Sensors and Actuators B 179 (2013) 336–346

Contents lists available at SciVerse ScienceDirect

Sensors and Actuators B: Chemical

journal homepage: www.elsevier.com/locate/snb

Systematic design of micro-resistors for temperature control by quasi-simultaneous heating and temperature sensing

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ARTICLE INFO

Article history:
Received 21 July 2012
Received in revised form 2 October 2012
Accepted 11 October 2012
Available online 26 October 2012

Keywords:
Temperature control
MEMS
Micro-hot-plates
Resistive heaters

ABSTRACT

A single resistor can be used for quasi-simultaneously heating and temperature sensing. For instance, in microsystems this strategy is often employed for temperature control, flow sensors, gas sensors, material characterization, hot-wire anemometers, bio-MEMS and more. However, the design of such integrated resistors is complex due to both many design parameters and several conflicting specifications so that until now there is no method for predicting, based on a given set of specifications, if a solution exists and, eventually, for systematic design of optimal devices. Here we determine a complete set of relations which allows to easily find if, for given specifications, a solution is possible and, if not, to identify which specifications must be relaxed. Moreover, even with very severe specifications, our relations offer insight for systematic design of optimal resistors for quasi-simultaneous heating and temperature sensing; illustrative design cases are reported.

1. Introduction

Since most physical, chemical, and, biological properties depend on temperature, temperature control is crucial for a multitude of applications (including microelectronic devices [1], pressure sensors [2,3], implantable systems [4], chemical sensors [5–9], bio-MEMS [10,11], memories [12], material science [13,14] and, more [15]). In particular, resistive heaters are widely used in micro-hot-plates [5–7,16,17], for instance for micro-reactors and for chemical sensors which typically require high temperatures; remarkably, bulk micromachining easily allows to selectively heat specific parts of the chip, without significantly affecting the temperatures of other on-chip elements such as electronic devices, other sensors/actuators and possibly, in future, integrated energy harvesters [18–20]. For temperature control, two distinct devices can be used for, respectively, temperature sensing and heating [9,21,22]. However, it is also possible to use a single device (e.g. a resistor) for both measuring and heating, simultaneously [2,23] or quasi-simultaneously [24]. In some cases the "single-device" strategy can be very convenient or even necessary because less wires are required and therefore, the temperature-controlled region can be much more compact (e.g. silicon cantilevers with integrated and extremely small heaters for data storage, scanning microscopy based on the measurement of thermo-physical parameters, and nanoscale manufacturing [25]); as an additional advantage, the "single-device" strategy eliminates the separation, and therefore the temperature difference, between the heater and the temperature sensing device, thus zeroing one of the possible errors of temperature control systems. However, clearly, using a single device for both temperature sensing and heating is more difficult; in fact, beside requiring a more complex electronic interface, if a single device must perform both temperature measurement and heating, its characteristics must simultaneously meet all the specifications for both temperature sensing and heating, whereas the characteristics of two distinct devices could be separately optimized.

Resistors have been widely used both as temperature sensors and as heaters in temperature control systems, including with "single-device" temperature control strategies. However, in general, the design of integrated resistors and of their electronic interface is not straightforward because of several design parameters and many conflicting specifications including resistor area, electro-migration [26], temperature measurement error, self-heating during measurement, heating power, duty cycle, measurement current, heating current and resistor geometry (width, length, and thickness). In literature very few papers have discussed, and only marginally, guidelines for the design of integrated heaters; in [27] it has been shown that there may be an optimum thickness of the heat spreading plate for minimizing the time response of the heater; in [26] it has been mentioned that, when designing the heater, the heating power and electro-migration must be simultaneously taken into account and that, when designing the temperature sensor, the minimum detectable voltage change of the electronic interface, the resistance at the reference temperature, the measurement current, and the temperature coefficient of

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http://dx.doi.org/10.1016/j.snb.2012.10.065

the resistor, all contribute to the minimum detectable temperature variation; clearly, such observations are insufficient for systematic design and even for predicting, for a given set of specifications, if a solution exists. Moreover, clearly, the situation is even much more intricate for "single-device" strategy.

