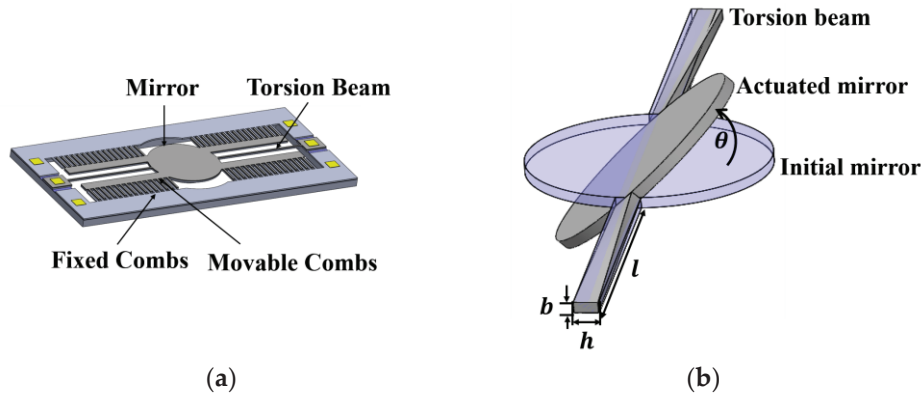


Figure 1. (a) Schematic diagram of electrostatic vertical comb-drive micromirror; (b) simplified model of one-dimensional micromirror torsion.

In the equation, GI_t is referred to as the torsional stiffness of the beam:

$$GI_t = G\beta hb^3 \quad (2)$$

$$G = \frac{E}{2(1 + \mu)} \quad (3)$$

where h represents the length of the long side of the torsion beam's cross section, b represents the length of the short side of the torsion beam's cross section. l is the length of the torsion beam, as shown in Figure 1b. E is Young's modulus of silicon material; μ is the Poisson's ratio. The factor β is related to the torsion of the rectangular cross-section beam and is dependent on the ratio of the side lengths h/b . Factor β for rectangular cross-section beam torsion is listed in Table 1.

Table 1. Factor β for rectangular cross-section beam torsion [27].

h/b	1.0	1.2	1.5	2.0	2.5	3.0	4.0	6.0	8.0	10.0	∞
β	0.141	0.166	0.196	0.229	0.249	0.263	0.281	0.299	0.307	0.313	0.333

The external torque for driving the torsional motion of the micromirror is provided by the vertical comb-drive structure. A change in the electric field between the comb-drive structures results in the generation of a force that induces the deflection of the micromirror. In the quasi-static torsional mode, the system attains a static equilibrium state wherein the torque experienced by the torsional beam T is the applied torque. M_e represents the external torque exerted by the combs.

$$T = M_e = F_Z d \quad (4)$$

where F_Z is the electrostatic force between the combs in the vertical direction, and d is the distance from the torsion beam to the end of the movable comb finger. As shown in Figure 2, the single-beam micromirror undergoes torsional motion due to the driving force provided by the comb-finger structure in the Z-direction.

The electric energy $W(\theta)$ between the interdigitated comb electrodes can be approximated as the energy stored in a capacitor [28]. The expression for this energy is:

$$W(\theta) = \frac{1}{2} C(\theta) V^2 \quad (5)$$

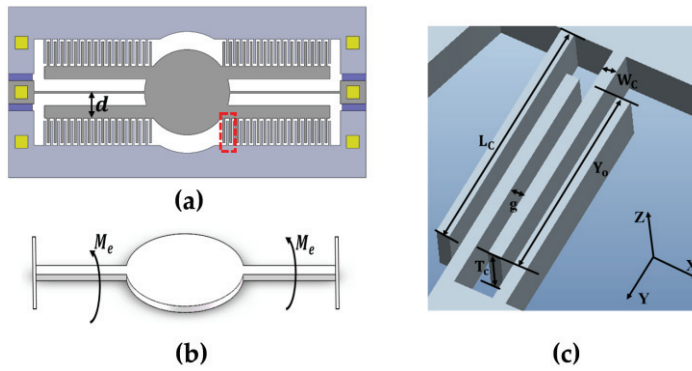


Figure 2. (a) The 1D schematic of an electrostatic comb-drive micromirror, including the comb-finger structure, mirror surface, and torsion beams; (b) the micromirror subjected to an external torque on one side; (c) the detailed schematic of the micromirror's comb-finger structure (The red box part in subfigure (a)), highlighting the specific structural parameters of the movable and fixed comb finger.

A movable comb finger is affected by the fixed comb finger on the left and right sides, therefore, the total electric energy generated by comb fingers become:

$$W_e(\theta) = NC(\theta)V^2 \quad (6)$$

where N is the number of comb fingers, V is the driving voltage, and $C(\theta)$ is the capacitance between the movable comb finger and the fixed comb finger. The electrostatic force between the comb fingers can be calculated using the derivative of the total electric energy with respect to the rotation angle, which is given by:

$$F_Z(\theta) = \frac{\partial W_e(\theta)}{\partial \theta} = N \frac{\partial C(\theta)}{\partial \theta} V^2 \quad (7)$$

where $\frac{\partial C(\theta)}{\partial \theta}$ is related to the rate of change in the overlap area between the movable and fixed comb fingers. The capacitance formula between the plate electrodes is given by:

$$C(\theta) = \frac{\epsilon}{g} A(\theta); \frac{\partial C(\theta)}{\partial \theta} = \frac{\epsilon}{g} \frac{\partial A(\theta)}{\partial \theta} \quad (8)$$

where ϵ is the permittivity of the air, $A(\theta)$ represents the overlapping area between the movable and fixed comb fingers, and g is the gap distance between the movable and fixed comb finger.

When a voltage is applied between the movable and fixed comb fingers, electrostatic forces are generated in the vertical direction. The rotation angle between the comb fingers is considered to be θ in this particular state. As the torsion beam of the micromirror twists, the overlap area between the movable and fixed comb fingers undergoes irregular changes [29–31], which can be categorized into the following situations, where the fringing field effects are small. As shown in Figure 3(a-1), in the presence of a dielectric layer with a thickness of T_o between the movable and fixed comb fingers, the dielectric layer acts as an insulating material, preventing direct contact between the comb fingers and creating a gap between them. Figure 3(a-2) shows the status of the comb fingers.

Case 1 is shown in Figure 3(b-1). Point C on the left side of the movable comb finger is in contact with the upper edge of the fixed comb finger, and the deflection angle θ_1 is given in Figure 3(b-2). There is no overlap area between the movable and fixed comb fingers, so the overlap area is zero.

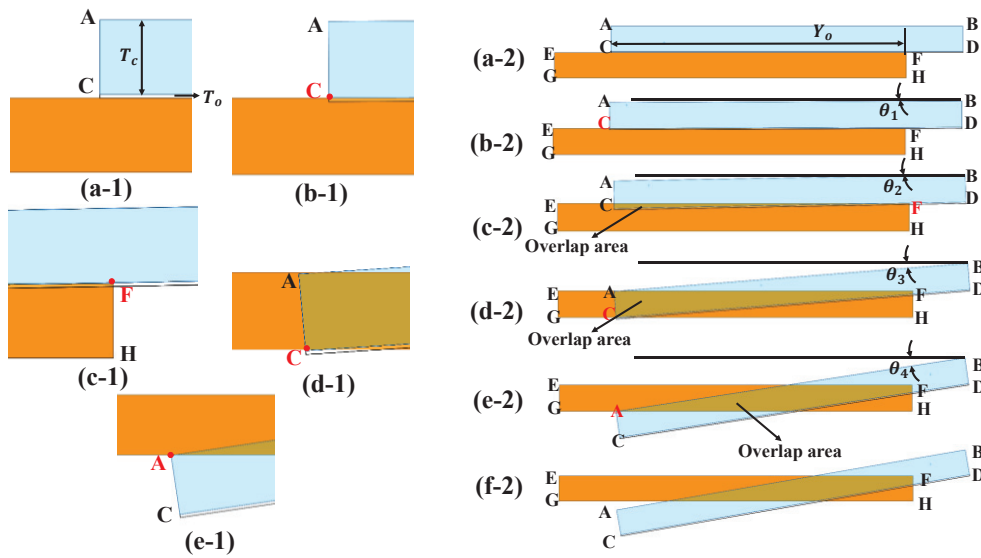


Figure 3. Changes in the overlap area of the comb finger. When the comb finger rotates around the central flexure beam with angle θ , the overlap area of the upper (fixed) and lower (movable) comb fingers changes. (a-1,a-2) are the initial state diagrams of the comb finger; (b-1,b-2) when the deflection angle is θ_1 , point C intersects the upper edge of the fixed comb finger, and the overlap area is zero; (c-1,c-2) when the deflection angle is θ_2 , point F intersects with the upper edge of the fixed comb finger, and the overlap area is a triangle; (d-1,d-2) when the deflection angle is θ_3 , point C intersects the lower edge of the fixed comb finger, and the overlap area is a quadrilateral; (e-1,e-2) when the deflection angle is θ_4 , point A intersects the lower edge of the fixed comb finger, and the overlap area is an irregular pentagon; (f-2) when point A intersects with the lower edge of the fixed comb finger, the overlap area becomes smaller.

The maximum angle of torsion is denoted by θ_1 :

$$\theta_1 = \tan^{-1} \left(\frac{T_o}{d + L_c} \right) \quad (9)$$

where d is the distance from the torsion beam to the end of the movable comb finger. L_c is the length of the comb finger. When $0 < \theta < \theta_1$, the resulting overlap area of a single finger gap is:

$$A(\theta) = 0 \quad (10)$$

The incremental area change in the single finger gap is:

$$\frac{\partial A(\theta)}{\partial \theta} = 0 \quad (11)$$

Case 2 is shown in Figure 3(c-1). The lower edge of the movable comb finger touches the right peak F of the fixed comb finger. As the angle continues to increase, the overlap area forms a triangle, and Figure 3(c-2) shows that the maximum angle at this time is θ_2 .

The maximum angle of torsion is denoted by θ_2 :

$$\theta_2 = \tan^{-1} \left(\frac{T_o}{d + L_c - Y_o} \right) \quad (12)$$

where Y_o is the overlap length of the movable and fixed comb fingers. When $\theta_1 < \theta < \theta_2$, the resulting overlap area of a single finger gap is:

$$A(\theta) = \frac{1}{2} \frac{[(d + L_c) \tan \theta - T_o]^2}{\tan \theta} \quad (13)$$

The incremental area change in the single finger gap is:

$$\frac{\partial A(\theta)}{\partial \theta} = \frac{1}{2} \left[\left(\frac{(d + L_c)}{\cos \theta} \right)^2 - \left(\frac{T_o}{\sin \theta} \right)^2 \right] \quad (14)$$

Case 3 is shown in Figure 3(d-1). When the movable comb finger's point C intersects the lower edge of the fixed comb finger, the maximum angle is θ_3 . As the angle continues to increase, the overlap area forms a quadrilateral, as shown in Figure 3(d-2).

The maximum angle of torsion is denoted by θ_3 :

$$\theta_3 = \tan^{-1} \left(\frac{T_o + T_c}{d + L_c} \right) \quad (15)$$

where T_c is the thickness of the comb finger. When $\theta_2 < \theta < \theta_3$, the resulting overlap area of a single finger gap is:

$$A(\theta) = \frac{1}{2}(L_c + d)^2 \tan \theta - (L_c + d)T_o - \frac{1}{2}(d + L_c - Y_o)^2 \cos \theta \sin \theta + (d + L_c - Y_o)(\cos \theta)T_o \quad (16)$$

The incremental area change in the single finger gap is:

$$\frac{\partial A(\theta)}{\partial \theta} = \frac{1}{2} \left[(L_c + d)^2 \sec^2(\theta) + (d + L_c - Y_o)^2 (\sin^2(\theta) - \cos^2(\theta)) - 2(d + L_c - Y_o)T_o(\sin \theta) \right] \quad (17)$$

Case 4 is shown in Figure 3(e-1). As the torsion angle continues to increase, the shape becomes an irregular pentagon before the movable comb finger's point A intersects the lower edge of the fixed comb finger. Figure 3(e-2) shows the diagram of the maximum angle θ_4 of this process.

The maximum angle of torsion is denoted by θ_4 :

$$\theta_4 = \sin^{-1} \left(\frac{T_o + T_c}{d + L_c - T_o} \right) \quad (18)$$

When $\theta_3 < \theta < \theta_4$, the resulting overlap area of a single finger gap is:

$$A(\theta) = \frac{1}{2}(L_c + d)^2 \tan \theta - (L_c + d)T_o - \frac{1}{2}(d + L_c - Y_o)^2 \cos \theta \sin \theta + (d + L_c - Y_o)(\cos \theta)T_o - \frac{1}{2}Y_o^2 \tan \theta + \frac{1}{2}T_c Y_o \quad (19)$$

The incremental area change in the single finger gap is:

$$\frac{\partial A(\theta)}{\partial \theta} = \frac{1}{2} \left[(L_c + d)^2 \sec^2(\theta) + (d + L_c - Y_o)^2 (\sin^2(\theta) - \cos^2(\theta)) - 2(d + L_c - Y_o)T_o \sin \theta \right] - \frac{1}{2}Y_o^2 \sec^2(\theta) \quad (20)$$

Case 5 is shown in Figure 3(f-2). As the twisting angle continues to increase, the shape becomes an irregular hexagon, and at this point, the overlap area starts to decrease. The capacitance decreases, and the driving force direction changes. In theory, this situation will not occur. The change in overlap area with the angle is a nonlinear function of the micromirror twisting angle, increasing from the rest position until $\theta = \theta_4$ (see Figure 3(f-2)).

Combining Equations (1)–(4) and (7)–(8), we calculate the mirror rotation angle θ :

$$\theta = \frac{TI}{G\beta hb^3} = \frac{\varepsilon NIV^2}{G\beta ghb^3} \frac{\partial A(\theta)}{\partial \theta} \quad (21)$$

2.2. Analysis of Micromirror Driving Stability

The analysis of the micromirror driving stability involves examining the equilibrium and dynamic behavior of the mirror under different operating conditions. Various factors would influence the stability of the micromirror, such as its mechanical structure, electricity,

and control aspects. This section mainly discusses and analyzes the mechanical stability of the micromirror.

The driving force of the micromirror in the vertical comb-drive structure is primarily provided by the electrostatic force between fixed and movable comb fingers. The driving force increases as the gap between the comb fingers decreases. However, due to the residual stress in fabrication or alignment deviation, the micromirror may experience small driving forces in directions other than the desired deflection direction. Besides the force generated in the deflection direction, there may be minor driving forces in other directions. The contact between the movable and fixed comb fingers results in a short-circuit failure of the device [29–31]. Several scenarios of structural instability are illustrated in Figure 4 as shown below. These models are established based on the micromirror being deflected to its maximum allowable angle, specifically at the position where the overlap area between the movable and fixed comb fingers is maximized.

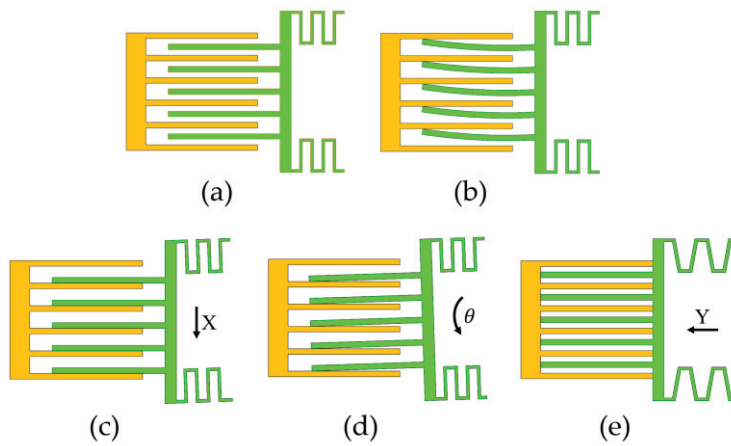


Figure 4. Several scenarios of structural instability. (a) In the initial state of the movable and fixed comb fingers, under normal conditions, the comb fingers are in force equilibrium, and the structure remains stable; (b) the comb fingers' lateral bending contact's instability model; (c) the comb fingers' lateral displacement contact's instability model; (d) the comb fingers' lateral rotational contact's instability model; (e) the comb fingers' lengthways displacement contact's instability model.

2.2.1. The Comb Fingers' Lateral Bending Contact's Instability Model

Figure 5 schematically illustrates one movable electrode comb finger placed between fixed electrode comb fingers. In this figure, V is the applied voltage across the electrodes, g is the gap between the fixed and movable comb fingers, T_c is the thickness of the upper layer of the movable comb finger, and Y_o is the length of the overlap of the comb finger. Assume that the movable electrode finger moves x in the X direction.

$$F_{\text{right}} = \frac{\varepsilon T_c Y_o}{2(g-x)^2} V^2; F_{\text{left}} = \frac{\varepsilon T_c Y_o}{2(g+x)^2} V^2 \quad (22)$$

where F_{right} and F_{left} represent the lateral force on the right side and left side, respectively, and ε is the permittivity of free space.

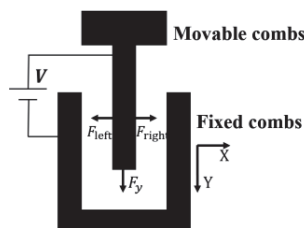


Figure 5. Schematic view of the comb finger's electrode.

In this case, the lateral forces on both sides of the comb finger are no longer balanced, and the electrostatic force F_x generated by both sides of the parallel plate is:

$$F_x = F_{\text{left}} - F_{\text{right}} = \frac{\epsilon T_c Y_0}{2(g-x)^2} V^2 - \frac{\epsilon T_c Y_0}{2(g+x)^2} V^2 \quad (23)$$

where T_c is the thickness of the movable comb fingers. A positive value of F_x proves that the suspended comb is unstable. It looks as if there was a “negative” spring. The equivalent “negative” spring constant k_e when the movable comb finger is placed at the center of the gap is

$$k_e = \frac{\partial F_x}{\partial x} = \frac{\epsilon T_c Y_0}{(g-x)^3} V^2 + \frac{\epsilon T_c Y_0}{(g+x)^3} V^2 \quad (24)$$

In this case, the spring constant of the movable comb finger bending along the X direction is given by:

$$k_{xc} = \frac{ET_c W_c^3}{12L_c^3} \quad (25)$$

where E is Young’s modulus, and W_c and L_c are the width and length of the comb finger, respectively.

The mechanical spring with spring constant k_{xc} keeps the position of the movable comb finger against the instability of the electrostatic force.

If

$$|k_e| > |k_{xc}| \quad (26)$$

the movable comb finger stays stable; otherwise, as shown in Figure 4b, the movable comb finger become unstable and touches the fixed comb fingers. The voltage at this point is referred to as the lateral side-instability voltage. At this stage, the electrostatic force between the comb finger undergoes a rapid change, resulting in the attraction between them. Hence, to keep the stable operation of the comb structure [32], Equation (26) must be satisfied.

When the driving voltage exceeds the so-called side-instability voltage V_1 , the comb drive becomes unstable leading to a side-sticking of the movable and fixed comb fingers. By combining Equations (23)–(26) the voltage at which the side instability occurs can be expressed as:

$$V_1 = \sqrt{\frac{EW_c^3(g-x)^3 + (g+x)^3}{12\epsilon L_c^3 Y_0}} \quad (27)$$

2.2.2. The Comb Fingers’ Lateral Displacement Contact’s Instability Model

When the elastic constant of the comb finger is large enough, it is difficult to generate the bending contact of the comb finger, and the generated force is transferred to the torsion beam through the root of the comb finger. As shown in Figure 6, the torsion beam anchored at both ends experiences axial forces (also known as tension or compression forces), resulting in axial expansion and contraction deformation. The left beam of the mirror extends due to tensile forces, while the right side undergoes compression forces, causing it to shorten, as illustrated in Figure 4c. As a result, lateral displacement occurs when the movable comb fingers make contact with the fixed comb fingers, creating variations in electrical potential that disrupt the circuit, leading to device instability and an eventual cessation of operation.

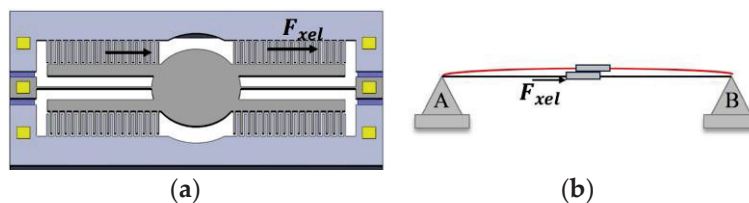


Figure 6. The comb fingers’ lateral displacement contact’s instability model. (a) The driving of one

side of the comb-finger structure results in a total force in the X direction; (b) a schematic diagram showing the deformation of the beam anchored at both ends due to axial forces.

The total force of the comb structure in the X direction is:

$$F_{xel} = \frac{N\varepsilon T_c Y_o}{2(g-x)^2} V^2 - \frac{N\varepsilon T_c Y_o}{2(g+x)^2} V^2 \quad (28)$$

where n is the number of movable comb finger.

The elastic constant of the torsion beam in the X direction is:

$$k_{xl} = \frac{2ET_l W_l}{l} \quad (29)$$

The analysis process is similar to that of the previous case. The equivalent “negative” spring constant k_{e2} when the torsion beam bends up or down is:

$$k_{e2} = \frac{\partial F_{xel}}{\partial x} = \frac{N\varepsilon T_c Y_o}{(g-x)^3} V^2 + \frac{N\varepsilon T_c Y_o}{(g+x)^3} V^2 \quad (30)$$

It must satisfy $|k_{e2}| > |k_{xl}|$. Finally, the instability voltage V_2 in this case is obtained by

$$V_2 = \sqrt{\frac{2EW_l [(g-x)^3 + (g+x)^3]}{n\varepsilon l T_c Y_o}} \quad (31)$$

2.2.3. The Comb Fingers’ Lateral Rotational Contact’s Instability Model

According to the Figure 7a, the lateral electrostatic force F_{xel} generated between the comb finger creates a torque on the torsion beam. In this case, the motion model of the beam is as shown in Figure 7b, where the distortion of the torsion beam causes deflection in the comb finger connected to the beam, resulting in contact between the movable and fixed comb fingers, leading to device instability. Meanwhile, the deflection angle of the movable comb finger is the same as that of the midpoint of the torsion beam.

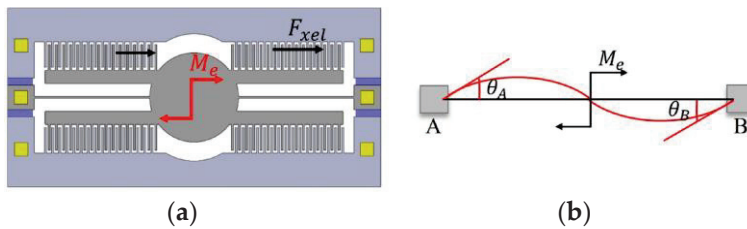


Figure 7. The comb fingers’ lateral rotational contact’s instability model. (a) Driving one side of the comb-finger structure generates a total force in the X direction; (b) a simplified diagram illustrating the bending deformation of the beam supported and anchored at both ends due to an external torque.

θ_A and θ_B represent the angular displacements of the cross sections at points A and B along the torsional beam, respectively. The deflection angle of the midpoint of the torsion beam can be calculated based on the applied torque and the torsional stiffness of the system:

$$\theta_o = \frac{\theta_A}{2} = \frac{\theta_B}{2} = \frac{M_e 2l}{12EI} = \frac{F_{xel} dl}{6EI} \quad (32)$$

where I is the moment of inertia, which is a measure of an object’s resistance to changes in its rotational motion. It quantifies how mass is distributed around the beam of rotation.

In the context of a torsion beam, the inertia moment refers to the moment of inertia of the beam itself, which depends on its geometry and mass distribution:

$$I = \frac{l \cdot W_l^3}{12} \quad (33)$$

The movable comb-finger structure is fixed on the mirror through connecting beams, so the movement of the comb fingers follow the deflection of the mirror structure. In other words, it experiences an angular displacement at the midpoint of the torsional beam. The angular displacement at the midpoint of the torsional beam represents the deflection angle θ of the movable comb fingers:

$$\theta = \tan\left(\frac{g-x}{d+L_c}\right) \quad (34)$$

When $\theta \geq \theta_o$, the movable comb finger touches the fixed comb finger and results the device fails. In this case, the voltage is referred to as the rotational pull-in voltage V_3 :

$$V_3 = \sqrt{\frac{2 \cdot \tan^{-1}\left(\frac{g-x}{d+L_c}\right) \cdot E \cdot l \cdot W_l^3}{n \varepsilon T_c Y_o \cdot \left[\frac{1}{(g-x)^2} - \frac{1}{(g+x)^2}\right] \cdot d \cdot 2L}} \quad (35)$$

2.2.4. The Comb Fingers' Longitudinal Displacement Contact's Instability Model

In this model, when a voltage is applied to the movable and fixed combs, the overlap area in the direction of the comb overlaps increases, resulting in the generation of electrostatic forces in the Y direction:

$$F_y = \frac{N \varepsilon T_c}{2g} V^2 \quad (36)$$

When the electrostatic force acts on the torsion beam through a mirror, the flexural model of the torsion beam is depicted in Figure 8b. According to the deformation formula for beams under simple loading in materials mechanics, with the ends of the beam securely fixed and subjected to a longitudinal force, the maximum deflection of the beam occurs at its midpoint:

$$\omega = \frac{F_y \cdot (2l)^3}{48EI} \quad (37)$$

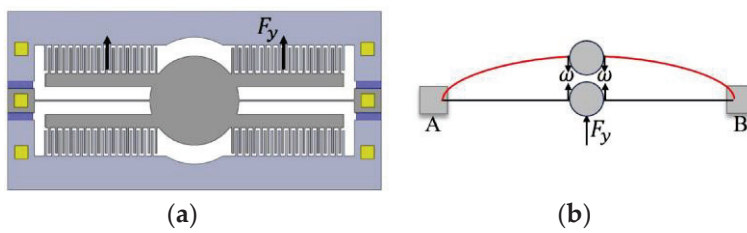


Figure 8. The comb fingers' longitudinal displacement contact's instability model. (a) Driving one side of the comb-finger structure generates a total force in the Y direction; (b) a simplified diagram of the deformation at the midpoint of an elastic beam fixed at both ends under the influence of a longitudinal force; ω represents the maximum deflection.

The displacement ($\Delta y = L_c - Y_o$) required for the device to fail after the contact between the movable and fixed comb fingers is established in the Y direction.

When $\omega \geq L_c - Y_o$, the contact between the movable comb-finger and the fixed comb-finger structure leads to the failure of the device.

By combining Equations (36) and (37), we obtain the longitudinal unstable voltage at that moment:

$$V_4 = \sqrt{\frac{L_c - Y_o E T_l W_l^3 g}{l^3 N \epsilon T_c}} \quad (38)$$

According to the above analysis, the structural parameters of the designed micromirror are determined. The geometric parameters and material properties [28] used in the MEMS fabrication of the micromirror are listed in Table 2.

Table 2. Geometric parameters and material properties.

Symbol	Parameters	Value
l	Length of torsion beam	1200 μm
W_l/b	Width of torsion beam	20 μm
T_l/h	Thickness of torsion beam	30 μm
d	Distance from the torsion beams to the end of the movable comb fingers	194 μm
g	Gap between fixed and movable fingers	5 μm
W_c	Width of comb fingers	5 μm
L_c	Length of comb fingers	280 μm
T_c	Thickness of movable comb fingers	10 μm
T_{c2}	Thickness of fixed comb fingers	20 μm
Y_o	Length of the overlap of the comb finger	240 μm
T_o	Thickness of dielectric layer	1 μm
N	Number of movable comb fingers	120
E	Young's modulus of silicon	130 GPa
μ	Poisson's ratio of silicon	0.28

When the given values are substituted into Equations (8)–(20), it is possible to calculate the micromirror incremental capacitance change ($\frac{\partial C(\theta)}{\partial \theta}$) versus the torsional angle under different driving voltages. To validate the structural design, a three-dimensional model was created based on the parameters mentioned above, and a finite element analysis was conducted using specialized software [33,34]. Finite element analysis (FEA) software is a tool used in engineering to simulate and analyze how structures behave under different conditions. It helps designers optimize designs by predicting factors like stress and deformation. We performed simulations via the finite element method (FEM). Figure 9a illustrates the simulation results of the structural displacement in the Z direction, Figure 9b shows the Z-direction displacement of the tips of the movable comb fingers, and Figure 9c demonstrates the incremental change in capacitance between the comb fingers. The device operated within the range of $0 < \theta < \theta_4$. Due to the idealized nature of theoretical calculations, where edge effects and other factors are not considered, the calculated values were slightly greater than the simulated values.

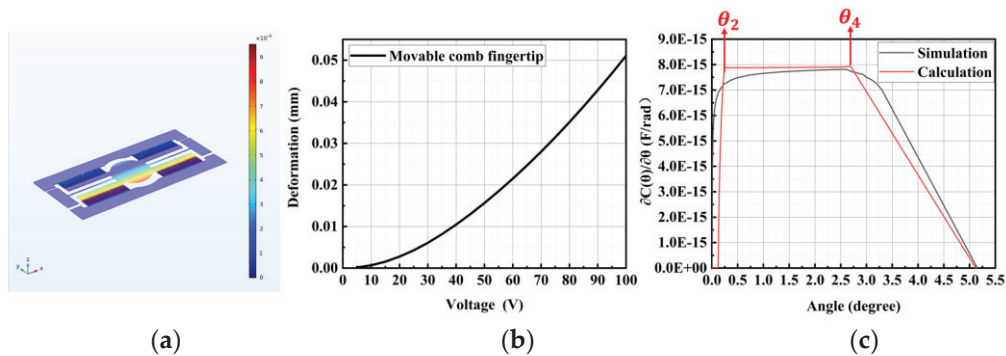


Figure 9. (a) The simulation result of the Z-direction displacement of the micromirror. (b) The displacement of the movable comb fingertip in the Z direction. (c) Incremental capacitance change $\frac{\partial C(\theta)}{\partial \theta}$ using finite element simulation and theoretical calculation for a limited angular range.

3. Fabrication

The fabrication process involved bonding two silicon-on-insulator (SOI) wafers together using a silicon-to-silicon wafer bonding technique. Subsequently, a substrate silicon layer was etched away, resulting in the formation of two silicon conducting layers. The silicon dioxide of the top layer was used as a mask, using a single photolithography step and a self-aligning process that created a height difference between the movable comb structure and the fixed comb structure of the micromirror. The fabrication process steps of the vertical comb-drive structure of the electrostatic micromirror are depicted in Figure 10. A multilayer SOI wafer was bonded together (Figure 10a). The first step involved photolithography on the top silicon dioxide layer to create a pattern, followed by dry etching to remove the oxide layer, acting as a mask for the third step to remove the top silicon layer as shown in Figure 10b. The second step encompassed a pattern formation and multiple stages of dry etching. This process included etching two oxide layers and two silicon layers sequentially to produce comb structures of equal height, as illustrated in Figure 10c. After removing the photoresist and cleaning, Figure 10d shows the deep silicon etching of the entire structure to form the unequal-height comb structures. The oxide layer structure was etched using a hydrofluoric acid (HF or BOE) solution to expose the second device layer (Figure 10e). The third step involved a front-side sputtering of metal material and using a wet-etching process to form electrodes (Figure 10f). Subsequently, a temporary bonding process was employed to etch the backside cavity, fully exposing the device layer and reducing the device mass to increase the deflection space (Figure 10g). The final step involved chip cleaning using a wet-cleaning process.

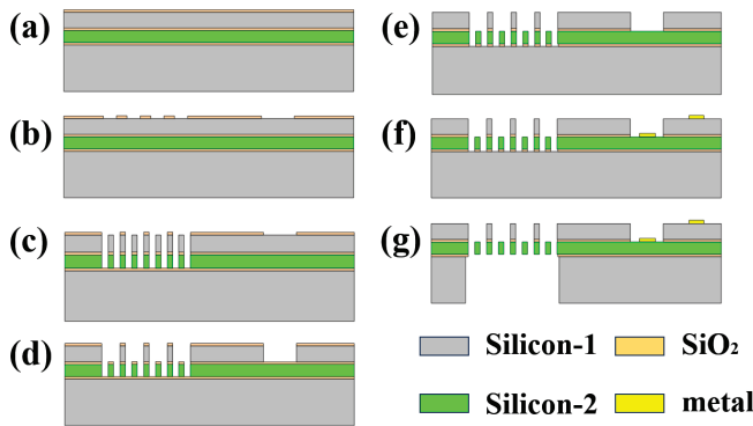


Figure 10. Simplified fabrication process flow for vertical comb drive with self-aligned gaps. A single lithography step defines the gap between the upper and lower fingers. Steps (a–g) show cross sections during micromirror fabrication processing.

The vertical comb-drive micromirror devices were fabricated and used to verify the correctness of the above model. Figure 11a shows the overall SEM image of the MEMS micromirror, with a mirror diameter of 0.8 mm. Figure 11b shows a symmetrical arrangement of vertical comb-finger structures on both sides of the torsion beam. This structure allowed the mirror to deflect at a certain angle in both left and right directions. A closed-up view of the vertical comb structure is shown in Figure 11c. In these images, the upper structure of the movable comb fingers is about 10 μm above the lower structure of the fixed comb fingers. Figure 11d shows the gaps on both sides of the movable comb fingers are symmetrical. The upper and lower comb fingers also aligned well.

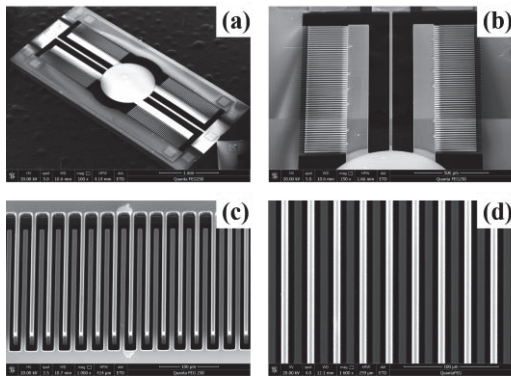


Figure 11. SEMs of fabricated MEMS micromirror: (a) whole device; (b) micromirror torsion beam and vertical comb-finger structure distribution; SEMs of self-aligning vertical comb fingers: (c) 3D view and (d) top view.

4. Experiments and Discussion

The rotation angle of the micromirror was determined by measuring the distance traveled by a laser beam reflected from the mirror surface when the micromirror was actuated. The static response of the MEMS mirror was measured using an experimental setup that involved a laser beam incident to the center of the mirror plate. The setup comprised a DC power supply, a laser, a MEMS mirror, and a screen, as shown in Figure 12a. The micromirror chip, after processing the wafer, was securely mounted onto a PCB (Printed Circuit Board) for ease of applying the driving voltage, as illustrated in Figure 12b. The experimental testing process was as follows: Firstly, the micromirror chip was securely fixed onto an optical platform, and a signal was applied using a direct current power source. Next, the laser beam was aligned with the mirror surface. Finally, after the mirror surface was deflected, the point projected onto the screen moved from A to B. By measuring the displacement of the laser point and the distance from the micromirror to the screen, the mechanical rotation angle of the micromirror was calculated.

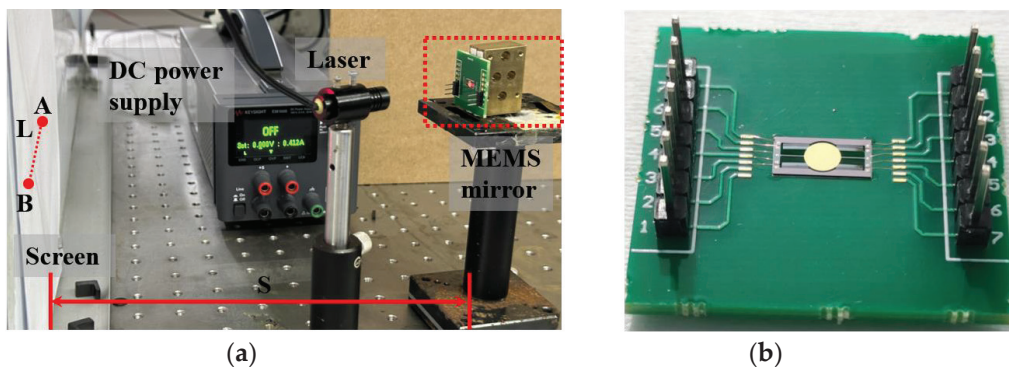


Figure 12. (a) Schematic of the optical measurement setup. L represents the length of the light ray on the screen, S is the vertical distance from the mirror to the screen. (b) Photo of an assembled MEMS micromirror.

As shown in Figure 13, when the driving voltage increased, the angle of deflection continuously increased. At the driving voltage of 60 V, the deflection angle reached 2.25° . When comparing these experimental results with the simulated and calculated results, it was found that the experimental results were slightly lower. This could be attributed to the fact that in reality, the gap between the comb fingers on the chip was larger than $5\ \mu\text{m}$, resulting in a smaller driving force than the ideal value, which in turn reduced the deflection angle. Continuing to increase the driving voltage, the angle increased slowly, reaching 2.6° at 80 V. The main reason for this phenomenon is that under a high voltage, the lateral force

increases, and under small disturbances, the vertical comb-drive structure is unbalanced, causing the torsion beam to bend and deform, resulting in the actual mechanical deflection angle of the micromirror being less than the ideal value. After further increasing the voltage, the torsion angle of the micromirror remained unchanged. The reason for this phenomenon is that under quasi-static deflection, the torsion angle of the micromirror is affected by the change in the overlap area between the comb fingers. When the angle reached the maximum allowable deviation angle between the comb fingers (2.6°), a further voltage application did not significantly change the angle of the micromirror. At that point, an obvious instability in the lateral rotation and movement of the comb finger was observed. Therefore, the stable working voltage should be less than 80 V.

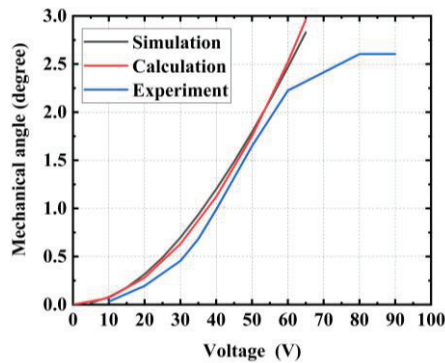


Figure 13. Static response of the MEMS micromirror.

As the voltage continued to increase, the quasi-static deflection angle of the micromirror remained unchanged. However, the increasing voltage introduced unbalanced forces in other directions on the micromirror's comb fingers. According to the equilibrium stability analysis, all parameters of the micromirror were substituted into Equations (27), (31), (35), and (38), respectively. The calculated unstable voltages were as follows: $V_1 = 313.12$ V, $V_2 = 34,846$ V, $V_3 = 129$ V, and $V_4 = 237$ V. The observation and testing were conducted under an optical microscope. As shown in Figure 14, the partial image of the comb fingers structure was observed. The blue dashed line represents the condition with no applied voltage, while the red solid line represents a slight displacement in the Y direction of the movable comb finger after applying a driving voltage of 80 V. This observation provides a clear explanation for the deviation of the actual deflection angle of the microscope from the theoretical and simulated values during testing within the voltage range of 60 V–80 V. However, since there was no contact between the comb finger, the micromirror continued to operate.

