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# 2 Array of Resonant Electromechanical Nanosystems: A

## 3 Technological Breakthrough for Uncooled Infrared

## 4 Imaging

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12 Abstract: Microbolometer is the most common uncooled infrared technique that allows to achieve 13 50mK-temperature resolution on the scene. However, this approach has to struggle with both the 14 self-heating inherent to the resistive readout principle and the 1/f noise. We present an alternative 15 approach that consists in using micro / nanoresonators vibrating according to a torsional mode, and 16 whose resonant frequency changes with the incident IR-radiation. Dense arrays of such 17 electromechanical structures were fabricated with a 12µm-pitch at low temperature allowing their 18 integration on CMOS circuits according to a post-processing method. H-shape pixels with 9 µm-19 long nano-rods and a cross-section of 250 × 30 nm<sup>2</sup> were fabricated to provide large thermal 20 responses, whose experimental measurements reached up to 1024 Hz/nW. These 21 electromechanical resonators featured a noise equivalent power of 140pW for a response time of 22 less than 1 ms. To our knowledge, these performance are unrivaled with such small dimensions. We 23 also showed that a temperature sensitivity of 20 mK within 100ms-integration time is conceivable 24 at a 12µm-pitch by co-integrating the resonators with their readout electronics and suggesting a 25 new readout scheme. This sensitivity could be reached at short-term by depositing on top of the 26 nano-rods a vanadium oxide layer having a phase-transition that could possibly enhance the 27 thermal response by one order of magnitude.

28 Keywords: nano resonator; electromechanical system array; bolometer

## 29 1. Introduction

30 Microelectromechanical Systems (MEMS) are either used according to a static mode 31 (accelerometers, micro-mirrors, RF switches, bimorph structures) or rather in dynamic way, when 32 the harmonic response of the system is requested. Although quartz crystals were first studied (and 33 continue to be widely used as time references), silicon is most likely the most widely used material 34 for MEMS and NEMS (nano-electromechanical systems) due to its excellent mechanical properties. 35 Indeed, advances in microelectronics on silicon wafers (increasingly large diameters - reliability and 36 reproducibility of manufacturing methods, SOI substrates) have strongly enabled the rise of silicon-37 based MEMS/NEMS. Beyond silicon, alternative materials are nowadays already industrialized such 38 as PZT and AlN, or envisaged such as GaAs, graphene, or aluminum, generally deposited on silicon 39 substrates (in particular for economic reasons). Finally, the intrinsic compatibility of the 40 manufacturing of silicon or silicon-based materials with a CMOS integration ("above-IC" approach 41 or co-integration approach) is an undeniable asset in order to realize a high-SNR system including 42 actuator / detector and compact readout circuit that is affordable and energy efficient.

The small size of NEMS makes them particularly sensitive to their external environment, while keeping very good frequency stability. In particular, silicon-based nanoresonators have already demonstrated their formidable potential in numbers of application domains. Nanoresonators are

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46 excellent physical sensors, especially for measuring forces [1, 2] or mass [3, 4], for gas detection [5, 6] 47 but also for measuring the temperature [7-10]. Ultimate sensitivity of the order of the yoctogram (10-48 <sup>24</sup>g - mass of a proton) has even been demonstrated with single-wall carbon nanotubes (CNT) [11, 12]. 49 More generally, nanoresonators can extend proteomics to high mass biomolecules (like complex of 50 proteins or virus) [3, 13]. There is also an intense activity in the NEMS community related to the study 51 of oscillators in their fundamental quantum mode using reciprocal interaction of optical micro cavity 52 with a mechanical resonator [14-16]. Examples of mechanical resonator applications are plethoric: 53 particle counting in fluid medium [17, 18], magnetometry [19-23], actuators [24] and RF filters [25], 54 biology [26, 27].

55 The measurement principle is quite simple: it consists in monitoring the frequency shift of a 56 NEMS kept at a given vibration (at a fixed controlled amplitude) using a closed loop circuit via a 57 "Phase-Locked Loop" (PLL) or a self-oscillating circuit. Any change in an external physical parameter 58 (temperature, pressure, force, acceleration ...) or on the surface of the material (adsorption of a gas, 59 molecule, etc.) will modify its stiffness or its mass, thus inducing a change in the resonant frequency, 60 which is continuously measured. Figure 1, adapted from [28], illustrates the measurement principle 61 when a frequency shift is caused by a mass landing on top of a nanocantilever. In this case, a 62 piezoresistive transduction is used to measure the mechanical oscillations of the cantilever (the 63 actuation being purely capacitive [29]). Indeed, it was shown that piezoresistive detection is highly 64 suitable at high frequency compared to capacitive readout in self-oscillation loop [30].

Neutral mass spectroscopy with such a system is now on-going and some papers have already demonstrated their interest for biomolecule analysis [3]. Beyond the ultra-sensitivity of NEMS, the overall analysis time has to be fast enough to make this technique a realistic solution. To tackle this key issue, the basic way consists in using a NEMS array for increasing the capture cross-section and hence speed up the analysis. In a recently published work, a NEMS array dedicated to mass sensing was realized with a frequency addressing, each NEMS having a slightly different resonance frequency labelling its position inside the array [31].

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Figure 1. Principle of mass measurement (adapted from [28]) – a) example of a nanoresonator on which
particles have landed; b) shift in frequency caused by the arrival of particles. Monitoring in real-time
the resonance frequency allow to deduce the amount of accreted mass; c) from the spectral perspective:
shift in the spectrum toward low frequencies.

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78 Nanoresonators can also be used for thermal sensing. Resonant NEMS arrays with a suitable 79 driving electronics could be a way to continue to decrease the pixel pitch of thermal imager keeping 80 constant the performance. Among the various current uncooled InfraRed Focal Plane Array (IRFPA) 81 technologies, microbolometer is the most common uncooled infrared device. It operates by 82 converting the heating of a thin suspended membrane due to an IR absorption into a variation of the 83 electrical resistance of a layer deposited on it. This layer is commonly made of a thin film of 84 semiconductor, using mostly Vanadium Oxide (VOX) or amorphous silicon (a-Si), because of their 85 high thermal sensitivity  $(1 / R \times dR / dT)$ , about 2-3 %/K. To reach a high thermal sensitivity, the 86 membrane has to be insulated from the substrate and is suspended above the readout integrated 87 circuit (ROIC) by thin and long insulation legs. Thermal insulation as high as 2 MK/W has for instance 88 been reported for microbolometers with 12 µm-pixel pitch [32]. Such a development has achieved 89 Noise Equivalent Temperature Differences (NETD) at a very low level around 50 mK (F/1 lens, 30 Hz 90 frame rate, 300 K background) [33], [34]. At ultra-small pixel pitch (less than 12µm), the performances 91 could be kept at the cost of a very high thermal insulation leading to a dramatic temperature increase 92 of the suspended plate that can be deleterious for the electrical properties of the sensing material [35]. 93 Both VOX and a-Si microbolometers can be affected [36]. The direct consequence is a persistent 94 afterimage when the scene is removing [37]. This is particularly true when the microbolometer is 95 exposed to a high temperature source such as fire, explosion or sun, and is sometime referred to as 96 "Sun Burn" effect [38]. Although a shutter-based non-uniformity correction can mitigate this effect, 97 latent image reappears rapidly after shutter operation and persists with a very long decay time. Thus, 98 there is a clear need of a new transduction method that can withstand high temperature exposure. In 99 parallel, this technique has to be compatible with a very large scale manufacturing for future 100 consumer markets that require small pixel-pitch and high resolution.