Here we choose an on–off temperature control strategy because of its extreme simplicity and effectiveness [21,22,24] when compared with conventional interfaces for hot-wire anemometers (e.g. see [28–30] for frequency compensation issues). Then, we consider a resistor used as both a heater and a temperature sensor by alternating temperature reading and heating cycles. Consequently, we determine a set of relations which allow to determine if a solution exists and, if not, to identify which specifications must be relaxed and how much. Moreover, we give a systematic method for designing the resistor and for setting the crucial parameters of the electronic interface [31], namely the heating and measurement currents, at values which satisfy a consistent set of given specifications.

In Section 2 we discuss the specifications and the electronic interface; in Section 3 we find auxiliary analytical relations for design. In Section 4, first, assuming the specifications are given, we determine inequalities which easily permit to verify if a solution exists and, if not, to identify which specifications must be relaxed and how much; later, we show how to design an optimal integrated resistor. Section 5 illustrates design examples. Conclusions are drawn in Section 6.

2. Specifications and electronic interface

The goal of this section is to identify and discuss all the specifications. The strategy used in the electronic interface critically affects both the accuracy and complexity of the temperature control system; in particular, the specifications (e.g. input equivalent voltage error of the electronic interface and power consumption during temperature measurement) can only be quantified after defining the interfacing strategy. Therefore, in this section, we choose the most convenient interfacing strategy and accordingly, discuss the correspondent specifications.

Since conventional, purely analog interfaces for hot-wire anemometers involve complex frequency compensation issues [28–30], we consider an on–off control temperature control strategy which, for a first order system, is always stable [21,22,24,31]. The clock period is assumed much smaller than the thermal time constant of the heater. In most practical cases this assumption is absolutely straightforward because, with typical thermal time constants, clock frequencies in the kHz range are sufficient; we however mention that, since downscaling reduces the thermal inertia, extremely small heaters may have time constants much smaller than 1 ms (e.g. [25]) and thus require higher clock frequencies and a more complex interface design (alternatively, open–loop strategies can be used, as in [25], at the heavy cost of poor robustness against disturbing signals, including environment temperature).

As to the resistor geometry, we refer to W, t, and L, as to their width, thickness, and effective length, respectively. The effective L length takes into account the different contribution of the contacts and of the corners; for instance, each 90° corner approximately corresponds to an effective number of squares equal to 0.56, so that each 90° corner of the serpentine will contribute $W \times 0.56$ to the effective length L.

We first define the following consistent set of specifications and, then, we will separately discuss all the specifications:

- minimum operating temperature, T_{MIN}, and maximum operating temperature, T_{MAX}, of the resistor,
- resistor material with given electrical resistivity, and temperature coefficient of the resistivity (though, for simplicity, we restrict to resistances with linear temperature dependences, the extension is obvious; moreover, if more materials are available, the extension is also obvious, see later),
- total resistor "top" area A,
- minimum and maximum values for the thickness (t_m and t_M, respectively) and the width (W_m and W_M, respectively) of the resistor (these values strongly depend on the available process),
- thermal resistance between the heater and the environment, R_{TH} (which is assumed to be constant for a constant top area, see later for discussion),
- input equivalent voltage error of the electronic interface [31] equal to ΔV_{ERR} (equivalent to the minimum detectable voltage change of the electronic interface in [26]),
- heating power above a minimum required value $P_{HEAT,m}$ (this specification is necessary both for allowing the heater to reach a sufficiently high temperature and for sufficiently high speed, see later for discussion),
- temperature measurement error ΔT_{ERR} below the maximum acceptable temperature measurement error $\Delta T_{ERR,M}$,
- maximum possible self-heating during temperature measurement $\Delta T_{MEAS,\max}$ below a maximum acceptable value $\Delta T_{MEAS,\max,M}$ for heaters whose minimum operating temperature is environment temperature or, alternatively, for heaters whose minimum operating temperature is much higher than environment temperature (see later), power consumption during temperature measurement, P_{MEAS}, below a maximum acceptable value $P_{MEAS,M}$,
- maximum current density (i.e. current density during heating) below a certain value J_{MAX} (for convenience, without loss of generality, this specification is separately considered even if, strictly, it could have been "included" in the resistor material specification, because a given material will have a maximum allowed current density for preventing significant electro-migration and aging),
- the maximum supply voltage V_{DD} is given (for simplicity, we neglect the voltage drop-out required by the current sources for proper operation; however, if the heating current source is designed with proper low-voltage techniques a large fraction of the supply voltage may be applied to the resistor), where we clearly refer to the maximum available DC voltage (DC voltages higher than the supply voltage can easily be generated, e.g. by charge pumps [32,33], with, of course, insurmountable limits given by the process, e.g. in a CMOS system if the voltage exceeds a certain value the gate oxide may be irreversibly damaged).