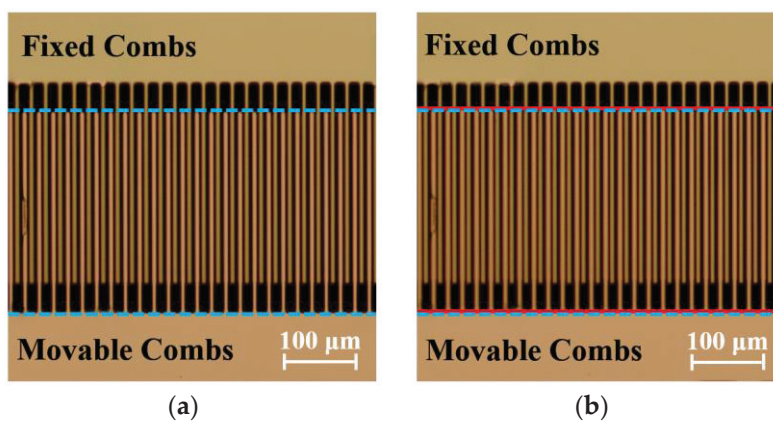


Figure 14. (a) The comb finger's image before applying the test voltage. (b) The comb finger's image

after applying the test 80 V voltage. The blue dashed line represents the condition with no applied voltage, while the red solid line represents a slight displacement of the movable comb finger after applying a driving voltage of 80 V.

With the continuous increase in voltage, when the voltage reached 129 V, the micromirror underwent a torsional motion in the horizontal direction, causing contact between the movable and fixed comb fingers. At that point, the micromirror became nonfunctional and in the lateral rotational contact instability state. After turning off the voltage, the micromirror returned to its original static state, as shown in Figure 15. The blue solid line frame represents the comb finger image before testing, while the red dashed line frame represents the comb finger image after testing. A comparison reveals that the movable comb finger shifted to the left and contacted with the fixed comb finger. This corresponded to the previously analyzed lateral torsional pull-in voltage values. Through testing, the stable operating voltage range of the micromirror was 0 V–80 V, and the device failed at 129 V. Utilizing this analysis provides a better understanding of the micromirror's unstable phenomenon and helps to confirm the unstable voltage in the operational mode.

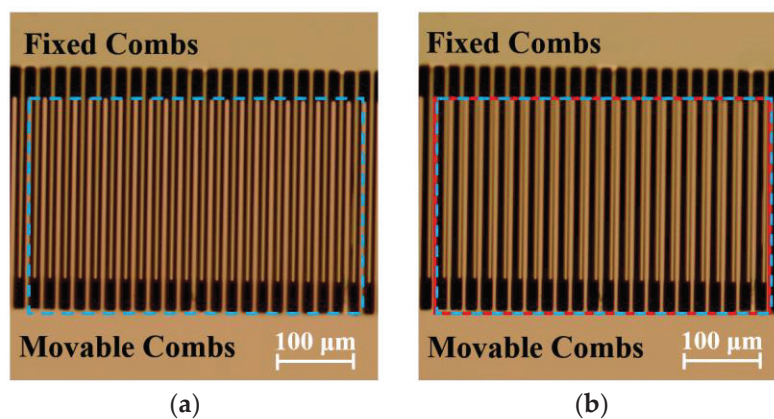


Figure 15. (a) The comb finger's image before applying the test voltage. (b) The comb finger's image after applying the test 129 V voltage. The blue solid line frame represents the comb finger image before testing, while the red dashed line frame represents the comb finger image after testing.

5. Conclusions

This paper established the principle of quasi-static deflection for an electrostatic micromirror with a vertical comb-driven structure and provided a detailed explanation of the forces generated by the unequal-height comb-finger structure. Four potential instability models that may manifest during the operation of the micromirror were explored in connection with its structure. The experimental testing effectively validated the analysis of the driving principles and stability model of the vertically comb-driven micromirror. This micromirror was driven by a direct current voltage, achieving a mechanical angle of 2.25° with 60 V@DC. When the driving voltage was below 60 V, stable operation was observed, and it was generally consistent with theoretical calculations and simulations. At a driving voltage of 80 V, the comb-finger structure was observed to undergo a longitudinal movement, and when the driving voltage reached 129 V, a comb-finger adhesion phenomenon occurred, leading to device failure. This failure voltage corresponded to the lateral torsional failure voltage.

In order to make the micromirror torsion angle larger during design, the torsion beam is usually a thin and elongated structure, which is more likely to produce unstable pull-in phenomena and cause device failure. The establishment of the above model can calculate the torsion angle and failure voltage during the design stage, thus greatly reducing the workload. Devices with errors introduced during the manufacturing process can also be analyzed to determine the stable operating voltage so that the micromirror can operate stably. This method can effectively ensure the long-term efficient and stable operation of the device.

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References

1. Wu, M.C.; Solgaard, O.; Ford, J.E. Optical MEMS for lightwave communication. *J. Light. Technol.* **2006**, *24*, 4433–4454. [CrossRef]
2. Wang, D.; Watkins, C.; Xie, H. MEMS mirrors for LiDAR: A review. *Micromachines* **2020**, *11*, 456. [CrossRef] [PubMed]
3. Petrak, O.; Schwarz, F.; Pohl, L.; Reher, M.; Janicke, C.; Przytarski, J.; Senger, F.; Albers, J.; Giese, T.; Ratzmann, L.; et al. Laser beam scanning based AR-display applying resonant 2D MEMS mirrors. In *Optical Architectures for Displays and Sensing in Augmented, Virtual, and Mixed Reality (AR, VR, MR) II*; SPIE: Bellingham, WA, USA, 2021; Volume 11765, pp. 15–32. [CrossRef]
4. Hah, D.; Huang, S.Y.; Tsai, J.C.; Toshiyoshi, H.; Wu, M.C. Low-voltage, large-scan angle MEMS analog micromirror arrays with hidden vertical comb-drive actuators. *J. Microelectromechanical Syst.* **2004**, *13*, 279–289. [CrossRef]
5. Maroufi, M.; Fowler, A.G.; Moheimani, S.R. MEMS for nanopositioning: Design and applications. *J. Microelectromechanical Syst.* **2017**, *26*, 469–500. [CrossRef]
6. Chen, C.; Lee, C. Design and modeling for comb drive actuator with enlarged static displacement. *Sens. Actuators A Phys.* **2004**, *115*, 530–539. [CrossRef]
7. Hung, A.C.L.; Lai, H.Y.H.; Lin, T.W.; Fu, S.G.; Lu, M.S.C. An electrostatically driven 2D micro-scanning mirror with capacitive sensing for projection display. *Sens. Actuators A Phys.* **2015**, *222*, 122–129. [CrossRef]
8. Winter, C.; Fabre, L.; Conte, F.L.; Kilcher, L.; Kechana, F.; Abelé, N.; Kayal, M. Micro-beamer based on MEMS micro-mirrors and laser light source. *Procedia Chem.* **2009**, *1*, 1311–1314. [CrossRef]
9. Wolffenbuttel, R. MEMS-based optical mini-and microspectrometers for the visible and infrared spectral range. *J. Micromechanics Microeng.* **2005**, *15*, S145–S152. [CrossRef]
10. De Dobbelaere, P.; Falta, K.; Gloeckner, S.; Patra, S. Digital MEMS for optical switching. *IEEE Commun. Mag.* **2002**, *40*, 88–95. [CrossRef]
11. Schenk, H.; Durr, P.; Haase, T.; Kunze, D.; Sobe, U.; Lakner, H.; Kuck, H. Large deflection micromechanical scanning mirrors for linear scans and pattern generation. *IEEE J. Sel. Top. Quantum Electron.* **2000**, *6*, 715–722. [CrossRef]
12. Carlen, E.T.; Heng, K.H.; Bakshi, S.; Pareek, A.; Mastrangelo, C.H. High-aspect ratio vertical comb-drive actuator with small self-aligned finger gaps. *J. Microelectromechanical Syst.* **2005**, *14*, 1144–1155. [CrossRef]
13. Tsou, C.; Lin, W.T.; Fan, C.C.; Chou, B.C. A novel self-aligned vertical electrostatic combdrives actuator for scanning micromirrors. *J. Micromechanics Microeng.* **2005**, *15*, 855–860. [CrossRef]
14. Ono, T.; Sim, D.Y.; Esashi, M. Micro-discharge and electric breakdown in a micro-gap. *J. Micromechanics Microeng.* **2000**, *10*, 445. [CrossRef]
15. Zhang, W.M.; Yan, H.; Peng, Z.K.; Meng, G. Electrostatic pull-in instability in MEMS/NEMS: A review. *Sens. Actuators A Phys.* **2014**, *214*, 187–218. [CrossRef]
16. Degani, O.; Socher, E.; Lipson, A.; Lejtner, T.; Setter, D.J.; Kaldor, S.; Nemirovsky, Y. Pull-in study of an electrostatic torsion microactuator. *J. Microelectromechanical Syst.* **1998**, *7*, 373–379. [CrossRef]
17. Chen, Y.C.; Chang, I.C.M.; Chen, R.; Hou, M.T.K. On the side instability of comb-fingers in MEMS electrostatic devices. *Sens. Actuators A Phys.* **2008**, *148*, 201–210. [CrossRef]
18. Hah, D.; Toshiyoshi, H.; Wu, M.C. Design of electrostatic actuators for MOEMS applications. *SPIE* **2002**, *4755*, 200–207. [CrossRef]
19. Nathanson, H.C.; Newell, W.E.; Wickstrom, R.A.; Davis, J.R. The resonant gate transistor. *IEEE Trans. Electron Devices* **1967**, *14*, 117–133. [CrossRef]
20. Hirano, T.; Furuhashi, T.; Gabriel, K.J.; Fujita, H. Design, fabrication, and operation of submicron gap comb-drive microactuators. *J. Microelectromechanical Syst.* **1992**, *1*, 52–59. [CrossRef]
21. Bochobza-Degani, O.; Elata, D.; Nemirovsky, Y. A general relation between the ranges of stability of electrostatic actuators under charge or voltage control. *Appl. Phys. Lett.* **2003**, *82*, 302–304. [CrossRef]
22. Pamidighantam, S.; Puers, R.; Baert, K.; Tilmans, H.A. Pull-in voltage analysis of electrostatically actuated beam structures with fixed-fixed and fixed-free end conditions. *J. Micromechanics Microeng.* **2002**, *12*, 458. [CrossRef]
23. Zhou, G.; Dowd, P. Tilted folded-beam suspension for extending the stable travel range of comb-drive actuators. *J. Micromechanics Microeng.* **2002**, *13*, 178. [CrossRef]

24. Huang, W.; Lu, G. Analysis of lateral instability of in-plane comb drive MEMS actuators based on a two-dimensional model. *Sens. Actuators A Phys.* **2004**, *113*, 78–85. [CrossRef]
25. Lee, D.; Solgaard, O. Pull-in analysis of torsional scanners actuated by electrostatic vertical combedrives. *J. Microelectromechanical Syst.* **2008**, *17*, 1228–1238. [CrossRef]
26. Tang, W.C.K. *Electrostatic Comb Drive for Resonant Sensor and Actuator Applications*; University of California: Berkeley, CA, USA, 1990.
27. Liu, H.W. *Mechanics of Materials I/5th Edition*; Higher Education Press: Beijing, China, 2011.
28. Tang, W.C.; Lim, M.G.; Howe, R.T. Electrostatic comb drive levitation and control method. *J. Microelectromechanical Syst.* **1992**, *1*, 170–178. [CrossRef]
29. Hah, D.; Patterson, P.R.; Nguyen, H.D.; Toshiyoshi, H.; Wu, M.C. Theory and experiments of angular vertical comb-drive actuators for scanning micromirrors. *IEEE J. Sel. Top. Quantum Electron.* **2004**, *10*, 505–513. [CrossRef]
30. Yeh, J.L.; Hui, C.Y.; Tien, N.C. Electrostatic model for an asymmetric combedrive. *J. Microelectromechanical Syst.* **2002**, *9*, 126–135. [CrossRef]
31. Li, Y.; Psychogiou, D.; Kühne, S.; Hesselbarth, J.; Hafner, C.; Hierold, C. Large stroke staggered vertical comb-drive actuator for the application of a millimeter-wave tunable phase shifter. *J. Microelectromechanical Syst.* **2013**, *22*, 962–975. [CrossRef]
32. Legtenberg, R.; Groeneveld, A.W.; Elwenspoek, M. Comb-drive actuators for large displacements. *J. Micromechanics Microeng.* **1996**, *6*, 320. [CrossRef]
33. Bahgat, A.S.; Zaki, A.H.; Abdo Mohamed, M.; El Sherif, A.F. Design and simulation of MEMS-actuated adjustable optical wedge for laser beam scanners. *J. Micro/Nanolithography MEMS MOEMS* **2018**, *17*, 015501. [CrossRef]
34. Chyuan, S.W. Computational simulation for MEMS combedrive levitation using FEM. *J. Electrostat.* **2008**, *66*, 361–365. [CrossRef]

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Article

Design of a Micro-Electro Mechanical System Quad Mass Gyroscope with Compliant Mechanical Amplification

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Abstract: In this work, a novel mechanical amplification structure for a MEMS vibratory gyroscope is proposed with the aim of improving their sensitivity. The scheme is implemented using a system of micromachined V-shaped springs as a deflection amplifying mechanism. The effectiveness of the mechanism is first demonstrated for a capacitive fully decoupled quad mass gyroscope. A proof of concept vertical-axis mechanically amplified gyroscope with an amplification factor of 365% has been designed, simulated and fabricated, and results from its evaluation are presented in this paper. Experimental results show that the natural frequency of the gyroscope is 11.67 KHz, and the full scale measurement range is up to $\pm 400^\circ/\text{s}$ with a maximum nonlinearity of 54.69 ppm. The bias stability is $44.53^\circ/\text{h}$. The experiment results show that this quad mass gyroscope's performance is a very potential new way of reaching the navigation grade in the future.

Keywords: MEMS gyroscope; quadruple mass gyroscope; mechanical amplification; large mechanical sensitivity; finite element analysis

1. Introduction

Micro-Electro Mechanical Systems (MEMS) and especially gyroscope sensors are part of a flourishing market ranging from consumer electronics to space applications. Although MEMS gyroscopes have been evolving for more than 25 years, there is still considerable research effort for further improving their performance [1,2]. The performance of MEMS gyroscopes is often limited to transducing the initial physical signal into a mechanical displacement or deformation, which in turn causes a change in capacitance or piezoresistive. Typically, the resulting electrical changes are quite small and need to be amplified, which is usually carried out using charge amplifiers or transimpedance amplifiers. However, noise and other disturbance sources at the front-end of the electronic amplifier such as parasitic capacitances as well as intrinsic noise sources of the amplifier itself are enhanced, which restricts the sensitivity and the resolution of gyroscope systems [3,4].

The differential structure designs with two out-of-phase vibrating proof masses to reduce the vibration and acceleration induced errors are reported [5–8]. A further increase in the rejection of external stimuli has been proposed by University of California (Irvine) with an $8 \times 8 \text{ mm}^2$ quadruple mass gyroscope (QMG), which can minimize frequency and damping mismatches and maximize the Q-factor [9–11]. QMG is fully dynamically balanced in forces and momentum and provides mechanical rejection of external vibrations and mechanical shocks along both the drive and sense axis. Moreover, Singapore Institute of Microelectronics presents a QMG with new working modals, which is similar to the wine-glass vibrating gyroscope [12]. The size of this QMG was $5.8 \times 5.8 \text{ mm}^2$. Southeast University proposed a symmetrical quadruple mass gyroscope with four hair masses at

$8 \times 8 \text{ mm}^2$. It realized an improvement in sensitivity and a differential detection of the Coriolis force [13]. Tsinghua University designed a QMG in $7.7 \times 7.7 \text{ mm}^2$ with a center support frame which had a similar working principle with ring-like gyroscopes [14]. The QMGs demonstrated above were operated under wine-glass working modes. However, compared to the working modes used in University of California (Irvine), the wine-glass type did not provide a symmetric working mode which should be an advantage of QMGs. It should be noticed that all kinds of QMGs result in a high rejection to external vibrations and shocks [15,16].

Instead of amplifying the pick-off signal in the electrical domain for the first time, we show a primary amplification in the mechanical domain using a new compliant mechanical amplifier. The Coriolis force signal is therefore amplified before electric noise disturbances can affect the analog signal [17]. Compliant mechanical amplifiers are well known for piezo-electric actuator applications. Generally, these amplifiers can be divided into flexure-based and lever-based mechanisms [18–20]. In recent studies, lever-type compliant amplifiers had a great combination with an accelerometer. The University of Southampton have proposed a capacitive accelerometer with a lever-type mechanical motion amplifier [21]. By observed the contrast between the change of capacitance from the proof mass and end of the lever, it showed a higher sensitivity without sacrificing bandwidth. The University of California, Berkeley, proposed a resonant accelerometer with a two-stage lever as a force amplifier [22]. All of the mechanisms mentioned above realized a great amplification effect. We innovatively combined the compliant mechanical amplifier with a quadruple mass gyroscope.

The system proposed in this work is a novel rhombus-like compliant mechanical amplification concept for a silicon micro-machined resonant quadruple mass gyroscope (QMG) to improve its performance with a smaller active area of $4 \times 5 \text{ mm}^2$. The scheme is implemented using a system of rhombus-type compliant amplifiers as a deflection amplifying mechanism. This results in a larger Coriolis deflection and a higher signal-to-noise ratio compared to conventional QMGs. It also has the supplementary benefit of reducing the energy leakage from drive mode to sense mode.

The paper is organized as follows: Section 2 gives the demonstration about the structural design of quad mass gyroscopes, which includes the basic structure and the rhombus-like compliant mechanical amplification structure design. The analysis of the amplification structure is briefly introduced. Section 3 confirms the validity of the structural design with Finite-Element Analysis (FEA). Section 4 describes the fabrication process of the quad mass gyroscope, which is based on cavity SOI and gold–silicon eutectic bonding. The experimental methods and results are presented in Section 5. Finally, Section 6 concludes the whole paper with a summary.

2. Design and Analysis

2.1. Structure and Working Principle

Figure 1 shows the schematic diagram of the mechanical design of our gyroscope. It is a symmetric structure with quad proof masses. Each proof mass is connected to the drive frame by four straight beams. Two masses on both sides are linked together with a U-shaped beam, which can provide a large range of displacement that is linear to the stress. Masses on different sides are connected with a balance leverage which can make sure masses move in opposite direction but at the same amplitude. The upper and lower masses are connected through V-shaped springs, and the other end of V-shaped springs are connected to the sense frame which is composed of the movable parts of comb capacitors. There are four anchors in left and right balance beams, respectively, and two anchors in the middle of this structure.

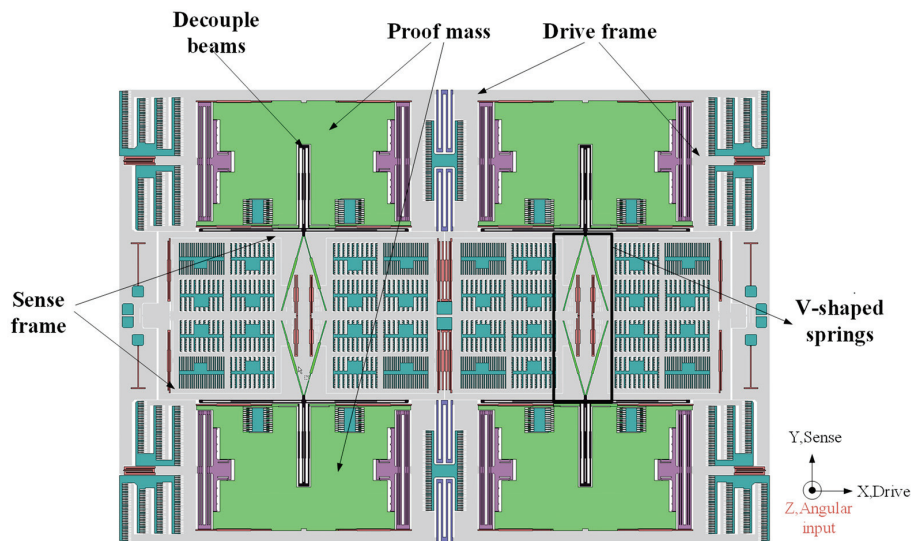


Figure 1. Structure of quad mass gyroscope.

The working principle of the gyroscope is explained as follows. The proof masses are electrostatically actuated to vibrate in the X direction as shown in Figure 2. Four masses were named as 1 to 4. For example, when the gyroscope worked, mass 1 would vibrate in a different direction to mass 2 and mass 3. The movement of each proof mass in the QMG is depicted in red arrows in Figure 2, and that was the basic drive mode of this QMG. When there is a rotation around the Z-axis, a Coriolis force would be induced. Then, it would apply on the proof mass to make it move in the Y direction, as the yellow arrows show in Figure 2.

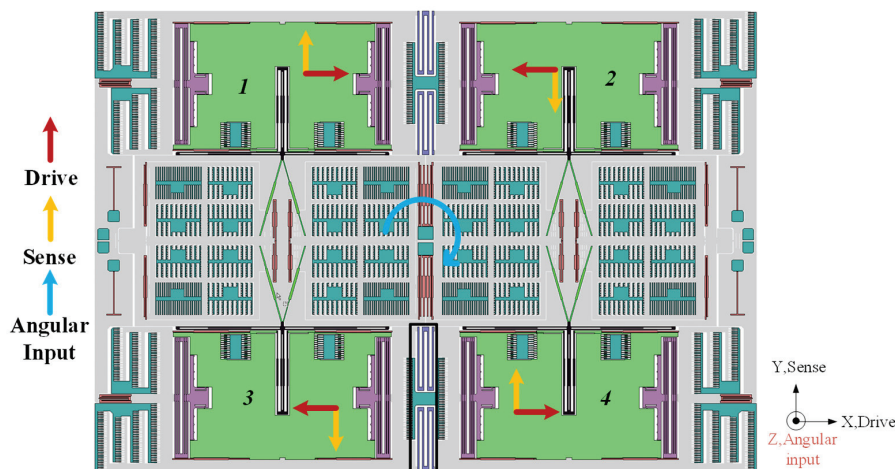


Figure 2. The basic working principle of quad mass gyroscope.

The displacement caused by the Coriolis force is then amplified by V-shaped springs (compliant mechanical amplification structure) which is used as a mechanical motion amplifier. Then, it changed the movement of the sense frame from the Y direction to X direction. Sensing capacitors connected to the sense frame can detect the movement. Because of the mechanical motion amplifier, the displacement is greatly increased. So, the mechanical sensitivity was literally improved. Moreover, the motion coupling of the sense and drive frame is only through the proof mass, the drive and sense frame movement can be perfectly decoupled due to a decouple spring design.

2.2. Design Strategy

The Coriolis effect is not obvious in the microscale resonator, and its signal is very weak. Therefore, it is quite necessary to design a gyroscope with high mechanical sensitivity.

Mechanical sensitivity is defined as the ratio of the displacement of the sense mode and the input angular velocity. The mechanical sensitivity of QMG can be expressed by (1).

$$y(t) = \frac{2Q_x\omega_x A_{drive}\Omega_z}{K_x\omega_y^2} \frac{1}{\sqrt{\left(1 - \frac{\omega_{drive}^2}{\omega_y^2}\right)^2 + \left(\frac{\omega_{drive}}{Q_y\omega_y}\right)^2}} \cos(\omega_{drive}t + \phi_y - \phi_x) \quad (1)$$

In Equation (1), Q_x and Q_y are quality factors of the drive and sense mode. ω_{drive} is the frequency of the drive force. ω_x and ω_y are the natural frequency of the drive and sense mode. A_{drive} is the amplitude of the drive mode. Ω_z is the input angular rate. K_x is the stiffness coefficient of the drive mode. When the sense mode is equal to the drive mode frequency, the mechanical sensitivity changes to (2).

$$y(t) = \frac{2Q_xQ_y A_{drive}\Omega_z}{K_x\omega_x} \sin(\omega_x t - \phi_y) = x_{drive} \frac{2Q_y\Omega_z}{\omega_x} \sin(\omega_x t - \phi_y) \quad (2)$$

It was shown clearly from Equations (1) and (2) that there is a large mechanical sensitivity need mode match, large Q value and large drive amplitude [23]. During the design state, the frequency tuning electrodes were reserved to the realized mode match. To obtain a large Q value, the fabrication process including eutectic bonding and integrated getter in the cavity of the CAP wafer to ensure a high vacuum level. Moreover, QMG's drive springs were folded into double U-shape springs which can provide a big movement of up to 20 μm . Despite the fact that the QMG has all of the factors considered above, in this work, we present a mechanical motion amplifier to increase the sensitivity of the Coriolis detection. The sensitivity amplification part was implemented in the front of electronic part, then the desired amplification will be achieved without the generation of additional noise.

2.3. Mechanical Motion Amplifier

Figure 3 shows the proposed design of the Mechanical motion amplifier which is basically simplified to V-shaped springs. As Figure 3 shows, d_y is the proof mass displacement caused by Coriolis force, and d_x is the displacement of sense frame, which is also the movement of the sensing capacitors' movable parts.

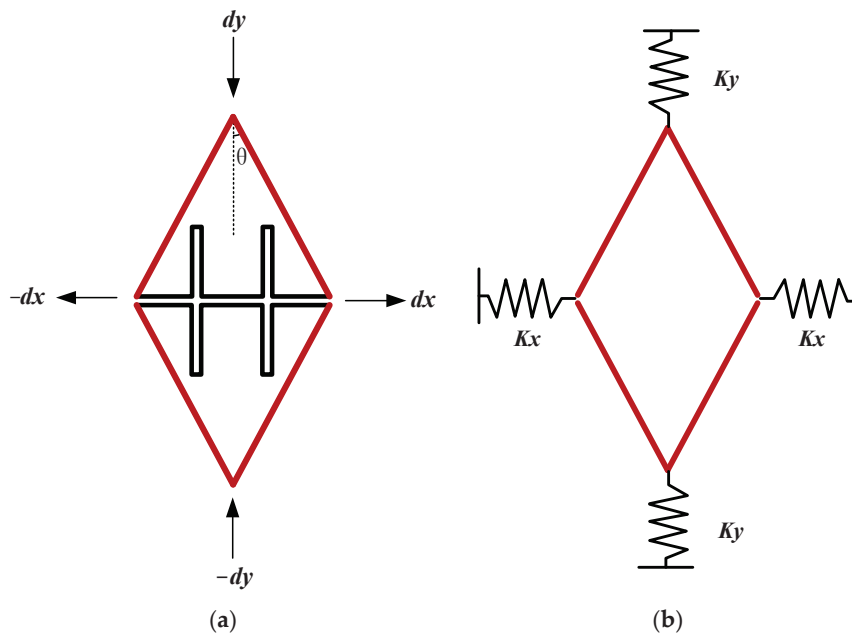


Figure 3. Simplified model of mechanical motion amplifier; (a) scheme and working principle of the V-shaped springs; (b) the simplified model for stiffness transformation analysis.

The angle θ is a parameter, which would decide the amplification factor (from d_y to d_x) of this mechanical motion amplifier. The amplification factor α can be calculated as in Equations (3) and (4).

$$\alpha = 1 / \tan \theta \quad (3)$$

$$d_x = \alpha d_y \quad (4)$$

To change the angle inside the V-shape springs, this amplifier could achieve different amplification factors. And the fold beam inside this V-shape springs is also carefully designed. When calculated, this quad mass gyroscope characteristic, the V-shape springs cannot provide enough stiffness to make the natural frequency of the sense mode match the drive mode which is firstly confirmed about 13 kHz.

According to Hooke's Law, the spring coefficient could be obtained by Equation (5). In this V-shaped spring, the spring coefficient in the X direction can be transferred into the Y direction as shown in Equation (6). Finally, considering the Y-axis as the sense direction, the equivalent stiffness of the sense direction can be calculated as Equation (7). Therefore, the fold beam can provide additional stiffness in the sense direction and match the sense mode and drive mode.

$$K_{spring} = F / X \quad (5)$$

$$K_X = \frac{F_x}{d_x} = \frac{\frac{1}{\alpha} F_y}{\alpha d_y} \quad (6)$$

$$K_s = \alpha^2 K_x + K_y \quad (7)$$

In Equation (7), K_x and K_y represent the stiffness coefficient of the beam connected to the V-shape springs. The angle θ should not be too big or too small. If θ was designed bigger than 25° , the amplification factor α would be smaller than 2, which cannot provide a satisfied effect on amplification. On the other hand, if θ was designed smaller than 10° , the stiffness of the X direction would be too small that the structure would be too fragile.

After optimization, the angle in this work was confirmed to be $\theta = 15^\circ$. The amplification factor of the mechanical motion amplifier was also simulated under the condition of steady analysis. As shown in Figure 4, in the simulation, a force which pretended to be the Coriolis force was given to the upper and lower end of this V-shaped spring. The result showed that in this amplifier, the displacement can be changed from that induced by Coriolis to the movement of sensing capacitors and be magnified 3.65 times, which is quite the same with calculation.

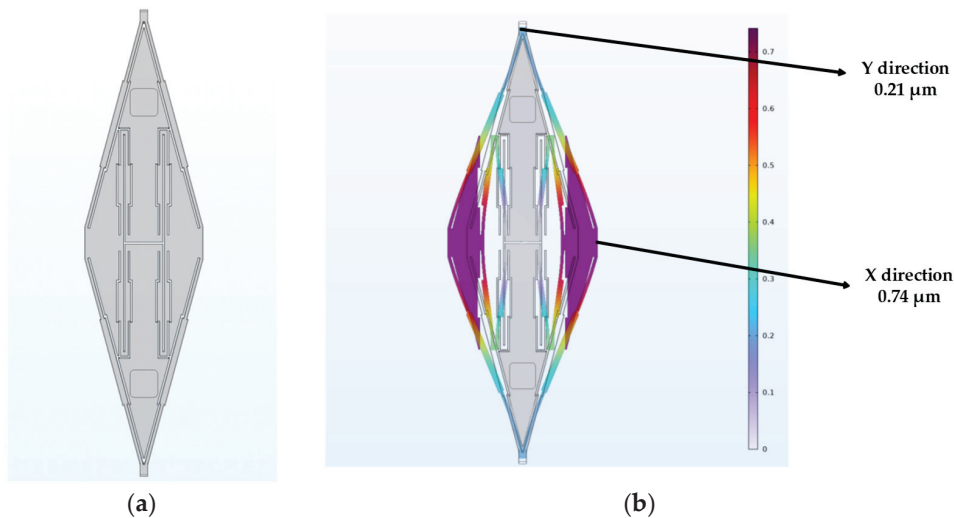


Figure 4. The results of steady simulation of mechanical motion amplifier. (a) The model of mechanical motion amplifier. (b) Displacement distribution of the Steady simulation.

2.4. Other Parameter Designs

The beams and the proof masses have the same thickness of 60 μm , the device size is 4 mm $\times$ 5 mm. The drive beams are four-fold beams with a width of 12 μm and length of 400 μm for each beam. The mechanical motion amplifier parameters are a width of 8 μm , length of 540 μm and angle of 15°, as discussed in Section 2.3. In the middle of the amplifier, the fold beam consists of a thinner and a thicker rectangle (Figure 4a). The thicker one's width is 15 μm , and the length is 230 μm ; this kind of composite rectangle design can precisely adjust the stiffness of the beams. This kind of design is used on almost all beams inside this QMG. More parameters about this gyroscope are as follows in Table 1.

Table 1. Main parameter of QMG structure.

Parameter	Drive Mode	Sense Mode
Resonant frequency	13,863 Hz	13,919 Hz
Spring constant	5590 N/m	5637 N/m
Come finger length	42 μm	7 μm
Come finger width	4 μm	4 μm
Nominal electrode overlap	18 μm	3 μm
Nominal capacitive gap	2 μm	2 μm
Number of fingers (all)	684	3812
Nominal capacitance (all)	6.54 pF	6.08 pF
Number of fingers (Drive)	492	1906
Nominal capacitance (Drive)	4.7 pF	3.04 pF
Number of fingers (Sense)	192	1906
Nominal capacitance (Sense)	1.84 pF	3.04 pF
Capacitive scale factor	0.102 pF/ μm	1.01 pf/ μm
Mass of proof mass	0.737 mg	0.627 mg

3. Finite Element Simulations

We used COMSOL, a commercial finite element program, to model this quad mass gyroscope. We considered a thickness of 60 μm for the whole structure which is the same as fabrication. The material properties of the silicon were chosen to be single-crystal (isotropic). We used free tetrahedral mesh to divide the structure with the maximum units of 50 μm and minimum units of 5 μm . Using the setting above, different kinds of Finite Element Analysis were carried out to observe the characteristics of this gyroscope.

First of all, the modal analysis was used to observe that the frequency of the QMG. The working modes of the gyroscope are simulated, and the results are shown in Figure 5. The modal frequency of the drive mode was 13.8 kHz and 13.9 kHz for the sense mode, which perfectly agreed with the design. Moreover, a harmonic response simulation was also carried out to check the decoupling design of the structure. In the simulation, a harmonic signal at the drive-mode natural frequency is given to the structure on the drive electrode, as shown in Figure 6a. The amplitude of the drive mode is 7.43 μm . Meanwhile, the observed amplitude of the sense mode is 5.5×10^{-3} μm . The amplitude ratio of the sense mode to the drive mode is 0.07%. In Figure 6b, a harmonic signal at the sense mode natural frequency is given to the gyroscope. The amplitude of the sense mode is 1.65 μm , and the amplitude of the sense mode is 0.8×10^{-3} μm . Then, the amplitude ratio of the drive mode to the sense mode is 0.04%. The results above show that the drive and sense movement can be literally decoupled.

The influence of linear acceleration to this structure was also carried out in COMSOL as shown in Figure 7. First, the gyroscope was vibrating in the drive mode, which is its normal working state. Then, linear acceleration ranging from 1 g to 10 g was loaded onto X-axis and Y-axis, respectively. In Figure 7a,b, the relationship between X-axis/Y-axis acceleration and the drive part displacement is depicted. The rest of the pictures are similar but show the displacement of the sense part.

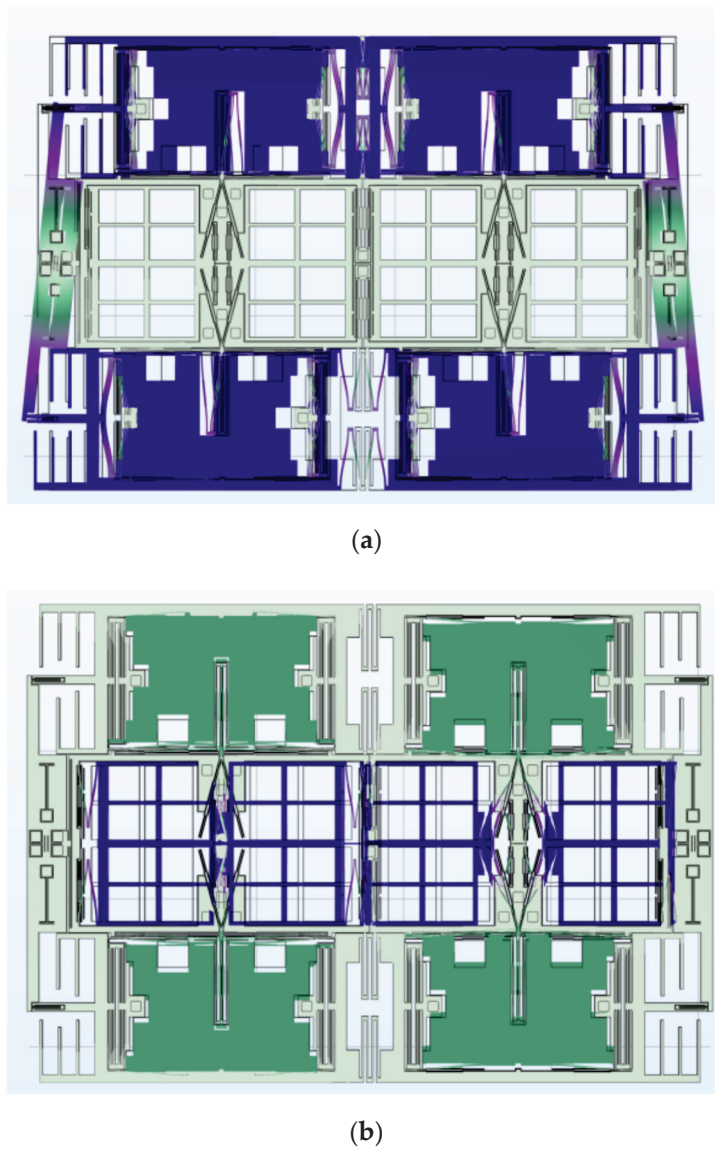


Figure 5. Modal analysis of gyroscope structure. (a) Drive mode. (b) Sense mode.

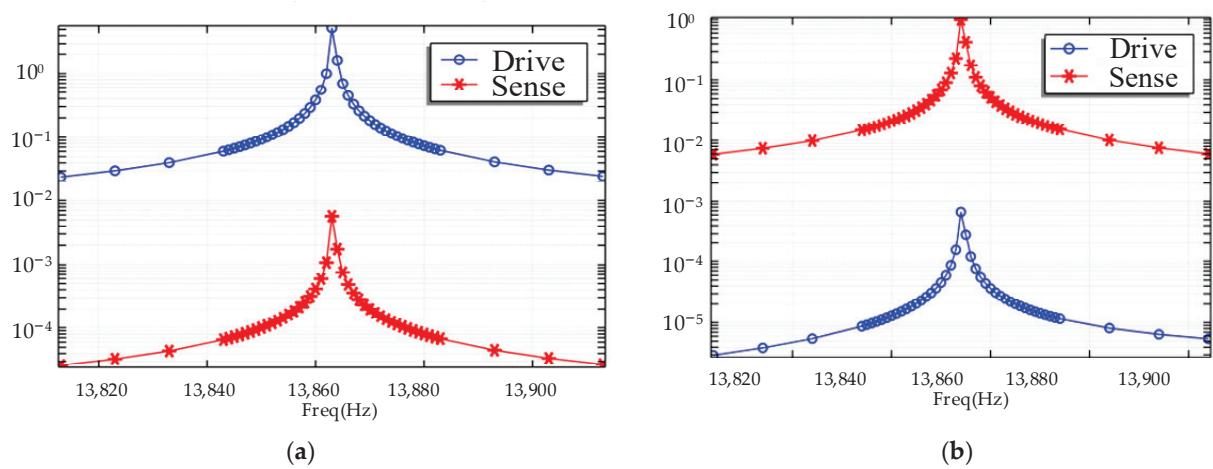


Figure 6. The results of harmonic response analysis. (a) Mechanical coupling from driving mode to sensing mode. (b) Mechanical coupling from sensing mode to driving mode.