101 In this context, we suggest a new transduction method based on high-frequency mechanical 102 nanoresonators designed to be ultra-sensitive to IR radiation. In a mid-term vision, this approach 103 intends to replace the current thermistor-based bolometers. The frequency stability of these 104 nanoresonators should allow to reach the fundamental phonon noise and better cope with the 105 thermal issues, while keeping a high resolution on the scene for the small pixel sizes. In principle, the 106 excellent frequency sensitivity of nano/micro resonators makes them perfect ultra-small thermal 107 sensors. However, two main questions must be raised: how can an efficient electrical transduction of 108 tiny mechanical displacement be realized without any self-heating? Is the frame rate fast enough with 109 such an approach to get a net image? This paper will provide some key insights using a CMOS-based 110 approach.

111 In a first level of answer, the transduction technique has to be properly chosen. Unlike the 112 conclusions done in previous papers [30], capacitive transduction is preferred to get rid of the self-113 heating that results in a background signal and additional noise. Ultra-small capacitance variation is 114 however more delicate to read out [39], and a specific buffer circuit has to be developed as we will 115 show later on in the paper. Many types of transductions have been suggested over the last ten years. 116 A second level of answer lies in the use of NEMS array collectively or individually actuated and 117 detected by a CMOS circuit that could be placed in its close vicinity. The CMOS circuit will actuate 118 the mechanical systems at their resonance frequencies and perform the addressing of a single NEMS 119 or a sub-array of NEMS to be addressed. The spatial proximity of the CMOS with the thermal sensors 120 drastically limits the effects of electrical parasitic coupling and attenuation. Limiting these effects 121 consequently maximizes at circuit input both the absolute value of the useful signal and its signal to 122 background ratio (SBR), resulting in maximizing the SNR at the circuit output. To conclude, this 123 juxtaposition has significant advantages, namely the compactness of the system and the unparalleled 124 electrical transduction efficiency.

#### 125 2. Design and fabrication of electromechanical resonator arrays

126 The basic geometry of the resonator is a suspended plate that experiences a torsional vibration 127 around a rotation axis. Several metallic electrodes are structured underneath for electrostatic 128 actuation and capacitive measurement of the paddle displacement. These electrodes and the paddle

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129 form a  $\lambda/4$ -resonant cavity centered at 8µm for enhancing the absorption of the incident IR-radiation. 130 Fabrication process and materials have to be temperature-compatible with monolithic integration in 131 order to simplify the manufacturing of the imagers and their integration on a CMOS readout circuit. 132 Thus, the materials must meet three main criteria: (i) good mechanical features, (ii) low thermal 133 conductivity, (iii) low-temperature deposition. The critical dimensions (as the width legs ensuring 134 the rotation and the thermal insulation of the plate) have to be well controlled during the fabrication 135 process to have the all pixels functional inside the imager.

136 The low-temperature fabrication process is inherited from classical bolometers (i.e. deposition 137 process < 300°C and above-IC compatible). First, a 300nm thick AlCu layer is deposited on a silicon 138 substrate and structured to form the transduction electrodes. A 2µm-deposition of a polyimide layer 139 is then realized and will constitute the sacrificial layer. The latter is opened to build up the metal 140 studs that will insure the mechanical support and the electrical connection with the electrical 141 connections below. Two silicon nitride (SiN) layers of 10nm encapsulate a titanium nitride (TiN) layer 142 which will act as an electrode as well as an absorber. The TiN thickness is defined to be impedance 143 matched with the vacuum ( $Z_0$ ~376  $\Omega$ ) and to get a direct absorption rate close to 50%. The  $\lambda$ /4-optical 144 cavity (2µm thick) between the aluminum-copper electrodes and the TiN layer allows to reach an 145 absorption efficiency of 80% over the 8-14µm wavelength range. Figure 2g) shows the spectral 146 absorption of such cavity with this specific SiN/TiN/SiN/a-Si tack in this wavelength range.

147 The encapsulation of TiN layer is performed for stress compensation reason and for protecting 148 it during the release step. An amorphous silicon of 150nm-thick (a-Si) is deposited on the top-SiN to 149 stiffen the plate. The plate has to be polarized through the legs via the thin TiN layer. Electrostatic 150 actuation is possible in these conditions since no high current is required. Electrical contacts are made

151 by opening the top layers (SiN and a-Si).

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The torsional mechanical eigenmode has always been addressed in every designs of pixels so far. Indeed, the advantages of the torsional mode are threefold: (i) this mode is less sensitive to the residual axial stress that could be different from one side to the other side of an array [40], [41], (ii) the dynamic range set by the onset of nonlinearity is higher for the torsional mode compared to

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163 flexural modes [40], [42], [43], since only the external fiber of the rods experiences a strain [44], (iii) 164 the paddle surface remains large compared to the overall resonant body and makes easy a capacitive 165 actuation. Resonator arrays of 666×520 pixels, with a 12µm-pitch have been fabricated. Figure 3a) 166 shows a SEM picture of a typical array. The first electromechanical tests were achieved using 167 polarization lines structured below the pixel. This interconnection can be observed on Fig 3a) and 168 enables to actuate and read out an array of 96x96 electromechanical pixels. An SEM zoom-in of a 169 typical H-shape pixel is presented in Fig 3b and c. The nano-rod length is 1.5µm for a cross-section 170 of 250nm x 180nm (width x thickness). The insulation arm length is 8.6µm. This design is the nominal 171 version of our electromechanical pixel. Other versions were however realized and some of them are 172 presented in Fig 4a)-d). These alternative versions will be reviewed in the next section: they were 173 thought to try to meet the best trade-off between an efficient thermal insulation and a large 174 mechanical dynamic range, these two key features being antagonist.