For clarity the last five specifications will be referred to as the "inequality specifications" because each of them involves an inequality; the key role of the "inequality specifications" for design will be discussed later. We also mention that though in MEMS the resistor thickness is generally a free parameter, in some cases, e.g. in standard CMOS processes, the resistor thickness is fixed; for generality we consider a variable thickness, but will also discuss the case of constant thickness (the constant thickness would be an additional specification).

2.1. Temperature measurement error

The accuracy [31] of a temperature control system can be quantified by the temperature measurement error ΔT_{ERR}, which, therefore, is a key specification; for convenience we separately consider the self-heating error, i.e. the variation of temperature induced by the measurement current (see later).

2.2. Heating power

The heating power, in combination with the thermal resistance, determines the maximum desired overheating, which is generally fixed by the application. However we stress that the heating power is also crucial for the speed of the heating process; for instance if the heating power is barely able to reach the maximum operating temperature, T_{MAX}, such an heating process would require a few thermal time constants before completing the transient.

2.3. Minimum and maximum operating temperatures

The minimum and maximum operating temperatures of the resistor are generally defined by the applications. At very high maximum temperatures, reliability issues, aging, and electro-migration phenomena will be exacerbated.

2.4. Resistor material

Here we assume that the resistor material is given; in fact, the resistor material is generally chosen based on its long-term stability (also taking into account the maximum operating temperature), high temperature coefficient of the resistivity, and ease of deposition. For instance, sputtered platinum has high stability when exposed to temperature cycles [21]. However the resistor material, besides determining the electrical resistivity at the reference temperature and the temperature coefficient of the resistivity, also defines the maximum allowed current density for preventing significant electro-migration and aging. In most literature the maximum current density refers to the value which guarantees the absence of fracture of the resistor under specified operating conditions for a given amount of time; however, for our target application, specifications should refer to the maximum allowed current density that also avoids significant aging (unfortunately, these data may be difficult to find); in the following, this maximum current density will be referred to as J_{MAX}.

We also mention that, though here we consider the resistor material as a given specification, obviously, if different materials are available, our approach can be repeated for each material in order to select the most convenient material.

For clarity, since integrated resistors are often fabricated by using thin film metals, we will consider metal resistors and will neglect the (generally weak) non-linear temperature dependence, so that the resistance R can be expressed as

$$R(T) = R_0[1 + \alpha(T - T_0)] \tag{1}$$

where R_0 is the resistance at the reference temperature T_0, and α is the temperature coefficient of R at the reference temperature T_0. With obvious modifications, though with an approximation, our analysis can be extended to the case of resistors with arbitrary temperature dependences (see discussion in Section 4).

2.5. Resistor "top" area

We also consider as a given specification the resistor "top" area WL, which is generally dictated by the given application; clearly, the total "active" area necessary for integrating the heater will be larger than WL (e.g. about twice larger for serpentine resistors with separation between different parallel tracks of the serpentine equal to the width).