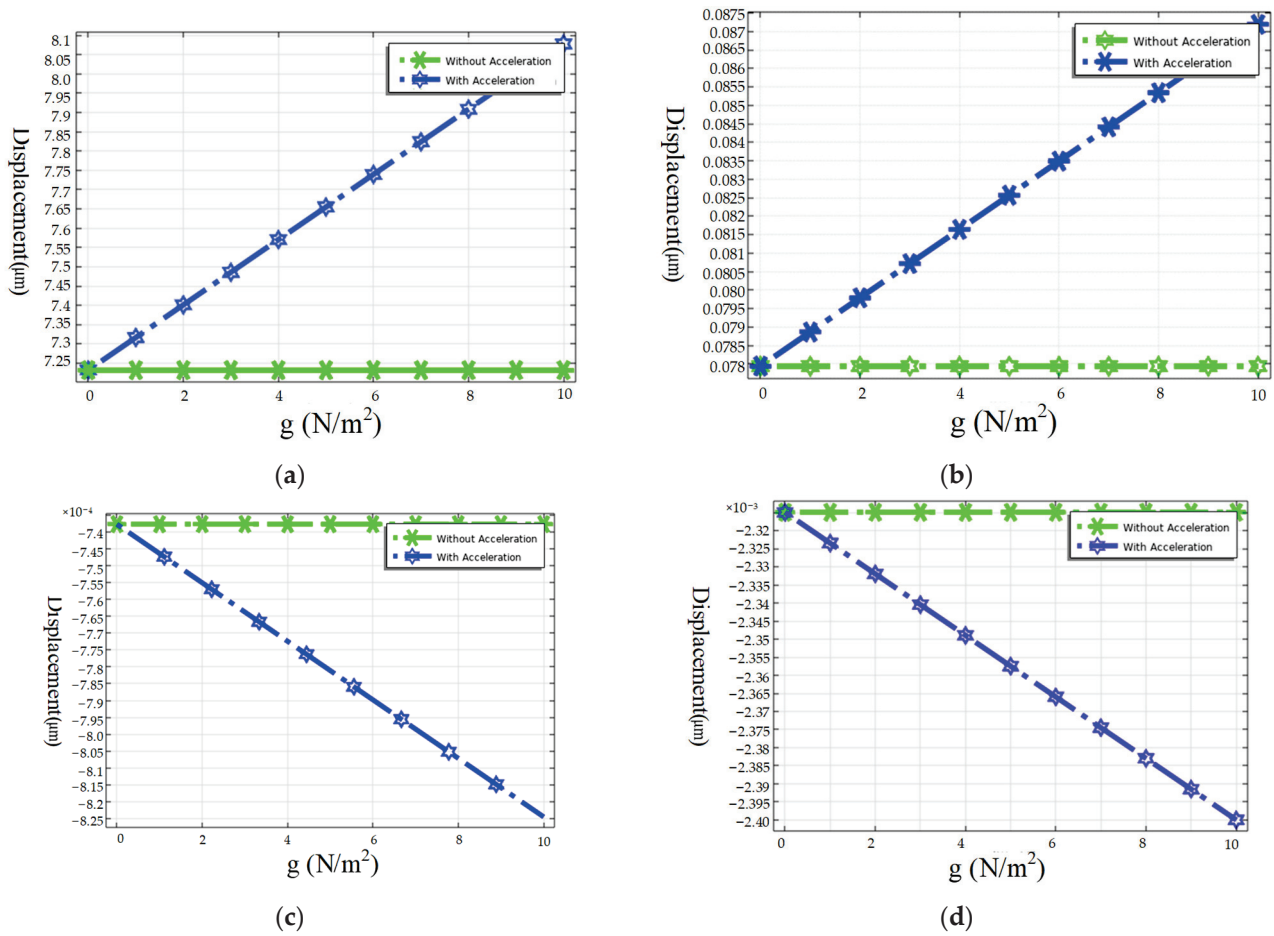


Figure 7. The results of linear acceleration analysis. (a,b) Drive parts' movement under X and Y direction acceleration, respectively. (c,d) Sense parts' movement under X and Y direction acceleration, respectively.

The simulation results showed that the acceleration had barely any effect on the change in sense capacitance. Though the drive part's movement under X-axis acceleration can reach about $0.8 \mu\text{m}$, the design of the differential sense capacitance makes sure it would not influence the output. Under other circumstances, the linear acceleration effect was calculated to be $0.03^\circ/\text{s/g}$, and this gyroscope had great resistance over linear acceleration.

4. Device Fabrication

The Gyroscope is divided into two parts. The device structure is fabricated with a silicon on insulator (SOI) wafer, and the CAP wafer is fabricated in a (100) orientated wafer.

Figure 8 depicts the whole fabrication process of this gyroscope. First, the back alignment marks of the CAP wafer are defined by mask1 and the Deep RIE process (Figure 8a). Then, it used mask2 to define thermal SiO_2 as a hard mask and etched $5 \mu\text{m}$ cavities using KOH wet etching (Figure 8b). After removing the hard mask, a 500 nm thermal oxide layer was grown as an insulator layer (Figure 8c). The first layer electrical connections are formed by metal sputtering Ti/W and wet etching of spare metal through the pattern of mask3 (Figure 8d). The insulator between the first and second metal layer is formed through PECVD and then used mask4 and SAMCO Rie to define the holes through two metal layers (Figure 8e). Finally, the bond ring and second metal layer are formed by metal sputtering and IBE with the pattern of mask5 (Figure 8f). The last step of the CAP wafer's fabrication was depositing the getters (Figure 8g).

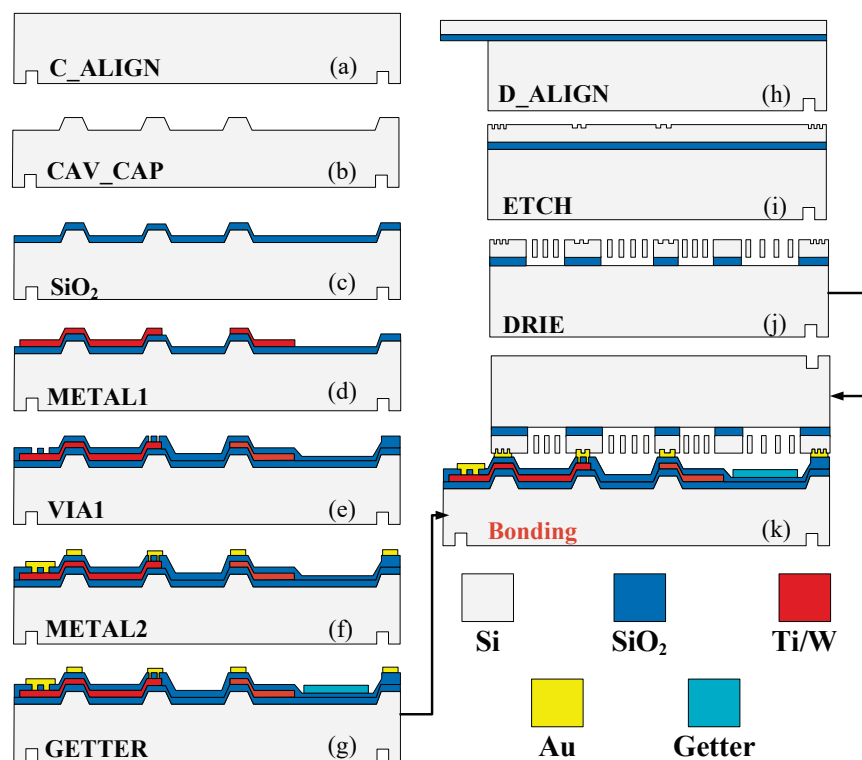


Figure 8. Fabrication process.

The device wafer is fabricated using the following steps. First, the back alignment is defined with mask7. The positions of the structure which are on top of the electrode are first released from the backside of the wafer by etching the handle silicon layer through to the buried oxide layer using Deep RIE with mask8 (Figure 8h). Then, a shallow trench of about 4 μm was defined with mask9 and etched using silicon dioxide as a mask (Figure 8i). Finally, the structure is etched through to the buried oxide layer by DRIE with mask10, then released by the Vapor HF etch (Figure 8j). The device is fully fabricated after the gold silicon eutectic bonding of the CAP wafer and the device wafer. The final device is shown in Figure 9.

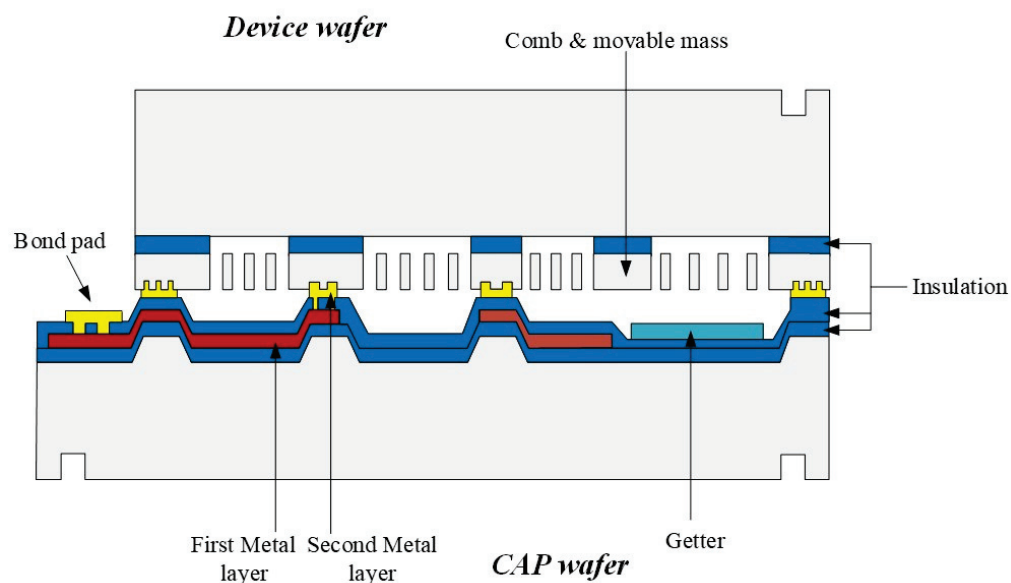


Figure 9. Device structure schematic.

The SEM images of the fabricated device are shown in Figures 10 and 11. Due to the device limitations, only half of the QMG can fit into an image. As Figure 10 shows, the drive capacitance finger length and overlap length were $42\text{ }\mu\text{m}$ and $4\text{ }\mu\text{m}$, respectively. The sense capacitance finger length and overlap length were $18\text{ }\mu\text{m}$ and $2\text{ }\mu\text{m}$, respectively. They were both fabricated as designed. However, the finger gap was $0.4\text{ }\mu\text{m}$ larger than designed because of the Deep RIE CD loss. The SEM image of the CAP wafer was shown in Figure 11. The width of the metal wire on top and the bonding ring was $20\text{ }\mu\text{m}$ and $120\text{ }\mu\text{m}$, respectively. All of the key dimensions of the CAP wafer were closely matched with the design values.

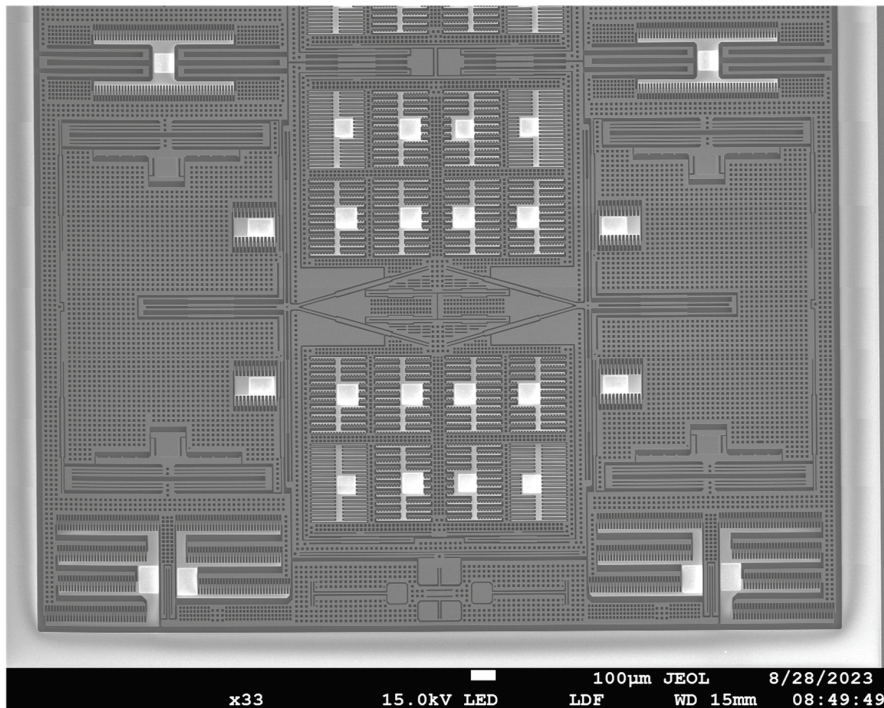


Figure 10. SEM image of fabricated gyroscope device wafer.

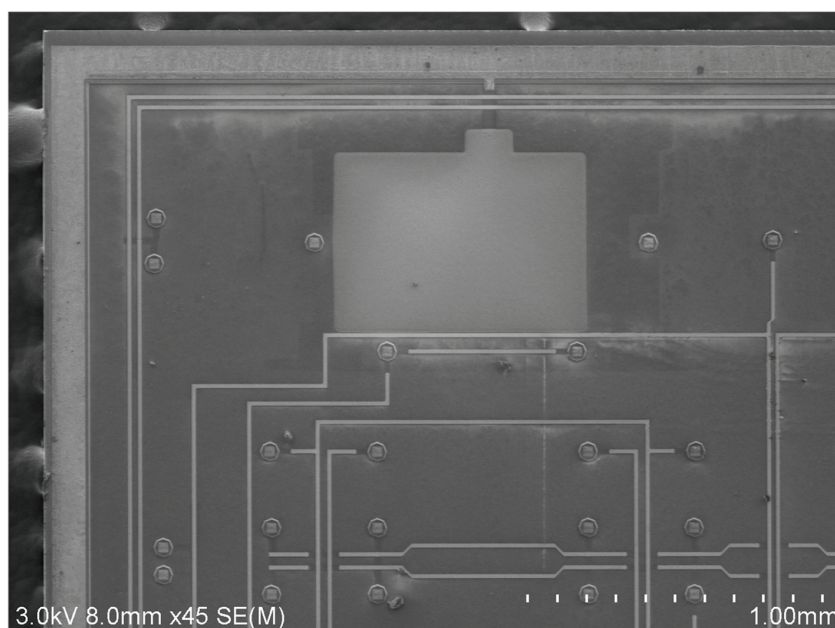


Figure 11. SEM image of fabricated gyroscope CAP wafer.

5. Test Results

5.1. Frequency Response Test

The frequency response and ring-down tests were performed in order to characterize the basic parameters of the QMG. The experimental circuit is shown in Figure 12. In order to minimize the influence of parasitic capacitance and external environment on weak capacitance signals, the Gyroscope is encapsulated separately into the ceramic package. Therefore, the sensor is in a relatively stable environment. Then, the frequency response test was carried on.

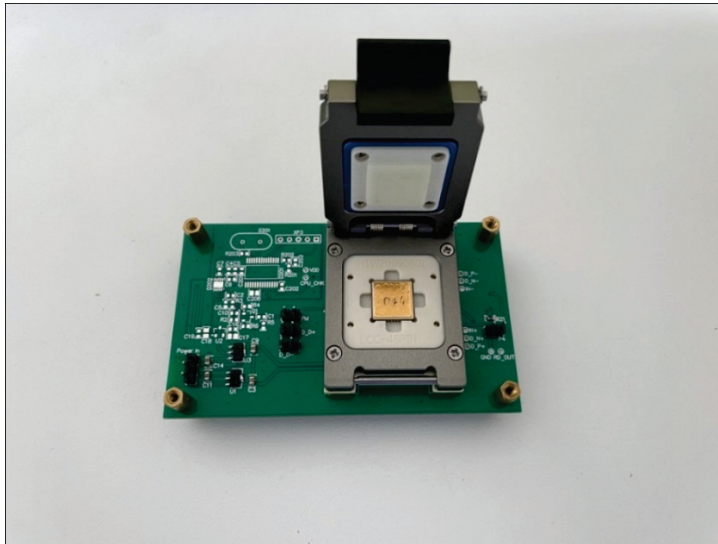


Figure 12. Photo of the experimental PCB.

To precisely get the ring-down signals, we use a capture board with a maximum sampling rate of 250 kHz. During the test, the sample rate was set to be 120 kHz, which is about ten times over the frequency of the gyroscope. The sample time was set to be 3 s. The total data volume is 360,000. The ring-down part of data was extracted and replotted into curves using MATLAB R2021a, as shown in Figure 13b.

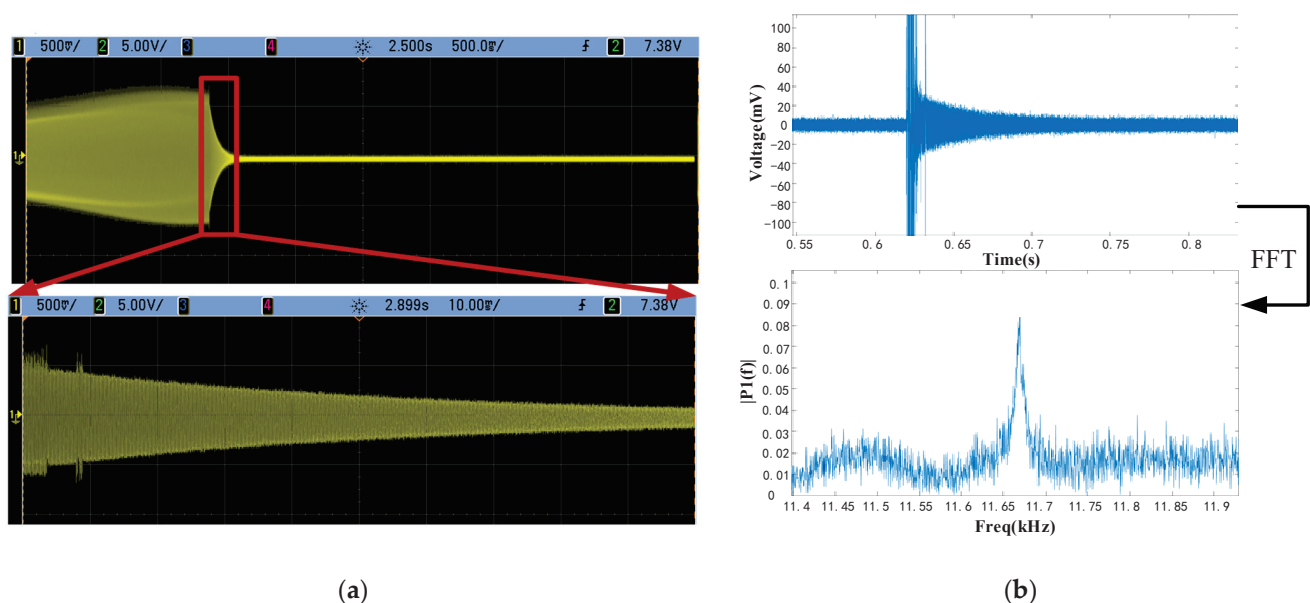


Figure 13. Results of ring-down test. (a) Ring-down test result on oscilloscopes, (b) ring-down data on MATLAB and FFT analysis of ring-down data.

The frequency characteristics were obtained by calculating the results of the ring-down test in MATLAB using the FFT function. After calculation, the accurate resonant frequency f was shown in Figure 13b. The resonant frequency obtained from the analysis is 11.67 kHz, which is basically within a reasonable range for the results of the structural simulation. The error was mostly caused by fabrication. During the Deep RIE process, the drive beams were etched more than designed, which caused a lower stiffness in the drive mode. This resulted in a lower natural frequency.

5.2. Gyroscope Characteristic Testing

The overall gyroscope measurement and control circuits consisted of two parts. As shown in Figure 14, the first one included a MEMS gyroscope, a CV circuit, a quadrature calibration voltage module and a frequency tuning module. In Figure 15, the other one includes an FPGA control circuit, signal demodulation circuits, AD/DA circuits, excitation single-to-dual circuits, a carrier generation circuit and a voltage reference module. Among them, the design of the C/V circuit is shown in Figure 16. The C/V analog interface circuit consisted of charge amplification circuits, high-pass filter circuits, diode demodulation circuits, low-pass filter circuits and anti-phase difference amplifier circuits.

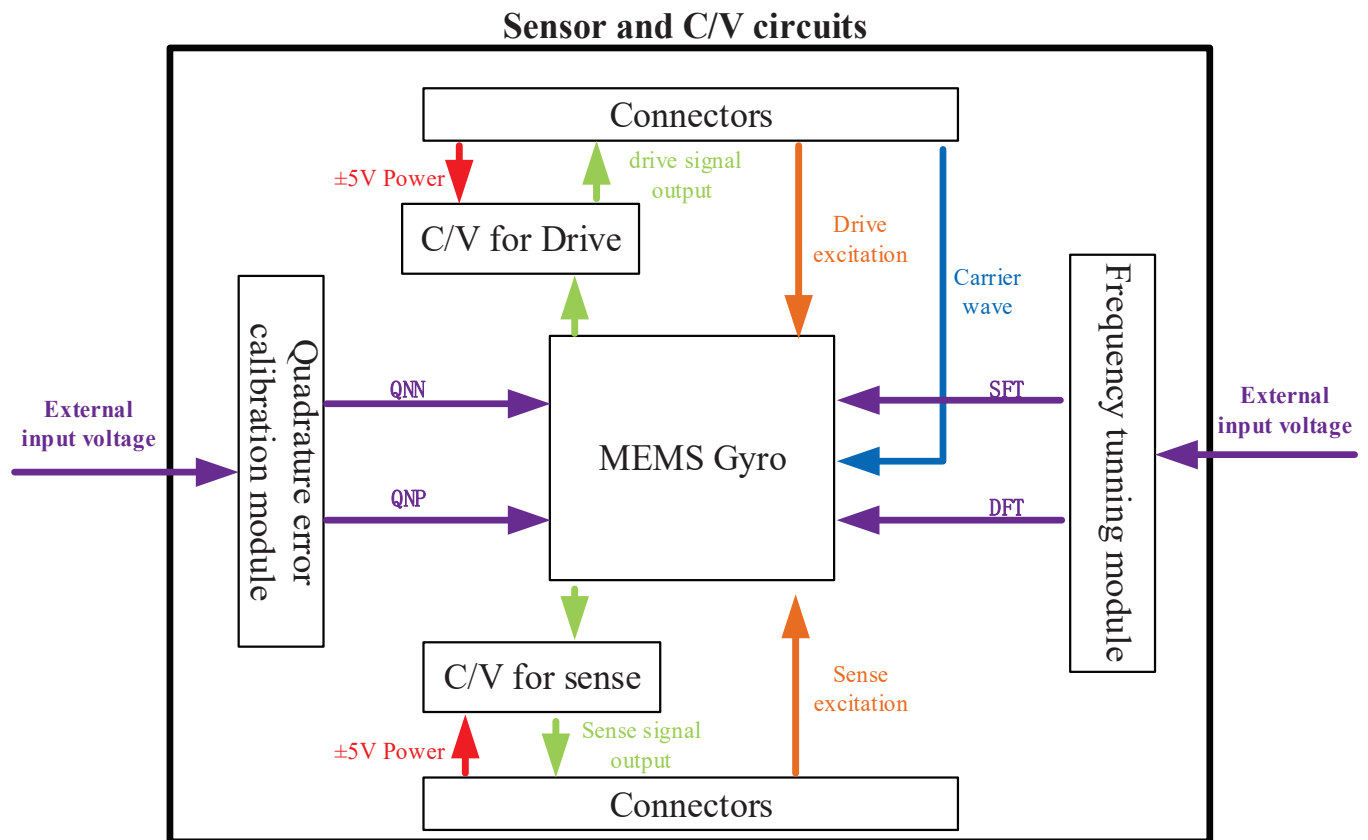


Figure 14. The diagram of gyroscope measurement and control circuit's daughter board.

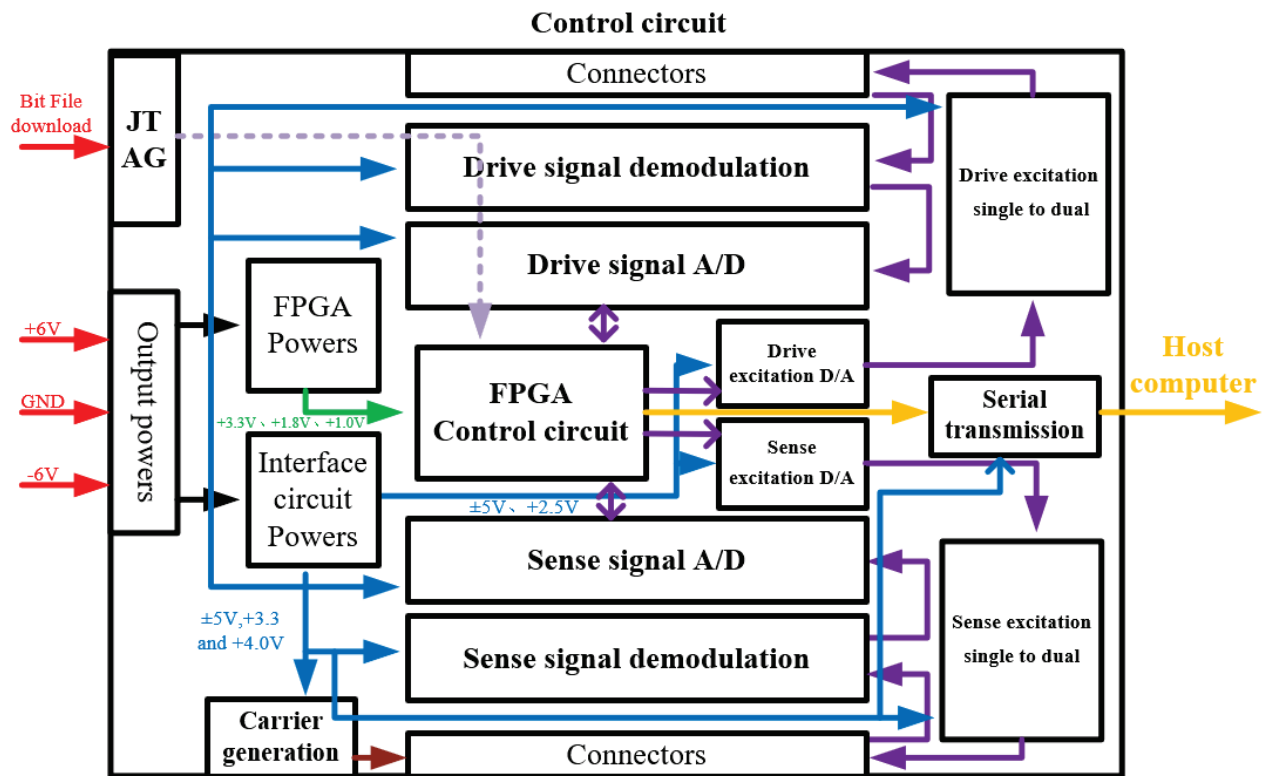


Figure 15. The diagram of gyroscope measurement and control circuit's mother board.

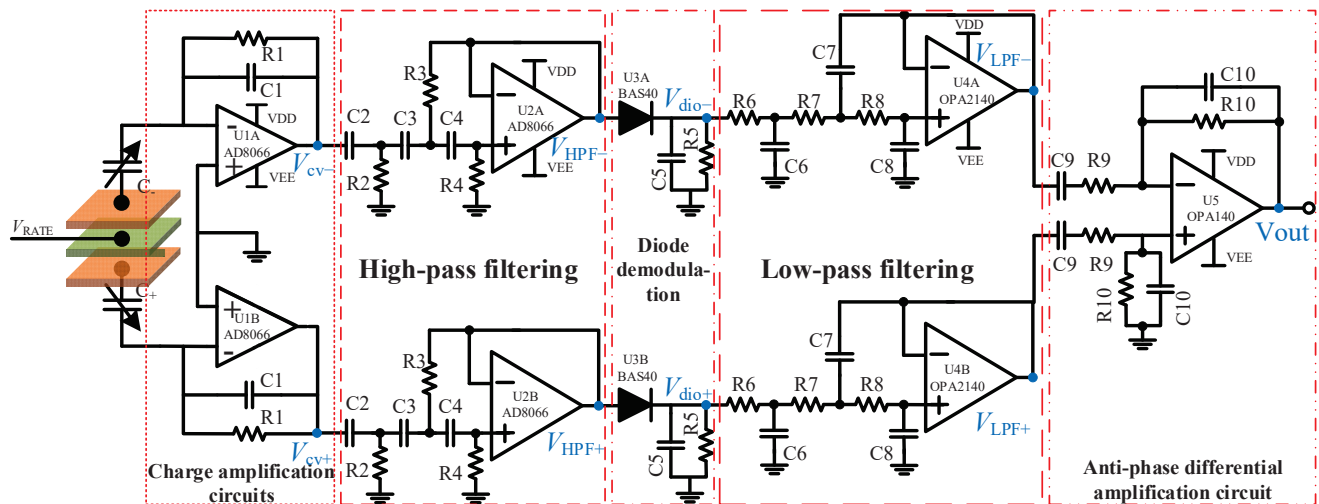


Figure 16. The diagram of gyroscope measurement and control circuit's C/V analog interface circuit.

When the gyroscope was working, the drive signal was modulated in a mother board and delivered to a daughter board, then input into the device through A/D circuits. From the drive mode output signal, the vibration information of the device can be obtained and transferred from the daughter board to mother board through C/V circuits. Moreover, it was analyzed by FPGA control circuits to output a feedback drive signal. This close-loop circuit ensured that the gyroscope was vibrating under a constant amplitude, constant frequency situation.

After an angular rate input, the gyroscope sense mode was excited. The output sense signal was also transferred into the FPGA control circuit. After analyzing this signal, the FPGA will output a feedback signal to cancel the vibrations in the sense mode. In our

circuits, The frequency tuning module and quadrature error calibration module were also designed to improve the performance of the gyroscope.

The C/V circuits in Figure 16 mean that the capacitance signal can change to a voltage signal. The charge amplification circuit can pick out the change in the charge change on capacitive plates and then turn it into a voltage signal. The high-pass filter circuit can filter out the carrier wave signal. Then, the diode demodulation circuit demodulates a useful signal from the carrier signal. Finally, two signals were sent to the differential amplification circuit to realize the elimination of common-mode signals and differential amplification of useful signals.

Using the circuits mentioned above, the Gyroscope characteristic tests were carried out. The first test was about the scale factor nonlinearity. A rate table is used to apply an angular rate to the QMG so that we can get its angular rate response. Figure 17 shows the measured angular rate response of the QMG. The input angular rate is from $-400^\circ/\text{s}$ to $400^\circ/\text{s}$. The X-axis data of Figure 17 are the angular rate from the rate table, which are used as a standard angular rate. The Y-axis data were the angular rate output of the QMG after normalization. The scale factor nonlinearity is 54.69 ppm in the full-scale range of $\pm 400^\circ/\text{s}$.

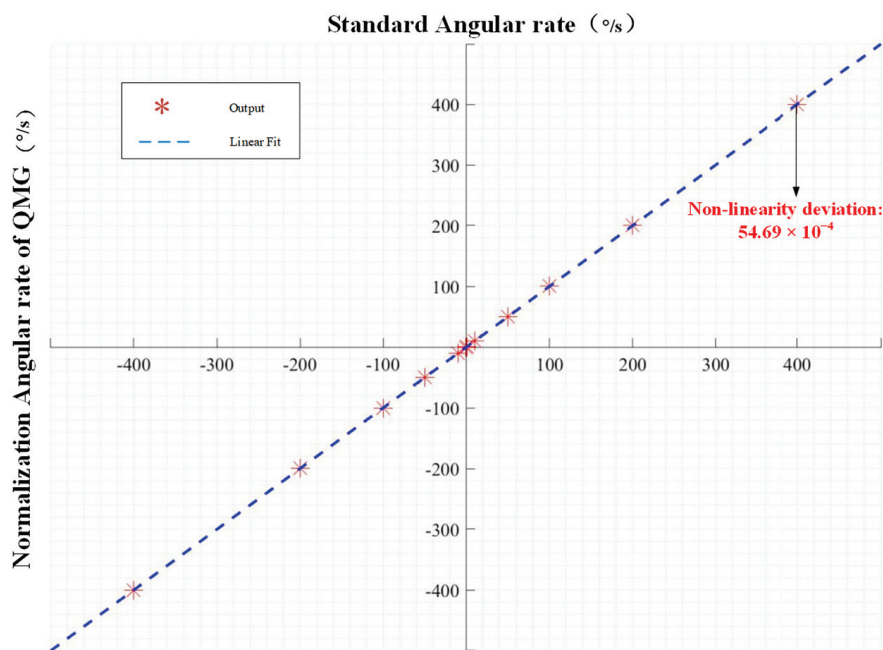


Figure 17. Measured angular rate output of Gyroscope. X-axis represents the normalized QMG response; Y-axis represents the angular rate of rate table. The scale factor nonlinearity is 54.69 ppm in the full range of $\pm 400^\circ/\text{s}$.

The zero rate output (ZRO) data are collected for 2 h at room temperature in order to calculate the bias stability of the QMG gyroscope system. The zero rate output of the QMG under constant temperature and pressure is shown in Figure 18. In Figure 18, the blue line is the raw data of the Gyroscope output, and the red line is the data being averaged for 10 s. As Figure 18 shows, the zero rate output data had a lot of undesired signals, which may have been induced by external vibration. By calculating the standard deviation of the averaging data, the bias stability was obtained to be $44.53^\circ/\text{h}$.

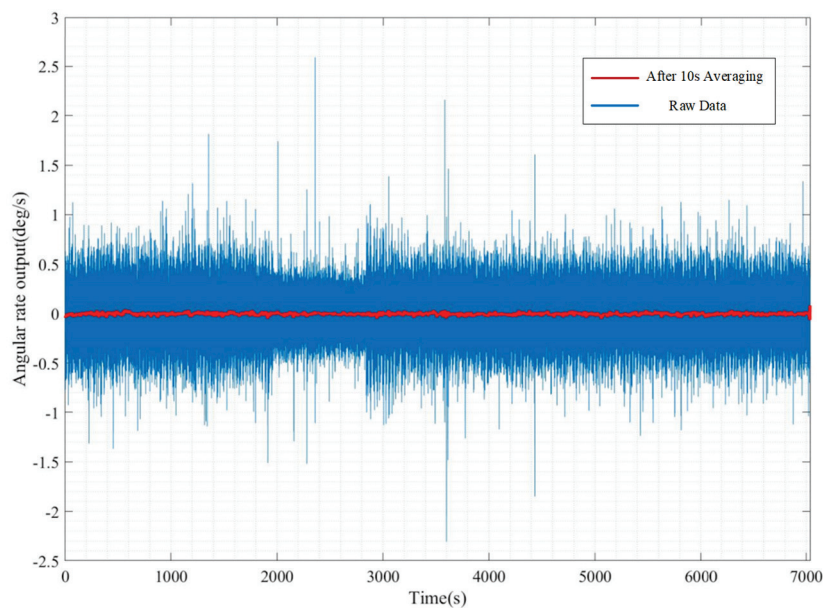


Figure 18. Angular rate output of the QMG system. The blue line is the raw data and the red line is the data being averaged for 10 s. After calculating, the bias stability was $44.53^\circ/\text{h}$.

6. Conclusions

In this paper, the design of a quadruple mass gyroscope with compliant mechanical amplification was successfully proposed, designed and fabricated. Due to the innovative mechanical amplification design, the Coriolis force signal is therefore amplified before the electric domain, which successfully enlarged the mechanical sensitivity of the QMG for 365%. The QMG can realize the decoupling of the drive mode and sense mode and is insensitive to external vibrations such as linear accelerations. Then, the gyroscope was fabricated through gold–silicon bonding of an SOI wafer and a Cap wafer. The prototype has shown a scale factor nonlinearity of 55.69 ppm. The bias stability of the QMG is calculated to be $44.53^\circ/\text{h}$. The QMG demonstrated in our work realized the combination of mechanical amplification mechanisms with the gyroscope. The mechanical amplification structure also provides a new way to design high-performance inertial sensors.

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Conflicts of Interest: The authors declare no conflicts of interest.

References

1. Xia, D.; Yu, C.; Kong, L. The Development of Micromachined Gyroscope Structure and Circuitry Technology. *Sensors* **2014**, *14*, 1394–1473. [CrossRef]
2. Zhao, C.; Montaseri, M.H.; Wood, G.S.; Pu, S.H.; Seshia, A.A.; Kraft, M. A review on coupled MEMS resonators for sensing applications utilizing mode localization. *Sens. Actuators A Phys.* **2016**, *249*, 93–111. [CrossRef]
3. Mohd-Yasin, F.; Nagel, D.J.; Korman, C.E. Topical Review: Noise in Mems. *Meas. Sci. Technol.* **2010**, *21*, 209–213. [CrossRef]
4. Khan, S.; Ananthasuresh, G.K. Improving the sensitivity and bandwidth of in-plane capacitive microaccelerometers using compliant mechanical amplifiers. *J. Microelectromech. Syst.* **2014**, *23*, 871–887. [CrossRef]

5. Sharma, A.; Zaman, M.F.; Zucher, M.; Ayazi, F. A 0.1° /HR bias drift electronically matched tuning fork microgyroscope. In Proceedings of the 2008 IEEE 21st International Conference on Micro Electro Mechanical Systems, Tucson, AZ, USA, 13–17 January 2008; pp. 6–9.
6. Zaman, M.F.; Sharma, A.; Hao, Z.; Ayazi, F. A Mode-Matched Silicon-Yaw Tuning-Fork Gyroscope with Subdegree-Per-Hour Allan Deviation Bias Instability. *J. Microelectromech. Syst.* **2008**, *17*, 1526–1536. [CrossRef]
7. Sharma, A.; Zaman, M.F.; Ayazi, F. A Sub- 0.2° /hr Bias Drift Micromechanical Silicon Gyroscope with Automatic CMOS Mode-Matching. *IEEE J. Solid-State Circuits* **2009**, *44*, 1593–1608. [CrossRef]
8. Trusov, A.A.; Schofield, A.R.; Shkel, A.M. Gyroscope architecture with structurally forced anti-phase drive-mode and linearly coupled anti-phase sense-mode. In Proceedings of the TRANSDUCERS 2009—2009 International Solid-State Sensors, Actuators and Microsystems Conference, Denver, CO, USA, 21–25 June 2009; pp. 660–663.
9. Asadian, M.H.; Askari, S.; Shkel, A.M. An Ultrahigh Vacuum Packaging Process Demonstrating Over 2 Million Q-Factor in MEMS Vibratory Gyroscopes. *IEEE Sens. Lett.* **2017**, *1*, 6500104. [CrossRef]
10. Trusov, A.A.; Atikyan, G.; Rozelle, D.M.; Meyer, A.D.; Zotov, S.A.; Simon, B.R.; Shkel, A.M. Flat is not dead: Current and future performance of Si-MEMS Quad Mass Gyro (QMG) system. In Proceedings of the 2014 IEEE/ION Position, Location and Navigation Symposium—PLANS 2014, Monterey, CA, USA, 5–8 May 2014; pp. 252–258.
11. Trusov, A.A.; Prikhodko, I.P.; Zotov, S.A.; Schofield, A.R.; Shkel, A.M. Ultra-high Q silicon gyroscopes with interchangeable rate and whole angle modes of operation. In Proceedings of the IEEE SENSORS, Waikoloa, HI, USA, 1–4 November 2010; pp. 864–867.
12. Wu, G.; Chua, G.L.; Singh, N.; Gu, Y. A Quadruple Mass Vibrating MEMS Gyroscope with Symmetric Design. *IEEE Sens. Lett.* **2018**, *2*, 2501404. [CrossRef]
13. Guo, X.; Yang, B.; Li, C. A stereoscopic symmetrical quadruple mass gyroscope. In Proceedings of the 2018 IEEE 4th Information Technology and Mechatronics Engineering Conference (ITOEC), Chongqing, China, 14–16 December 2018; pp. 476–479.
14. Zhou, B.; Zhang, T.; Yin, P.; Chen, Z.; Song, M.; Zhang, R. Innovation of flat gyro: Center Support Quadruple Mass Gyroscope. In Proceedings of the 2016 IEEE International Symposium on Inertial Sensors and Systems, Laguna Beach, CA, USA, 22–25 February 2016; pp. 42–45.
15. Gao, Y.; Huang, L.; Ding, X.; Li, H. Design and implementation of a dual-mass MEMS gyroscope with high shock resistance. *Sensors* **2018**, *18*, 1037. [CrossRef] [PubMed]
16. Gao, Y.; Li, H.; Huang, L.; Sun, H. A lever coupling mechanism in dual-mass micro-gyroscopes for improving the shock resistance along the driving direction. *Sensors* **2017**, *17*, 995. [CrossRef] [PubMed]
17. Pedersen, C.B.W.; Seshia, A. On the optimization of compliant force amplifier mechanisms for surface micromachined resonant accelerometers. *Micromech. Microeng.* **2004**, *14*, 1281–1283. [CrossRef]
18. Schmitt, P.; Hoffmann, M. Engineering a Compliant Mechanical Amplifier for MEMS Sensor Applications. *J. Microelectromech. Syst.* **2020**, *29*, 214–227. [CrossRef]
19. Lai, L.-J.; Zhu, Z.-N. Design, modeling and testing of a novel flexure-based displacement amplification mechanism. *Sens. Actuators A Phys.* **2017**, *266*, 122–129. [CrossRef]
20. Chen, G.; Ma, Y.; Li, J. A tensural displacement amplifier employing elliptic-arc flexure hinges. *Sens. Actuators A Phys.* **2016**, *247*, 307–315. [CrossRef]
21. Zeimpekis, I.; Sari, I.; Kraft, M. Characterization of a Mechanical Motion Amplifier Applied to a MEMS Accelerometer. *J. Microelectromech. Syst.* **2012**, *21*, 1032–1042. [CrossRef]
22. Su, S.X.P.; Yang, H.S.; Agogino, A.M. A resonant accelerometer with two-stage microleverage mechanisms fabricated by SOI-MEMS technology. *IEEE Sens. J.* **2005**, *5*, 1214–1223. [CrossRef]
23. Zaman, M.F. Degree-Per-Hour Mode-Matched Micromachined Silicon Vibratory Gyroscopes. Ph.D. Thesis, Georgia Institute of Technology, Atlanta, GA, USA, 2008.