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Figure 3. SEM pictures of an array of electromechanical pixels fabricated with a low-temperature process: a) large field view of an array; only the central 96x96 array is connected to electrical pads;
 pixels above the connection wires have been removed to avoid any cross-talk – b) Zoom-in on the center of the array – c) SEM picture of a typical H-shape pixel; nano-rod length= 1.5µm, width= 250nm and thickness= 180nm (insulation arm length= 8.6µm)



Figure 4. SEM pictures of alternative versions derived from the nominal design: a) butterfly-shape
pixel with longer rods – b) simple pixel without insulating legs – c) H-shape pixel with thinner nano
rod for enhancing the thermal insulation; rod-thickness= 30nm – d) Zoom of the legs (Fig 4c)) attached
to a stud that acts as mechanical anchor and that provides the electrical contact with the lines
underneath

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189 For the sake of clarity, a short introduction to the key mechanisms and noise sources is presented 190 below. We do not aim at detailing the thermo-electromechanical equations describing the overall 191 interactions between mechanics and IR-light. We rather give key expressions for catching up this 192 approach that may constitute a new paradigm in the field of IR-imager. The overall measurement 193 system including a single pixel is depicted in Fig 5. The expressions of parameters shown in this 194 figure are detailed step by step below for a comprehensive vision. The further expressions are 195 appropriate under a small-displacement (small deflection angle) assumption. The polarization of the 196 pixel is set to keep the angular vibration in its linear range at the chosen torsional resonance 197 frequency.  $V_B$  is the bias voltage applied on the paddle (through the studs), and  $V_{AC}$  is the sinusoidal 198 polarization applied on the actuation electrode through the capacitance  $C_a$ . This signal can be applied 199 with an external RF-source, in particular for the first electromechanical characterizations, but can 200 come from the feedback loop in the case of a closed loop. The actuation frequency f is swept to 201 measure the electromechanical response and the resonance frequency  $f_0$ .



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202 Figure 5. Synopsis of the open-loop measurement chain: The red box corresponds to a single 203 electromechanical pixel that translates the incident IR-radiation  $P_{inc}$  on the scene into a resonance 204 frequency shift; the blue box corresponds to the close by electronics that converts the mechanical 205 oscillations into an electrical signal;  $V_{pol}$  is the polarization of the pixel;  $\theta(f)$  is the angular oscillation 206 of the paddle around the rods;  $C_d(f)$  is the induced capacitance variation used to read out the signal; 207  $V_{out}(f)$  is the output signal supplied by the buffer.  $C_p$  is the total capacitance due to amplifier 208 input capacitance and parasitic capacitances between the electrical connections and the 209 ground.

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211 Basically, the electromechanical pixel converts the incident IR-optical power  $P_{inc}$  into a 212 resonance frequency shift  $\Delta f$  according to a sensitivity  $R_f$ , which depends on both the thermal 213 conductance of the paddle insulation (through insulation legs between the torsional rods and the 214 plate) and the temperature coefficient of frequency:

$$\Delta f = \frac{\alpha_{\rm T} f_0 \beta \eta}{G|1 + j 2 \pi v \tau_{\rm th}|} P_{\rm inc} = f_0 R_f P_{\rm inc}$$
(1)

215 where  $\tau_{th} = C/G$  is the thermal time constant of the sensor,  $C = \left(\frac{\partial U}{\partial T}\right)_V$ , the thermal capacitance 216 at constant volume, *G* the thermal conductance,  $\alpha_T$  the temperature coefficient of frequency (TCF) 217 (typically  $-60 \ ppm/^\circ C$  for silicon),  $\beta$  the pixel fill factor,  $\eta$  the bolometer absorption,  $f_0$  the 218 resonance frequency and  $\nu$  the frame rate of electronic readout. The thermal conductance is mainly 219 due to the thermal conductance of heat through the legs. The other sources of thermal leaks – 220 radiative and heat conductance through air are negligible.

The capacitance variations can be calculated from geometrical considerations. After cumbersome mathematical manipulation, the final expressions can be approximated as:

$$C_a(\theta) \approx -C_0 \frac{\theta_{max}}{\theta} \ln\left(1 - \frac{\theta}{\theta_{max}}\right),\tag{2}$$

$$C_d(\theta) \approx C_0 \frac{\theta_{max}}{\theta} \ln\left(1 + \frac{\theta}{\theta_{max}}\right)$$
 (3)

223  $C_0 = \varepsilon_0 \frac{L_p W_p/2}{g}$  and  $\sin(\theta_{max}) = \frac{g}{W_p/2}$ .  $C_0$  and  $\theta_{max}$  are respectively the capacitance value at rest, and 224 the maximum deflection angle. The deflection angle is directly computed from the dynamic equation:

$$J\theta + b\theta + \kappa\theta = T_e$$

$$T_e = \frac{1}{2} \frac{dC_a}{d\theta} V_{pol}^2 + \frac{1}{2} \frac{dC_d}{d\theta} V_B^2$$
(4)

225  $T_e$  is the electrostatic torque.  $J = \frac{M_p W_p^2}{L_r}$  is the moment of inertia of the paddle, assuming the inertia 226 moment of rods is negligible.  $\kappa = \frac{2GI_r}{L_r}$  is the rod torsional stiffness.  $G = \frac{E}{2(1+\nu)}$  is the shear modulus 227  $I_r$  is the torsional quadratic moment of the rectangular suspended rods<sup>1</sup>. *E* and  $\nu$  are the equivalent 228 Young modulus and Poisson ratio of the stack. Under the assumption of a linear regime (and small 229 deflection amplitude of the paddle), the angle can be rewritten in the Fourier space:

$$\theta(f) = \frac{T_e}{J} \frac{1}{f_0^2 - f^2 + j\frac{ff_0}{Q}}$$
(5)

The table 1 presents the values of the main features of equations (1)-(5) for our typicalelectromechanical pixel. Some parameters are compared with data from literature.

<sup>1</sup> 
$$I_r = w_r t_r^3 \left( \frac{1}{3} - 0.21 \frac{t_r}{w_r} \left( 1 - \frac{1}{12} \frac{t_r^4}{w_r^4} \right) \right), w_r > t_r$$

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**Table 1.** Key parameters presented in the equations (1)-(5) for our device compared with an advanced resistive bolometer and MEMS bolometer: temperature sensitivity corresponds to  $1/f \times \partial f / \partial T$  for resonant thermal sensor and  $1/R \times \partial R / \partial T$  for resistive one.

	This work (Fig. 3c)	Bolometer [32]	Resonant MEMS [9]
Maximal angle $\theta_{max}$ (°)	21	N.A	-
Inertial moment J (kg.m <sup>2</sup> )	$3,9 \times 10^{-25}$	N.A	$1.5 \times 10^{-27}$
Stiffness $\kappa$ (N.m)	$1,8 \times 10^{-11}$	N.A	$6.8 \times 10^{-13}$
Resonant frequency (MHz)	1.1	N.A	-
Onset of linearity $\theta_c(^\circ)^2$	13.5	N.A	-
Quality Factor Q	1800	N.A	1555
Capacitance at rest $C_0$ (fF)	0.185	N.A	N.A
Pitch [µm]	12	12	5
Thermal Conductance G [W/K]	$5 \times 10^{-8}$	$5 \times 10^{-9}$	$1.5 \times 10^{-8}$
Thermal Capacity C [J/K]	$26 \times 10^{-12}$	$80 \times 10^{-12}$	$3 \times 10^{-12}$
Thermal constant $\tau_{th}$ [ms]	0.5	16	0.2
Temperature sensitivity [/°C]	0.01 %	3.6 %	0.0092 %