2.6. Thermal resistance between the heater and the environment

The temperature differences among different parts of the heaters are typically small. In fact, typical resistor materials (e.g. platinum) have very high thermal conductivity; moreover, even in case of rather thin resistors deposited on thin membranes (e.g. micro-hot-plates require excellent thermal insulation between the heaters and bulk, so that the membrane and metal layer must be sufficiently thin) a key specification for the layout design is achieving a sufficient temperature uniformity in the active region [6,34]. Therefore, as a first approximation, we can assume the temperature across the heater is constant; with such assumption the thermal resistance, R_{TH}, between the heater and the environment can be defined as the ratio between $T_H - T_{ENV}$ and the heating power, where T_H is the heater temperature (constant along all the heater region) and T_{ENV} is the environment temperature. As to dynamic analyses, in most cases integrated resistors can be accurately described by means of first-order thermal systems (i.e. there is a single dominant pole).

The total thermal resistance between the micro-heater and the environment is almost independent on the thickness of the heater, t, because both the width, W, and the length, L, are generally much larger than the thickness, so that only negligible heat can flow through the "lateral" walls of the heater. On the contrary, most heat flows from the heater toward the substrate through the "bottom" area (equal to WL) or toward the environment through the "top" area (also equal to WL), so that, for a given "top area" WL the thermal resistance between the micro-heater and the environment is approximately constant. This result applies to both the two most important types of micro-heaters: in the first type the heater is in excellent thermal contact with the bulk (e.g. the heater is deposited on a substrate without etching for enhancing thermal insulation) due to the large thickness of the bulk and to the typically high thermal conductivity of the substrate (e.g. silicon); in the second type, such as membrane or micro-bridge type micro-hot-plates or resistively heated cantilevers, there is very high thermal insulation from the bulk.

In conclusion, since we assume the "top" heater area WL is given, we consider the thermal resistance between the resistor and the environment as a given specification.

2.7. Current-driven measurement and input equivalent voltage error

Clearly, different strategies are possible for measuring temperature dependent resistances; for instance, a reference voltage can be applied to a resistor and the resulting current may be measured; alternatively, by duality, a reference current can be injected into the resistor and the resulting current may be measured. In all cases, for a given supply voltage V_{DD}, the voltage across the resistor will obviously be lower than V_{DD}; therefore, the maximum power which can be dissipated into the resistor (i.e. the maximum heating power) will be lower than $(V_{DD})^2/R$, so that small resistance values are generally necessary in order to achieve significant heating power with reasonably low supply voltages. However, with small resistance values, series parasitic resistances may result in large errors unless 4-wires techniques [31] are used. For this reason, we restrict our attention to injecting a reference current and measuring the resulting voltage which very easily allows to take advantage of the 4-wires technique.

As we shall discuss in Section 3, the error of the resulting electronic interface can be accurately represented by an input equivalent voltage error (which will be generally dominated by the input offset and noise voltages of the instrumentation amplifier) which is, in practice, independent on both the resistance and the measurement current (for the typically small heater resistance values the errors induced by the very low input currents of typical instrumentation amplifiers and by thermal noise of the resistors will generally be negligible). In discrete electronic interfaces even assuming an ideal instrumentation amplifier, the input equivalent voltage error would still easily be in the order of tens of μV due to interferences and spurious signals; even in this case, this error will be, in practice, independent on both the resistance and the measurement current.

2.8. Current-driven heating

Similar to measuring temperature dependent resistances, there are also different strategies for heating. In particular, heating can result from both injecting a desired heating current into the resistor or, by duality, applying a desired voltage to the resistor. In some cases the heating power must be measured (e.g. thermal flow sensors); in such cases, the current-driving approach may again be more convenient because the heating power in the microheater does not depend on series parasitic resistances. Here, without loss of generality, we will consider current-driven heating.

2.9. On–off temperature control and power consumption during the temperature measurement

We consider a first-order thermal system, which is almost always an accurate approximation for micro-fabricated resistors. We choose an on–off temperature control strategy because of its simplicity and effectiveness [21,22,24], with the additional advantage that the average heating power in a certain amount of time can be estimated by simply counting the on-cycles and the off-cycles in that period, a characteristic which easily allows extracting information on the flow speed [24].