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Article

Control Software Design for a Multisensing Multicellular Microscale Gas Chromatography System

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Abstract: Microscale gas chromatography (μ GC) systems are miniaturized instruments that typically incorporate one or several microfabricated fluidic elements; such systems are generally well suited for the automated sampling and analysis of gas-phase chemicals. Advanced μ GC systems may incorporate more than 15 elements and operate these elements in different coordinated sequences to execute complex operations. In particular, the control software must manage the sampling and analysis operations of the μ GC system in a time-sensitive manner; while operating multiple control loops, it must also manage error conditions, data acquisition, and user interactions when necessary. To address these challenges, this work describes the investigation of multithreaded control software and its evaluation with a representative μ GC system. The μ GC system is based on a progressive cellular architecture that uses multiple μ GC cells to efficiently broaden the range of chemical analytes, with each cell incorporating multiple detectors. Implemented in Python language version 3.7.3 and executed by an embedded single-board computer, the control software enables the concurrent control of heaters, pumps, and valves while also gathering data from thermistors, pressure sensors, capacitive detectors, and photoionization detectors. A graphical user interface (UI) that operates on a laptop provides visualization of control parameters in real time. In experimental evaluations, the control software provided successful operation and readout for all the components, including eight sets of thermistors and heaters that form temperature feedback loops, two sets of pressure sensors and tunable gas pumps that form pressure head feedback loops, six capacitive detectors, three photoionization detectors, six valves, and an additional fixed-flow gas pump. A typical run analyzing 18 chemicals is presented. Although the operating system does not guarantee real-time operation, the relative standard deviations of the control loop timings were $<0.5\%$. The control software successfully supported >1000 μ GC runs that analyzed various chemical mixtures.

Keywords: portable GC; embedded systems; C#; GUI; Python; I2C; SMBus; GPIO; JSON; firmware; middleware

1. Introduction

Microscale gas chromatography (μ GC) systems have shown promise for use in in situ detection and measurement of volatile organic compounds (VOCs) since their introduction in the 1970s [1,2]. A μ GC system incorporates a number of fluidic components for gas-phase flow management. The major fluidic components in μ GC systems include preconcentrators, separation columns, detectors, pumps, and valves [3]. Typical operation involves two steps that constitute an analytical run: first, a sampling step is performed, in which the VOC molecules are drawn into the system using a sampling pump and adsorbed in the preconcentrator; subsequently, the VOC molecules are thermally desorbed and transported through the separation column by a carrier gas flow, during which the VOC molecules

are spatiotemporally separated. Further downstream, the detector responds to the eluting analytes and forms a chromatogram, in which the location and magnitude of each peak indicate the species and quantity of a VOC, respectively.

The performance of μ GC systems can be enhanced using complex architectures with a larger number of fluidic components. Some μ GC systems incorporate multiple complementary detectors to enhance chemical recognition. These detectors are based either on a single sensing principle (e.g., capacitive detection [4], chemi-resistive detection [5], or resonance-based detection [6]) or on a mix of sensing principles with higher orthogonality [7]. A 2D μ GC system typically incorporates two columns of complementary separation mechanisms to enhance the separation; typically, it requires either a thermal modulator [8] or a flow modulator [9]. The recently reported progressive cellular architecture [7,10] incorporates multiple sets of preconcentrators and columns, with each set tailored for a different range of VOC volatilities.

In a μ GC system, the diverse fluidic components all require specialized controls. For example, the preconcentrators, separation columns, and detectors require separate temperature controls, as they are operated at different temperatures and with different heating rates. The heating control is usually provided by either pulse-width modulation (PWM) of a fixed voltage or by controlling an analog voltage. This control is preferably implemented in a servo-controlled (closed-loop) manner using temperature measurements from either thermistors or thermocouples connected to analog-to-digital converters (ADCs) [11,12]. Capacitive detectors are typically read out using capacitance-to-digital converters (CDCs) [4,7,10,13]; photoionization detectors are read out of amplifiers and ADCs [7,14]; resonance-based detectors are read out by phase-locked loop circuits [6] and ADCs. The valves used in μ GC systems are typically active valves, either latching or non-latching. Latching valves are actuated using a short voltage pulse, the polarity of which controls whether the valve is opened or closed. In contrast, non-latching valves need sustained voltage levels but with an additional complexity favored for certain models, in which the sustained voltage should be lower than the initial actuating voltage. The commonly used miniature piezoelectric or motor-driven diaphragm pumps are controlled by on/off power and sometimes in combination with a tunable actuation frequency.

Within an analytical run, all the μ GC components must be continuously and concurrently controlled. One of the challenges is that the data rates of different elements can vary widely. For example, the data rates of detectors vary based on the sensing principles and the readout electronics; higher data rates are generally favored. Other components may differ even more widely in data rates. The control software must accommodate the heterogeneous timing of individual components, with some components operating in closed-loop control.

The control software must also implement fail-safe protocols to handle hardware error cases, such as accidental overheating and thermal runaway. Otherwise, these error cases may cause irreversible damage to the μ GC system.

To achieve standalone operation, all these hardware controls must be provided by an embedded computer (EC). Conversely, to facilitate user interaction, a user interface is often preferred on an external computer that can communicate with the EC.

This work presents advances in control software using a representative μ GC system that is based on a multisensing progressive cellular architecture (MPCA) [7]. This MPCA system is an example of a highly complex μ GC system with a large number and variety of components that require concurrent software control. The MPCA system incorporates eight elements that require individual temperature control, six detectors that require capacitive readout, three detectors that require voltage readout, six latching valves, and three pumps with different control requirements. This work investigates a multithreaded software architecture for the concurrent control of the MPCA system, with specific considerations regarding thread implementation, thread timing, temperature and flow control, communications between the EC and electronics, data saving and storage, low-power operation, and the user interface (UI) that communicates with the EC.

2. Design and Implementation

2.1. MPCA System Overview

The MPCA system incorporates a monolithic microfabricated MPCA chip in addition to multiple miniature valves, pumps, and pressure sensors. The MPCA chip integrates three μ GC cells, each containing a preconcentrator, a separation column, and three complementary detectors (Figure 1). The three cells are tailored to different ranges of analyte volatility, primarily by altering the sorbent materials in the preconcentrators and adjusting the stationary phase thicknesses in the separation columns. The three detectors within each cell are based on different sensing principles, providing complementary responses that enhance chemical recognition. Two of the detectors are capacitive detectors (CapDets) that use coplanar interdigitated thin-film metal electrodes with a sensing polymer of controlled thickness. The absorption of incoming analytes causes changes in both the thickness and dielectric constant of this layer, leading to changes in the detector capacitance. The first capacitive detector (CapDetA) primarily responds to polymer swelling, which produces only positive changes to the capacitance for all analytes. In contrast, the response of the second capacitive detector (CapDetB) is dominated by the change in dielectric constant, which may produce positive or negative changes to the capacitance depending on the dielectric constant of the analytes relative to that of the polymer itself. The third detector is an arrayed integrated photoionization detector (AiPD), which uses biased coplanar electrodes to draw a current from the photoionization of the analyte molecules. The photoionization is caused by ultraviolet radiation. The AiPDs across the three cells share a vacuum ultraviolet (VUV) lamp. Additionally, the MPCA chip incorporates a carrier gas filter, which dehumidifies ambient air to supply the carrier gas. For each preconcentrator, each separation column, the carrier gas filter, and the ensemble of all the detectors, a pair of on-chip heater and thermistor are incorporated to support closed-loop temperature control.

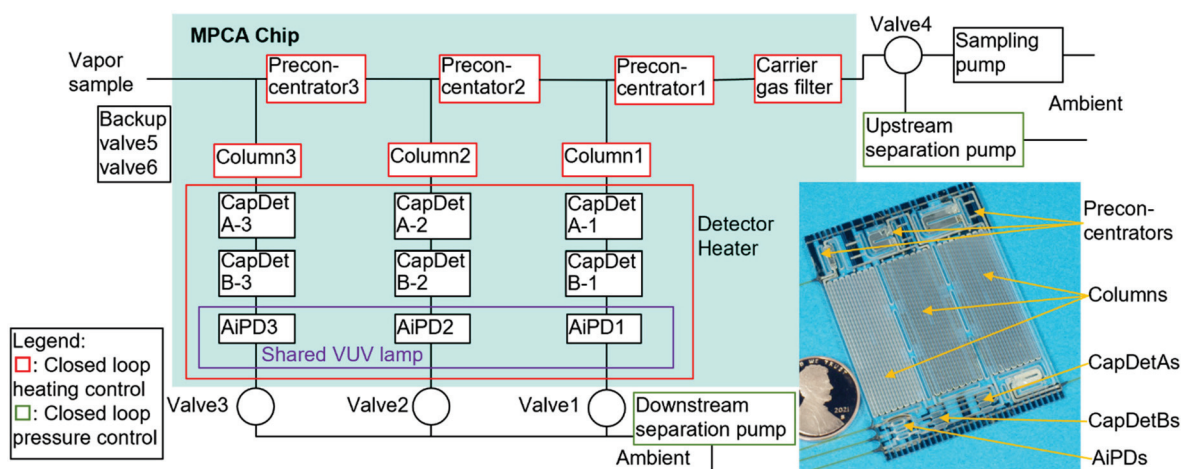


Figure 1. MPCA fluidic architecture and photo of the chip.

The MPCA system operation includes a sampling step, three sequential separation steps (one for each cell), and multiple purging steps. During the sampling step, the flow is generated using a sampling pump and routed using the valves. The analyte samples are transported through the preconcentrators sequentially, first through the one with the least adsorptive sorbent and then through those with progressively more adsorptive sorbents. The analytes are adsorbed by the preconcentrators, with the lower-volatility chemicals adsorbed by the upstream preconcentrator, and the more volatile chemicals breaking through the upstream preconcentrator and being adsorbed by downstream preconcentrators. During the separation step for each cell, the corresponding preconcentrator is heated to desorb the chemicals, which are then carried forward by the carrier gas flow and separated by the corresponding column and subsequently detected by the detectors. The carrier gas flow during separation is generated using both an upstream separation pump and a downstream

separation pump, both of which may be used in conjunction with a pressure sensor to support closed-loop flow control. Prior to the next sampling step, purging steps may be used to reduce humidity and chemical carryover; these purging steps involve combinations of component heating and carrier gas flow. To support the MPCA system operation and continuing development, the control software must provide users with enough flexibility to modify operation sequences and parameters.

2.2. Electronic Interfaces

The EC serves as both the controller for the MPCA system electronics and the interface with the laptop UI (Figure 2). The EC may be embodied as either a microcontroller [11] or a single-board computer [5]. Microcontrollers are highly compact and inexpensive, but their limitations in computational power, visualization, and storage resources increase the burden on software development. In contrast, single-board computers provide a favorable balance between device size, cost, computational power, and ease of software development. The control and readout functions in this work can be supported by either a microcontroller or a single-board computer. However, from the perspective of a fully autonomous system for continuous, near-real-time in situ analysis, the post-processing of raw data requires extensive computation [15,16]. An alternative to the EC is a data acquisition (DAQ) device [17]; it can be easily programmed and is attractive for proof of concept. However, without a built-in controller, the control software for a DAQ device typically resides on an external computer, which does not always facilitate standalone operation. This work uses a Raspberry Pi (RPi) series single-board computer [18] as the EC. The RPi series includes various models with different computational power, power consumption, and sizes. Most models have the same pinout and the same operating system, which facilitates upgrades to newer models. For this work, the control software is built in the RPi Model 3B+ [18], which was the latest model available at the start of this work.

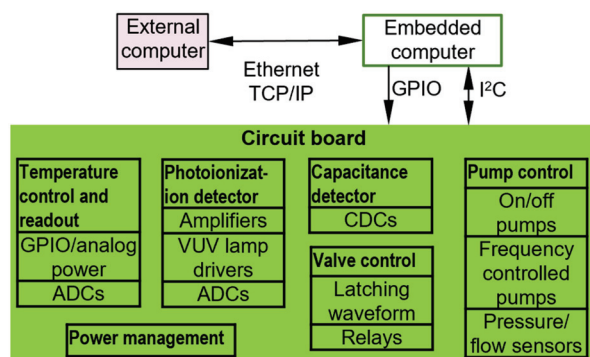


Figure 2. Block diagram of the electronic interface between components.

The electronic interfaces used within the MPCA system are selected for their compatibility with the general-purpose input/output (GPIO) pins and/or the inter-integrated circuit (I^2C), which are readily available on the EC. A GPIO pin can be easily programmed to provide simple on/off control or to provide square waveforms with controllable frequency and duty cycles. An I^2C bus, operating with just two wires, may be used to communicate with tens of integrated circuit (IC) chips. This capability makes the I^2C more suitable for the MPCA system than other protocols, such as the serial peripheral interface (SPI) and the universal asynchronous receiver/transmitter (UART).

The component heating is controlled by pulse-width modulation (PWM) for certain elements and by analog voltage control for other elements. The PWM-based circuit uses a relay (#CS326, Coto Technology, North Kingstown, RI, USA) controlled by a PWM signal from the EC (Figure 3a). By varying the duty cycle of a 24 V square wave, the PWM signal controls the average output voltage using just one GPIO pin, while eliminating additional ICs. However, for certain components that are sensitive to electrical interference, an analog voltage control circuit is desired. This is implemented using a digital potentiometer (#

MCP45HVX1, Microchip Technology, Chandler, AZ, USA) that controls the output voltage of a buck converter (# TPS54061DRBR, Texas Instruments, Dallas, TX, USA) (Figure 3b). In the MPCA system, the heating of the preconcentrators, columns, and carrier gas filter uses the PWM-based circuit, whereas the heating of the ensemble of the detectors uses the analog voltage control circuit.

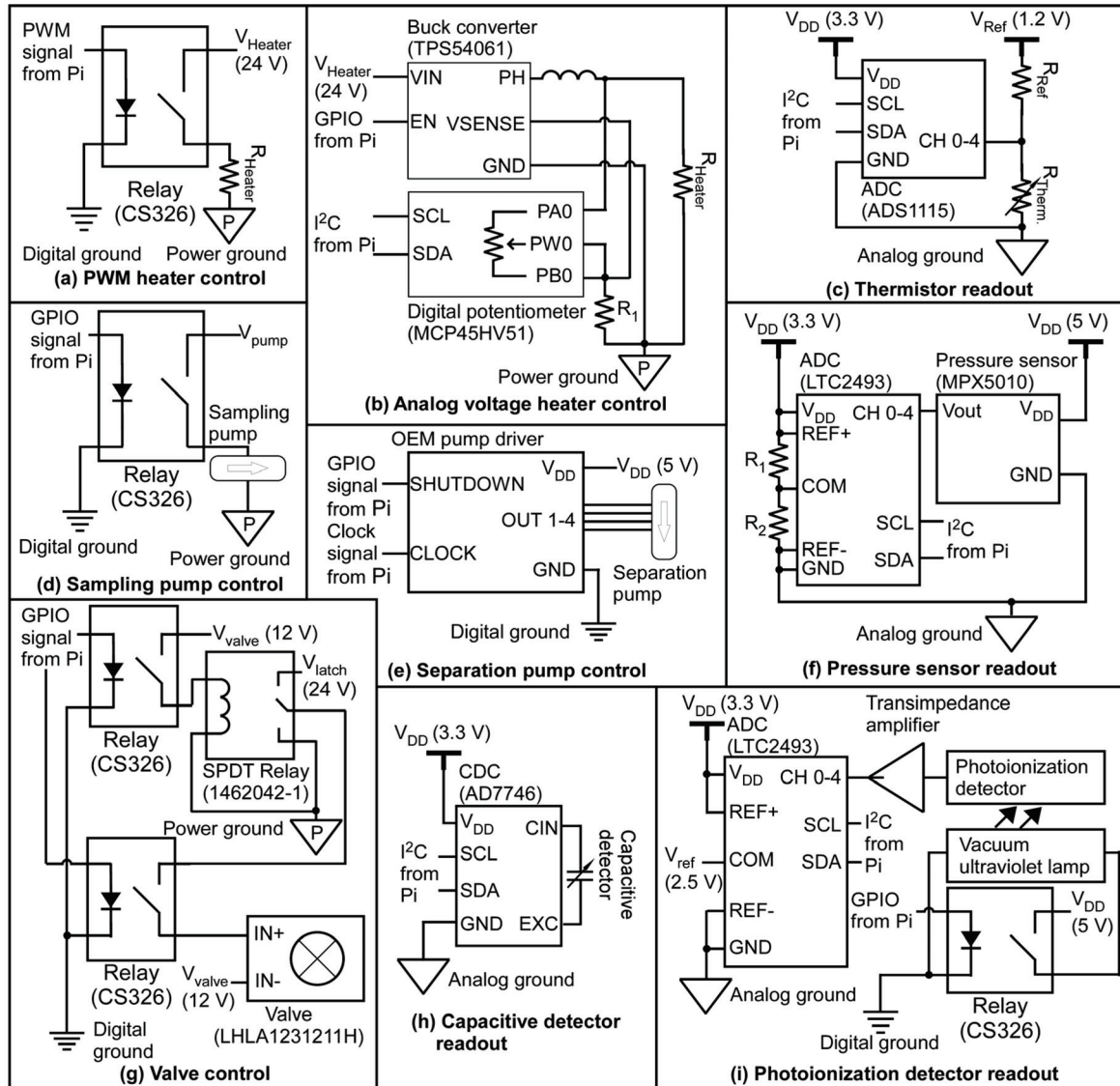


Figure 3. Circuit schematics of μ GC system hardware.

For temperature measurements, each on-chip thermistor is read out using a voltage divider coupled to an ADC (#ADS1115, Texas Instruments, Dallas, TX, USA) (Figure 3c). This ADC provides 16-bit resolution at a data rate of 8–860 Hz. It contains a converter core that is shared by four input channels in a time-multiplexed manner. Therefore, the eight on-chip temperature channels can be read out using two chips of this ADC, maintaining the same resolution but at $\frac{1}{4}$ of the data rate compared to an individual channel. These are necessary for the closed-loop temperature control of the MPCA system.

The sampling pump is a mini motor-driven pump (#NMP03KPDC-L, KNF Neuberger Inc., Trenton, NJ, USA) that is controlled using a relay (Figure 3d). The separation flow requires more accurate and adjustable control. The two separation pumps are piezoelectric micropumps (#mp6-air, Servoflo Corporation, Lexington, MA, USA), each driven using an original equipment manufacturer (OEM) driver circuit. The driver circuit receives

on/off control from a GPIO pin; it also receives a square wave with a variable frequency of 0–1000 Hz to control the flow rate of the separation pump (Figure 3e). To implement closed-loop control of the separation flow rate, two differential pressure sensors (#MPX5010DP, NXP Semiconductors, Eindhoven, The Netherlands) are used to measure upstream and downstream pressure heads generated by the two separation pumps; these two pressure sensors are directly read out using a high-resolution ADC (#LTC2493, Linear Technology, Milpitas, CA, USA) (Figure 3f), which is rated with a 24-bit resolution at a data rate of 7.5 Hz. It contains a converter core shared by four time-multiplexed input channels. Therefore, the two pressure sensors can be read out using one chip of this ADC, maintaining the same resolution but at half of the data rate (i.e., 3.75 Hz) compared to an individual channel.

Six latching solenoid valves (#LHLA1231211H, The Lee Company, Westbrook, CT, USA) are used to route gas flow within the system. A single-pole double-throw relay (#1462042-1, TE Connectivity, Bevaix, Switzerland) is used to select the voltage level between that required for latching and unlatching a valve. Additionally, multiple relays (#CS326, Coto Technology, East Greenwich, RI, USA) are used to separately control the actuation of individual valves (Figure 3g).

Each of the two CapDet elements in the active cell is read out using a separate high-resolution capacitance-to-digital converter (CDC) (#AD7746ARUZ, Analog Devices Inc., Wilmington, MA, USA) (Figure 3h); the CDC is rated for a root mean square (RMS) input-equivalent noise level of 4 aF and a peak-to-peak noise of 27 aF, at a data rate of 9.1 Hz. A limitation of this CDC is its single fixed I²C address, which allows for only one CDC chip on each I²C bus. This limitation is overcome by using an I²C address translator (#LTC4316, Linear Technology, Milpitas, CA, USA) to assign different I²C addresses to the additional CDC chips. All the CDC chips are set in continuous conversion mode, operating at the slowest data rate (9.1 Hz) to achieve the highest available resolution. The excitation signal is set at 32 kHz frequency and ± 1.65 V peak-to-peak amplitude. The built-in capacitance offset is set at 13.344 pF, which provides an effective measurement range of 9.248 pF to 17.440 pF. More details of the CDC chip connectivity are described in Supporting Information Section S1.

For the AiPDs, the shared VUV lamp is controlled using on/off power. The current generated by each AiPD is amplified and converted into a voltage signal using a low-noise transimpedance amplifier circuit. This voltage is read out using a high-resolution ADC (#LTC2493, Linear Technology, Milpitas, CA, USA), which is the same model as that for the pressure sensors (Figure 3i). Because of the sequential (i.e., non-concurrent) operation of the separation steps in the three cells, the three AiPDs can be read out using three channels of this ADC chip without time multiplexing. Therefore, each AiPD can be read out at a nominal data rate of 7.5 Hz (which is twice as fast as for the pressure sensors).

To power the MPCA system, both 24 V and 12 V supplies are needed, each of which can be sourced from either a wall plug power supply or a battery pack (Figure 4). The 12 V supply is dedicated to valve actuation. In contrast, the 24 V supply is used directly for component heaters (Figure 3a,b) and for valve actuation (Figure 3g). Additionally, the 24 V supply is used to provide 2 V for sampling pump actuation using a DC-DC converter (LMR14010A, Texas Instruments, Dallas, TX, USA). The 24 V supply is used to provide a 5 V supply, i.e., V_{Digital} (5 V), using another DC-DC converter (R-745.0, RECOM Power, Gmunden, Austria). The V_{Digital} (5 V) is used to provide a 3.3 V supply, i.e., V_{Digital} (3.3 V), using a voltage regulator (NCP51460, Onsemi, Scottsdale, AZ, USA). These digital voltage supplies are used for digital electronics. Additionally, the V_{Digital} (5 V) is treated using a Pi filter to provide another low-noise 5 V supply, i.e., V_{Analog} (5 V), which is used to provide another low-noise 3.3 V supply, i.e., V_{Analog} (3.3 V). These analog voltage supplies supply the data converters. The V_{Analog} (5 V) is also converted to 1.2 V and 2.5 V reference voltages for the ADC chips using two different power regulators (LM4041, Texas Instruments, Dallas, TX, USA and LT6657, Analog Devices Inc., Wilmington, MA, USA, respectively).

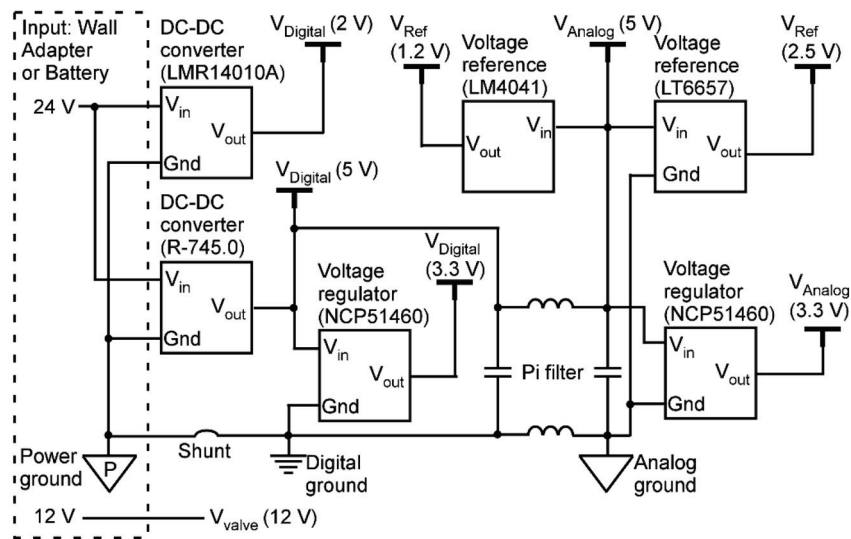


Figure 4. Schematic illustration of the power management circuit.

2.3. Control Software Implementation

In the MPCA system, the control software executed by the EC is programmed in Python 3.7.3. As the default language for the RPi, Python is compatible with a wide range of built-in and online open-source libraries, easing control software development. The main libraries that were used for this work include the open-source pigpio library [19] for GPIO pin control, the open-source SMbus library [20] for I²C communication, the Python built-in multithreading library for achieving multithreading, and the Python built-in JSON library for handling operation method files in JavaScript Object Notation (JSON) format. An operation method file contains all the necessary information for automating an analytical run. Each analytical run can be divided into up to eight run steps (such as a sampling step, a separation step, or a purging step). The operation methods file contains parameters for thermal, fluidic, detector, and timing controls of each run step. The libraries for interfacing with the electronics, namely SMbus and pigpio, are open-source libraries with core functions implemented in C programming language [19,20], which provides the necessary performance for the control software. The specific challenges and solutions are described in the following sub-sections.

2.3.1. Multithreading Architecture

In each run step of the MPCA system operation, multiple fluidic components must be controlled concurrently. There are two mainstream approaches for providing concurrency: multiprocessing and multithreading (Figure 5). Multiprocessing provides true parallelism [21], where multiple tasks are executed simultaneously using multiple central processing unit (CPU) cores. It is most suitable for tasks that require intensive and parallel CPU usage. In contrast, multithreading achieves concurrency through context switching [22], where only one task is active at a time while the others are waiting. Multithreading is better suited for the MPCA system because all the readout ICs require relatively long data conversion intervals (of 15.6–146.9 ms) between measurements; increasing the computational power does not benefit the performance. In addition, multithreading is compatible with single-core CPUs that are commonly used in low-power single-core microprocessors (e.g., the RPi Model Zero [23]) that are favored for ultra-portable systems. Therefore, for this work, multithreading was chosen to provide concurrency.

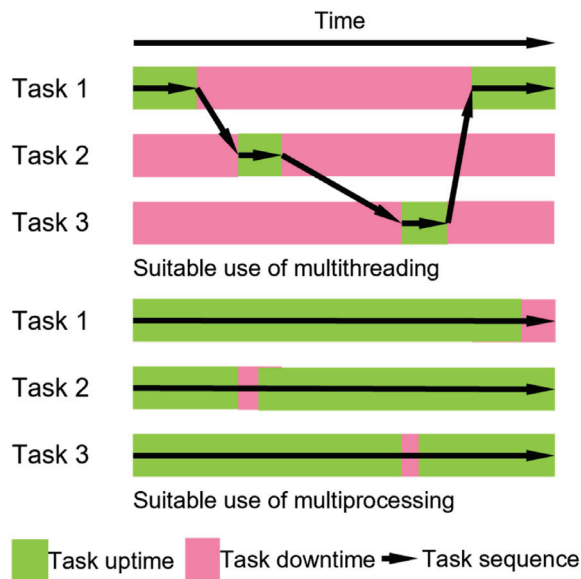


Figure 5. Temporal flow examples of multithreading and multiprocessing.

To align multithreading with the MPCA system hardware, the control software is divided into six threads: the main thread (Th0), the temperature control thread (Th1), the capacitive detector readout thread (Th2), the valve control thread (Th3), the pump control thread (Th4), and the AiPD readout thread (Th5). Each thread operates using its own dedicated library that is specifically designed to interface with and control the corresponding electronics. The main thread manages the other threads, handles communication with the UI, monitors overall progress, saves readout data, and provides centralized control to start or stop all the threads as needed (Figure 6).

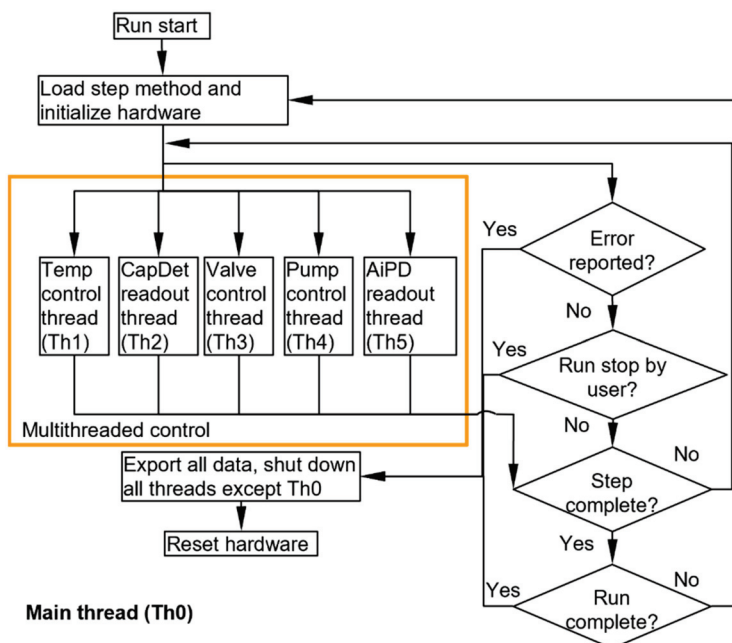


Figure 6. Flowchart of the main thread Th0.

2.3.2. Concurrent Threads Th1–Th5

The Th1 thread manages the closed-loop controls of all the eight temperature channels that are read using two ADC chips (Figure 7). In a control loop, it first initiates data conversion on one channel of each ADC chip and then enters sleep mode, awaiting the data

conversion while allowing other threads to progress. Once the prescribed data conversion interval has elapsed, the Th1 thread reads the output data. If a channel is designated to perform active heating, its temperature data are used to compute the parameter that controls the heating power (as described in Section 2.3.4). The process is repeated for the four channels of each ADC chip, and the total duration of the four-channel process is recorded as the current loop execution time. The Th1 thread then re-enters sleep mode for a duration equal to the difference between the predetermined loop cycle time (which is 100.0 ms) and the execution time (which is typically between 62.5 ms and 100.0 ms). The Th1 thread also monitors any anomalous temperature readings, which indicate system malfunction that may lead to hardware damage if left uncorrected. In this process, for each temperature channel, if anomalous temperature readings occur consecutively over an excessive count (here set at three consecutive readings), this thread terminates all the heating and temperature readout actions; it also alerts the Th0 thread, which then terminates all the threads.

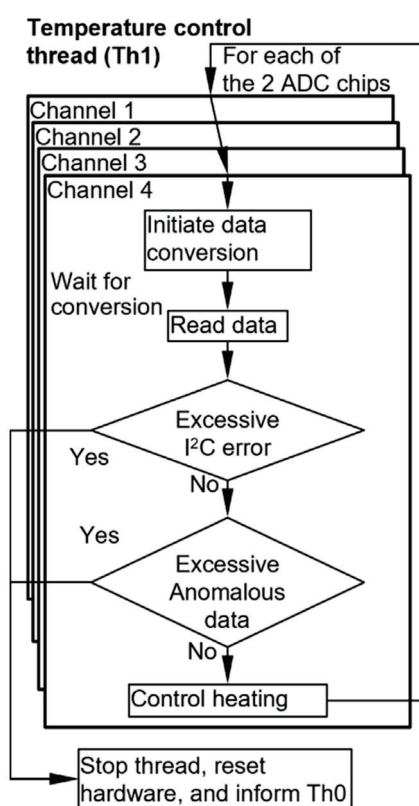


Figure 7. Flowchart of the temperature control thread.

The Th2 thread manages the readout of the two CapDet elements in each cell. The two CDC chips for CapDet readout are programmed to operate in the continuous data conversion mode, which automatically initiates a new data conversion immediately after the preceding one completes (Figure 8). After initiation, each CDC chip is read at a fixed interval, i.e., cycle time, of 110.0 ms, which is slightly longer than the conversion interval of 109.6 ms (Table 1).

The Th3 thread controls the six latching valves (Figure 9). The valves are actuated sequentially using voltage pulses of 50.0 ms duration. The cycle time of Th3 is set at one second in order to reduce the computational burden (Table 1).

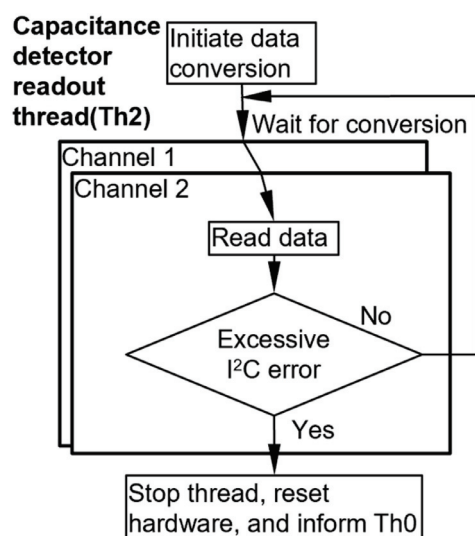


Figure 8. Flowchart of the capacitive detector readout thread.

Table 1. The IC model and conversion time used for each type of readout.

Readout Type	IC Model	OCCT (ms)	Time Multiplexed?	ECCT (ms)	PLCT (ms)
Temperature	ADS1115	15.6	Yes (4 channels)	62.4	100.0
CapDet	AD7746	109.6	No	109.6	110.0
Pressure head	LTC2493	146.9	Yes (2 channels)	293.8	400.0
AiPD	LTC2493	146.9	No	146.9	200.0

Abbreviations. OCCT: one channel conversion time; ECCT: effective channel conversion time (depending on whether multiple channels are time-multiplexed); PLCT: programmed loop cycle time.

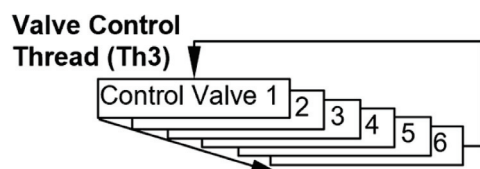


Figure 9. Flowchart of the valve control thread.

The Th4 thread controls the three pumps (Figure 10). For the sampling pump, the thread sends an on/off signal to start or stop the sampling pump. Then, for the separation pumps, it initiates data conversion for pressure head readout on one of the two ADC channels and then enters sleep mode. After the prescribed data conversion interval has elapsed, the Th4 thread reads the output pressure head from this channel and computes the actuation frequency required for its corresponding separation pump control (as described in Section 2.3.4). This process is then repeated for the other ADC channel. To accommodate the total required conversion time of both channels, which is 293.8 ms, the loop cycle time of the thread is set at 400.0 ms (Table 1). This is a much longer loop cycle time, leaving enough margin for any algorithmic processing that may potentially be required.

The Th5 thread controls the AiPD power supply and readout of the AiPD (Figure 11). In its loop, it first sends an on/off signal to the VUV lamp. Next, it initiates data conversion on the ADC and then enters sleep mode. After the prescribed data conversion interval, the Th5 thread reads the output data. To accommodate the data conversion time of 146.9 ms, the cycle time of the thread is set at 200.0 ms (Table 1). Although this cycle time can be reduced, it is set at 200 ms for consistency with the Th4 thread, where the same ADC model is used but with two time-multiplexed input channels.

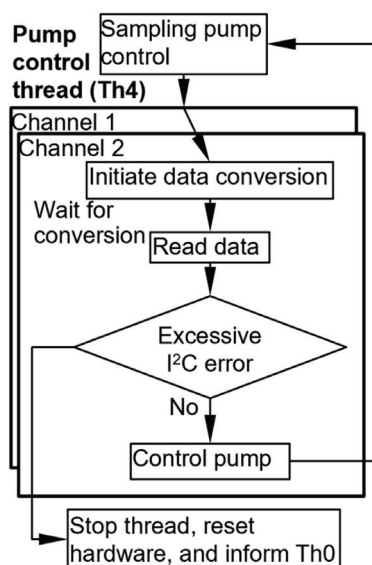


Figure 10. Flowchart of the pump control thread.

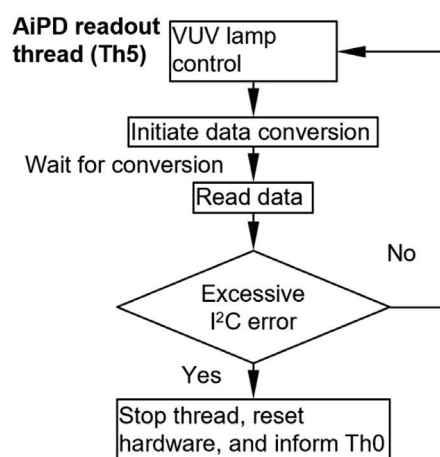


Figure 11. Flowchart of the AiPD output voltage readout thread.