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237 At this stage we have to struggle with a strong signal attenuation due to a capacitive bridge formed by parasitic capacitances from metallic pads, connections and input impedance of the final 238 readout electronics board:  $\frac{V_B \Delta C}{(C_0 + C_p)}$ . The order of magnitude of an expected capacitance 239 240 variation  $\Delta C$  is around 10 aF for a  $C_0 \sim 200$  aF and  $C_p \sim 10 pF$ . In this condition, the output signal is 241 divided by a factor ~106. This attenuation of the signal can be deleterious to get a high-enough signal 242 to background ratio (SBR) to initiate a self-oscillation within a closed-loop. A way to address this issue can be a use of a semitone-actuation (at  $f_0/2$ ):  $V_{pol} = V_{AC} cos\left(\frac{2\pi ft}{2}\right) - V_B$ . In this case the 243 244 electrostatic torque is proportional to  $V_{AC}^2/2$  which reduces the coupling between the actuation signal 245 and the output signal. A differential measurement can also be added to further improve the SBR. In 246 this scheme, two identical pixels are used to cancel out the common modes. A more complex 247 approach based on the down-mixing method [45] can be used to get rid of the parasitic capacitances. 248 In particular, the bias voltage is no more constant and is modulated:  $V_B = V_{B0} \cos(2\pi f t + \Delta f)$  where 249  $\Delta f \ll f$ . A comparison between different readout modes is shown in Tab 2. We notice a quite strong 250 improvement of the SBR. However, in the best cases the signal to noise ratio (SNR) was lower than 251 20dB, which did not guarantee a functional closed-loop.

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**Table 2.** Preliminary measurement of the *SBR* for different transduction strategies (direct semitone, direct 1f, differential and down-mixed);  $V_{B0} = 10V$ ,  $V_{DC} = 10V$  and  $f_0 = 1MHz$ ,  $\Delta f = 10kHz$ ;  $V_{AC0} =$ 4.2*V* for semitone actuation and  $V_{AC0} = 0.5V$  for 1-f and 2f actuations.

Transduction method	Volta	SBR (dB)	
	V <sub>AC</sub>	V <sub>B</sub>	
1 <i>f</i> -actuation	$V_{AC0}cos(2\pi ft)$	$V_{B0}$	-33
f/2-actuation	$V_{AC0}cos\left(rac{2\pi ft}{2} ight)$	$V_{B0}$	-13
f/2-actuation / differential mode	$V_{AC0}cos\left(\frac{2\pi ft}{2}\right)$	$V_{B0}$	2
f/2-actuation / down-mixing mode	$V_{AC0}cos\left(\frac{2\pi ft}{2}\right)$	$V_{B0}\cos(2\pi ft+\Delta f)$	22
<i>f</i> -actuation / down-mixing mode	$V_{AC0}cos(2\pi ft) + V_{DC}$	$V_{B0}\cos(2\pi ft+\Delta f)$	20
2 <i>f</i> -actuation / down-mixing mode	$V_{AC0}cos(4\pi ft + \Delta f)$	$V_{B0}\cos(2\pi ft+\Delta f)$	22

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<sup>2</sup> This value is computed by solving nonlinear dynamic equation [46].

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In conclusion, to reach higher SBR (40dB can be considered as a suitable value for the PLL) a dedicated off-chip buffer to cope with the tiny capacitance variation was developed and placed in the close vicinity of the pixels under test. Its schematic is presented in Fig 5. The capacitance variation is read through an intermediate circuit that measures the charge carrier variation, the applied voltage being kept constant. The current is read with a feedback capacitance  $C_{C2V}$  instead of a resistor to minimize the background, as shown in Fig 5. Thus, the output voltage is proportional to this feedback

263 capacitance as  $V_{out} = V_{pol} \frac{\delta C(\theta(f))}{C_{C2V'}}$  where  $\delta C(\theta_r)$  is the capacitance variation resulting from

- 264 the motion of the paddle.  $C_{C2V}$  must be chosen as low as possible to maximize the output signal but
- should be high enough to avoid unwanted effects due to parasitic capacitances (and  $k_BT/C$  noise).
- 266 A 1pF feedback capacitance in parallel with a 10  $G\Omega$  resistor are set to prevent from any saturation
- of the output signal caused by DC current.  $C_d$  and  $C_p$  are the sensor capacitance and the total input
- capacitance of the amplifier, respectively. This electrical scheme corresponds to a first-order filter[47]:

$$V_{out} = V_{pol} \frac{C_d}{C_{C2V}} \frac{1}{\left(1 - j\frac{f_c}{f}\right) \left(1 + j\frac{f}{f_{co}}\frac{C_{C2V}}{C_p}\right)}$$
(5)

270  $f_{CO}$  is the high-pass cut-off frequency of the amplifier and  $f_c = \frac{1}{2\pi R_{FB}C_{C2V}}$  the low-pass frequency at 271 -3dB of the *RC* filter. The improvement brought by the buffer circuit is illustrated in Fig 6. With our 272 buffer, the SBR was kept quite constant while the SNR is strongly improved growing up to 42dB (to 273 be compared with 20dB without buffer).



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Figure 6. Enhancement provided by the buffer circuit: a) Synopsis showing the set-up to characterize the pixels in a down-mixed readout scheme (in open loop or in a closed-loop: red part) – b) Comparison of the output signal between a semitone down-mixing approach with the buffer circuit and without the buffer circuit (the polarization voltages are explained in Tab 2) – c) Typical output signals for the three down-mixing approaches with the buffer circuit. At resonance the capacitance variation is close to 10 aF. The SNR is larger than 40dB for the three cases

284 The output voltages at resonance were high and clean enough to embed the electromechanical 285 pixels into a closed-loop. At this stage, the pixels of the array of 96x96 pixels (see Fig 3a) & b)) have 286 been tested using an external closed-loop based on a down-mixed Phase locked loop (PLL) scheme. 287 The latter is shown in red in Fig 6a). To do so, the output phase signal  $\Delta \varphi$  (demodulated at  $\Delta f$ ) is 288 the input signal of a digital proportional-integral-derivative controller (PID). The output  $f_r$ 289 corresponds to the frequency applied to the pixel. The PID parameters were set according to the 290 Ziegler-Nichols method [48], hence modifying the bias voltage  $V_B$  applied on the pixel. This voltage 291 enables to control the effective stiffness [46].

292 Next section will provide typical electromechanical results and the thermal sensitivity of our 293 electromechanical system. The noise sources of such a system are presented. Based on these 294 measurements, new readout schemes of large pixel arrays are suggested to achieve a compact CMOS

circuit beneath the imager.