In an on–off temperature control system, when heating is required, it would be possible to measure the temperature by taking advantage of the heating current flowing through the resistor; however, when heating is not required, in order to close the control loop the resistance must still be regularly measured, which requires to inject a measurement current into the resistor. Therefore, it is necessary to have both a heating and a measurement current; it is then convenient to quasi-simultaneously use the resistor as an heater and as a temperature sensor by alternating temperature reading cycles and heating cycles; in practice, each "control period" will be divided into a temperature measurement sub-cycle and an heating sub-cycle; we will refer to the duty cycle as to the ratio between the time duration of the heating sub-cycle and the total "control period". During the temperature measurement sub-cycle, a measurement current is injected into the resistor and the resistor voltage is converted into a digital signal (e.g. instrumentation amplifier and analog to digital converter, see later), which can then be acquired by a digital system (e.g. microcontroller); a suitable current source, instrumentation amplifier, and analog to digital converter must be chosen in order to not degrade the temperature measurement accuracy. With a much simpler approach, a voltage comparator followed by a flip-flop is sufficient to close the loop [22,24].

The duty cycle d is the ratio between the time duration of the heating sub-cycle and the total "control period"; therefore, by definition, the duty cycle is smaller than 1. The duty cycle must be sufficiently large for two reasons. First, if the duty cycle is too small, for a given heating current, the *rms* value of the heating current becomes significantly smaller than the heating current itself, thus limiting the maximum overheating and the speed of the heating process. Second, for a given measurement current, if the measurement current is injected into the resistor for a small fraction of the total "control period" (i.e. duty cycle close to 1), self-heating due to the measurement current is minimized because the *rms* value of the measurement current becomes much smaller than the measurement current level.

On the other hand, the duty cycle should not be too close to 1, otherwise, since the temperature measurement time cannot be too small (for a given desired accuracy), the "control period" would become too large and, therefore, the temperature ripple would also be large. In conclusion, extreme values of the duty cycle are not convenient as very large duty cycles will result in excessive temperature ripples and very small duty cycles will result in limited and slow overheating as well as in excessive self-heating induced by the measurement current.

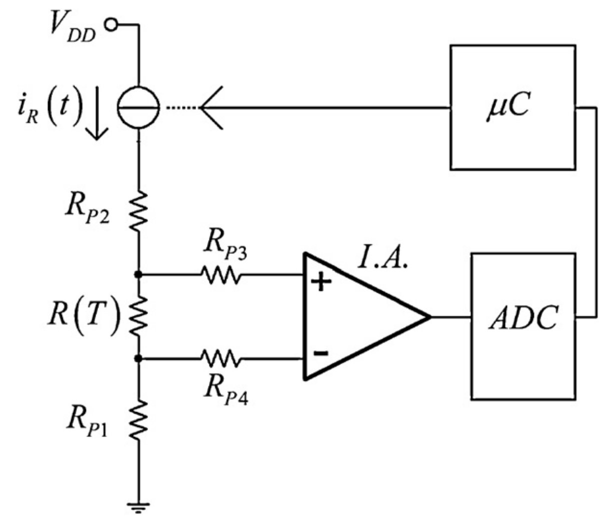

Fig. 1. A circuit with analog-to-digital converter (ADC) and microcontroller (μC) for temperature control of a resistor by simultaneous or quasi-simultaneous heating and temperature sensing; the micro-controller dynamically determines the current to be injected into the resistor.