Within each run step in the operation method, the Th2–Th5 threads are programmed to be active only when their specific functions are required. Otherwise, these threads enter sleep mode. For example, during the sampling step, where users typically do not need detector readouts, both the Th2 and Th5 threads enter sleep mode. When in sleep mode, a thread performs no action other than checking periodically—once per cycle time—to determine whether it needs to resume activity. Sleep mode conserves computational power and energy expenditure.

2.3.3. Thread Timing

The Raspberry Pi OS, which is the main operating system on RPi, is not a real-time operating system (RTOS). Consequently, programmed events are not always executed at precisely scheduled time points [24]. The timing errors can accumulate and consequently affect the accuracy of the chromatograms. To address this, two solutions are implemented. Firstly, the loop cycle time in each thread is set to be sufficiently longer than the time required by the hardware (Table 1), as described above. Secondly, all readout data are timestamped relative to the start of each run step. This allows back-end chromatogram processing (although not performed for this work) to interpolate the data, thereby compensating for any accumulated timing error, obtaining time-corrected chromatograms.

2.3.4. Control of Temperature and Flow

In this work, all the temperature and pressure head channels are closed-loop controlled to follow user-defined transient temperature and pressure head profiles. After reading out the temperature or pressure head data, the corresponding output control is computed using a proportional–integral–derivative (PID) algorithm. For a parameter x under control, a unified transfer function of the control is as follows:

$$C = C_{\max}(P \cdot \Delta x_n + I \cdot \Delta x_{\Sigma} + D \cdot (\Delta x_n - \Delta x_{n-1})) \quad (1)$$

where C is the control signal strength; $C_{\max}$ is the maximum available signal strength; P , I , D are the proportional, integral, and derivative coefficients, respectively; Δx_n is the difference between the target value of x and the measured value of x , both at the current (i.e., most recent) measurement time, n . The measurement time is discretized in 0.1 s intervals; Δx_{Σ} is the accumulated difference between the target value of x and the measured value of x since the beginning of each run step; Δx_{n-1} is the difference between the target value of x and measured value of x , both at the preceding (i.e., second most recent) measurement time. In temperature control, x represents the temperature. Further, for the PWM-based control, C represents the output duty cycle, with a maximum available value (i.e., $C_{\max}$) at 100%. For the analog voltage control via a potentiometer, C represents the output resistance, whereas $C_{\max}$ represents the maximum available output resistance of the potentiometer. In pressure head control, which effectively controls the flow rate, C represents the actuation frequency for a separation pump, $C_{\max}$ represents the maximum available actuation frequency, and x represents the pressure head.

The square waves required for PWM and for pump actuation may be generated using either hardware PWM or software PWM to provide the timing pulses. Hardware PWM supports precise timing control over a large range of frequency (up to 30 MHz) [18]. In contrast, software PWM is less precise in timing, because its pulses are generated using a software clock. Additionally, it can only support a lower frequency range (up to 40 kHz) [18]. Despite the superiority of the hardware PWM, on the RPi, it is limited to only 4 GPIO pins, which are insufficient for all the channels in the MPCA system. Considering the relatively large thermal time constants in the heating of the MPCA components (typically >0.5 s), all the temperature control channels are set to use software PWM. This software PWM operates at a frequency of 50 Hz with an 8-bit duty cycle resolution. Conversely, for the two separation pumps that require 0–1000 Hz frequency control, hardware PWM is selected. It provides a square wave at 50% duty cycle with a frequency control resolution down to a millionth of the set frequency. In this work, both types of PWM are managed by the pigpio library.

2.3.5. Communication between the EC and the Electronics

As I²C bus is shared by the EC and multiple electronic components; the communication on this bus may be affected by power glitches, noise, and interference. However, I²C communications use a handshake protocol, so the Smbus used in this work can detect a failed handshake and report an error. If left unaddressed, accumulated I²C errors can prematurely end a thread without implementing proper hardware controls. This is especially risky during component heating, which may eventually damage the hardware irreversibly. Nevertheless, tolerating a limited number of I²C errors can sometimes be beneficial, especially at early stages of hardware development. Therefore, in the threads that use I²C, namely Th1, Th2, Th4, and Th5, an error-handling protocol is implemented as follows. Each thread continuously monitors its cumulative I²C errors. If the accumulated number of I²C errors in a thread exceeds a pre-determined threshold (here set at 3), the I²C communication is deemed unreliable. As a result, this thread terminates its associated hardware operation and alerts the main thread, which then shuts down all the other threads and saves the data.

2.3.6. Data Saving and Storage

The control software is designed to automatically export all raw data, including timestamps and values of the readout data points, into a comma-separated value (CSV) file at the end of each run step. CSV files are compatible with a wide range of commonly used software, such as Microsoft Excel, MATLAB, and Python. After an analytical run, the raw data files that are generated can be readily used for automated peak detection [16] and subsequently for automated chemical detection [15]. Additionally, after each run, the control software automatically saves as a metadata file, the operation method file that contains the entire set of operation parameters used for the run (which is further described in Section 2.3.8). This file also serves as a record of the exact operating parameters used for the run, which is important for systems where the operating parameters are frequently adjusted by users. For clear and convenient record keeping, all data and metadata files of a particular run are saved in a dedicated folder that is uniquely named with the hardware serial number and the date and time of the run.

2.3.7. Low-Power Mode

For applications with constrained power sources, e.g., small-size batteries or solar panels, μ GC systems must operate within a limited power budget. Therefore, this work also investigates software approaches to reduce the power consumption of the MPCA system while maintaining essential hardware control capabilities. Table 2 shows various options for reducing power consumption by turning off certain RPi functions. Among these, the most significant power saving is achieved by turning off the Ethernet and USB functions. By implementing a low-power mode with all these options active, a total reduction in power consumption of 1.6–1.8 W is achievable.

Table 2. Normal-mode and low-power mode power consumption of the RPi (Model 3B+ in this work).

Functionalities	Normal Mode	Low Power Mode Savings
Bluetooth + Wi-Fi	On	Off, 0.2 W reduction
Soundboard	On	Off, 0.2 W reduction
Ethernet + USB	On	Off, 1.2 W reduction
HDMI	On	Off, 0.15 W reduction
Energy usage (idle state)	2.8–2.9 W	1.6–1.8 W reduction 1.1–1.2 W total consumption

In the low-power mode, the MPCA system must be able to perform the analytical runs without Ethernet communication to the UI. Upon starting, the control software executes a script to terminate Ethernet and USB functions, then operates the MPCA system based on the previously transferred operation method file. After the prescribed runs are complete, the control software executes another script to restore Ethernet and USB functions, allowing users to configure the system for the next series of runs. Turning off the other RPi functions in the low-power mode does not affect the system function in any way.

2.3.8. UI Design and Communications to the EC

The user interface (UI) of the MPCA system allows users to configure the operation method and to start and stop the system operation; it also provides a real-time display of the data. The UI operates in a Windows (Microsoft Corporation, Redmond, WA, USA) environment using the Windows Presentation Foundation (WPF) [25], which is a proven framework with strong open access library support. The UI is programmed in C# and Extensible Application Markup Language (xaml), which are the primary languages for WPF. In the UI, the main window contains plots for the temperature profiles, pressure head profiles, and the chromatograms. The plotting of real-time readout data (Figure 12) is implemented using LiveCharts [26]. To manage resource demands of plotting, each data stream is accompanied by a checkbox, which allows users to select whether to plot the data

stream. After each run step, the plots are automatically cleared and prepared for the next run step.

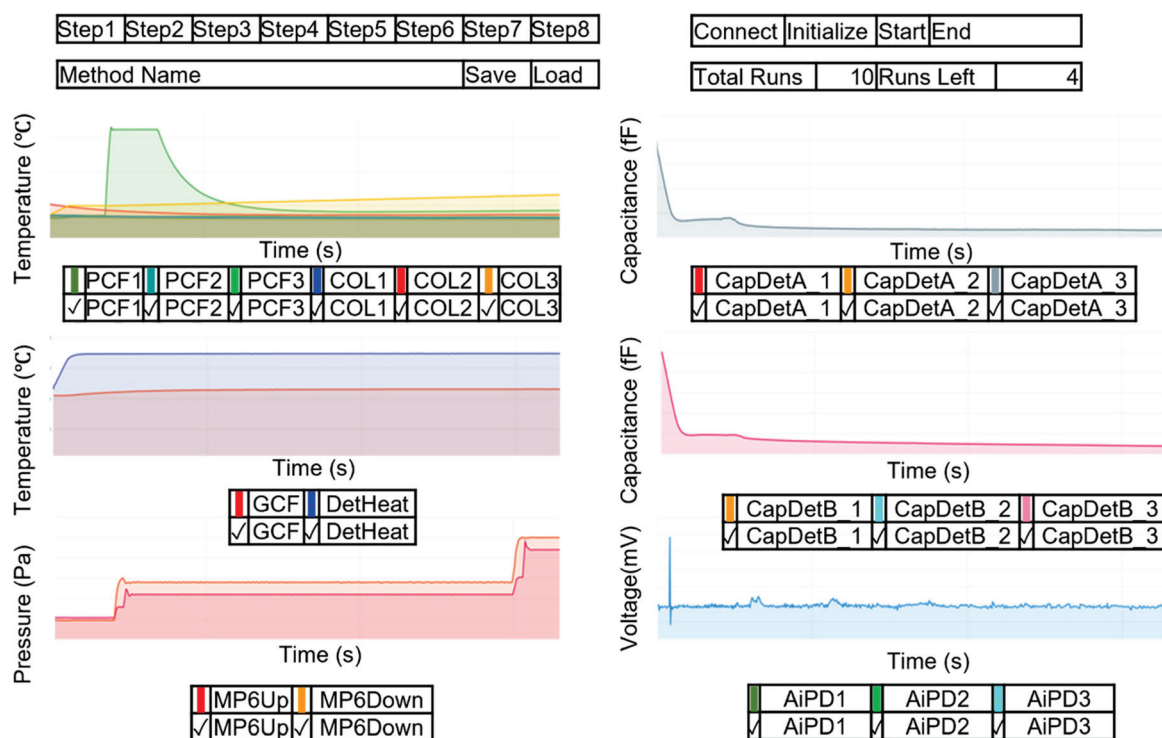


Figure 12. Graphical representation of the main UI window with real-time plotting of data.

The method configuration page allows the user to customize the operation parameters of each run step (Figure 13). This page opens when users click the corresponding button of a run step in the main UI window. The option to save the current method for future use is also included. More details about the UI are described in Supporting Information Section S2.

The operation parameters of all the run steps set up in the UI are arranged into an operation method file in the JSON format [27], which is both machine-readable and comprehensible to users. The widely used open-source Newtonsoft library is used to generate this file from the parameters entered into the UI. For the MPCA system, this file is typically larger than the TCP/IP receive buffer in the EC. Therefore, for transmitting it to the EC, this file is first segmented into smaller packets, which are then individually sent to the EC and finally reassembled into the full operation method file. The EC then reads the operation method file and prepares for initiating the MPCA system operation. During automated runs, the EC sequentially executes each run step based on the operation method file; in this process, the system operation is controlled by the EC and does not rely on the UI. The operation method file is also saved by the EC as the metadata for the corresponding MPCA system run.

In the normal operation mode, as opposed to low-power mode, real-time data transfer between the EC, and the UI is performed. This includes sending the readout data from the RPi to the UI and sending the user commands from the UI to the EC. For the readout data, the UI issues requests for the latest real-time data by sending a simple string to the EC. To avoid overburdening the UI, the frequency of these requests is programmed to vary inversely with the number of data streams already plotted. The EC then replies with the readout data in a string, which are parsed by the UI. The user commands include those to initialize the TCP/IP connection, to start the runs, and to stop the runs. These commands are sent to the EC in the form of a simple string immediately after being issued from the UI.

	Ramp Start Time (s)	Ramp End Time (s)	Heating End Time (s)	Initial Temp. (°C)	Ramp Target Temp. (°C)	Enabled
Preconcentrator1	0	0	0	0	0	False
Preconcentrator2	15	20	35	20	165	True
Preconcentrator3	0	0	0	0	0	False
Column1	0	0	0	0	0	False
Column2	20	398	500	30	70	True
Column3	0	0	0	0	0	False
Carrier Gas Filter	0	0	0	0	0	False
Detector Heater	1	5	500	20	40	True
Test Chip Heater	0	0	0	0	0	False

	Valve Open Time	Valve Close Time	Enable
Valve1	-1	-1	False
Valve2	1	498	True
Valve3	-1	-1	False
Valve4	-1	2	True
Valve5	-1	3	True
Valve6	-1	-1	False

	AiPD Start Time (s)	AiPD End Time (s)	Enable
AiPD1	0	0	False
AiPD2	0	500	True
AiPD3	0	0	False

	Lamp Start	Lamp End	Enable
AiPD Lamp	0	500	True

	Cap Start Time	Cap End Time	Enable
CapDetA_1	0	0	False
CapDetA_2	0	500	True
CapDetA_3	0	0	False
CapDetB_1	0	0	False
CapDetB_2	0	500	True
CapDetB_3	0	0	False

Knudsen Pump	Start time (s)	-1	-1	-1	-1	Enable	False
	End time	-1	-1	-1	-1	Closed Loop	
	Freq/Press	-1	-1	-1	-1		

MP6 pump Upstream	Start time (s)	20	260	-1	-1	Enable	True
	End time	-1	498	-1	-1	Closed Loop	True
	Freq/Press	500	1700	-1	-1		

MP6 pump Downstream	Start time (s)	20	260	-1	-1	Enable	True
	End time	260	498	-1	-1	Closed Loop	True
	Freq/Press	900	2000	-1	-1		

Total Time (s)	500
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Enable Step	True
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Figure 13. Graphical representation of UI for customizing the operation method. The entries in blue are representative values filled by a user for the Cell2 separation step.

3. Experimental Validation

The control software has been deployed in multiple generations and multiple units of the MPCA system [7] and has supported >1000 analytical runs, which have analyzed >40 different VOCs in various mixtures. Additional results of chemical analyses conducted using this control software have been previously reported [7]. The details of the system setup for experimental evaluation are described in Supporting Information Section S3. The performance of the control and readout can be assessed by the readout data from a typical run, as described below.

The efficacy of the closed-loop control was confirmed by verifying the measured temperature and pressure head profiles during the separation steps. For example, for the separation step performed in Cell2, Preconcentrator2 was set to ramp from 20 °C to 165 °C from 15 s to 20 s, to maintain a steady temperature at 165 °C from 20 to 35 s, and then to cool down naturally; Column2 was set to ramp from 20 °C to 70 °C from 20 s to 398 s and to maintain at 70 °C till 500 s; the detectors were set to heat up to 40 °C by 20 s and to maintain this temperature until 500 s. In a representative set of results, all these components adhered closely to their programmed temperature profiles (Figure 14a). Preconcentrator2 reached 93% of the target temperature at 20 s and remained within $\pm 1\%$ of the target temperature during the maintaining period. Column2 tightly followed its target temperature profile, showing an average deviation of only 0.38 °C. The region of the detectors also tightly followed its target temperature profile, showing an average deviation of only 0.02 °C.

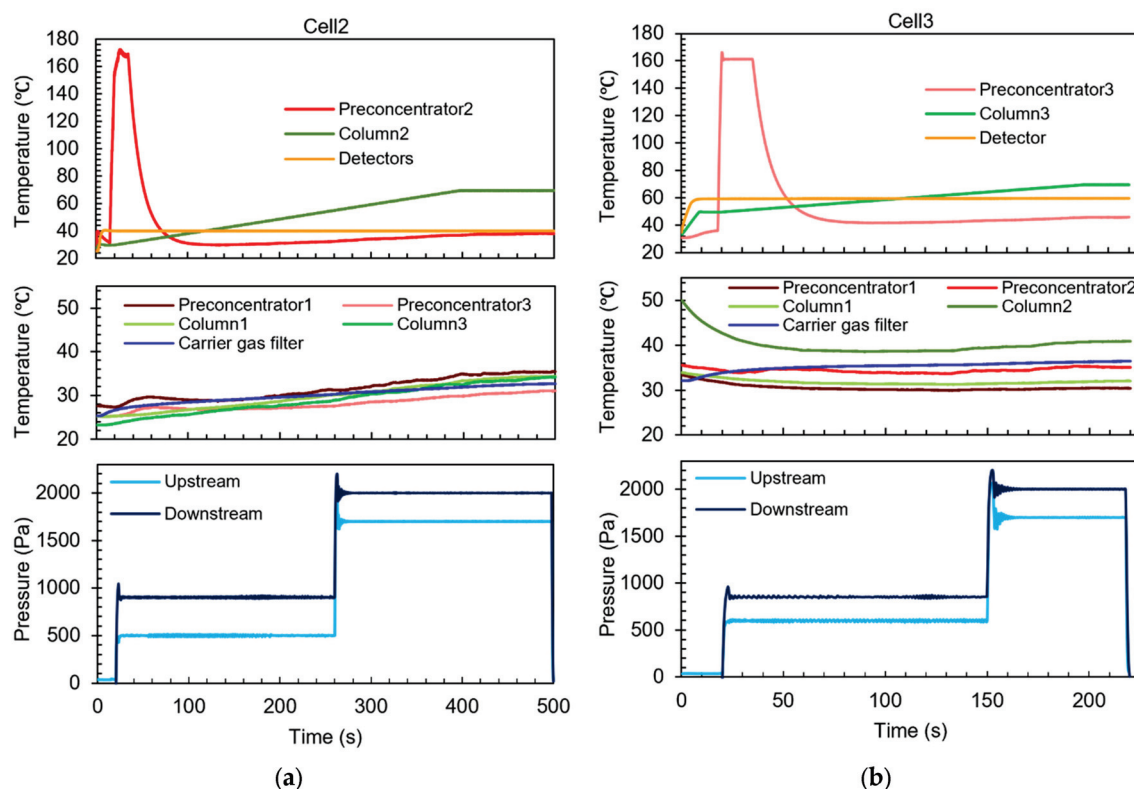


Figure 14. Typical temperature and flow control results in the MPCA system during (a) the Cell2 separation step and (b) the Cell3 separation step.

A parasitic temperature rise was observed in the unheated components, including Preconcentrator1, Preconcentrator3, Column1, Column3, and the carrier gas filter. This phenomenon was anticipated, because both the heated and unheated components were monolithically integrated in close proximity on the same MPCA chip, which had incomplete thermal isolation. Nevertheless, this parasitic temperature rise did not adversely affect the system operation.

For the separation step performed in MPCA Cell2, each flow control loop, both the upstream and downstream, was programmed for a two-stage flow rate pattern: an initial modest flow rate, followed by a stronger flow. The initial flow was used to provide an amenable flow rate for separation, whereas the stronger flow was used to expedite the elution of the low-volatility chemicals in the cell. For example, the upstream loop was set to provide a pressure head at 500 Pa during 20–260 s and 1700 Pa during 260–500 s. From the pressure readout, the upstream pressure head stabilized by 30 s into a range of 485–513 Pa, which was within $\pm 3.0\%$ of the target pressure. Following the scheduled increase at 260 s, the upstream pressure head stabilized by 275 s within a range of 1704–1696 Pa, i.e., within $\pm 0.3\%$ of the target pressure. Similar control performance was also obtained for the downstream loop (Figure 12a).

For the separation step in Cell3, the operation parameters were set up in a similar manner as those for Cell2, albeit with slightly different values. Both heating and pressure head controls in Cell3 provided similar performance as Cell2 (Figure 14b). The unheated components also showed unremarkable temperature responses. For example, Column2 started at an elevated temperature because it was heated in the preceding separation step in Cell2; as it was not actively heated during the current separation step in Cell3, it gradually cooled down during this step.

The effective control of the MPCA system was further validated by the chromatograms generated in analytical runs. In one such run, the MPCA system was tested with a mixture that included 18 detectable analyte chemicals, including cyclohexane, pentanal, heptane,

pinacolyl alcohol, methyl isobutyl ketone (MIBK), toluene, butyl acetate, ethylbenzene, m-xylene, o-xylene, 1-chloroheptane, nitrobenzene, mesitylene, decane, 2-nonanone, undecane, and dodecane. All the chemicals were prepared at a concentration of 200 parts per billion (ppb), except for o-xylene, which was present at an unspecified concentration. The sampling time was set by the user to be 10 min.

After the run, the exported data successfully provided an ensemble of chromatograms (Figure 15). The analyte chemicals presented various degrees of separation, such as pairs with complete baseline separation (e.g., between 1-chloroheptane and nitrobenzene), partial separation (e.g., between ethylbenzene and m-xylene), and complete overlap (e.g., between pinacolyl alcohol and MIBK). It was evident that the chromatograms obtained from the three different types of detectors were complementary in nature, even within each cell. For example, in Cell3, nonane produced relatively small positive CapDetA and AiPD peaks and a relatively large negative CapDetB peak, whereas ethylbenzene produced relatively small positive CapDetA peaks, almost no CapDetB peak, and a relatively large AiPD peak.

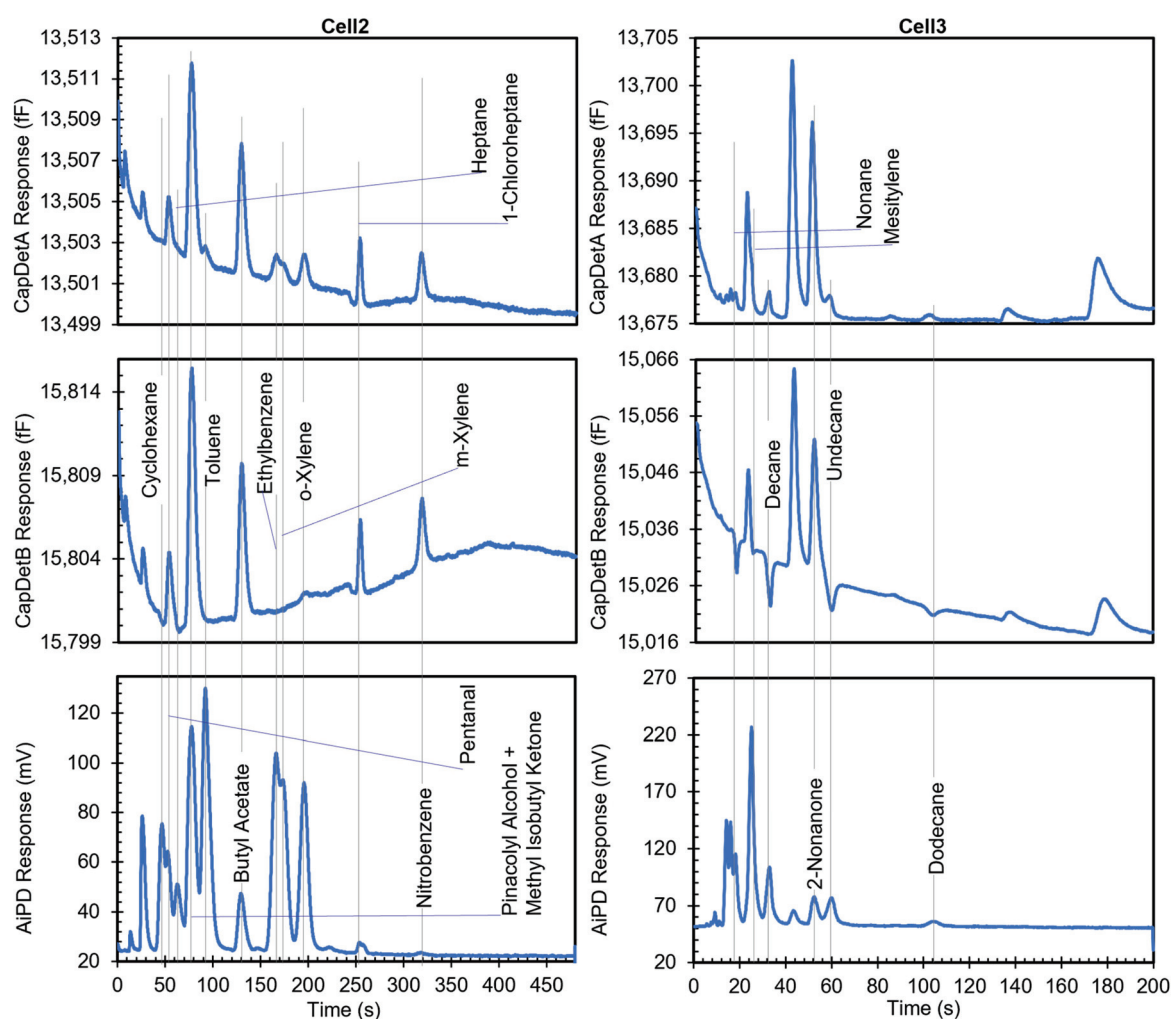


Figure 15. Chromatogram of a mixture containing cyclohexane, pentanal, heptane, pinacolyl alcohol, toluene, methyl isobutyl ketone, butyl acetate, ethylbenzene, m-xylene, o-xylene, 1-chloroheptane, nitrobenzene, nonane, mesitylene, decane, 2-nonanone, undecane and dodecane. Currently, the chromatograms are only presented for Cell2 and Cell3, as Cell1 is still under hardware development.

Most peaks showed peak widths > 10 s, which were much larger than the data acquisition rate of any detector. Therefore, each peak was well profiled by at least 50 data points. Even for those peaks that were partially separated within a narrow time window (e.g., the three peaks in Cell3 from 17 s to 20 s), their profiles were well represented by > 5

data points per peak. The available time resolution provided accurate and undistorted acquisition of the peak profiles, allowing recognition and quantification of the chemicals.

To examine the data acquisition rate more closely, readout data streams in the Cell2 separation step are used here as a case study. In this step, the data stream of each readout transducer type (i.e., temperature, CapDet, pressure head, or AiPD) contained >1200 data points. For each readout type, the actual time interval between measurements (TIBM) was obtained as the difference in the recorded timestamps between every two consecutive data points (Figure 16). Across all the readout types, the standard deviation of TIBM was <0.5% of the mean TIBM value, demonstrating timing stability (Table 3). This low variation in TIBM is desired, as it contributes to repeatable control of temperature and flow from run to run, which, in turn, contributes to repeatable system results. Additionally, the mean TIBM values for all readout types closely matched the intended cycle times, with <0.1% discrepancy in the temperature readout, <0.5% discrepancy in the CapDet readout, and <1.2% discrepancy in the pressure head and AiPD readouts. These discrepancies are further discussed in Section 4.

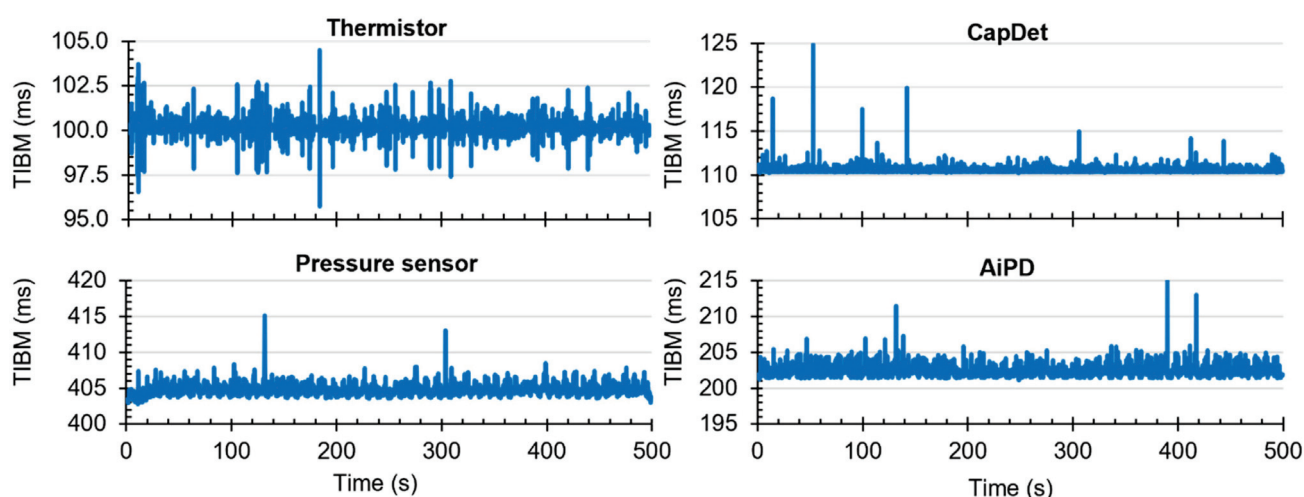


Figure 16. The typical experimentally measured time interval between measurement (TIBM) for each type of readout over time (measured during the MPCA Cell2 separation step). The number of data points in each plot is greater than 1200.

Table 3. Actual time interval between measurements (TIBM) of all the readout component types (measured at the peak software workload step, i.e., the Cell2 separation step).

Readout Type	Intended Loop Cycle (ms)	Actual Mean TIBM (ms)	TIBM Standard Deviation (ms)
Temperature	100.0	100.1	0.4
CapDet	110.0	110.5	0.4
Pressure head	400.0	404.7	1.0
AiPD	200.0	202.3	0.9

The noise level of each readout was also experimentally measured and compared to the rated noise in the converter datasheet (Table 4). The measured RMS noise values for the temperature, CapDet, pressure head, and AiPD readouts were 32.0 μ V, 53.0 aF, 66.9 μ V, and 107.1 μ V, respectively.

Table 4. Noise data of the detectors.

Readout Type	Noise (RMS)		Typical Noise Waveform (over 5 s Duration)	
	Measured	Datasheet		
Temperature	32.0 μV (0.04 $^{\circ}\text{C}$)	62.5 μV *		64.1 μV (0.85 $^{\circ}\text{C}$)
CapDet	53.0 aF	4.0 aF		219.8 aF
Pressure head	66.9 μV (0.15 Pa)	0.6 μV *		249.3 μV (0.55 Pa)
AiPD	107.1 μV	0.6 μV *		400.1 μV

* Per datasheet, this number is constant over a range of data rates and corresponds to the voltage for the least significant bit.

The measured noise of the temperature readout was comparable to that rated in the datasheet. According to the datasheet, the voltage resolution that corresponds to the least significant bit (LSB) of the ADC, under the settings in this work, is 62.5 μV , which aligns with the rated RMS noise. As shown by the measured noise waveform (Table 4), the measured voltage of the temperature readout typically toggled between two discretized values with an increment that corresponded to an LSB of the ADC. This observation not only confirmed the datasheet ratings but also indicated that the temperature readout resolution was limited by the LSB or the quantization error of the ADC.

In contrast, the CapDets measured much higher noise than that rated in the CDC datasheet. Similarly, the pressure head and AiPD readouts (both using the same ADC model) measured noise levels that significantly exceeded that rated in the datasheet (Table 4). These observations indicated that the resolutions of these readouts were not limited by the quantization errors of the ADC and CDC models. Instead, they appeared to be limited by the overall noise of the system, which possibly resulted from various sources, such as the noise in the reference voltage, noise in the power supply, interference, and the amplifier. Additionally, the elevated noise observed in the CDC readings was likely caused by the large baseline capacitances of the CapDets.

4. Discussion and Conclusions

Beyond the demonstrated functionalities, the control software may be improved in several ways. Firstly, ringing behaviors observed in the pressure head profiles likely resulted from the relatively long loop cycles (0.4 s) used for controlling the separation pumps. This problem can be mitigated by using an ADC model with a faster data rate.

Secondly, the recorded loop cycle times for pressure sensor and AiPD readouts were slower than the targets by 1% for all the data points. This was most likely because the control software assumed the computation times of these threads were negligible in the pressure sensor control thread (Th4) and the AiPD readout thread (Th5), whereas, in fact, the computation times of these threads were not negligible and caused some delay. In contrast, this issue was less pronounced in the temperature control thread, likely because of the implementation of a variable cycle wait time based on the measured computation time. In the capacitance readout thread Th2, the behavior of slower loop cycle time was also less significant, because the CDC chips operated in the continuous conversion mode, requiring only two I²C read operations per loop. For the pressure sensor control thread (Th4) and the AiPD readout thread (Th5), the approach of using variable cycle wait time can be implemented in the future.

In conclusion, the experimental results demonstrate that the control software reported in this work is successful in using multithreading to provide concurrent control of all

aspects of the μ GC system operation, including temperature control, flow control, capacitive detector readout, AiPD readout, and user interface. The efficacy of the control software is reflected by the generated temperature profiles, pressure head profiles, and chromatograms, all aligning with the user-defined operation method and expectations. Despite the use of a non-real-time operating system, the variations in TIBM are small and show no adverse impact on system control. The raw chromatogram data acquired can be readily processed, first using an automated peak detection algorithm [16] and then using an automated chemical recognition algorithm [15], culminating in a fully automated VOC analyzer.

With the demonstrated functionality for the MPCA system, the control software reported in this work shows great potential for applications in other μ GC systems. Its modular thread design can be easily expanded or adjusted to support different hardware configurations. In the future, real-time data transfer between the RPi and the UI can be implemented wirelessly, e.g., via Bluetooth or Wi-Fi. Additionally, the control software can be tested and assessed on a lower-power, single-core processor (e.g., the RPi Zero) to further reduce power consumption.

Supplementary Materials: The following supporting information can be downloaded at: <https://www.mdpi.com/article/10.3390/mi15010095/s1>, Section S1: Additional content on the CDC setup; Section S2: Additional contents on the graphical user interface; Section S3. Experimental Evaluation Setup.

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Conflicts of Interest: The authors declare no conflicts of interest.

References

1. Carle, G.C.; Donaldson, R.W.; Terry, S.C.; Wise, K.D. *Microminiature Gas Chromatograph*; NASA Tech Briefs, Ames Research Center: Moffett Field, CA, USA, 1972.
2. Terry, S.C.; Jerman, J.H.; Angell, J.B. A gas chromatographic air analyzer fabricated on a silicon wafer. *IEEE Trans. Electron Devices* **1979**, *26*, 1880–1886. [CrossRef]
3. Regmi, B.P.; Agah, M. Micro gas chromatography: An overview of critical components and their integration. *Anal. Chem.* **2018**, *90*, 13133–13150. [CrossRef] [PubMed]
4. Qin, Y.; Gianchandani, Y.B. A fully electronic microfabricated gas chromatograph with complementary capacitive detectors for indoor pollutants. *Microsyst. Nanoeng.* **2016**, *2*, 15049. [CrossRef] [PubMed]
5. Wang, J.; Nuño, N.; Nidetz, R.; Peterson, S.J.; Brookover, B.M.; Steinecker, W.H.; Zellers, E.T. Belt-mounted micro-gas-chromatograph prototype for determining personal exposures to volatile-organic-compound mixture components. *Anal. Chem.* **2019**, *91*, 4747–4754. [CrossRef] [PubMed]
6. Li, M.; Myers, E.B.; Tang, H.X.; Aldridge, S.J.; McCaig, H.C.; Whiting, J.J.; Simonson, R.J.; Lewis, N.S.; Roukes, M.L. Nano-electromechanical resonator arrays for ultrafast, gas-phase chromatographic chemical analysis. *Nano Lett.* **2010**, *10*, 3899–3903. [CrossRef] [PubMed]
7. Liao, W.; Winship, D.; Lara-Ibeas, I.; Zhao, X.; Xu, Q.; Lu, H.-T.; Qian, T.; Gordenker, R.; Qin, Y.; Gianchandani, Y.B. Highly Integrated μ GC Based on a Multisensing Progressive Cellular Architecture with a Valveless Sample Inlet. *Anal. Chem.* **2023**, *95*, 2157–2167. [CrossRef] [PubMed]

8. Collin, W.R.; Bondy, A.; Paul, D.; Kurabayashi, K.; Zellers, E.T. $\mu\text{GC} \times \mu\text{GC}$: Comprehensive Two-dimensional gas chromatographic separations with microfabricated components. *Anal. Chem.* **2015**, *87*, 1630–1637. [CrossRef] [PubMed]
9. Huang, X.; Li, M.W.-H.; Zang, W.; Huang, X.; Sivakumar, A.D.; Sharma, R.; Fan, X. Portable comprehensive two-dimensional micro-gas chromatography using an integrated flow-restricted pneumatic modulator. *Microsyst. Nanoeng.* **2022**, *8*, 115. [CrossRef] [PubMed]
10. Liao, W.; Zhao, X.; Lu, H.-T.; Byambadorj, T.; Qin, Y.; Gianchandani, Y.B. Progressive Cellular Architecture in Microscale Gas Chromatography for Broad Chemical Analyses. *Sensors* **2021**, *21*, 3089. [CrossRef] [PubMed]
11. Garg, A.; Akbar, M.; Vejerano, E.; Narayanan, S.; Nazhandali, L.; Marr, L.C.; Agah, M. Zebra GC: A mini gas chromatography system for Trace-level determination of Hazardous Air Pollutants. *Sens. Actuators B Chem.* **2015**, *212*, 145–154. [CrossRef]
12. McCartney, M.M.; Zrodnikov, Y.; Fung, A.G.; Levasseur, M.K.; Pedersen, J.M.; Zamuruyev, K.O.; Aksenov, A.A.; Kenyon, N.J.; Davis, C.E. An easy to manufacture micro gas preconcentrator for chemical sensing applications. *ACS Sens.* **2017**, *2*, 1167–1174. [CrossRef] [PubMed]
13. Patel, S.; Mlsna, T.; Fruhberger, B.; Klaassen, E.; Cemalovic, S.; Baselt, D. Chemicapacitive microsensors for volatile organic compound detection. *Sens. Actuators B Chem.* **2003**, *96*, 541–553. [CrossRef]
14. Li, M.W.; Ghosh, A.; Venkatasubramanian, A.; Sharma, R.; Huang, X.; Fan, X. High-sensitivity micro-gas chromatograph–photoionization detector for trace vapor detection. *ACS Sens.* **2021**, *6*, 2348–2355.
15. Xu, Q.; Qin, Y.; Gianchandani, Y.B. A Rule-Based Automated Chemical Recognition Algorithm for a Multi-Cell Multi-Detector Micro Gas Chromatograph. *Separations* **2023**, *10*, 555. [CrossRef]
16. Zhao, X.; Aridi, R.; Hume, J.; Subbiah, S.; Wu, X.; Chung, H.; Qin, Y.; Gianchandani, Y.B. Automatic peak detection algorithm based on continuous wavelet transform for complex chromatograms from multi-detector micro-scale gas chromatographs. *J. Chromatograms A* **2024**, *1714*, 464582. [CrossRef] [PubMed]
17. Zhou, M.; Lee, J.; Zhu, H.; Nidetz, R.; Kurabayashi, K.; Fan, X. A fully automated portable gas chromatography system for sensitive and rapid quantification of volatile organic compounds in water. *RSC Adv.* **2016**, *6*, 49416–49424. [CrossRef]
18. Raspberry Pi Foundation. Raspberry Pi Model 3B+. Available online: <https://static.raspberrypi.org/files/product-briefs/Raspberry-Pi-Model-B-plus-Product-Brief.pdf> (accessed on 20 July 2023).
19. Pigiop Library. The Pigiop Library. Available online: <http://abyz.me.uk/rpi/pigpio/> (accessed on 13 July 2023).
20. SMBus Protocol Summary. Available online: <https://kernel.org/doc/Documentation/i2c/smbus-protocol> (accessed on 6 September 2023).
21. Python Documentation. Multiprocessing—Process-Based Parallelism. Available online: <https://docs.python.org/3/library/multiprocessing.html?highlight=multiprocessing#module-multiprocessing> (accessed on 11 July 2023).
22. Python Documentation. Threading—Thread-Based Parallelism. Available online: <https://docs.python.org/3/library/threading.html> (accessed on 11 July 2023).
23. Lady Ada. Introducing the Raspberry Pi Zero. Available online: <https://cdn-learn.adafruit.com/downloads/pdf/introducing-the-raspberry-pi-zero.pdf> (accessed on 21 July 2023).
24. Hambarde, P.; Varma, R.; Jha, S. The survey of Real Time Operating System: RTOS. In Proceedings of the 2014 International Conference on Electronic Systems, Signal Processing and Computing Technologies, Nagpur, India, 9–11 January 2014.
25. Microsoft Learn. Windows Presentation Foundation for .NET 7 Documentation. Available online: <https://learn.microsoft.com/en-us/dotnet/desktop/wpf/?view=netdesktop-7.0> (accessed on 11 July 2023).
26. LiveCharts. LiveCharts Simple, Flexible, Powerful and Open Source Data Visualization for .Net. Available online: <https://v0.livcharts.com/> (accessed on 11 July 2023).
27. JSON. Introducing Json. Available online: <https://www.json.org/json-en.html> (accessed on 13 July 2023).