## **296 3. Results**

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297 *3.1. Electromechanical characterizations* 

First, the electromechanical responses of the pixels of the 96x96 array were measured in open loop. The variations of the resonance amplitude as a function of the voltages  $V_{B0}$  and  $V_{AC0}$  have been

loop. The variations of the resonance amplitude as a function of the voltages  $V_{B0}$  and  $V_{AC0}$  have been verified for the f/2 and 2f actuation schemes, up to the onset of non-linearity (i.e.  $\theta_c \sim 17^\circ$ ). Figures 7a)

werified for the f/2 and 2f actuation schemes, up to the onset of non-linearity (i.e.  $\theta_c \sim 17^\circ$ ). Figures 7a) and b) correspond to a f/2-actuation showing, as expected, a quadratic variation of the output voltage

302 at resonance with  $V_{ACO}$  and a linear variation with  $V_{BO}$  respectively. Figures 7c) and d) correspond

to the 2f-actuation showing a linear variation of the output voltage with  $V_{AC0}$  and a quadratic

304 variation with  $V_{B0}$ , which is expected too.



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**Figure 7.** Electromechanical response when frequency is swept around resonance for different actuation schemes: a) Amplitude versus f and  $V_{AC}$  for a semitone actuation ( $V_B = 10 V$ ); b) Amplitude versus f and  $V_B$  for a semitone actuation ( $V_{AC} = 6.5V$ ); c) Amplitude versus f and  $V_{AC}$ for a 2f-actuation ( $V_{B} = 10 V$ ); d) Amplitude versus f and  $V_B$  for a 2f-actuation ( $V_{AC} = 1.8V$ ).

The main electromechanical features for the torsional mode ( $f_0$ , Q and  $V_{out}$ , the maximum output voltage corresponding to the onset of nonlinearity of the deflection angle  $\theta_c$ ) were measured on every pixels of a 96x96 array with the 2f-actuation scheme:

- 313  $f_0 = [1.05 MHz 1.2 MHz];$
- $314 \quad \bullet \quad Q = [1600 2500];$
- 315  $V_{out} = [100 \,\mu V 350 \mu V]$

The range of resonance frequencies is coherent with the fabrication process dispersion of 10% on the torsional rod width (length and thickness variations are negligible). The variations on the quality factor and the maximum voltage are rather more sensitive to the mechanical anchoring of the insulation legs and the over-etching effect between the edge and the center of the array explains this difference.

The frequency dispersion does not impact the closed-loop performance and has a tiny impact on the thermal response (see equation 1). However, the quality factor and the dynamic range have a larger impact on the noise floor level, as we will see later in section 3.2. It means that the pixel on the

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324 edges of the future imager will be slightly less sensitive compared to the others. To go further on this 325 topic, the next section will address the noise of the readout chain and the thermal performance of 326 such pixels.

327 *3.2. Thermal characterizations* 

328 Let's go back to equation 1 that gives the thermal response to an incident IR-radiation. The 329 frequency shift is proportional to the temperature coefficient of frequency  $\alpha_T$  and inversely 330 proportional to the thermal conductance *G* of the material stack of rods and insulation legs.

331 3.2.1. TCF & G

332 Using the closed loop (Fig 5a)), we have implemented systematic TCF measurements on typical 333 devices (Fig 3c) and Fig 4a) to c)). In order to carry out a large number of measurements within a 334 reasonable time, the devices were tested on an automatic probe-station dedicated to 200mm wafers. 335 The latter was heated with a hotplate to have a temperature variation between 0° and 20° above the 336 ambient temperature. Beyond this limit, the closed-loop did not track the frequency shift anymore. 337 The measurements were performed with a coupled Peltier-Pt sensor controlled by a Proportional 338 Integral Derivative controller (PID) to know the chamber temperature (down to 0.1°C-accuracy). The 339 statistics are summarized in Tab 3.

340 341

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**Table 3.** TCF measured on different types of pixels: mean and standard deviation per wafer; Thermal conductance of rods & legs *G* (computed from material properties and geometry measured by SEM)

			0	
Pixel types	$\langle \alpha_T \rangle$ (ppm/°C)	$\sigma_{\alpha_T}$	<b>G</b> (W/K)	$\langle \alpha_T \rangle / G$
Typical (Fig 3c))	55.4	14.6	$5 \times 10^{-8}$	$1.11 \times 10^{9}$
Butterfly (Fig 4a))	45.2	3.6	$3.10 \times 10^{-8}$	$1.46 \times 10^9$
Typical with thin nano-rod (Fig 4c))	86.2	16.4	$1.8 \times 10^{-8}$	$4.79 \times 10^{9}$

343

The TCF of a typical pixel (55.4ppm/°C) is in good agreement with the theoretical value of 48ppm found by Finite Element Method simulation (FEM) done on our stack. In the case of pixel with thin nano-rods, the axial internal stress is higher and its variation with temperature reinforces the TCF (same sign of variation). In the meantime, thermal insulation is quite enhanced. Thus a global improvement of the thermal response should be expected with this kind of pixel.

349

350 3.2.2. Thermal response

351 The thermal response was measured and compared with the theoretical values computed by FEM. 352 Measurements were performed with the readout chain shown in Fig 5a) in closed-loop. The device 353 under test (a pixel array) was placed into a vacuum chamber and a blackbody source (RCN 1200 from 354 HGH Infrared Systems set at 1200 °C) was positioned in front of it. An 8-12 µm-filter was put between 355 our chamber and this source to control the incident power. The optical bench was aligned thanks to 356 a visible laser. The optical set-up was calibrated using a Fourier Transform Infrared instrument 357 (FTIR). In particular, the spectral response of the filter according to the spectral luminance of a perfect 358 blackbody at 1200°C was measured. A photometric computation (knowing optical apertures and 359 relative distances between the optical blocks) was used to determine the incident optical power. We 360 considered that the source is a Lambertian black body with a monochromatic luminance described 361 by Planck's law. The aperture of the source and the chamber window were close enough to neglect 362 the atmosphere absorption in the estimation of incident power (d = 2.5 cm). The frequency response 363 of our typical pixel to IR incident pulses (17nW peaks) is presented in Fig 8a). Thermal responses up 364 to  $R_f$ =1050/W were extracted with the best devices ( $f_0 = 1.15MHz$ ). Assuming a fill factor  $\beta$ =0.8, an 365 efficiency  $\eta$ =0.8 in the 8-12 µm window and considering the measured TCF,  $\alpha_T$ =-76 ppm/°C, a 366 theoretical thermal response  $R_f = 950/W$  was expected, which is very close to the observed 367 sensitivities. In a second experiment, the incident IR flux on a pixel was changed by varying the 368 distance between the window and the IR-source. The frequency shift was then the measured for 369 optical powers varying from 2.5nW to 16nW. The experimental results and their linear fit are

presented in Fig 8b). A thermal response of  $R_f$ =1350/W is extracted from the slope considering the resonance frequency mentioned above. Above 8nW, the relationship between the IR-flux and the thermal frequency shift is no more linear. To increase the incident power, the source was moved close to the window, which turned into its heating. This effect lowers its transmittance making the estimation of the thermal response wrong (namely  $R_f$ =1050/W).