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Article

Structural Design and Optimization of the Milling Force Measurement Tool System Embedded with Thin-Film Strain Sensors

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Abstract: A milling force measurement tool system is designed with an elastic beam structure, which is divided into a two-end ring hoop compression sensor mode and a two-end square hoop compression sensor mode to improve the strain sensitivity. A simplified mechanical model of the elastic beam is established, and the relationship between the strain and force of the elastic beam under the action of three cutting force components is investigated, which can act as a guide for subsequent milling force measurement tool system calibration tests. Thin-film strain sensors occupy a central position in the milling force measurement tool system, which consists of a substrate, transition layer, insulating layer and resistance grid layer. The resistance grid layer has a particularly significant effect on the thin-film strain sensor's performance. In order to further improve the sensitivity of thin-film strain sensors, the shapes of the substrate, the transition layer, the insulating layer and the resistance grid layer are optimized and studied. A new thin-film strain sensor is designed with a resistance grid beam constructed from an insulating layer and a resistive grid layer double-end-supported on the transition layer. The flow of the wet-etching process of thin-film strain sensors is studied and samples are obtained. The surface microforms of the sensor samples are observed by extended depth-of-field microscopy, confocal microscopy and atomic force microscopy. It can be seen that the boundary of the resistance grid layer pattern is tidy and has high dimensional accuracy, thus enabling the basic achievement of the expected effect of the design. The electrical performance of the samples is tested on an experimental platform that we built, and the results show that the resistive sensitivity coefficient of the samples is increased by about 20%, to 51.2%, compared with that of the flat thin-film strain sensor, which fulfils the design's requirements.

Keywords: thin-film strain sensor; cutting force; wet etching process; performance characterization

1. Introduction

With the gradual development of MEMS, the advantages of intelligence, integration and miniaturization are gradually becoming clear, and it thus provides a good method for the measurement of cutting force. So far, developing a means of assessing the milling force that arises in traditional cutting remains important and difficult. In order to study the effect of milling force on the milling process, the measurement of milling force is very important. In order to measure milling force accurately, many scholars have undertaken much work. Adolfsson and Stahl [1] developed a milling force sensor and proposed a new way to measure the milling force; with the measurement equipment they developed, it is possible to directly measure the cutting forces acting on each individual cutting edge. Andreas Albrecht et al. [2] measured the milling force by measuring the displacement of a rotating spindle, and this could compensate for the effect of interference during milling force measurement. G. Byrne and G.E. O'Donnell [3] monitored the cutting forces that arise in the online machining process with two ring piezoelectric sensors. Ihsan Korkut [4] designed a data acquisition system and connected it to a strain dynamometer

to accurately measure cutting forces. In order to improve the accuracy of milling force measurement, the force-measuring element is studied. Süleyman Yıldız et al. [5] designed an octagonal ring milling dynamometer to measure cutting forces using strain gauges. Martin B. Jun et al. [6] used a ring force sensor to measure milling forces. G. Totis et al. [7] designed and manufactured a new type of rotary dynamometer for the online monitoring of the dynamic information related to cutting forces. Qiaokang Liang et al. [8] designed an E-type membrane force sensor and measured forces and moments on three axes. You Zhao et al. [9] measured cutting forces through two interleaved octagonal ring elastic elements. Shanshan Hu et al. [10] designed a floating beam-type six-axis elastomer, which offers guidance in the design of elastomer structure. In order to improve the sensitivity of the milling force-measuring element, strain sensors have also been studied. Xudong Cheng and Xiaochun Li [11] designed and fabricated a PdCr thin-film strain sensor, and embedded it in a metal structure to measure milling forces. Maximilian Mathis et al. [12] used a NiCr-C thin-film strain gauge to reduce the influence of strain deviation on sensor accuracy. Much research has been done on the preparation of thin-film strain sensors. Ruyuan Ma [13] studied the factors affecting the surface microstructure of TiN films, and explored the optimal process parameters. Vijay Petley et al. [14] studied the interfacial shear stress and stress distribution of NiCr films. Christine Taylor and Suresh K. Sitaraman [15] studied NiCr thin-film resistance grids of different sizes. Li Wang et al. [16] proposed a total suspension structure based on AlN. F. Samaeifar et al. [17] designed a platinum microhot plate with a suspended membrane structure. Madhu Santosh Ku Mutyala et al. [18] studied methods to improve the sensitivity of thin-film strain sensors in microcantilever beams. Muhammad Rizal et al. [19] proposed a four-claw structured force-measuring instrument, which achieved good strain, but showed room for optimization in terms of structural details and strain sensor performance.

A cutting tool system for measuring milling force based on thin-film strain sensors is proposed in this paper. Here, the resistive grid beam is supported on the double end of the thin-film strain sensor on the transition layer to increase strain and sensitivity.

2. Structural Design and Strain Principle of Milling Force Measurement Tool System

As shown in Figure 1, the milling force measurement tool system with an elastic beam structure consists of a shank, a link block, four elastic beams, a tool holder, a milling tool, and a double-end-supported thin-film strain sensor. Each elastic beam consists of a horizontal beam and vertical beam. The thin-film strain sensors are mounted on both the horizontal beam and the vertical beam, transferring the strain signal generated by the milling forces to the resistance grid beam.

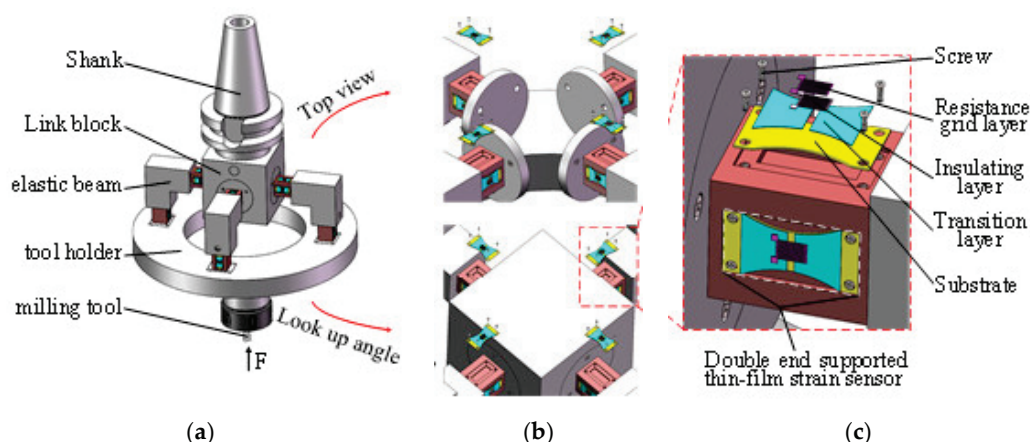


Figure 1. Milling force measurement tool system. (a) Tool system structure diagram; (b) elastic beam structure; (c) double-end-supported thin-film strain sensor.

2.1. Structural Design

Figure 2 shows a milling force measurement tool system in the two-end ring hoop compression sensor mode. The system has twenty four double-end-supported thin-film strain sensors, and a protective round sleeve is installed on the outermost part. The deformation of the double-end-supported thin-film strain sensor caused by the milling forces causes a change in the strain value in the resistance grid layer of the sensor. Figure 2d shows the sensor fixed on a two-end ring hoop.

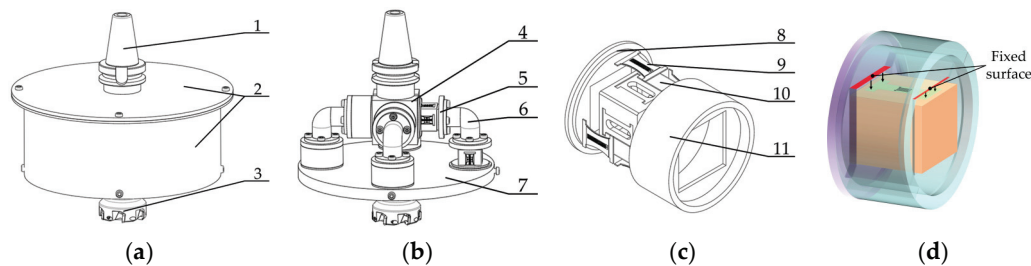


Figure 2. Milling force measurement tool system in the two-end ring hoop compression sensor mode. (a) Overall structure; (b) main structure; (c) two-end ring hoop compression sensor structure; (d) sensor in fixed mode. 1, shank; 2, protective round sleeve; 3, milling tool; 4, link block; 5, transfer disc; 6, connecting round beam; 7, tool holder; 8, annular hoop A; 9, double-end-supported thin-film strain sensor; 10, elastic square sleeve; 11, annular hoop B.

Figure 3 shows the milling force measurement tool system in two-end square hoop compression sensor mode. The structure of the tool system is similar to that shown in Figure 2, except that the sensor is fixed in a different way. As shown in Figure 3c, the sensor is fixed with square hoops at two ends.

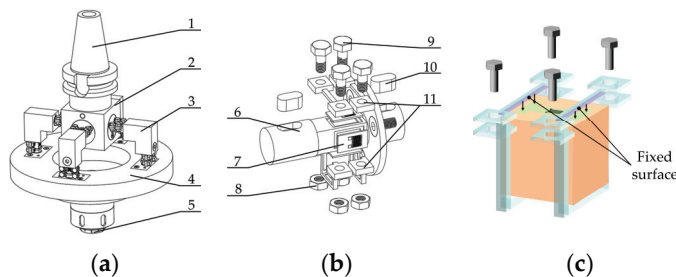


Figure 3. Milling force measurement tool system in the two-end square hoop compression sensor mode. (a) Overall structure; (b) two-end square hoop compression sensor structure; (c) diagram of sensor in fixed mode. 1, shank; 2, link block; 3, connecting square beam; 4, tool holder; 5, milling cutter chuck; 6, elastic horizontal beam; 7, double-end-supported thin-film strain sensor; 8, nut; 9, bolt; 10, key; 11, square hoop.

The milling force measurement tool system with a cross-beam structure consists of a shank, a force-measuring unit, and a milling cutter, as shown in Figure 4. The force-measuring unit has an internal cross-beam strain structure, similar to that shown in Figure 3.

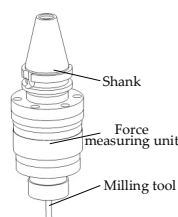


Figure 4. Milling force-measurement tool system for the cross-beam structure.

2.2. Strain Principle of Milling Force Measurement Tool System

As shown in Figure 5, the elastic beam is identical in shape to the cantilever beam. When a force is applied to the free end of the cantilever beam, a change results in the strain value on the sensor. The stress–strain relationship of the cantilever beam can be defined as

$$\varepsilon = \frac{\sigma}{E} = \frac{6FL}{Edh^2} \quad (1)$$

where ε is the strain, σ is the stress, E is the elastic modulus, F is the force on the free end of the cantilever beam, L is the moment length, d is the width of the cantilever beam, and h is the thickness of the cantilever beam.

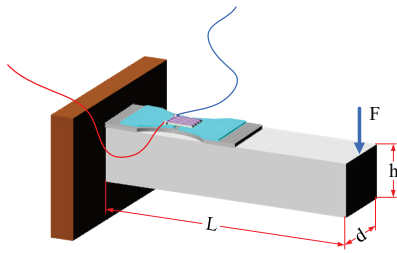


Figure 5. Schematic diagram of cantilever beam.

The link block, four elastic beams and tool holder (Figure 1) constitute the elastomer model that is subjected to three-directional cutting force, as shown in Figure 6. As the installation positions of the elastic beam and the sensor are symmetrical, a simplified model can be established by selecting one of the four elements of the elastic beam.

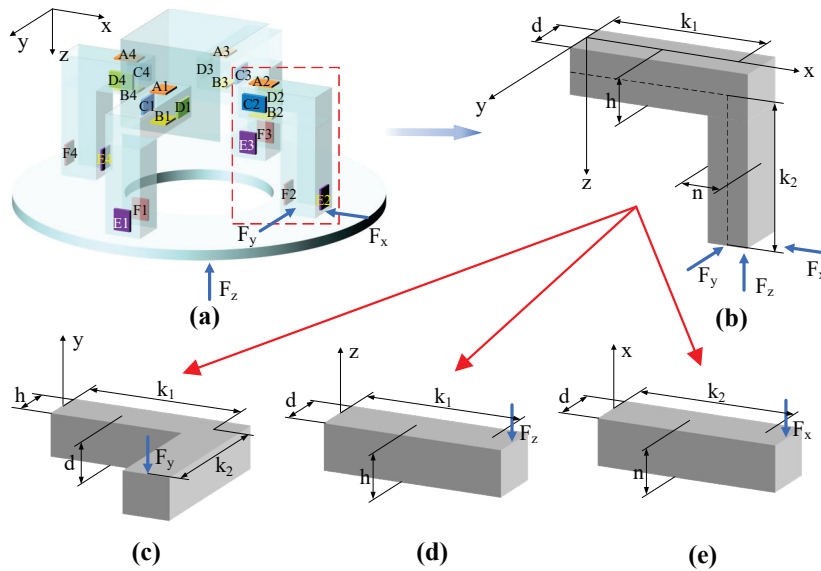


Figure 6. Elastomer model under three-directional cutting force. (a) Elastomer model and sensor distribution diagram; (b) schematic diagram of an elastic beam subjected to three-directional cutting force; (c) y-axis view; (d) z-axis view; (e) x-axis view.

The stress–strain relationship of one of the four elements of the elastic beam under the action of three-directional cutting force is here analyzed.

Moment equilibrium occurs on the neutral axis, so the arm of the moment gives the distance from the horizontal beam to the vertical beam, expressed as

$$L_k = (k_1^2 + k_2^2)^{\frac{1}{2}} \quad (2)$$

when the horizontal beam is subjected to a circumferential force, F_y , as shown in Figure 6c. By substituting Equation (2) into Equation (1), the relationship between the circumferential force F_y and the strain can be illustrated as

$$\varepsilon_{F_y} = \frac{3F_y}{2Ehd^2}(k_1^2 + k_2^2)^{\frac{1}{2}} \quad (3)$$

When the axial force F_z acts on the bottom of the horizontal beam, as shown in Figure 6d, the arm of moment is k_1 , and the relationship between F_z and the strain is

$$\varepsilon_{F_z} = \frac{3F_z k_1}{2Edh^2} \quad (4)$$

The corresponding relationship between the radial force F_x and the strain is

$$\varepsilon_{F_x} = \frac{3F_x k_2}{2Edn^2} \quad (5)$$

The strain gauge factor is a fundamental parameter reflecting strain sensitivity, and can be expressed as [20]

$$GF = \frac{\Delta R/R}{\Delta L/L} = \frac{\Delta R/R}{\varepsilon} \quad (6)$$

where R is the original resistance value of the resistance grid layer, L is the original length, and ε is the strain detected by the double-end-supported thin-film strain sensor.

The arrangement of thin-film strain sensors applied for detecting milling force is described in Figure 6a. The thin-film strain resistance sensors A1, A2, A3 and A4 are subjected to compressive stress, while the resistance sensors B1, B2, B3 and B4 are subjected to tensile stress, and these are all connected together to form a Wheatstone bridge circuit that can be used in detecting F_z , as shown in Figure 7a. The circumferential force F_y is detected by C1, C2, C3 and C4 when tensile stress occurs in these thin-film strain sensors, as shown in Figure 7b, while D1, D2, D3 and D4 are subjected to compressive stress. The radial force F_x is applied to the four beams with thin-film strain resistance sensors, as shown in Figure 7c. Tensile stress occurs on the thin-film strain resistance sensors E1, E2, E3 and E4, while thin-film strain resistance sensors F1, F2, F3 and F4 are subjected to compressive stress. The following formula is used to obtain the output voltage

$$\frac{U_o}{U_i} = GF\varepsilon \quad (7)$$

where U_i is the input voltage, U_o is the output voltage, GF is the strain gauge factor, and ε is the strain.

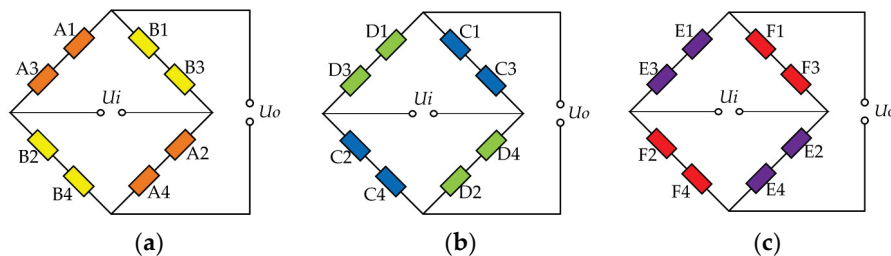


Figure 7. Wheatstone bridge connection diagram. (a) Wheatstone bridge for axial force F_z ; (b) Wheatstone bridge for circumferential force F_y ; (c) Wheatstone bridge for radial force F_x .

When the double-end-supported thin-film strain sensors are set up in a linear system, the relationship can be described as:

$$U = [\varepsilon]F \quad (8)$$

$$\begin{bmatrix} U_{F_y} \\ U_{F_z} \\ U_{F_x} \end{bmatrix} = \begin{bmatrix} \varepsilon_{11} & \varepsilon_{12} & \varepsilon_{13} \\ \varepsilon_{21} & \varepsilon_{22} & \varepsilon_{23} \\ \varepsilon_{31} & \varepsilon_{32} & \varepsilon_{33} \end{bmatrix} \begin{bmatrix} F_y \\ F_z \\ F_x \end{bmatrix} \quad (9)$$

where F is the vector of the input component force value $[F_y \ F_z \ F_x]$, $[\varepsilon]$ is the strain compliance matrix, and U is the vector of the output voltage value. Using the calibration platform, the output voltage generated by the double-end-supported thin-film strain sensor under three-directional cutting force loading can be measured, and when this process is combined with the above theoretical deduction, the milling force measurement tool system can be subjected to a calibration test.

3. Design and Preparation of Double-End-Supported Thin-Film Strain Sensor

3.1. Design of Double-End-Supported Thin-Film Strain Sensor

3.1.1. Structural Design of Double-End-Supported Thin-Film Strain Sensor

The double-end-supported thin-film strain sensor is composed of a substrate and thin-film system. The substrate's material and structure, and the thin-film system, are the main factors affecting the sensitivity of the thin-film strain sensor. The material of the elastic beam is 40Cr Steel (Shanghai Jiaoguan Industrial Co., Ltd., Shanghai, China), and the material of the substrate is stainless steel (Shanghai Baowei Industrial Co., Ltd., Shanghai, China).

A stepped-type substrate with good strain sensitivity has been designed, as shown in Figure 8a. But this structure has many right-angles, and stress concentrations occur at right-angles. The strain sensitivity values of the H-type substrate (Figure 8b) and the arc-type substrate (Figure 8c) are second only to that of the stepped-type substrate. The surface strain distribution of these two substrates was increased, but the degree of reduction in their stiffness was small. Therefore, combining these two substrates, an H-type substrate with a transitional arc has been designed, as shown in Figure 8d.

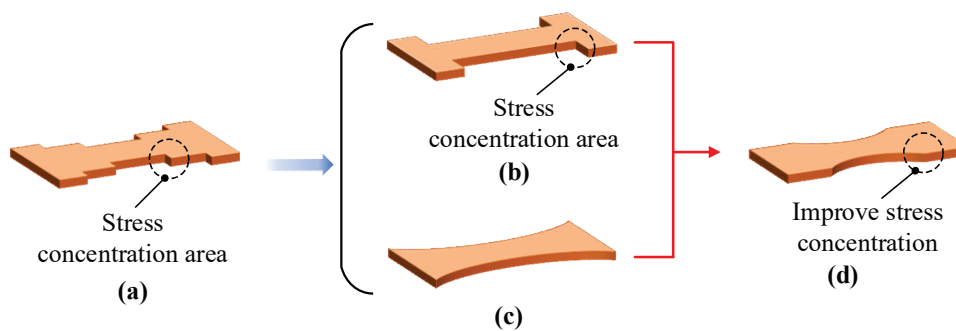


Figure 8. Optimization of the substrate. (a) Stepped-type substrate; (b) H-type substrate; (c) arc-type substrate; (d) arc transition H-type substrate.

The thin-film system shown in Figure 9 is composed of a transition layer, an insulating layer and a resistance grid layer. Because the Si_3N_4 thin-film has a high dielectric constant and good density, it is suitable for use as an insulating material in microelectronics applications, so the material of the insulating layer is Si_3N_4 . Previous studies have shown that the capacity for binding between the insulating layer and the substrate was not good, so a transition layer was added between the substrate and the insulating layer to ensure the insulating layer could grow well on the substrate's surface. The film–substrate binding capacity, using a TiN thin-film and 304 stainless steel substrate, is better; the film-binding capacity is also good when using a TiN thin-film and a Si_3N_4 thin-film. Therefore, TiN was selected as the material of the transition layer thin-film. The material and structure of the resistance grid layer seriously affect the performance of the sensor. When using nichrome ($\text{Ni}_{80}\text{Cr}_{20}$) as the resistance grid layer, it can work normally at high temperatures below 450°C .

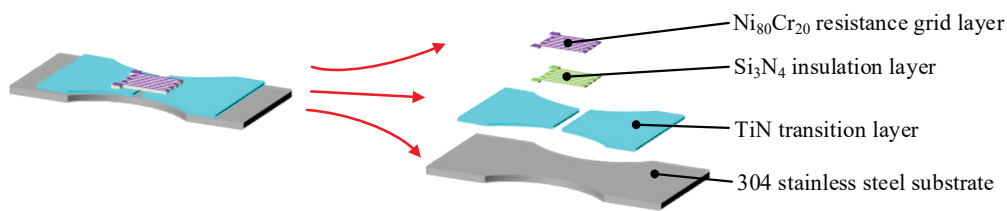


Figure 9. Double-end-supported thin-film strain sensor.

In order to improve the sensitivity of the thin-film strain sensor, a double-end-supported thin-film strain sensor is designed with the structure of an ordinary thin-film strain sensor. As shown in Figure 9, a strain gap was established on the transition layer. The insulating layer and resistance grid layer have the same shape, while the two layers overlap to form a composite resistance grid. As shown in Figure 10, the resistance grid is composed of electrodes, long grids and short grids. As shown in Figure 11, the long grid of the composite resistance grid is arranged on the strain gap, and is perpendicular to the strain gap.

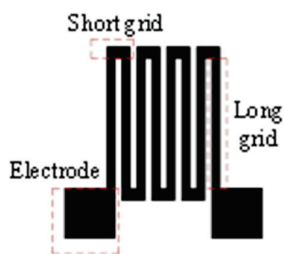


Figure 10. Resistance grid.

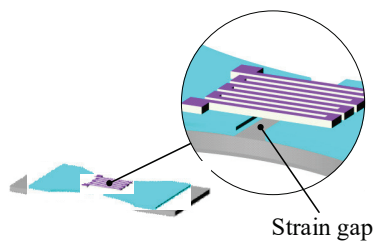


Figure 11. Double-end-supported structure.

3.1.2. Dimensional Design of Double-End-Supported Thin-Film Strain Sensor

The total length of the substrate of the double-end-supported thin-film strain sensor is 30 mm, and its total width is 16 mm. The sensing area is a rectangle of 16 mm × 8 mm, and the radius of the eight transition arcs is about 2 mm.

The long grid's size is 2.1 mm × 0.05 mm, the short grid's size is 0.2 mm × 0.1 mm, the thickness of the resistance grid layer is 1 μm, and the resistance grid comprises 10 long grids and nine short grids. The width of the strain gap is about 0.1 mm. The depth of the strain gap is about 500 nm.

3.2. Preparation Process of Double-End-Supported Thin-Film Strain Sensor

Figure 12 shows the preparation process of the double-end-supported thin-film strain sensor. The preparation process includes cleaning, magnetron sputtering deposition, photolithography, ion beam etching and wet etching. The pretreatment of the substrate, the removal of the photoresistor after ion beam etching and the removal of residual etching liquid after wet etching are all cleaning processes. The preparation of the transition layer, the sacrificial layer, the insulating layer and the sensitive layer comprise the magnetron sputtering coating process. Finally, the double-end-supported thin-film strain sensor is completed by cleaning it.

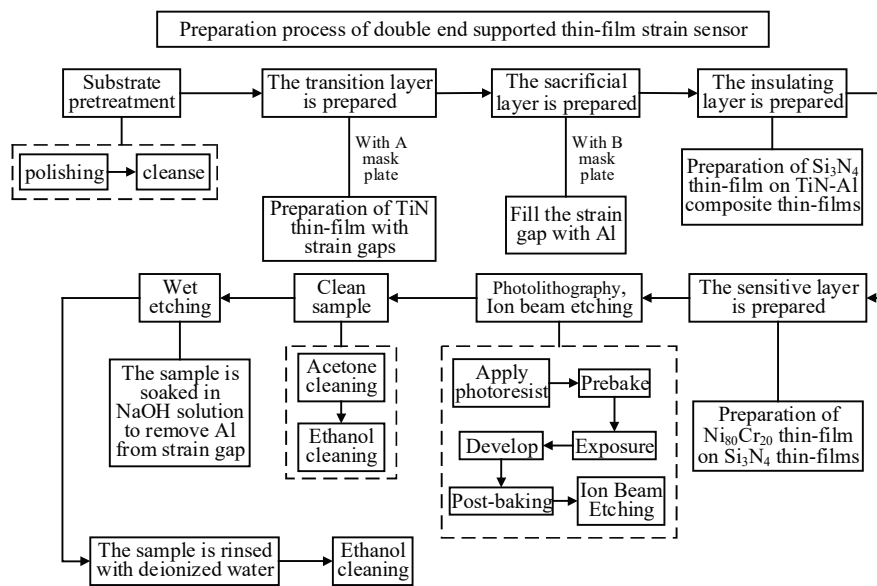


Figure 12. Flowchart of preparation process of double-end-supported thin-film strain sensor.

Figure 13 shows the pretreatment process applied to the substrate. Firstly, the substrate needs to be polished. See Table 1 for the detailed description of these steps.

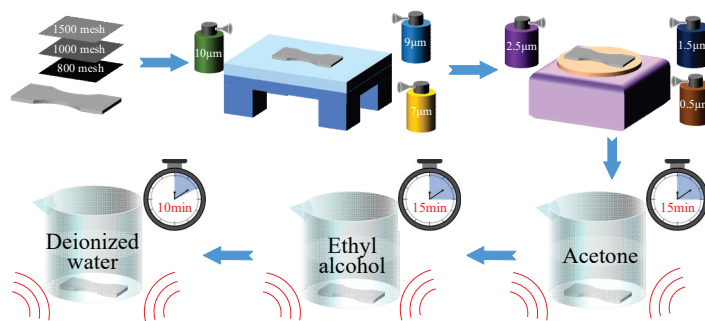


Figure 13. Pretreatment of substrate.

Table 1. Polishing procedure.

	Equipment	Polishing Material	Treatment Method	Time
Step One		Sandpaper (Suzhou Suboli Grinding Materials Co., Ltd., Suzhou, China)	800 mesh 1000 mesh 1500 mesh	Hand Grinding
Step Two	High-Strength Cast Iron HT200-300 Grinding Platform(Botou Bochuang Mechanical Equipment Manufacturing Co., Ltd., Cangzhou, China)	Diamond Spray Polish (Shenzhen Libaolixin Technology Co., Ltd., Shenzhen, China)	10 µm 9 µm 7 µm 5 µm 3 µm	Grind 30 min
Step Three	P-1 Type Metallographic Sample Polishing Machine (Shanghai Suoyan Testing Instrument Co., Ltd., Shanghai, China)	Diamond Spray Polish (Shenzhen Libaolixin Technology Co., Ltd., Shenzhen, China)	2.5 µm 1.5 µm 0.5 µm	Polishing 1 h

After finishing polishing, the surface roughness R_a of the substrate will be less than $0.2 \mu\text{m}$, thus meeting the requirements of the thin-film. Next, the substrate is cleaned to remove fine particles and any oil film that might remain on the surface of the substrate. The detailed steps of this procedure are shown in Table 2.

Table 2. Cleaning procedures.

	Equipment	Cleaning Solution	Time
Step One	Ultrasonic Cleaning Machine (Wenzhou Dongda Environmental Protection Equipment Co., Ltd., Wenzhou, China)	Acetone	15 min
Step Two		Anhydrous Ethanol	15 min
Step Three		Deionized Water	10 min

Figure 14 shows the magnetron sputtering deposition process. First, the transition layer was prepared, and the TiN thin-film with a strain gap was prepared by magnetron sputtering, using as a base the A-mask plate. The thickness of the TiN thin-film was about 500 nm. Due to the presence of the A-mask plate, blocking some of the TiN thin-film from being generated, a strain gap was formed, and the depth of the strain gap was about 500 nm. Next, the sacrificial layer is prepared. The A-mask plate of the previous step was removed and the B-mask plate was installed. The Al sacrificial layer was applied to fill in the strain gap using magnetron sputtering technology, and the thickness of this sacrificial layer was about 500 nm. Since the size of the gap of the B-mask plate corresponded to that of the strain gap, Al was deposited in the strain gap via the gap in the B-mask plate. Next, the insulating layer was prepared, the B-mask plate of the previous step was removed, and the Si_3N_4 thin-film was constructed on top of the TiN-Al composite thin-film by magnetron sputtering. The thickness of the Si_3N_4 thin-film was about 500 nm. Finally, the sensitive layer was prepared, and the $\text{Ni}_{80}\text{Cr}_{20}$ thin-film was generated on top of the Si_3N_4 thin-film by magnetron sputtering. The thickness of the $\text{Ni}_{80}\text{Cr}_{20}$ thin-film was $1000 \text{ nm} \pm 50 \text{ nm}$. Table 3 shows the parameters of the sputtering process.

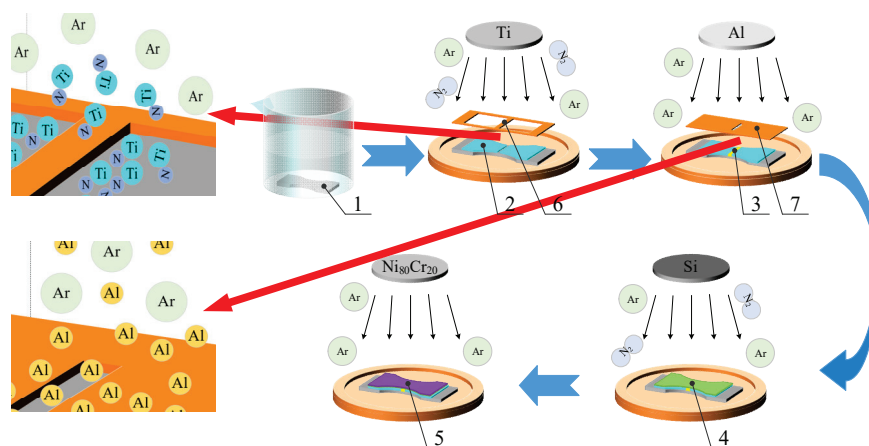


Figure 14. Magnetron sputtering deposition process. 1, substrate; 2, TiN thin-film; 3, Al sacrificial layer; 4, Si_3N_4 thin-film; 5, $\text{Ni}_{80}\text{Cr}_{20}$ thin-film; 6, A-mask plate; 7, B-mask plate.

Table 3. Sputtering process parameters.

	Deposition Power	Pressure	Argon Flow Rate	Nitrogen Flow Rate	Time
TiN	100 W	1 Pa	50 sccm	3 sccm	380 s
Al	100 W	1 Pa	50 sccm		2.5 h
Si_3N_4	140 W	1.2 Pa	70 sccm	10 sccm	2 h
$\text{Ni}_{80}\text{Cr}_{20}$	100 W	1 Pa	70 sccm		1000 s

As depicted in Figure 15, the photolithography process was carried out. An AZ4620 photoresistor was applied on the surface of the sample and homogenized using the homogenizing machine. After homogenizing, the thickness of the photoresistor on the surface of the sample was about 1.2 μm . The sample was prebaked and then exposed. The sample was immersed in the developer solution and studied with a microscope after development. After development had proceeded normally, the sample was post-baked and the photolithography process thus ended. Next, the sample was etched with an ion beam. The $\text{Ni}_{80}\text{Cr}_{20}$ thin-film and Si_3N_4 thin-film were etched onto the $\text{Ni}_{80}\text{Cr}_{20}\text{-Si}_3\text{N}_4$ composite resistance grid. After etching, the sample was cleaned to remove the residual photoresistor remaining on the surface. The details of the cleaning process are shown in Table 2. Finally, the wet etching process was carried out, and the Al in the strain gap was removed with a NaOH solution. After etching, the sample was put into the ultrasonic cleaning machine, and then cleaned with deionized water and ethanol. After drying, a double-end-supported thin-film strain sensor was obtained. Table 4 shows the parameters of the above process in detail.

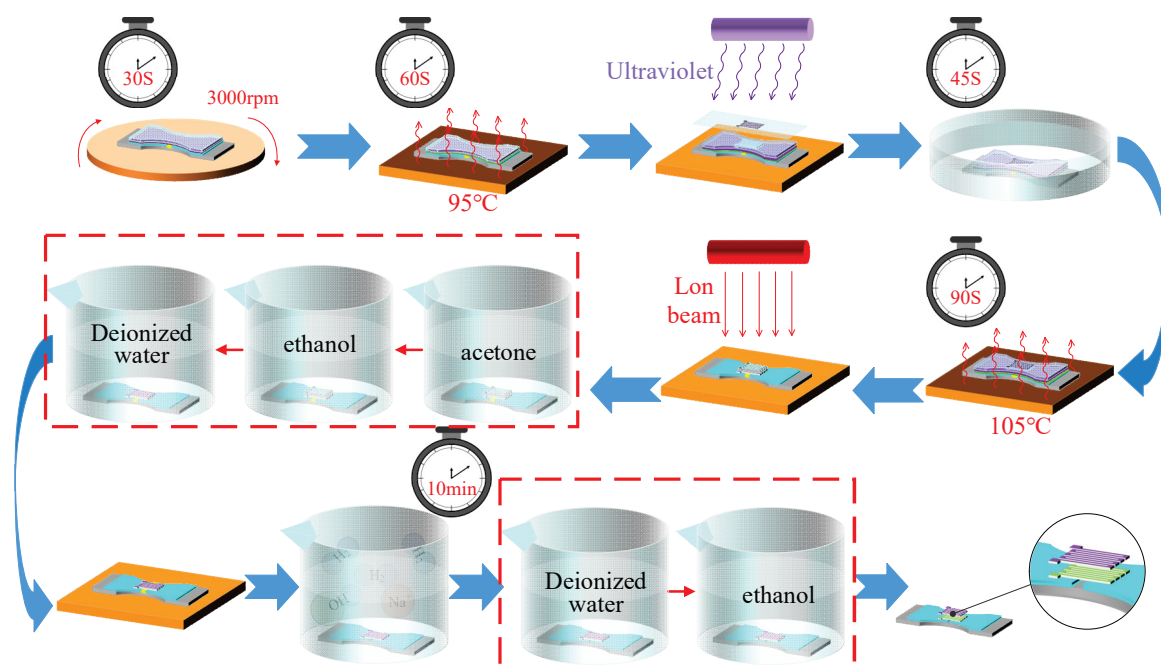


Figure 15. Photolithography, ion beam etching and wet etching processes.

Table 4. Detailed process parameters.

Photolithography				
Technology	Equipment		Conditions	
Apply Photoresistor (JSR Corporation)	Cee200XCB (Beijing Saidekesi Electronics Co., Ltd., Beijing, China)	AZ4620 Photoresist	3000 r/min	30 s
Prebake	EH-20B Anti-corrosion Electric Heating Plate (Shenzhen Sanli Technology Co., Ltd., Shenzhen, China)		95 °C	60 s
Exposure	EVG 610 (Beijing Yake Chenxu Technology Co., Ltd., Beijing, China)	Film Paper Mask Plate	400 mJ/cm ²	
Develop	Organic Hood (Shanxi Yitong Laboratory Equipment Co., Ltd., Taiyuan, China)	Developer Solution		45 s

Table 4. Cont.

Photolithography			
Technology	Equipment	Conditions	
Post-baking	EH-20B Anti-corrosion Electric Heating Plate (Shenzhen Sanli Technology Co., Ltd., Shenzhen, China)	105 °C	90 s
Ion Beam Etching			
Equipment	Time		
Ion Beam Etching Equipment (Beijing Midea Technology Co., Ltd., Beijing, China)	1 h 30 min		
Wet etching			
Etching Solution	Time		
3 mol/L NaOH Solution	10 min		

4. Characterization of Microstructure and Electrical Properties of Double-End-Supported Thin-Film Strain Sensor

4.1. Characterization of Microstructure of Double-End-Supported Thin-Film Strain Sensor

As shown in Figure 16, the resistance grid was observed using extended depth-of-field microscopy. The boundary of the resistance grid and the etching pattern were determined with high accuracy. The resistance grid, transition layer and strain gap on the double-end-supported thin-film strain sensor's surface could be clearly seen. As shown in Figure 17, a $128.437 \mu\text{m} \times 128.257 \mu\text{m}$ cavity region consisting of a resistance grid, a transition layer and a strain gap was observed using the Confocal Microscope. The maximum height of the thin-film system was $1.962 \mu\text{m}$, indicating that the distance from the bottom of the strain gap to the surface of the resistance grid was about $2 \mu\text{m}$, which is basically consistent with the thickness of the designed thin-film system.

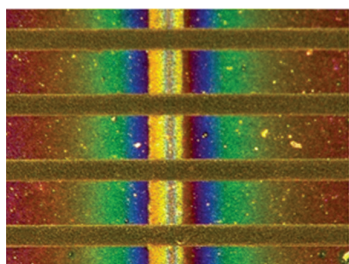


Figure 16. Microscopic morphology of double-end-supported thin-film strain sensor.

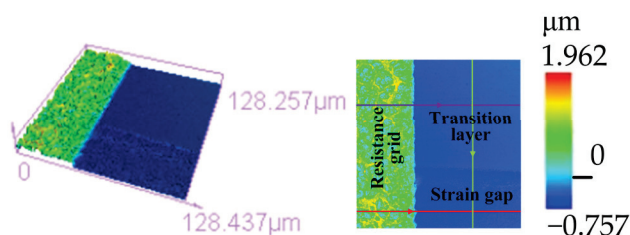


Figure 17. Height diagram of double-end-supported thin-film strain sensor.

In Figure 18, the scanned regions between the transition layer and the strain gap (Figure 18a), the resistance grid and the transition layer (Figure 18b), and the resistance

grid and the strain gap (Figure 18c) were all determined, and three sets of height waveform diagrams were obtained.

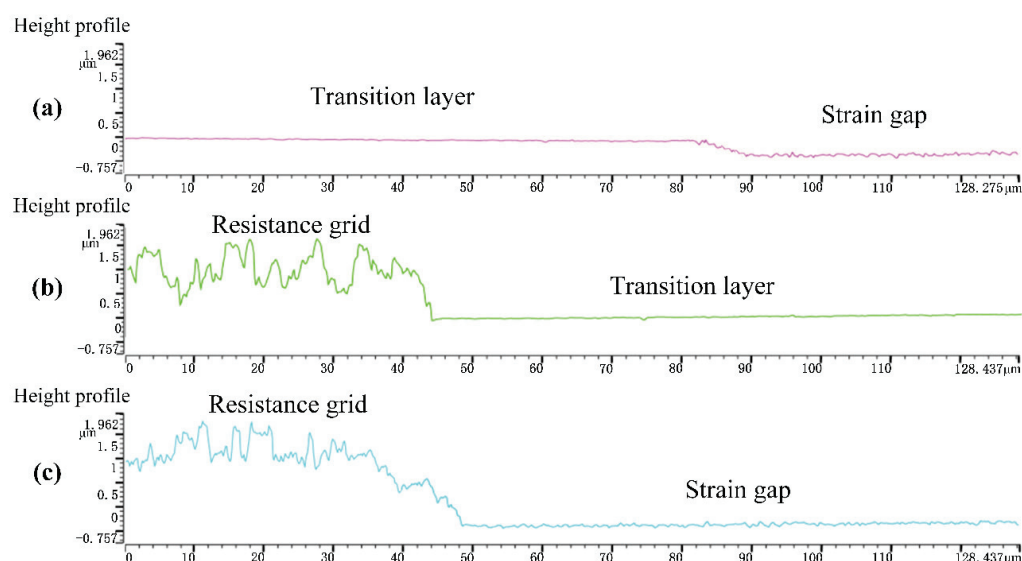


Figure 18. Height waveform diagram. (a) Transition layer to strain gap; (b) resistance grid to transition layer; (c) resistance grid to strain gap.

Figure 18a shows the height waveform diagram of the transition layer to the bottom of the strain gap, which was scanned in the direction shown by the green line in Figure 17. The height of the transition layer changed smoothly, indicating that the surface of the transition layer was relatively flat and the surface quality of the thin-film was good. When the scan reached about 82 μm , the resulting height waveform diagram began to decline, with a decline of about 450 nm, indicating that the depth of the strain gap was about 450 nm. At the same time, the height waveform diagram of the strain gap shows a small range with a sawtooth structure, indicating that a small amount of Al remained in the strain gap. The duration of wet etching can be extended to further reduce the residual Al without damaging the double-end-supported structure. Figure 18b shows a height waveform diagram between the resistance grid and the transition layer (in the direction shown by the purple line in Figure 17). The height waveform diagram of the resistance grid reflects a wide range, from 0.46 μm to 1.962 μm , with a roughly 1.5 μm difference in altitude between the resistance grid and the transition layer, indicating that when the surface roughness of the resistance grid is larger, the surface quality of the resistance grid will be affected. Photolithography and ion beam etching also affect the roughness of the resistance grid. Figure 18c shows that there is about 1.95 μm of altitude difference between the resistance grid and the bottom of the strain gap (along the direction shown by the red line in Figure 17).

Figure 19 shows the results of microscopic morphology performed using an Atomic Force Microscope. The two-dimensional diagrams in Figure 19a,c show that cracks of different degrees appeared on the resistance grid and transition layer. It is thus inferred that the thermal stress generated in the process of prebaking and post-baking may cause cracks to emerge on the surface of the thin-film. A comparison of the three-dimensional diagrams shown in Figure 19a,c indicates that the flatness of the resistance grid surface is worse than that of the transition layer. There are very densely positioned height differences shown in the three-dimensional diagram in Figure 19b, and the microscopic morphologies of the surfaces with different heights are different. When these findings are combined with the above analysis of the height waveform diagram, it can be inferred that the residual Al remaining in the strain gap creates a region with very dense height differences, as shown in the three-dimensional diagram. In Figure 19b, the relatively smooth ribbon-shaped area is the stainless-steel substrate.

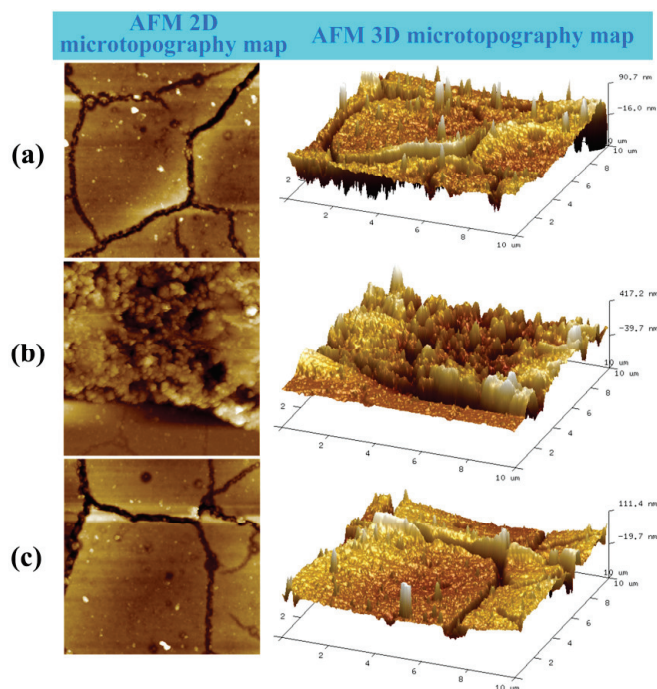


Figure 19. Micromorphology observation derived using AFM. (a) Resistance gate; (b) strain gap; (c) transition layer.

4.2. Electrical Performance Test of Double-End-Supported Thin-Film Strain Sensor

As shown in Figure 20, a test platform was built to determine the electrical performances of the ordinary thin-film strain sensor and the double-end-supported thin-film strain sensor, individually. Figure 20a shows a diagram of the Wheatstone bridge connection, wherein a resistance grid and three external resistors with the same resistance value are connected to form a quarter Wheatstone single bridge circuit; the input voltage here was 10 V, and the tension of 0–400 N was applied. As shown in Figure 20c, the two sensors are clamped horizontally and vertically, respectively. Eight groups of data were measured and subjected to fitting analysis, and finally, Figures 21 and 22 have been obtained.

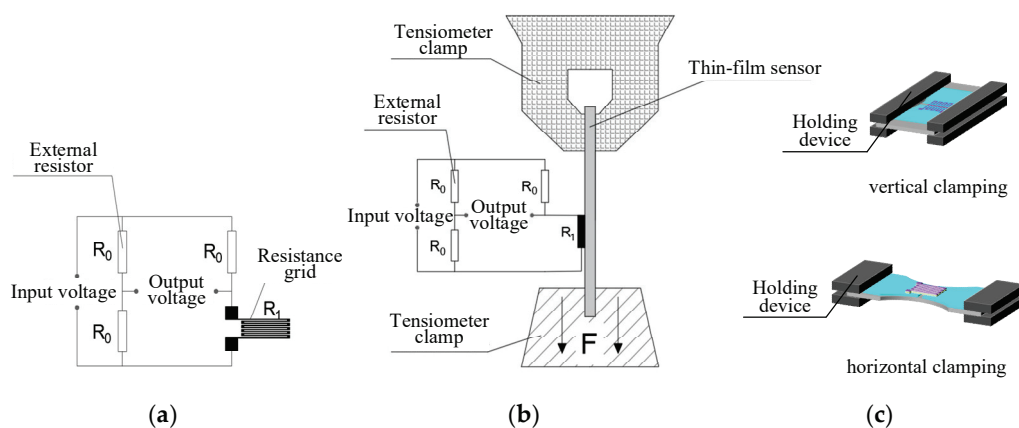


Figure 20. Electrical performance test. (a) Wheatstone bridge connection diagram; (b) test platform schematic diagram; (c) clamping mode.

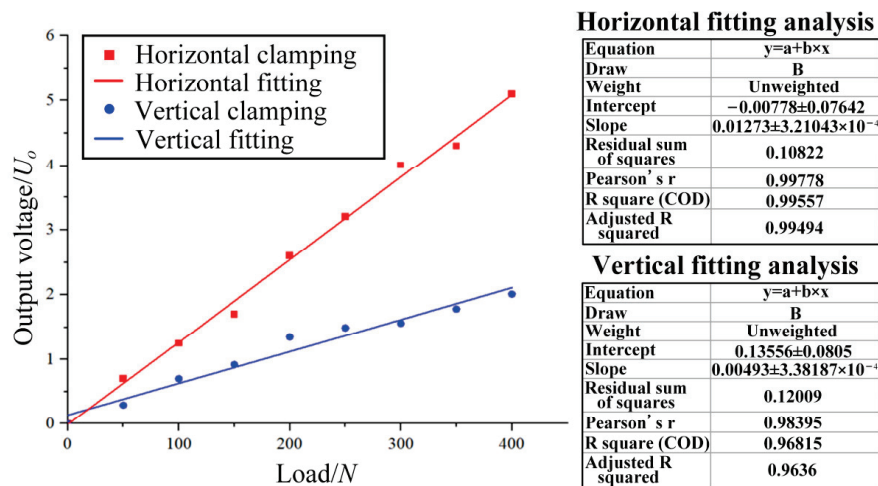


Figure 21. Relation between load and output voltage of double-end-supported thin-film strain sensor.

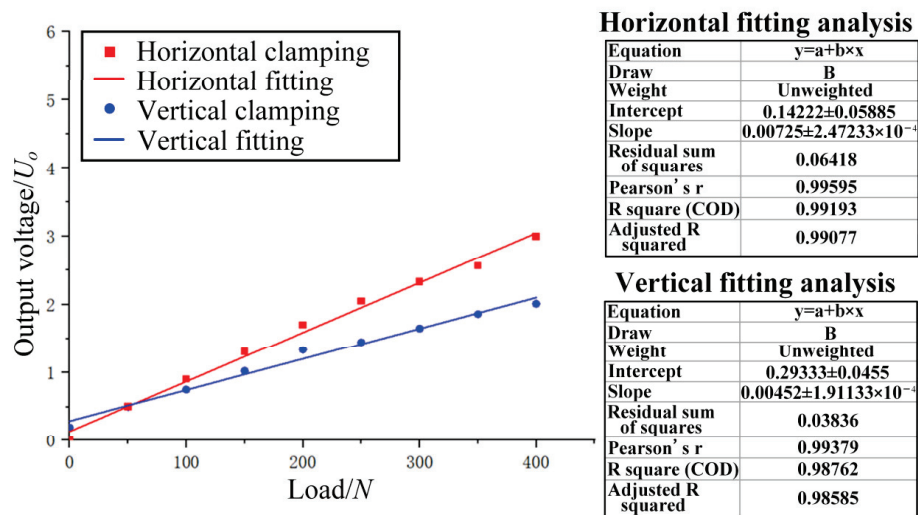


Figure 22. Relation between load and output voltage of ordinary thin-film strain sensor.

The fitting analysis in Figure 21 shows that the fitting value R^2 is close to 1, indicating that the linear error of the fitting result is small, and the double-end-supported thin-film strain sensor has good linearity. Under the same load, the voltage generated using horizontal clamping is larger than that generated using vertical clamping, meaning horizontal clamping produces greater strain.

As shown in Figure 22, the ordinary thin-film strain sensor achieved good linearity, but the slope of the linear fitting curve produced when it was subjected to horizontal clamping is significantly less notable than that of the double-end-supported thin-film strain sensor. By comparing the slopes of the linear fitting curves of the two kinds of sensors, we can see that the sensitivity of the double-end-supported thin-film strain sensor was about 51.2% greater compared to the ordinary thin-film strain sensor.

According to Figure 21 and Equation (7), it can be concluded that the GF value of the double-end-supported thin-film strain sensor was approximately 2.46. According to Figure 22 and Equation (7), it can be concluded that the GF value of the ordinary thin-film strain sensor was approximately 1.73. As regards the gauge factor (GF), the sensitivity of the double-end-supported thin-film strain sensor was about 42.2% greater than that of the ordinary thin-film strain sensor.

4.3. Experimental Results

A comparative experimental study of cutting was conducted between the milling force measurement tool system with a cross-beam structure (Figure 4) and the Kistler dynamometer, under the same experimental conditions, as shown in Figure 23.

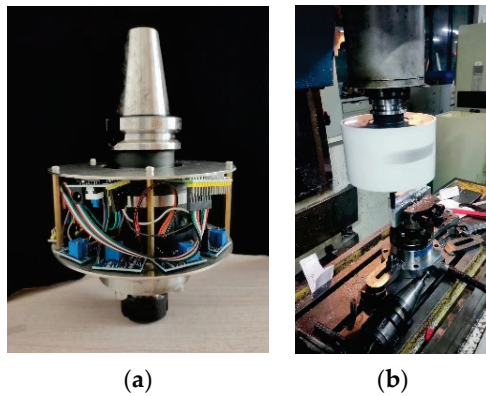


Figure 23. Cutting experiment applied to the milling force measurement tool system with a cross-beam structure. (a) Milling force measurement tool system with a cross-beam structure; (b) cutting test site diagram.

Experiment conditions: XK713 CNC milling machine (Wuhan Fourth Machine Tool Factory, Wuhan, China). Tools: $\varnothing 10$ mm carbide end mills. Workpiece material: 1060 aluminum alloy. The milling state was dry cutting, with a radial depth of the cut (a_e) of 0.2 mm, an axial depth of the cut (a_p) of 20 mm, a spindle speed of 1080 rpm and a feed speed of 80 mm/min.

In Figure 24, plots of the three-directional cutting force signal yielded over time by the milling force measurement tool system and the Kistler dynamometer are presented. From these, a table of the average values of the three-dimensional forces has been obtained, which can be used to compare between the force measuring system and the Kistler force measuring instrument, as shown in Table 5.

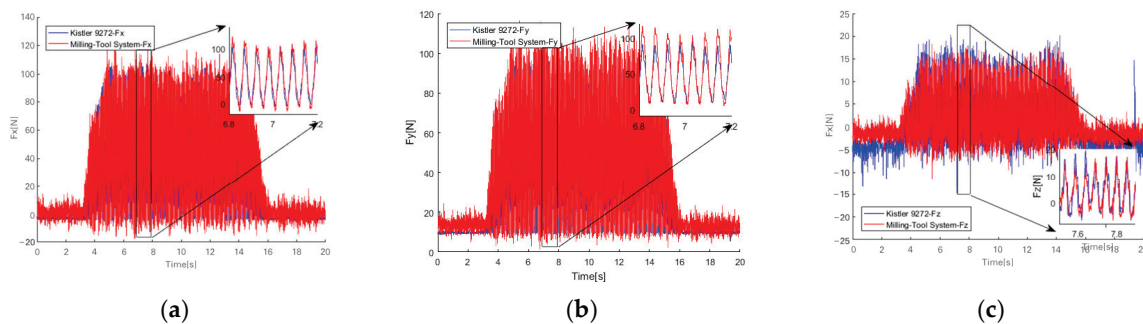


Figure 24. Plots of three-directional cutting force signals over time yielded by the milling force measurement tool system and the Kistler dynamometer. (a) Radial force F_x ; (b) circumferential force F_y ; (c) axial force F_z .

Table 5. Comparative analysis of the average values of the three-dimensional forces.

	F_x	F_y	F_z
Kistler dynamometer	106.85	108.5	20.32
Milling force measurement tool system	121.49	116.22	18.22
Deviation value	13.70%	7.12%	10.33%

As illustrated in Table 5, the experimental results verify the validity of the milling force measurement tool system with a cross-beam structure.

5. Conclusions

1. In this paper, three kinds of double-end-supported thin-film strain sensors and milling cutter elastomers with an integrated structure are designed. The two milling force measurement tool systems used different sensor fixation methods;
2. We designed a double-end-supported thin-film strain sensor and refined its preparation process. All parameters of the preparation process, including cleaning, magnetron sputtering deposition, photolithography, ion beam etching process and wet etching, have been determined. Finally, a double-end-supported thin-film strain sensor was prepared;
3. By means of extended depth-of-field microscopy, confocal microscopy and atomic force microscopy, the double-end-supported thin-film strain sensors were studied. The results show that the resistance grid's pattern boundary is regular, indicating higher etching accuracy. The transition layer and strain gap on the sensor's surface have been clearly illustrated. The double-end-supported thin-film strain sensor met the design requirements;
4. The electrical performance of the double-end-supported thin-film strain sensor was tested. The test results show that the sensitivity of the double-end-supported thin-film strain sensor was about 51.2% greater than that of the ordinary thin-film strain sensor. Regarding the gauge factor (GF), the sensitivity of the double-end-supported thin-film strain sensor was about 42.2% greater compared to those of the ordinary thin-film strain sensors;
5. We conducted cutting tests on the milling force measurement tool system with a cross-beam structure. This system showed deviations of 13.7% in average radial force F_x , 7.12% in average circumferential force F_y and 10.33% in average axial force F_z compared to the Kistler dynamometer. The experimental results verify the validity of this system.

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Data Availability Statement: The data used to support the findings of this study are available from the corresponding author upon request.

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References

1. Adolfsson, C.; Ståhl, J.-E. Cutting force model for multi-toothed cutting processes and force measuring equipment for face milling. *Int. J. Mach. Tools Manuf.* **1995**, *35*, 1715–1728. [CrossRef]
2. Albrecht, A.; Park, S.S.; Altintas, Y.; Pritschow, G. High frequency bandwidth cutting force measurement in milling using capacitance displacement sensors. *Int. J. Mach. Tools Manuf.* **2005**, *45*, 993–1008. [CrossRef]
3. Byrne, G.; O'donnell, G. An integrated force sensor solution for process monitoring of drilling operations. *CIRP Ann.* **2007**, *56*, 89–92. [CrossRef]
4. Korkut, I. A dynamometer design and its construction for milling operation. *Mater. Des.* **2003**, *24*, 631–637. [CrossRef]
5. Yıldız, S.; Ünsaçar, F.; Sağlam, H.; Işık, H. Design, development and testing of a four-component milling dynamometer for the measurement of cutting force and torque. *Mech. Syst. Signal Process.* **2007**, *21*, 1499–1511. [CrossRef]

6. Jun, M.B.; Ozdoganlar, O.B.; DeVor, R.E.; Kapoor, S.G.; Kirchheim, A.; Schaffner, G. Evaluation of a spindle-based force sensor for monitoring and fault diagnosis of machining operations. *Int. J. Mach. Tools Manuf.* **2002**, *42*, 741–751. [CrossRef]
7. Totis, G.; Wirtz, G.; Sortino, M.; Veselovac, D.; Kuljanic, E.; Klocke, F. Development of a dynamometer for measuring individual cutting edge forces in face milling. *Mech. Syst. Signal Process.* **2010**, *24*, 1844–1857. [CrossRef]
8. Liang, Q.; Zhang, D.; Song, Q.; Ge, Y.; Cao, H.; Ge, Y. Design and fabrication of a six-dimensional wrist force/torque sensor based on E-type membranes compared to cross beams. *Measurement* **2010**, *43*, 1702–1719. [CrossRef]
9. Zhao, Y.; Zhao, Y.; Ge, X. The development of a triaxial cutting force sensor based on a MEMS strain gauge. *Micromachines* **2018**, *9*, 30. [CrossRef]
10. Hu, S.; Wang, H.; Wang, Y.; Liu, Z. Design of a novel six-axis wrist force sensor. *Sensors* **2018**, *18*, 3120. [CrossRef]
11. Cheng, X.; Li, X. Development of metal embedded microsensors by diffusion bonding and testing in milling process. *J. Manuf. Sci. Eng.* **2008**, *130*, 061010. [CrossRef]
12. Mathis, M.; Vollberg, D.; Langosch, M.; Göttel, D.; Lellig, A.; Schultes, G. Creep adjustment of strain gauges based on granular NiCr-carbon thin films. *J. Sens. Sens. Syst.* **2021**, *10*, 53–61. [CrossRef]
13. Ma, R.; Wu, W.; He, Z.; Cheng, Y.; Liu, L.; Zhao, Y. Construction and Characterization of TiN/Si₃N₄ Composite Insulation Layer in TiN/Si₃N₄/Ni₈₀Cr₂₀ Thin Film Cutting Force Sensor. *Micromachines* **2021**, *12*, 1476. [CrossRef]
14. Petley, V.; Raman, K.T.; Kumar, S.; Jain, R.; Rao, G.M. Stress evolution mechanism of NiCr thin films on titanium alloy. *J. Alloys Compd.* **2017**, *695*, 2253–2260. [CrossRef]
15. Taylor, C.; Sitaraman, S.K. In-situ strain measurement with metallic thin film sensors. In Proceedings of the 2012 IEEE 62nd Electronic Components and Technology Conference, San Diego, CA, USA, 29 May–1 June 2012.
16. Wang, L.; Chiu, Y.; Gong, D.; Ma, S.; Yang, Y.; Li, H.; Lee, H.; Liu, H.; Jin, Y. Fabrication Process and Performance Analysis of AlN based Piezoelectric Micromachined Ultrasonic Transducer with a Suspended Structure. In Proceedings of the 2019 IEEE 14th International Conference on Nano/Micro Engineered and Molecular Systems (NEMS), Bangkok, Thailand, 11–14 April 2019; pp. 514–519.
17. Samaeifar, F.; Afifi, A.; Abdollahi, H. Simple fabrication and characterization of a platinum microhotplate based on suspended membrane structure. *Exp. Tech.* **2016**, *40*, 755–763. [CrossRef]
18. Mutyala, M.S.K.; Bandhanadham, D.; Pan, L.; Pendyala, V.R.; Ji, H.-F. Mechanical and electronic approaches to improve the sensitivity of microcantilever sensors. *Acta Mech. Sin.* **2009**, *25*, 1–12. [CrossRef]
19. Rizal, M.; Ghani, J.A.; Nuawi, M.Z.; Haron, C.H.C. Development and testing of an integrated rotating dynamometer on tool holder for milling process. *Mech. Syst. Signal Process.* **2015**, *52*, 559–576. [CrossRef]
20. Figliola, R.S.; Beasley, D.E. Theory and design for mechanical measurements. *Meas. Sci. Technol.* **2001**, *12*, 1743. [CrossRef]

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Article

Fabrication and Characterization of Photovoltaic Microgenerators Using the Complementary Metal Oxide Semiconductor Process

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Abstract: This study develops a photovoltaic microgenerator based on the complementary metal oxide semiconductor (CMOS) process. The photovoltaic microgenerator converts the absorbed light energy into electrical energy using the photovoltaic effect. The material for the photovoltaic microgenerator is silicon, and its structure consists of patterned p–n junctions. The design of the photovoltaic microgenerator utilizes a grid-like shape, forming a large-area p–n junction with a patterned p-doping and N-well structure to enhance the photocurrent and improve the device's performance. The photovoltaic microgenerator is fabricated employing the CMOS process with post-processing step. Post-processing is applied to enhance the microgenerator's light absorption and energy-conversion efficiency. This involves using wet etching with buffered-oxide etch (BOE) to remove the silicon dioxide layer above the p–n junctions, allowing direct illumination of the p–n junctions. The area of the photovoltaic microgenerator is 0.79 mm². The experimental results show that under an illumination intensity of 1000 W/m², the photovoltaic microgenerator exhibits an open-circuit voltage of 0.53 V, a short-circuit current of 233 μ A, a maximum output power of 99 μ W, a fill factor of 0.8, and an energy-conversion efficiency of 12.5%.

Keywords: photovoltaic microgenerator; energy-conversion efficiency; fill factor; MEMS; CMOS

1. Introduction

With the rapid advancement of technology, the human demand for energy has grown exponentially. Currently, the primary energy source is derived from the combustion of fossil fuels, leading to substantial emissions of greenhouse gases and severe environmental consequences. To alleviate this burden, the promotion of alternative green energy sources has become paramount. Green energy taps into the Earth's inexhaustible natural resources and is renewable, encompassing solar, hydro, wind, and geothermal power. Among these renewable sources, solar energy stands out for its unique capacity to transform sunlight into electricity, relying solely on light exposure for power generation [1–6]. Despite the array of available green energy sources, solar energy remains a prominent contender. However, most commercially available solar generators are characterized by their large size, rendering them unwieldy and lacking in mobility. Photovoltaic microgenerators improve these limitations and have many applications. For instance, in the work of Qiu et al. [7], nanocarbon tubes and solution processes were utilized to fabricate flexible, fiber-like perovskite microsolar generators. These were seamlessly integrated with textile technology to produce electronic textile products capable of powering garments autonomously. Similarly, Hyun et al. [8] integrated photovoltaic microgenerators with portable light-emitting diode (LED) lights, enabling daytime energy absorption and nighttime illumination. This microphotovoltaic system converts light energy into electricity, powering portable LED lights for both indoor and outdoor use.

Photovoltaic generators were created using microelectromechanical system (MEMS) technology, offering benefits such as size reduction, seamless integration with IC devices, and suitability for large-scale manufacturing. Several photovoltaic microgenerators have been developed using MEMS and CMOS technologies. For instance, Sakai et al. [9] fabricated organic solar cells utilizing heterojunction organic materials through screen printing. The materials used in the organic solar cell included the conjugated polymers as donors and the derivatives of fullerenes as acceptors. The results indicated that under AM1.5G solar irradiation, an open-circuit voltage of 0.821 V and an energy-conversion efficiency of 2.35% were achieved. However, this cost-effective screen-printing method often leads to uneven material distribution, resulting in variable performance among different solar cells and, ultimately, lower-than-anticipated efficiency. Arima et al. [10] employed the CMOS process to develop solar cells. They established two distinct p–n junctions by doping p-type and n-type silicon onto the solar chip, subsequently serially connecting these junctions. The results showed that an irradiance of 2000 lx led to an output voltage of 0.83 V and an energy-conversion efficiency of 2.6%. Marwick et al. [11] employed sapphire CMOS technology to manufacture photovoltaic generators. Sapphire, serving as an insulating substrate, was utilized to enhance the transmission of light from the rear. By serially connecting multiple sets of photodiodes, a high output voltage was attained. The materials used in the photodiodes were doped p-type and n-type silicon. The device revealed that an irradiance of 5000 lx resulted in a power output of 60 mW. Across the range of wavelengths from infrared to ultraviolet, the photovoltaic generator demonstrated an energy-conversion efficiency that was greater than 1%. Despite transitioning to transparent sapphire material from the traditional substrate, the energy-conversion efficiency for the device was not high. Law et al. [12] manufactured cascaded photodiodes using the CMOS process. The materials employed in the photodiodes were doped p-type and n-type silicon. The study employed second- and third-level cascading, yielding open-circuit voltages of 0.84 V and 1.3 V accompanied by short-circuit currents of 1.25 nA and 1.12 nA and conversion efficiencies of 0.3% and 0.06%, respectively, under an irradiance of 6.37 mW/cm². Although the study solely compared second and third-level cascading of photodiodes and suggested the superiority of third-level over second-level cascading, it failed to discuss the design of p–n junctions. Consequently, improvement in energy-conversion efficiency is still necessary. Horiguchi et al. [13] produced lightly doped triple-layer solar cells using CMOS technology. The interconnected triple-layer light doping augmented the output voltage and mitigated leakage current issues. Experiments demonstrated that under an irradiance of 31,000 lx, a substantial open-circuit voltage of 1.3 V, coupled with a short-circuit current of 6 µA and an energy-conversion efficiency of 7.5%, was achieved. Despite achieving a remarkable open-circuit voltage through series connection, the current enhancement was inadequate and, thus, still had a low energy-conversion efficiency. Hung et al. [14] fabricated interdigitated back-contact (IBC) solar cells using the CMOS process. By significantly reducing the thickness of the silicon substrate, the energy-conversion efficiency was elevated from 4% to 15%. Under conditions involving a wavelength of 980 nm and an irradiance of 10 mW/m², an open-circuit voltage of 0.57 V and an output power of 159 µW were obtained. While thinning the silicon substrate indeed improved performance substantially, the technical complexity and cost of this approach have limited its widespread adoption. Hung et al. [15] manufactured photovoltaic chips by selectively removing portions of the substrate using CMOS processes. That study managed to enhance the output voltage of back-illuminated photovoltaic chips by employing a substrate removal approach and stacking photovoltaic structures. Under an intensity of 40 mW, an open-circuit voltage of 2.05 V, a short-circuit current of 2.6 mA, and an energy-conversion efficiency of 11% were achieved. Hung et al. [16] developed photovoltaic microgenerators, segregating photovoltaic units on the chip to form photovoltaic modules. The materials used in the photovoltaic microgenerators were doped p-type and n-type silicon. Experimental results revealed an open-circuit voltage of 0.5 V for a single photovoltaic unit. The series connection of 25 units resulted in an open-circuit voltage of 12.5 V, a short-circuit current of 11 µA, and an energy-conversion efficiency of

0.5% under an irradiance of 6 mW/mm^2 . Despite attaining a substantial 12.5 V output voltage, the 0.5% energy-conversion efficiency was insufficient to enable effective self-power generation. Therefore, increasing the output current of the microgenerator became imperative. Yan et al. [17] used MEMS technology to fabricate thermoelectric–photovoltaic generators, combining thermoelectric microgenerators with photovoltaic counterparts. Photovoltaic elements adopted interdigitated electrode structures, incorporated beneath the thermoelectric structure. The materials of the photovoltaic microgenerators were doped p-type and n-type silicon. The experimental results demonstrated a fill factor of 0.792 and an energy-conversion efficiency of 4.45% when the photovoltaic component was illuminated with an irradiance of 100 mW/cm^2 . While the interdigitated electrode structure contributed to enhance photovoltaic performance, the stack configuration of the thermoelectric–photovoltaic generator led to performance interference. Sichao et al. [18] developed thermoelectric–photovoltaic microgenerators incorporating metal heat sinks through the MEMS process. This integration combined thermoelectric microgenerators with photovoltaic counterparts. The photovoltaic microgenerator employed a back-contact structure, allowing for autonomous selection of the illuminated surface. The materials utilized in the photovoltaic microgenerators were doped p-type and n-type silicon. Under illumination with an irradiance of 100 mW/cm^2 , the photovoltaic microgenerator achieved an energy-conversion efficiency of 5.446%. While the choice of illuminated surface was feasible, challenges associated with structural integration gave rise to mutual interference, subsequently undermining the photovoltaic performance and energy-conversion efficiency. Collectively, the aforementioned literature underscores the recent trend in photovoltaic microgenerators. These studies have explored altered silicon substrate structures, embraced back-illuminated configurations, and endeavored to fuse photovoltaic structures with alternative power-generation arrangements to achieve heightened output voltages. Nevertheless, it is evident that the performance of photovoltaic microgenerators can also be enhanced by means of patterned designs and alterations in the process materials through doping. This study focuses on the patterned design of p–n junctions within a photovoltaic microgenerator to enhance its performance.

Various microsensors [19–23], actuators [24–29], and devices [30–34] have been fabricated using the CMOS-MEMS technique. Photovoltaic microgenerators created through this technique offer improved efficiency, compactness, versatility, and cost-effectiveness, making them well-suited for a wide range of applications [35–39], including self-powered sensors, wearable devices, and internet of things (IoT). In this study, the same technique is employed to design a photovoltaic microgenerator. Additionally, a post-processing step is utilized to etch the silicon dioxide layer of the photovoltaic microgenerator, thereby enhancing its performance. In this paper, the design of the structure, simulation of its characteristics, the fabrication process, and the testing results of the photovoltaic microgenerator will be detailed.

2. Design of the Photovoltaic Microgenerator

A photovoltaic generator is a semiconductor device that converts sunlight directly into electrical energy through the photovoltaic effect. The fundamental principle behind the operation of a photovoltaic generator involves the interaction of light with semiconductor materials to generate an electric current. In this study, the material of the photovoltaic generator is silicon. When sunlight, which is composed of photons, strikes the surface of the silicon, it transfers energy to the atoms within the material. This energy is sufficient to dislodge electrons from their positions in the atoms, creating electron–hole pairs. The silicon material is doped with impurities to create a p–n junction. This is a boundary between a region with excess electrons (n-type) and a region with excess holes (p-type). The difference in charge concentration creates an electric field across the junction. The p–n junction absorbs photons, generating electrons and holes. Electrons are directed toward the n-type region, while holes are directed toward the p-type region. This charge separation leads to the accumulation of a voltage difference across the p–n junction. When an external

electrical circuit is connected to the photovoltaic generator, the separated charges flow through the circuit to recombine on the other side of the p–n junction. This movement of charges constitutes an electric current, which can be used to power electronic devices or stored in batteries.

Figure 1 illustrates the three-dimensional structure of the photovoltaic microgenerator. By incorporating distinct p-type and n-type doping into a silicon substrate, the photovoltaic microgenerator gives rise to intricate semiconductor configurations that involve p-type doping, n-type light doping (N-well), and n-type doping. These intricate configurations are structured to yield the photovoltaic microgenerator's functional framework. In this research, the patterned design is characterized by a mesh format, setting it apart from the conventional interdigitated shape layouts. As depicted in Figure 1, an extensive n-type lightly doped structure is seamlessly integrated into the silicon substrate. This strategic employment of light doping serves a dual purpose: it mitigates leakage currents while concurrently amplifying the performance of the p–n junctions. Moreover, a mesh-patterned p-type doping is ingeniously embedded within the region featuring n-type light doping. This innovative amalgamation of attributes—encompassing p-type doping/n-type light doping as well as p-type silicon substrate/n-type light doping p–n junctions—effectively enhances the power-generation capabilities of the photovoltaic microgenerator.

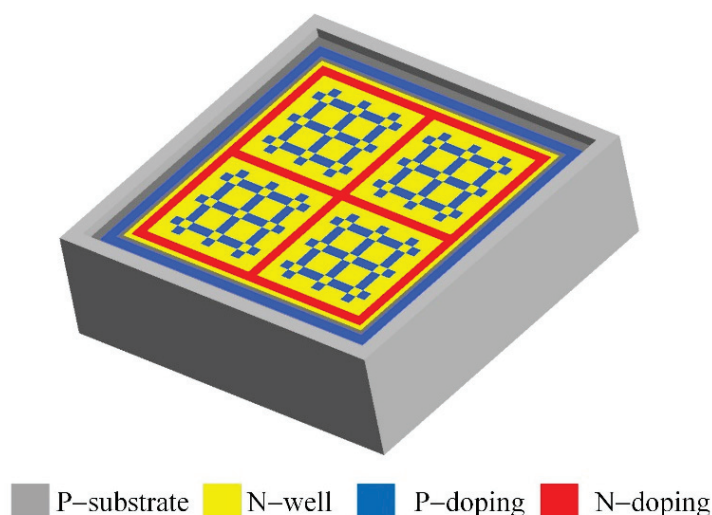


Figure 1. Three-dimensional structure of the photovoltaic microgenerator.

This research utilized the Sentaurus TCAD 2021 semiconductor device-simulation software developed by Synopsys to analyze the photovoltaic microgenerator. As shown in Figure 1, the model of the photovoltaic microgenerator is established, and the TCAD simulation process comprises three main stages: modeling and configuring the material parameters, meshing the model, and performing the computational analysis. In the initial phase, a three-dimensional model is constructed employing the Sentaurus structure-editor (SDE) tool, a component of the TCAD software. SDE delineates the structure of the model with fixed coordinates, delineating its dimensions, doping parameters across diverse layers, positioning the electrodes, and defining the Gaussian distribution to replicate the diffusion phenomena encountered during the actual chip fabrication. The subsequent phase encompasses mesh generation, entailing the determination of the model's mesh dimensions. Following the completion of meshing, the simulation's boundary conditions and environmental variables are established through command and parameter files. The command file governs environmental variables encompassing temperature, irradiance, light positioning, and wavelength. On the other hand, the parameter file manages the specification of the materials and physical property parameters. Once these variables are configured, the third step involves computing and analyzing the outcomes.

Figure 2 depicts the simulated voltage–current characteristics of the photovoltaic microgenerator under varying irradiance levels. Maintaining a constant wavelength and temperature, the simulation explores the voltage–current relationship across irradiance levels ranging from 100 W/m² to 1000 W/m², at 100 W/m² intervals. The simulations were carried out at a temperature of 25 °C and a wavelength of 600 nm. As illustrated in Figure 2, at an irradiance of 100 W/m², the open-circuit voltage is 0.48 V, while the short-circuit current registers at 27.5 µA. As the irradiance escalates to 1000 W/m², the open-circuit voltage elevates to 0.55 V, accompanied by a rise in the short-circuit current to 248 µA. The simulated results show that higher irradiance levels increase the current values of the photovoltaic generator.

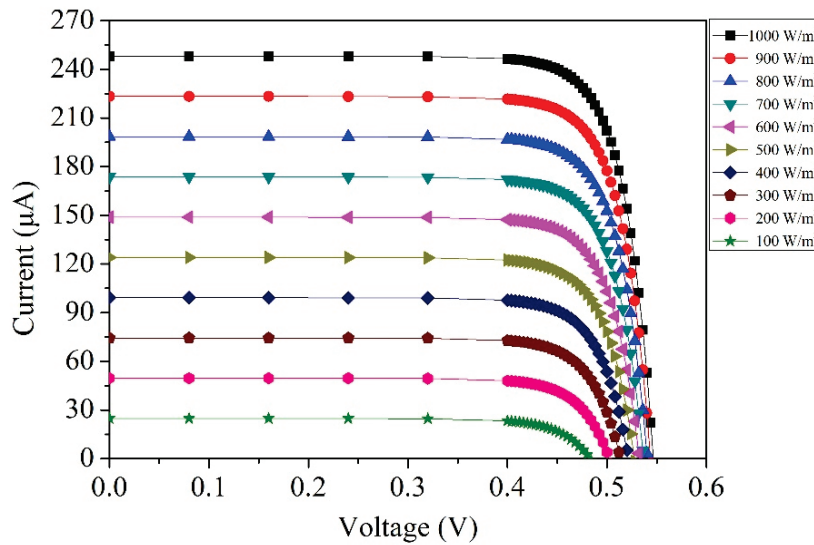


Figure 2. Simulated voltage–current characteristics of the photovoltaic microgenerator.