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**Figure 8.** Measurement of the thermal response: a) Resonance frequency shift induced by incident IR pulses (peaks of 17nW) – b) Frequency shift when the incident power is varied from 2 to 16nW

Similar experimental thermal responses were extracted on other pixels of a same array. From an array to another, experimental  $R_f$  varied from 700/W to 1350/W showing some dispersion attributed to the fabrication process.

383

384 3.2.3. Response time

385 The response time was measured with a Helium Neon laser (@633nm) from a commercial Polytec 386 vibrometer. Since the needed integration time was too short (180 µs) compared to the PLL time 387 constant, this experiment was performed in open loop scheme (see Fig 6a)). We verified that the 388 response time is not limited by the electrical low-pass filter from our measurement set-up by setting 389  $\tau$ = 50µs. The optical power was set to remain in the linear dynamic range (± 13.5° for the typical 390 design). The 10-90 % method was used to extract the fall time  $t_r$ , and the response time of the first 391 order low-pass filter  $\tau$  ( $\tau = t_r/ln(9)$ ). Doing so, we extracted a response time of 430µs, which is close 392 to the theoretical value computed with the thermal equations (500  $\mu$ s). The resonant 393 electromechanical pixels can follow quite fast event in the scene. They have faster response than 394 current resistive pixels.



396Time (ms) - arbitrary origin397Figure 9. Resonance frequency according to the acquisition time. A 1mW red laser is focalized onto the398pixel under test (typical Fig 3c)). Insert: Full data from our measurement of response time. The average399frequency jump is estimated as 110Hz. Then, the response time is extracted from one event fall time400(red).

#### 401 3.3. Noises and temperature sensitivity

402 The performances of the electromechanical pixels were estimated through the Noise Equivalent 403 Power (*NEP*) or the Noise Equivalent Temperature Difference (*NETD*). The *NEP* is defined as the 404 incident power on the sensor surface with a *SNR* of 1. This corresponds to the minimum measurable 405 frequency shift:

406

$$NEP = \frac{1}{R_f} \frac{\langle \delta f^2 \rangle^{1/2}}{f_0} = \frac{\sigma_y}{R_f}$$
(6)

407  $<\delta f^2 >^{1/2}$  is the rms frequency fluctuation for a given bandwidth and  $\sigma_y$  is the quadratic deviation 408 of the instantaneous relative frequency.

409 Current "column" or "rolling-shutter" readout schemes should be implemented with our 410 resonator array. The column readout, which is the current approach with CMOS circuit, requires a 411 60Hz frame rate, which sets a pixel integration bandwidth at 7kHz for 190 pixels per column, for 412 instance [49]. However, this readout scheme may induce a lag effect leading to an image distortion 413 when the scene moves faster than the frame rate. Single-pixel readout is a solution to remove this 414 effect. In this case, the integration time corresponds to the full frame rate, i.e. 50Hz, increasing thereby 415 the SNR. As the capacitive detection does not suffer from self-heating issue, it is possible to use a 416 longer integration time without material degradation even for small pitch below 12µm. This is why 417 the NEP of our sensor is estimated for three noise bandwidths,  $f_{BW}$  =7 kHz, 50 Hz and 10 Hz.

418 Let's get back to few computational and theoretical considerations to understand and estimate the 419 different noise sources contributing to  $\sigma_y$ . The overall  $\sigma_y$  is the quadratic sum of these noise 420 contributions that are considered as uncorrelated:  $\sigma_y = \sqrt{\sum_i \sigma_{yi}^2}$ , where *i* corresponds to: (1) the 421 thermomechanical noise, which is due to the coupling with the ambient thermal bath, (2) the readout 422 electronics' noise and (3) the phonon noise.

423  $\sigma_{y}$ , whose main origin is the thermomechanical noise, is inversely proportional to the *SNR* [28, 424 50]:

$$\sigma_y^{Th} = \frac{1}{2Q} \frac{\sqrt{\langle \theta_n^2 \rangle}}{\theta_c} = \frac{1}{2Q} \frac{1}{SNR}$$
(7)

- 426  $\langle \theta_n^2 \rangle$  is the thermomechanical deflection noise. The above expression shows that the linear range and 427 the quality factor must be as high as possible to get a stable oscillator. This noise can be estimated
- 428 through the Parseval-Plancherel theorem:  $\langle \theta_n^2 \rangle = \sqrt{\int_{f_0 \frac{\Delta f}{2}}^{f_0 + \frac{\Delta f}{2}} S_{\theta}(f) df}$ . Using the dissipation-fluctuation
- 429 theorem, the power spectral density of the thermochemical noise can be written as [51]:  $S_{\theta}(f) = 430 \quad (4\pi k_B T/Q) \kappa f_0^3/((f_0^2 f^2)^2 + (f f_0/Q)^2)$ , with:  $k_B$  the Boltzmann constant, and *T* the ambient
- 431 temperature.
- 432 When  $f \ll f_0/Q$  (i.e. when the readout bandwidth is smaller than the mechanical response time, in
- 433 other words for  $f_{BW} = 50 Hz$ ), this expression is simplified:  $\langle \theta_n^2 \rangle = \frac{2k_B TQ}{\pi \kappa f_0} \sqrt{f_{BW}}$ . Interestingly, at fast
- 434 integration time (i.e.  $f_{BW} = 7 \ kHz$ ),  $\langle \theta_n^2 \rangle = k_B T/\kappa$ , which corresponds to the equipartition energy
- 435 theorem.
- 436 Similarly,  $\sigma_y$ , whose origin is the readout electronics noise, is expressed as:

$$\sigma_{y}^{elec} = \frac{1}{2Q} \frac{\sqrt{\langle v_{n}^{2} \rangle}}{V_{0ut}}$$
(8)

- 437  $\langle v_n^2 \rangle$  is the readout electronics noise generated by the buffer circuit (see Fig. 5).
- 438 The fundamental source of noise for a thermal conductance higher than the radiation conductance
- 439 [45]) should be the phonon noise resulting from the random exchange of heat between the sensor and
- 440 the thermal bath through the mechanical anchors. At thermodynamic equilibrium, the temperature
- 441 fluctuations due to this fundamental phenomenon can be written as:

$$\sigma_{y}^{phonon} = \alpha^{2} < \overline{\Delta T} >^{2} = \begin{cases} \frac{4\alpha^{2}k_{B}T^{2}}{G_{th}} \Delta f \ if \ f_{BW} < f_{th} \\ \frac{\alpha^{2}k_{B}T^{2}}{C_{th}} if \ f_{BW} > f_{th} \end{cases}$$
(9)

442

443 with  $f_{th} = 1/4\tau_{th}$  the thermal cut-off frequency.

444

445 The direct phase or frequency fluctuations can be expressed according to a sum of frequency sources 446 with the spectral power density:  $S_y(f) = Kf^{\alpha}$  (-4 <  $\alpha$  < 2).