Output power stands as a critical performance metric for photovoltaic generators. The output power (P_o) of a photovoltaic microgenerator can be expressed as follows [40]:

$$P_o = I_o V_o \quad (1)$$

where I_o represents the output current of the photovoltaic microgenerator, and V_o is its output voltage. When the output current and voltage of the photovoltaic microgenerator are known, Equation (1) becomes instrumental in computing its output power. The simulation results of current and voltage for the photovoltaic microgenerator are depicted in Figure 2. By substituting these simulated outcomes into Equation (1), it becomes feasible to calculate the output power of the photovoltaic microgenerator, as depicted in Figure 3. The simulated output power of the photovoltaic microgenerator is shown in Figure 3. At an irradiance of 100 W/m², with an output voltage of 0.43 V, the photovoltaic microgenerator achieves a peak output power of 8.5 µW. As the irradiance increases to 1000 W/m², with an output voltage of 0.47 V, the photovoltaic microgenerator reaches its maximum output power of 109 µW.

The fill factor (FF) of a photovoltaic microgenerator is a measure of its ability to convert sunlight into electricity efficiently. It quantifies the quality of the current–voltage (I - V) characteristics of the photovoltaic microgenerator. The FF of the photovoltaic microgenerator is defined as [14]:

$$FF = \frac{P_{max}}{V_{oc} I_{sc}} \quad (2)$$

where P_{max} is the maximum output power of the photovoltaic microgenerator, V_{oc} is the open-circuit voltage of the photovoltaic microgenerator, and I_{sc} is the short-circuit current of the photovoltaic microgenerator. A higher fill factor indicates better performance and

efficiency in converting sunlight into usable electrical power. As shown in Figures 2 and 3, at an irradiance of 1000 W/m^2 , the photovoltaic microgenerator has an open-circuit voltage of 0.55 V , a short-circuit current of $248 \text{ }\mu\text{A}$, and a maximum output power of $109 \text{ }\mu\text{W}$. By substituting these values into Equation (2), the evaluated FF of the photovoltaic microgenerator is 0.8 .

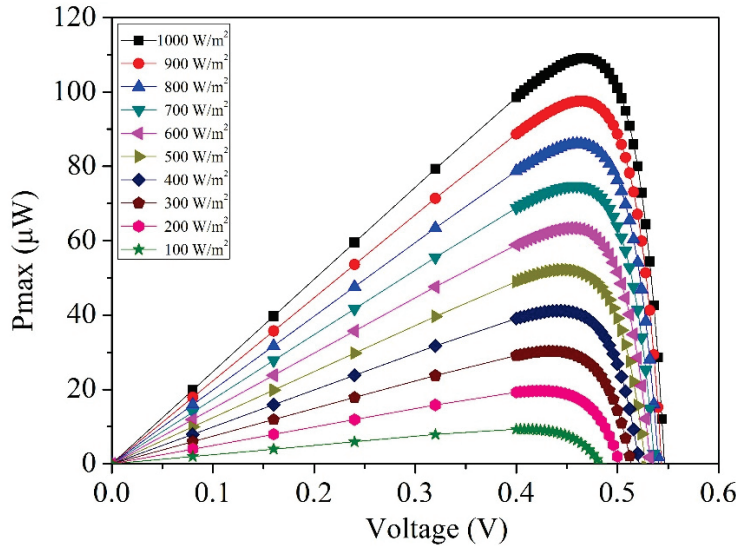


Figure 3. Simulated output power of the photovoltaic microgenerator.

The energy-conversion efficiency of a photovoltaic microgenerator refers to the ratio of the electrical power output generated by the photovoltaic device to the incident solar energy (light) input from the sun. It represents the effectiveness of the photovoltaic microgenerator in converting sunlight into usable electrical energy. The energy-conversion efficiency (η) of the photovoltaic microgenerator is expressed as [14]:

$$\eta = \frac{P_{max}}{P_{in}} \times 100\% = \frac{P_{max}}{AE} \times 100\% \quad (3)$$

where P_{max} is the maximum output power of the photovoltaic microgenerator, P_{in} is the input power of the incident light, A is the area of the photovoltaic microgenerator, and E is the irradiance of the incident light. As shown in Figure 3, at an irradiance of 1000 W/m^2 , the maximum output power of the photovoltaic microgenerator is $109 \text{ }\mu\text{W}$. The area of the photovoltaic microgenerator is 0.79 mm^2 . By substituting these values into Equation (3), the evaluated energy-conversion efficiency of the photovoltaic microgenerator is 13.8% .

3. Fabrication of the Photovoltaic Microgenerator

The photovoltaic microgenerator is produced utilizing the standardized $0.18 \text{ }\mu\text{m}$ CMOS process of Taiwan Semiconductor Manufacturing Company (TSMC). TSMC executed the fabrication of the photovoltaic microgenerator in accordance with the layout design depicted in Figure 1, utilizing the CMOS process. Figure 4 illustrates the process flow of the photovoltaic microgenerator. The oxide layer is initially present as part of the standard CMOS process. Its primary purpose is to act as a dielectric layer or a passivation layer. The passivation layer protects the underlying structures from environmental factors and potential contaminants. Figure 4a presents the cross-sectional structure of the photovoltaic microgenerator after the completion of the CMOS process. Notably, a silicon dioxide layer overlays the p–n junctions. To mitigate potential photon energy losses due to the reflections, refractions, or absorptions caused by the silicon dioxide layer—factors that could potentially impede the power generation efficiency of the photovoltaic microgenerator—this study employs a post-processing technique to remove the silicon dioxide layer.

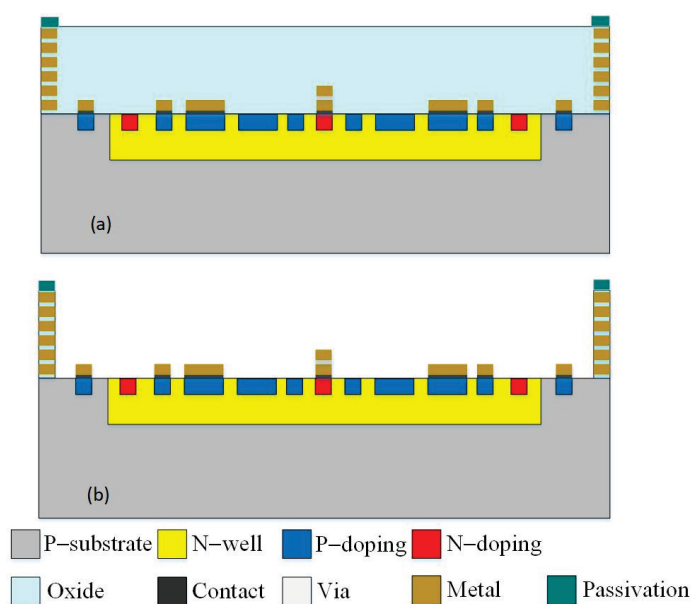


Figure 4. Cross-sectional structure of the photovoltaic microgenerator: (a) completion of the completion of the CMOS process; (b) etching the silicon dioxide layer.

Figure 4b shows the cross-sectional constitution of the photovoltaic microgenerator following the removal of the silicon dioxide layer. Post-processing involves the utilization of a BOE solution to etch the silicon dioxide layer until the p–n junctions are exposed. The thickness of the oxide layer was monitored using ellipsometry measurements to ensure the complete removal of the oxide layer during the etching process. The measured thickness of the oxide layer was about 8 μm . This configuration, as illustrated in Figure 4b, enables direct illumination of the p–n junctions by incident light, enhancing the overall efficiency. Figure 5a illustrates the optical image of the photovoltaic microgenerator following the completion of the CMOS process. In Figure 5b, a scanning electron microscope (SEM) image of the photovoltaic microgenerator after silicon dioxide layer etching is depicted. The SEM image distinctly reveals substantial depressions on the surface of the photovoltaic microgenerator’s structure resulting from the etching process. Figure 5c provides a localized SEM image that accentuates the concave characteristics resulting from the etching process, thus confirming the elimination of the silicon dioxide layer. The etching process is carefully optimized to minimize any adverse effects on the photovoltaic microgenerator chip. The SEM images (Figure 5) and the testing of the photovoltaic microgenerator demonstrate that the etching process does not compromise the structural integrity or performance of the photovoltaic microgenerator. The removal of the oxide layer can potentially expose the underlying material to external influences, which might reduce the device’s reliability. Figure 5d displays the image of the photovoltaic microgenerator chip subsequent to the wire-bonding procedure. The chip was affixed to a printed circuit board (PCB), and a wire bonder was employed to establish connections between the pads on the photovoltaic microgenerator chip and the PCB, utilizing aluminum wires.

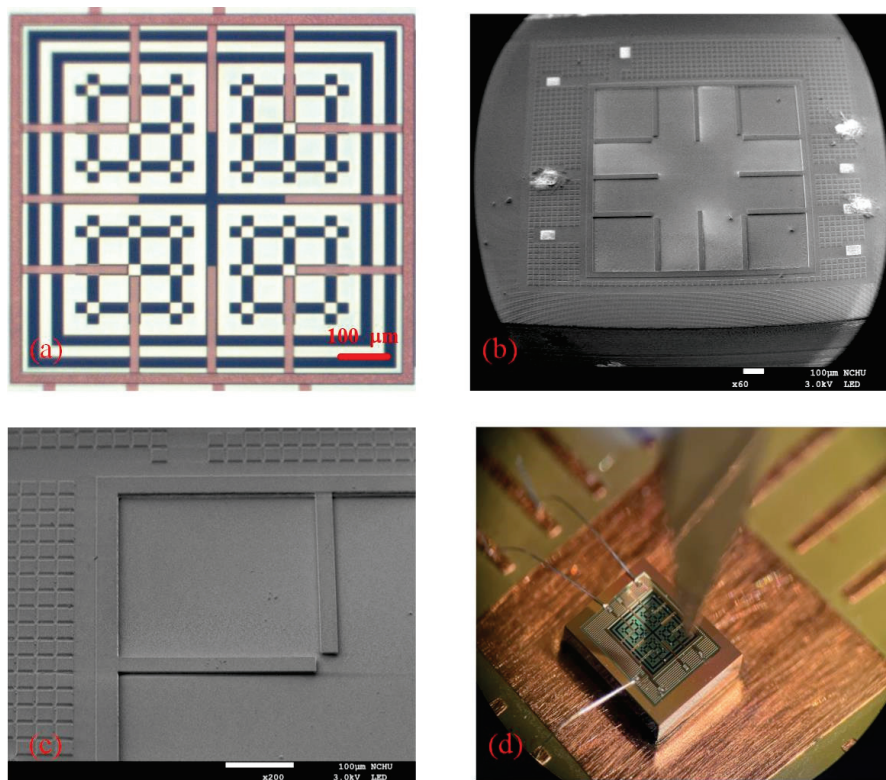


Figure 5. The photovoltaic microgenerator: (a) optical image of following the completion of the CMOS process; (b) scanning electron microscope (SEM) image after etching the silicon dioxide layer; (c) localized SEM image after post-processing; (d) image of the photovoltaic microgenerator chip during the wire bonding.

4. Results

The performance assessment of the photovoltaic microgenerator involved the utilization of a testing chamber, an optical power meter, and a digital multimeter. Within the testing chamber, a tungsten filament lamp served as the incident light source, offering adjustable irradiance through integrated control knobs. The tungsten filament lamp (TFC 250 W 120 V, Taiwan Fluorescent Lamp Co., Taipei, Taiwan) has a spectral response of approximately 380–780 nm. The optical power meter was utilized for calibrating the illumination intensity within the chamber. The digital multimeter was deployed to measure the output voltage and current generated by the photovoltaic microgenerator. The photovoltaic microgenerator was set within the testing chamber, and the intensity of the incident light source was regulated by manipulating the irradiance control knobs. The optical power meter was employed to monitor the irradiance level of the incident light source, whereas the digital multimeter recorded the output voltage and current of the photovoltaic microgenerator.

To comprehend the performance differences between photovoltaic microgenerators featuring etched and unetched silicon dioxide layers, tests were performed on both the photovoltaic microgenerator with an unetched silicon dioxide layer and the one with an etched silicon dioxide layer. Figure 6 illustrates the current–voltage curves of the photovoltaic microgenerators with etched and unetched silicon dioxide layers. These measurements were taken at room temperature with an incident light-source irradiance of 1000 W/m^2 .

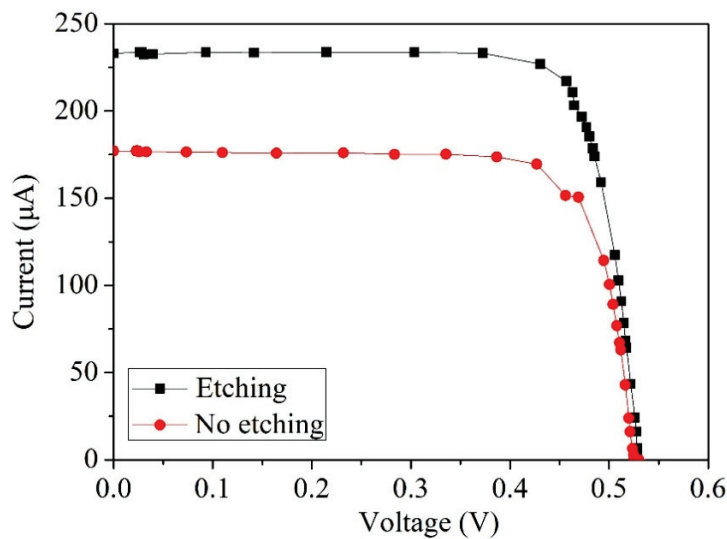


Figure 6. Current–voltage characteristics of photovoltaic microgenerators with etched and unetched silicon dioxide layers.

As shown in Figure 6, the short-circuit current of the photovoltaic microgenerator with an unetched silicon dioxide layer is 175 μA . In comparison, the photovoltaic microgenerator with an etched silicon dioxide layer exhibits a 25% increase in short-circuit current. This increase can be attributed to the fact that the incident light can directly illuminate the p–n junctions without being obstructed or interfered with by the silicon dioxide layer in the case of the photovoltaic microgenerator with an etched silicon dioxide layer. Consequently, the short-circuit current of the photovoltaic microgenerator with an etched silicon dioxide layer is enhanced. The influence of the silicon dioxide layer on light absorption in the photovoltaic microgenerator is primarily related to the optical properties and surface interactions. While SiO_2 itself is relatively transparent in the visible and near-infrared regions due to its high bandgap, its presence can still impact light absorption in a few ways. Firstly, the silicon dioxide layer can cause reflection and refraction of the incident light, leading to losses in the overall light absorption. Additionally, any surface imperfections or roughness induced by the silicon dioxide layer might contribute to scattering or diffraction of the light, thereby reducing the efficiency of light capture by the photovoltaic material. It is important to note that while SiO_2 may not directly absorb light, its presence can indirectly influence the overall light absorption efficiency of the photovoltaic microgenerator through these optical phenomena.

The measurement results from Figure 6 indicate that the performance of the photovoltaic microgenerator with the etched silicon dioxide layer surpasses that of the one with an unetched silicon dioxide layer. Consequently, all subsequent tests were conducted using the photovoltaic microgenerator with the etched silicon dioxide layer. Figure 7 illustrates the voltage–current curves of the photovoltaic microgenerator measured at various irradiance levels. These measurements were carried out at intervals of 200 W/m^2 , encompassing irradiance levels ranging from 200 W/m^2 to 1000 W/m^2 . As depicted in Figure 7, at an irradiance of 200 W/m^2 , the open-circuit voltage is 0.49 V, accompanied by a short-circuit current of 38 μA . With an increase in irradiance to 1000 W/m^2 , the open-circuit voltage ascends to 0.53 V, while the short-circuit current reaches 233 μA .

Figure 8 illustrates the comparison between the simulated and measured open-circuit voltages of the photovoltaic microgenerator. The simulated results are based on the open-circuit voltage data obtained from the photovoltaic microgenerator under different irradiance levels, as shown in Figure 2. On the other hand, the measured results are gathered from the open-circuit voltage data recorded under varying irradiance levels, as depicted in Figure 7. As depicted in Figure 8, at an irradiance of 200 W/m^2 , the simulated open-circuit voltage is 0.51 V, while the measured open-circuit voltage is 0.49 V. When the irradiance

increases to 1000 W/m^2 , the simulated open-circuit voltage is 0.55 V , and the measured open-circuit voltage is 0.53 V . The comparison of simulated and measured open-circuit voltages of the photovoltaic microgenerator reveals an error percentage of approximately 4%. Figure 9 displays the comparison between the simulated and measured short-circuit currents of the photovoltaic microgenerator. The simulated results are derived from the short-circuit current data of the photovoltaic microgenerator under different irradiance levels, as depicted in Figure 2. Similarly, the measured results are acquired from the short-circuit current data observed under varying irradiance levels, as shown in Figure 7. As shown in Figure 9, the short-circuit current exhibits a proportional relationship with light irradiance. With increasing irradiance, the short-circuit current demonstrates linear growth. At an irradiance of 1000 W/m^2 , the simulated short-circuit current is $248 \mu\text{A}$, while the measured short-circuit current is $233 \mu\text{A}$. The comparison of simulated and measured short-circuit currents of the photovoltaic microgenerator reveals an error percentage of approximately 6%. The difference between the simulated and experimental data in Figures 8 and 9 may be attributed to factors including the material doping concentration, manufacturing variations, and environmental influence.

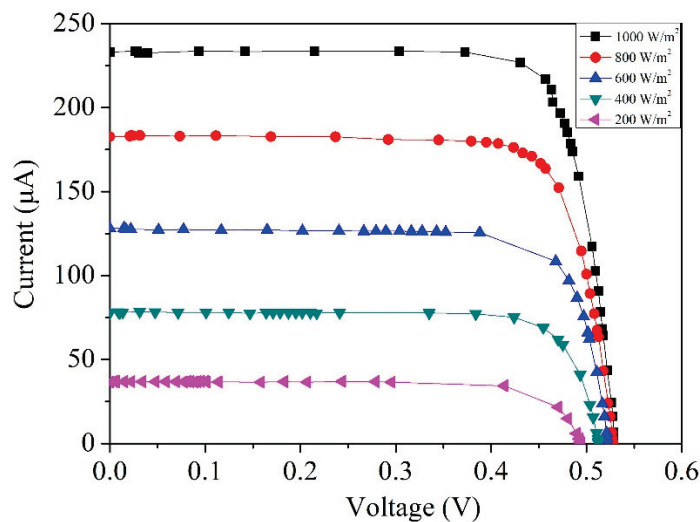


Figure 7. Measured voltage–current characteristics of the photovoltaic microgenerator.

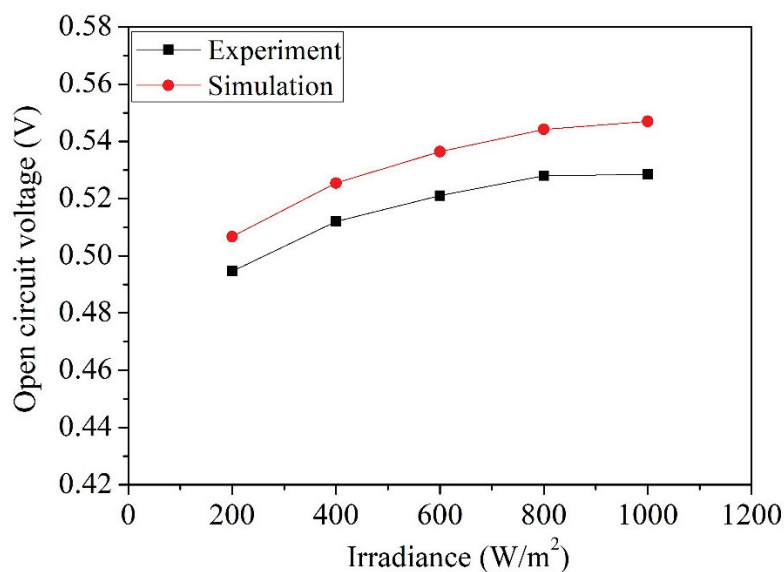


Figure 8. Open-circuit voltage of the photovoltaic microgenerator as a function of the irradiance.

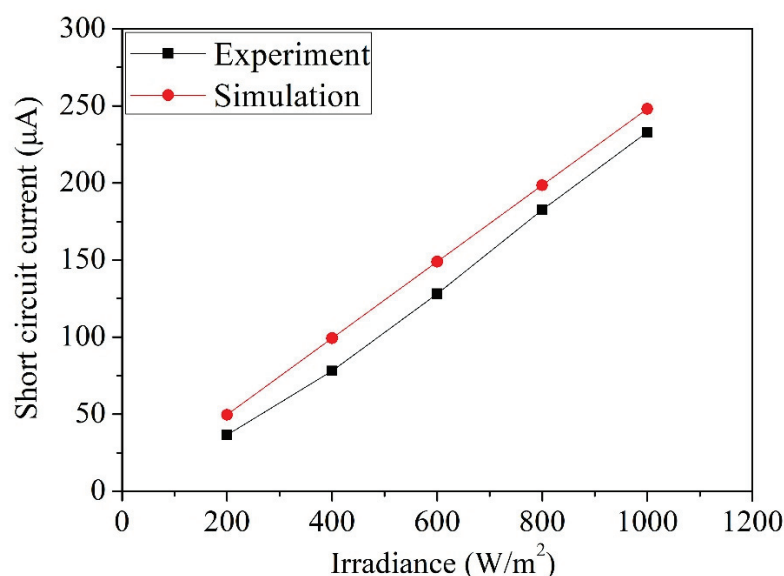


Figure 9. Short-circuit current of the photovoltaic microgenerator as a function of the irradiance.

By substituting in the measured current and voltage results of the photovoltaic microgenerator from Figure 7 into Equation (1), the calculated measured output power of the photovoltaic microgenerator can be determined. These results are showcased in Figure 10. As shown in Figure 10, at an irradiance level of 200 W/m^2 and an output voltage of 0.42 V , the photovoltaic microgenerator reaches a peak output power of 11 μW . As the irradiance level escalates to 1000 W/m^2 , the output voltage increases to 0.47 V , leading to a maximum output power of 99 μW for the photovoltaic microgenerator. Figure 11 depicts the comparison between the simulated and measured maximum output power of the photovoltaic microgenerator under different irradiances. The simulated results are based on the maximum output power data obtained from the photovoltaic microgenerator under different irradiance levels, as shown in Figure 3. On the other hand, the measured results are gathered from the maximum output power data recorded under varying irradiance levels, as depicted in Figure 10. The simulated outcomes depicted in Figure 11 show that at an irradiance level of 1000 W/m^2 , the simulated maximum output power of the photovoltaic microgenerator is 109 μW . A comparison between the simulated and measured maximum output power of the photovoltaic microgenerator reveals an error percentage of 10%.

As shown in Figures 7 and 10, at an irradiance of 1000 W/m^2 , the photovoltaic microgenerator has an open-circuit voltage of 0.53 V , a short-circuit current of 233 μA , and a maximum output power of 99 μW . By substituting these values into Equation (2), the measured FF of the photovoltaic microgenerator is 0.8. As depicted in Figure 11, at an irradiance of 1000 W/m^2 , the maximum output power of the photovoltaic microgenerator is 99 μW . The area of the photovoltaic microgenerator is 0.79 mm^2 . By substituting these values into Equation (3), the measured energy-conversion efficiency of the photovoltaic microgenerator is 12.5%.

Table 1 summarizes the performance of various photovoltaic microgenerators. The energy-conversion efficiencies of the photovoltaic microgenerators presented by Sakai et al. [9], Arima et al. [10], Law et al. [12], Hung et al. [16], and Yan et al. [17] are lower than 5%. Horiguchi et al. [13] achieved an energy-conversion efficiency of 7.5%, while Hung et al. [14] achieved 15%. In this study, the energy-conversion efficiency is 12.5%. Compared to those photovoltaic microgenerators [9–18], the energy-conversion efficiency in this study exceeds that of Sakai et al. [9], Arima et al. [10], Law et al. [12], Horiguchi et al. [13], Hung et al. [15], Hung et al. [16], and Yan et al. [17].

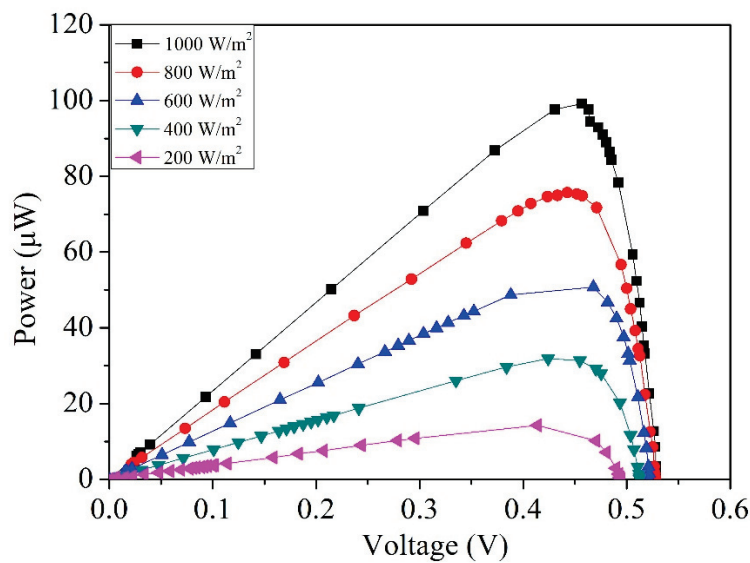


Figure 10. Measured output power of the photovoltaic microgenerator at varying voltages.

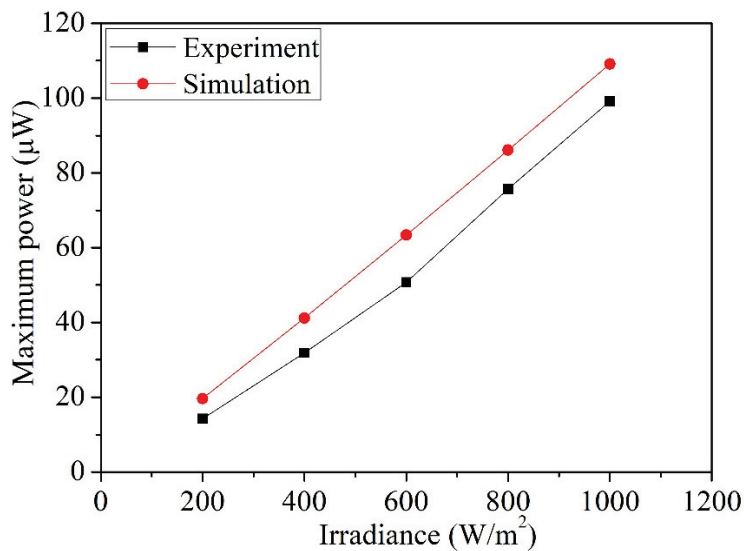


Figure 11. Maximum output power of the photovoltaic microgenerator at different irradiance levels.

Table 1. Summary of performance for various photovoltaic microgenerators.

Authors	Open-Circuit Voltage (V)	Short-Circuit Current	Energy-Conversion Efficiency (%)
Sakai et al. [9]	0.821	5.1 mA	2.35
Arima et al. [10]	0.83	0.4 μ A	2.6
Law et al. [12]	0.84	1.25 nA	0.3
Horiguchi et al. [13]	1.3	6 μ A	7.5
Hung et al. [14]	0.57	35 mA	15
Hung et al. [15]	2.05	2.6 mA	11
Hung et al. [16]	0.5	11 μ A	0.5
Yan et al. [17]	0.53	10.6 mA	4.45
Sichao et al. [18]	0.53	15.3 mA	5.45
This work	0.53	233 μ A	12.5

5. Conclusions

The photovoltaic microgenerator has been successfully manufactured using the standard 0.18 μ m CMOS process, coupled with subsequent post-processing treatment. TCAD

was employed to simulate the performance of the photovoltaic microgenerator. The simulated outcomes revealed that, at a temperature of 25 °C and a wavelength of 600 nm, under an incident light irradiance of 1000 W/m², the photovoltaic microgenerator had an open-circuit voltage of 0.55 V, a short-circuit current of 248 µA, and a maximum output power of 109 µW. Following the completion of the CMOS process, a wet-etching post-processing step was implemented to eliminate the surface silicon dioxide layer of the photovoltaic microgenerator. The incident light could directly illuminate the p–n junctions, thereby enhancing the output current of the photovoltaic microgenerator. The experimental results indicated that the photovoltaic microgenerator with the etched silicon dioxide layer exhibited a higher short-circuit current—approximately 25% greater than that of the photovoltaic microgenerator with the unetched silicon dioxide layer. Moreover, measurements demonstrated that at an irradiance of 1000 W/m², the photovoltaic microgenerator with the etched silicon dioxide layer showcased an open-circuit voltage of 0.53 V, a short-circuit current of 233 µA, a maximum output power of 99 µW, a fill factor of 0.8, and an energy-conversion efficiency of 12.5%.

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References

- Gu, Q.; Li, S.; Gong, W.; Ning, B.; Hu, C.; Liao, Z. L-SHADE with parameter decomposition for photovoltaic modules parameter identification under different temperature and irradiance. *Appl. Soft Comput.* **2023**, *143*, 110386. [CrossRef]
- Liao, Q.; Li, S.; Xi, F.; Tong, Z.; Chen, X.; Wan, X.; Ma, W.; Deng, R. High-performance silicon carbon anodes based on value-added recycling strategy of end-of-life photovoltaic modules. *Energy* **2023**, *281*, 128345. [CrossRef]
- Zhang, J.; Zhong, A.; Huang, G.; Yang, M.; Li, D.; Teng, M.; Han, D. Enhanced efficiency with CDCA co-adsorption for dye-sensitized solar cells based on metallosalophen complexes. *Sol. Energy* **2020**, *209*, 316–324. [CrossRef]
- Rehman, M.A.; Park, S.; Khan, M.F.; Bhupal, M.F.; Nazir, G.; Kim, M.; Farooq, A.; Ha, J.; Rehman, S.; Jun, S.C.; et al. Development of directly grown-graphene–silicon Schottky barrier solar cell using co-doping technique. *Int. J. Energy Res.* **2022**, *46*, 11510–11522. [CrossRef]
- Dheyab, A.B.; Wali, L.A.; Alwan, A.M.; Naseef, I.A. Perfect incorporation of AuNPs on the p-n+ porous silicon for highly-efficient solar cells. *Optik* **2019**, *198*, 163317. [CrossRef]
- Wali, L.A.; Dheyab, A.B.; Alwan, A.M. Study the influence of shell thickness in bimetallic Ag core&Au shell configurations integrated in bare Si PN junction solar cells. *Mater. Sci. Eng. B* **2023**, *288*, 116210.
- Qiu, L.; Deng, J.; Lu, X.; Yang, Z.; Peng, H. Integrating perovskite solar cells into a flexible fiber. *Angew. Chem. Int. Ed.* **2014**, *53*, 10425–10428. [CrossRef] [PubMed]
- Hyun, J.H.; Ha, D.S.; Han, D.D.; Kyung, C.M. A solar energy harvesting system for a portable compact LED lamp. In Proceedings of the IECON 2015—41st Annual Conference of the IEEE Industrial Electronics Society, Yokohama, Japan, 9–12 November 2015; pp. 001616–001621.
- Sakai, J.; Fujinaka, E.; Nishimori, T.; Lto, N.; Adachi, J.; Nagano, S.; Murakami, K. High efficiency organic solar cells by screen printing method. In Proceedings of the Conference Record of the Thirty-First IEEE Photovoltaic Specialists Conference, Lake Buena Vista, FL, USA, 3–7 January 2005; pp. 125–128.
- Arima, Y.; Ehara, M. On-chip solar battery structure for CMOS LSI. *IEICE Electron. Express* **2006**, *3*, 287–291. [CrossRef]
- Marwick, M.A.; Andreou, A.G. Photo-battery fabricated in silicon on sapphire CMOS. *Electron. Lett.* **2008**, *44*, 766–767. [CrossRef]
- Law, M.K.; Bermak, A. High-voltage generation with stacked photodiodes in standard CMOS process. *IEEE Electron Device Lett.* **2010**, *31*, 1425–1427. [CrossRef]
- Horiguchi, F. Integration of series-connected on-chip solar battery in a triple-well CMOS LSI. *IEEE Trans. Electron Device* **2012**, *59*, 1580–1584. [CrossRef]
- Hung, Y.J.; Chuang, T.Y.; Chun, C.L.; Cai, M.S.; Su, H.W.; Lee, S.L. CMOS-enabled interdigitated back-contact solar cells for biomedical applications. *IEEE Trans. Electron Device* **2014**, *61*, 4019–4024. [CrossRef]
- Hung, Y.J.; Cai, M.S.; Su, H.W. High-voltage generation in CMOS photovoltaic devices by localized substrate removal. *IEEE Electron Device Lett.* **2016**, *37*, 754–757. [CrossRef]

16. Hung, Y.J.; Cheng, Y.C.; Cai, M.S.; Lu, C.H.; Su, H.W. High-voltage 12.5-V backside-Illuminated CMOS photovoltaic mini-modules. *IEEE J. Electron. Device Soc.* **2018**, *6*, 135–138. [CrossRef]
17. Yan, J.; Liao, X.; Ji, S.; Zhang, S. MEMS-based thermoelectric–photoelectric integrated power generator. *J. Microelectromech. Syst.* **2019**, *28*, 1–3. [CrossRef]
18. Sichao, J.; Liao, X. Researches on MEMS thermoelectric-photoelectric integrated energy harvester with metal heat sink. *Microelectron. J.* **2020**, *96*, 104702.
19. Krayden, A.; Shlenkevitch, D.; Blank, T.; Stolyarova, S.; Nemirovsky, Y. Selective sensing of mixtures of gases with CMOS-SOI-MEMS sensor dubbed GMOS. *Micromachines* **2023**, *14*, 390. [CrossRef]
20. Shen, W.C.; Shih, P.J.; Tsai, Y.C.; Hsu, C.C.; Dai, C.L. Low-concentration ammonia gas sensors manufactured using the CMOS-MEMS technique. *Micromachines* **2020**, *11*, 92. [CrossRef]
21. Yang, M.Z.; Dai, C.L.; Wu, C.C. A zinc oxide nanorod ammonia microsensor integrated with a readout circuit on-a-chip. *Sensors* **2011**, *11*, 11112–11121. [CrossRef]
22. Guo, Y.; Ma, H.; Lan, J.; Liao, Y.; Ji, X. Impact of various thermistors on the properties of resistive microbolometers fabricated by CMOS process. *Micromachines* **2022**, *13*, 1869. [CrossRef]
23. Dai, C.L.; Kao, P.H.; Tai, Y.W.; Wu, C.C. Micro FET pressure sensor manufactured using CMOS-MEMS technique. *Microelectron. J.* **2008**, *39*, 744–749. [CrossRef]
24. Teo, A.J.T.; Li, K.H.H. Realization of three-dimensionally MEMS stacked comb structures for microactuators using low-temperature multi-wafer bonding with self-alignment techniques in CMOS-compatible processes. *Micromachines* **2021**, *12*, 1481. [CrossRef] [PubMed]
25. Dai, C.L.; Peng, H.J.; Liu, M.C.; Wu, C.C.; Hsu, H.M.; Yang, L.J. A micromachined microwave switch fabricated by the complementary metal-oxide semiconductor post-process of etching silicon dioxide. *Japan. J. Appl. Phys.* **2005**, *44*, 6804–6809. [CrossRef]
26. Perelló-Roig, R.; Verd, J.; Bota, S.; Segura, J. A Tunable-gain transimpedance amplifier for CMOS-MEMS resonators characterization. *Micromachines* **2021**, *12*, 82. [CrossRef] [PubMed]
27. Dai, C.L.; Chen, Y.L. Modeling and manufacturing of micromechanical RF switch with inductors. *Sensors* **2007**, *7*, 2660–2670. [CrossRef] [PubMed]
28. Zaman, A.; Alsolami, A.; Wei, M.; Rivera, I.; Baghelani, M.; Wang, J. Lateral extensional mode piezoelectric ZnO-on-Nickel RF MEMS resonators for back-end-of-line integration. *Micromachines* **2023**, *14*, 1089. [CrossRef] [PubMed]
29. Cheng, Y.C.; Dai, C.L.; Lee, C.Y.; Chen, P.H.; Chang, P.Z. A circular micromirror array fabricated by a maskless post-CMOS process. *Microsyst. Technol.* **2005**, *11*, 444–451. [CrossRef]
30. Peng, S.W.; Shih, P.J.; Dai, C.L. Manufacturing and characterization of a thermoelectric energy harvester using the CMOS-MEMS technology. *Micromachines* **2015**, *6*, 1560–1568. [CrossRef]
31. Lin, C.H.; Song, C.H.; Wen, K.A. Multi-function microelectromechanical systems implementation with an ASIC compatible CMOS 0.18 μm process. *Micromachines* **2021**, *12*, 314. [CrossRef]
32. Hu, Y.C.; Dai, C.L.; Hsu, C.C. Titanium dioxide nanoparticle humidity microsensors integrated with circuitry on chip. *Sensors* **2014**, *14*, 4177–4188. [CrossRef]
33. Lee, A.H.; Lee, J.; Laiwalla, F.; Leung, V.; Huang, J.; Nurmikko, A.; Song, Y.K. A scalable and low stress post-CMOS processing technique for implantable microsensors. *Micromachines* **2020**, *11*, 925. [CrossRef]
34. Yang, M.Z.; Dai, C.L.; Hong, J.Y. Manufacture and characterization of high Q-factor inductors based on CMOS-MEMS techniques. *Sensors* **2011**, *11*, 9798–9806. [CrossRef] [PubMed]
35. Chen, Z.; Law, M.K.; Mak, P.I.; Martins, R.P. A single-chip solar energy harvesting IC using integrated photodiodes for biomedical implant applications. *IEEE Trans. Biomed. Circuits Syst.* **2017**, *11*, 44–53. [CrossRef] [PubMed]
36. Ballo, A.; Grasso, A.D.; Palumbo, G. A bulk current regulation technique for dual-branch cross-coupled charge pumps. *IEEE Trans. Circuits Syst. II Express Briefs* **2022**, *69*, 4128–4132. [CrossRef]
37. Kuo, P.H.; Wong, O.Y.; Tzeng, C.K.; Wu, P.W.; Chiao, C.C.; Chen, P.H.; Chen, P.C.; Tsai, Y.C.; Chu, F.L.; Ohta, J.; et al. Improved charge pump design and ex vivo experimental validation of CMOS 256-pixel photovoltaic-powered subretinal prosthetic chip. *IEEE Trans. Biomed. Eng.* **2020**, *67*, 1490–1504. [CrossRef] [PubMed]
38. Ballo, A.; Grasso, A.D.; Palumbo, G. Fully on-chip charge pump-based boost converter in 65-nm CMOS for single solar cell powered IC. In Proceedings of the 2023 IEEE International Symposium on Integrated Circuits and Systems, Monterey, CA, USA, 21–25 May 2023; pp. 1–5.
39. Chiang, T.Y.; Dai, C.L.; Lian, D.M. Influence of growth temperature on the optical and structural properties of ultrathin ZnO films. *J. Alloys Compd.* **2011**, *509*, 5613–5626. [CrossRef]
40. Sun, M.; Liao, X. Modeling of the photoelectric–thermoelectric integrated micropower generator. *IEEE Trans. Electron. Device* **2021**, *68*, 4509–4515. [CrossRef]

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