The orders of magnitudes of the noises and their consequence on the frequency stability and *NEP* are summarized in Tab 4 below, for the two considered bandwidths. We notice that the readout electronics noise drastically degrades the performance of such a system. In comparison, the *NEP* of a classical resistive 12µm-pitch pixel is around 30pW. This performance can be reached in principle if the electronics noise is minimized. We also mention the *NEP* for another integration time ( $f_{BW}$  = 1 *Hz*) that is rather used for gas or mass measurements. If a new readout strategy could be defined with 1s-integration time, the performance would even be better than the current bolometers.

454

455 **Table 4.** Theoretical frequency stability and NEP for the nominal pixel with our electronics and 2f-456 down-mixing readout scheme:  $R_f = 1050 / W$ ,  $\theta_c = 13.5^\circ$ ,  $V_n = 10 nV / \sqrt{Hz}$ , Q = 2500 and  $V_{out} =$ 457  $320 \mu V$ 

'	0Ξ0 μι								
	Noise cources	$f_{BW} = 50 Hz$		$f_{BW} = 7 \ kHz$				$f_{BW} = 1Hz$	
indise sources		$X_n$	$\sigma_y$	NEP	$X_n$	$\sigma_y$	NEP	NEP	
	Thermodynamic	$3.6 \times 10^{-6} rad$	$6.3 \times 10^{-9}$	6 pW	$1.2 \times 10^{-5}$ rad	$2.6 \times 10^{-8}$	25 pW	0.85 <i>pW</i>	
	Electronics	70.7 <i>nV</i>	$8.9 \times 10^{-8}$	85 pW	836.7 <i>nV</i>	$1 \times 10^{-6}$	1000  pW	12 pW	
	Phonon	-	$5,8 \times 10^{-9}$	5.5  pW		$1,8  imes 10^{-9}$	17 pW	0.8 <i>pW</i>	
- '									-

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Experimental measurements of the frequency stability were achieved to verify our assumption and think about a specific readout strategy of our electromechanical array. To this end, the frequency stability was measured in closed-loop by estimating the Allan deviation  $\sigma_A$  of the output signal according to the integration time  $\tau$ . This deviation is a usual tool to estimate the stability of oscillators [52], in particular to characterize their long-term drift. However, the main types of noises can easily



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Figure 10. Noise characterization achieved on the typical electromechanical pixel – The amplitude at 469 resonance is set at 320µV: a) Allan deviation measurement; the red hexagon indicates the frequency 470 deviation for  $f_{BW} = 7 \ kHz$  ( $\sigma_A = 3.5 \times 10^{-6}$ ) and the red disk the one for  $f_{BW} = 50 \ Hz$  ( $\sigma_A = 3 \times 10^{-6}$ ) 471  $10^{-7}$ ); a plateau is reached between 50ms and 200ms integration time ( $\sigma_A = 1.5 \times 10^{-7}$ ); beyond 200ms 472 a strong drift effect can be observed - b) Spectral power density measurement achieved on the same 473 device. The Allan deviation has a  $\tau^{-\frac{1}{2}}$  slop at short integration time corresponding to the signature of 474 a White amplitude noise. Beyond 50ms integration time, the Allan deviation presents a plateau, which 475 shows a 1/f frequency noise. These two noises can also be distinguished on the power spectral density: 476 the slope of -10 dB/decade corresponds to this plateau.

478 The Allan deviations measured in closed loop are presented in Fig 10a) for the typical pixel. Between 70  $\mu$ s and 50 ms  $\sigma_A$  drops with a  $\tau^{1/2}$  slope showing that white noise is the main 479 480 contributor in this interval. As shown in Tab. 4, this trend is mainly attributed to our readout 481 electronics, whose amplitude level was measured around  $40nv/\sqrt{Hz}$ . A plateau at  $1.5 \times 10^{-7}$ , 482 appears between 50ms and 200ms. This 1/f-noise is well above the noise floor normally set by the 483 thermomechanical white noise and phonon noises (close to few  $10^{-8}$  for the two noise sources). 484 Supplementary experiments were achieved to try to understand the origin of this 1/f-noise. In 485 particular, the Allan deviation was measured for different actuation voltages  $V_{AC}$  to increase the 486 maximum output voltage and improve the SNR. We had demonstrated that the plateau is 487 independent of the SNR. We believe that this noise floor is fully inherent in pure frequency 488 fluctuations, whose origin is not clearly identified. Similar noise signature has been reported as 489 an anomalous phase noise (APN) for flexural nanoresonators [53]. This fundamental noise would 490 only be relevant for small vibrating bodies, which is the case for the nano rods used in the pixel. 491 In first conclusion, the stability limit of our torsional resonators is set by the APN, and has to be 492 considered as the fundamental limit of our resonant sensors. Even for a 1s integration time, the 493 NEP would be stuck around 100 pW. In the discussion section below, we try to figure out this 494 issue.

495

### 496 4. Discussion

First, for the sake of clarity, the Noise Equivalent Temperature Difference (NETD), which is
basically the lowest temperature variation detectable on the scene, is computed for our devices. *NETD* is directly proportional to the *NEP* and thus to the frequency stability:

$$NETD = \frac{4F^2}{\pi A_p \Phi_{\lambda_1 \to \lambda_2} \left(\frac{\Delta L}{\Delta T}\right)_{300K_{\lambda_1 \to \lambda_2}}} \times \frac{\sigma_y}{R_f}$$
(10)

500 *F* is the optical aperture (usually F = 1),  $A_p$  is the pixel area,  $\Phi$  and  $(\Delta L/\Delta T)$  are the optical 501 transmission and the luminance variation with scene temperature around 300 K, both evaluated in 502 the  $[\lambda_1; \lambda_2]$  range. In the 8-14µm range,  $\Phi$  is usually close to one, and  $(\Delta L/\Delta T)$  is evaluated as 503 0.84W/m<sup>2</sup>/sr/K [54].

NETD was computed from the experimental Allan deviations and thermal responses with equation 10. A figure of merit,  $FOM = NETD \times \tau_{th}$ , is usually introduced to evaluate in a glance the quality of a microbolometer technology [55]. Actually this *FOM* avoids any dependence of the sensor performance to the thermal conductance. These two parameters are shown in Tab 5 for our electromechanical components and a current resistive pixel used as a reference.

509

510 Table 5. Comparison between our pixels and a classical resistive bolometer for three integration bandwidths –
 511 All electromechanical devices are set at their onset of nonlinearity

		1 III ciccuit	meenamean	ue vices ui	e set ut then	onset of nonlinearity	
Pixel	$ au_{th}$ (ms)	R <sub>f</sub> (/W <sub>inc</sub> )	$NEP$ $f_{BW}$ $= 10 Hz$ (nW)	$NEP$ $f_{BW}$ $= 50 Hz$ (nW)	$NEP f_{BW} = 7 kHz (nW)$	NETD (K) (FOM – K.ms) f <sub>BW</sub> = 10 Hz	NETD (K) (FOM – K.ms) f <sub>BW</sub> = <b>50</b> Hz
Typical	0.5	1050	0.14	0.19	2.4	1.5 (0.75)	2 (1)
Butterfly	0.8	1011	0.47	1.1	12	4.9 (3.96)	11.6 (9.28)
Thin rod	2.8	3555	1.3	3	35	13.7 (38.3)	-
Resistive pixel [1]	16	-	-	0.05	-	-	0.05 (4)

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513 A quick insight in Tab 5 shows that the NETD of our components cannot compete with the 514 temperature performance of a resistive pixel. Experimental NEP at 50Hz and 7kHz bandwidths are 515 close to values obtained through equations (6), (8) and (10) considering  $R_f = 1050 / W$  and the 516 experimental value of the electronics noise,  $40 \ nV / \sqrt{Hz}$ . The NETD of 2K at 50Hz and 1.5K at 100ms 517 – with a sub-millisecond response time for 8-12 µm incident radiation – were extracted. The usual 518 integration times correspond to the White noise region of our resonator (see Fig 10a)) and the NEP 519 can be expressed in terms a single power density of 30  $pW/\sqrt{Hz}$ . At long integration time (at 10Hz), 520 the NETD is set by the APN. In principle, improvement of the electronics or thermomechanical noises 521 will not positively influence this limit for long integration times. This analysis tells us that the readout 522 scheme per column at 7 kHz currently used for resistive bolometer is not suitable for our approach. 523 From these first conclusions, and if we look at the equation (10), few clear improvement ways

524 can be proposed:

<u>Frequency stability and matrix readout strategy</u>

526 50Hz integration bandwidth requires an improvement of the noise amplitude of our buffer 527 electronics close to the pixel. Lower amplitude noise level can be reached by using a self-528 oscillating electronics requiring only few transistors unlike PLL circuits. Moreover, our 529 electronics was realized close to the pixel but was not done through an ASIC fabricated 530 underneath the electromechanical pixels. The low-temperature fabrication process presented 531 above has already been used to manufacture resistive bolometer imagers on top of CMOS 532 circuits (ROIC) by post-process [37, 56] and it should be straightforward to reuse this approach 533 in our case. As mentioned in the introductive section of the paper, a co-integration of the readout 534 electronics at pixel level will reduce the parasitic capacitance down to a few fF and will decrease 535 the electrical noise down to a theoretical level of  $10nV/\sqrt{Hz}$  or even  $5nV/\sqrt{Hz}$ . This approach 536 makes the down-mixing detection scheme unnecessary, leading to a much simpler measurement 537 chain than the strategy presented here.  $\sigma_y$  will be decreased by a factor 8 with a self-oscillating 538 IC (gain of a factor 4 on the absolute noise, and gain of a factor 2 on the output voltage with a 539 the direct detection (see equation (8)). Thus the electronics noise will get lower than the

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fundamental *APN* ( $\sigma_y = 1.5 \times 10^{-7}$ ) for a 700Hz integration bandwidth. This conclusion leads us to suggest a new readout scheme consisting of reading 700/50=14 pixels during a 50Hz frame rate which allows a larger area for the co-integrated readout. These two straightforward improvements allow to get at a *FOM* close to 0.75 for a  $f_{BW} = 50Hz$  (global shutter approach), which is an encouraging element.

<u>Thermal response</u>

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At the end, the noise floor level will be set by the APN, whatever the electronics and the readout strategy. An improvement of the signal through the thermal insulation 1/G is much trickier in our case. Indeed this would require long and thin rods / insulations legs, and would lower the onset of nonlinearity of  $\theta_c$  (see equation (7)), leading to a degradation of the SNRs and thereby of the frequency stability  $\sigma_y$ .

552 A simple look at equation (10) demonstrates that the thermal response  $R_f$  has to be increased 553 through the TCF (the temperature coefficient of resistance of a resistive pixel is around 2%, to be 554 compared to 50ppm in our case). To date, we did not observe a major difference of the 555 experimental TCF values ranging from -35 ppm/°C to -100 ppm/°C. Unfortunately, the highest 556 TCF occurs with soft devices, which were not suitable for IR sensing as explained above.

557 We will thus focus the discussion on increasing the TCF of our devices. Some interesting 558 works have already shown a TCF up to 1000 ppm/°C, improving thereby the TCF by a factor 20 559 [57], [58]. In particular, the 1<sup>st</sup> order phase transition of diverse materials has been used to obtain 560 Young's Modulus that are highly sensitive to temperature [59], [60]. Following this line, we are 561 manufacturing similar 12µm-pitch electromechanical pixels including VO2 material on top of 562 our pixel. This material was deposited in its amorphous state by reactive deposition (Ion Beam 563 Deposition) and annealed at 400°C to get the crystalline state. The process temperature is kept 564 low enough to be used in a post-process of a CMOS circuit. Raman characterizations were done 565 to verify the crystallization obtained with this method. The resonators were designed to keep 566 the mechanical features of our current typical pixel (Fig 3c)). Nano-indentation measurements 567 were performed on full layer to extract the Young's modulus of our VO<sub>2</sub> layer (177GPa for the 568 crystalline state and 80 GPa for the amorphous state). A thickness of 80nm was then chosen with 569 1.5µm long and 300nm wide torsional rods. In a first version, both the rods and the plate are 570 covered with the VO<sub>2</sub> layer, and in a second version the VO<sub>2</sub> layer is only left on the rods. An 571 example of fabricated devices is shown in Fig 11. The TCF measurements and then the frequency 572 stability are on-going. We expect an improvement of one order of magnitude in the thermal 573 response (the mechanical features and the thermal insulation being kept constant compared to 574 the standard pixel).

#### 575 5. Conclusion

576 Our electro-optical measurements show that our current electromechanical resonant pixels cannot 577 compete with the best 12µm pitch resistive bolometer in terms of NETD. Three major straightforward 578 improvement can be done: 1) the buffer and the electronics readout including the addressing circuit 579 has to be included into a ROIC directly beneath the imager as the bolometer. Doing so, the electronics 580 noise and the parasitic capacitance will get negligible regarding the other sources of noises. The noise 581 level will be set by the APN that is the fundamental limit of such an approach. This limitation can be 582 overcome by increasing by a factor 10 or 100 the frequency response. We showed in this paper one of 583 the more promising way to reach this goal by integrating phase transition material on top of the rods. 584 The first realizations demonstrated that we were able to reproduce the same device without thermal 585 features degradations. The optimizations of the pixel is on-going. With both the improvements of the 586 frequency response of a factor 10, and a pixel wise readout or at least a readout of 14 in a same frame 587 rate, the NEP would be lower than 20 pW (i.e. NETD < 180 mK). The FOM would drop to 0.09, value 588 to be compared with the current technology (FOM=0.05). Based on this projection, we believe that the 589 uncooled IR sensors based on nanomechanical resonators will experience a new interest for small

590 pitches below 12  $\mu$ m.

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592 593

593 **Figure 11.** Torsional resonator design: a) Schematics of the design – b) SEM image of a typical pixel 594 with VO<sub>2</sub> layer on top of both the plate and torsional rods, plus partially on the insulation legs.

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