Fig. 1 shows a general circuit for temperature control of a resistor by simultaneous or quasi-simultaneous heating and temperature sensing; the instrumentation amplifier allows to perform a 4-wires resistor measurement, thus removing the errors due to the parasitic resistors R_{P1}, R_{P2}, R_{P3}, and R_{P4}. Obviously, in a CMOS implementation, dynamic techniques must be used for compensating the input offset and low frequency noise of the instrumentation amplifier [31] and, eventually, its gain error [35–37]. The micro-controller (μC) dynamically determines the current to be injected into the resistor based on the measured resistance, which is obviously related to the measured temperature; this more general circuit can also implement an on–off temperature control strategy if the microcontroller sets the current i_R to the measurement current value during the measurement sub-cycle and to zero or to the heating current during the heating sub-cycle (depending on the comparison between the desired temperature and the measured temperature); the loop will then control the resistor temperature with an oscillating ripple amplitude that decreases when decreasing the "control period" and can therefore be made very small by choosing a sufficiently high clock frequency.

However, in many cases, a simpler implementation is possible; as an example, Fig. 2 shows a system for the temperature control of a resistor with positive temperature coefficient (e.g. a metal resistor) by quasi-simultaneous heating and temperature sensing; the comparator determines if, in the next heating cycle, the heating current will be injected into the resistor (though, strictly, an ideal current source can not be placed in series with an open switch, which is another ideal current source with zero current value, for simplicity, we use current sources in series with switches for graphically illustrating that the current sources are disabled when the correspondent switch control signal is low); the voltage generated across the heater is compared with the voltage generated across a reference resistor. As we shall discuss in Section 3, as a key advantage of this strategy, provided the current sources are very well matched (in a switched-capacitor, autozero [31] CMOS implementation a single switched current source can

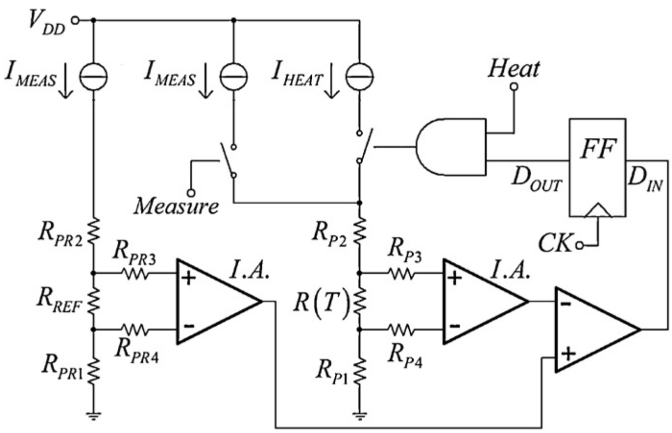

Fig. 2. A very simple circuit for temperature control of a resistor with positive temperature coefficient by quasi-simultaneous heating and temperature sensing; the comparator determines if, in the next heating cycle, the heating current will be injected into the resistor; the current sources in series with the switches graphically illustrate that the current sources are disabled when the switch control signal is low.

be used for generating both the currents to be injected into both the heater and the reference resistor, thus almost zeroing any current mismatch), the unavoidable uncertainty on the measurement current will not give significant error. We stress that, in striking contrast with other purely analog solutions [28–30], with first-order thermal system the circuit shown in Fig. 2 has no stability issues and is also very convenient for CMOS implementation, with the very few digital components occupying almost no area, consuming almost no power and not significantly affecting the accuracy. In Fig. 2 the reference resistor R_{REF} sets the desired temperature; clearly, due to the typically significant spread of resistor values, for very high accuracy both an external reference resistor (with accurate value, very low temperature coefficient, and possibly located in a region where temperature variations are small) and post-fabrication calibration may be required. If different temperatures are desired at different instants (e.g. this approach has been proposed for modulating the sensitivities of chemical sensors on micro-hot-plates) either different reference resistors should be used, either an electronic tuning can be implemented (e.g. the ratio between the currents injected into the heater $R(T)$ and into the reference resistor can be dynamically changed, similar to [22]).

Obviously, the overheating induced by the measurement operation should be very small. Though this consideration is general, it is useful to distinguish between heaters whose minimum operating temperature is environment temperature (or very close to environment temperature) and heaters whose minimum operating temperature is much higher than environment temperature. In fact, in the first case, the specification on small overheating induced by the measurement is generally given in terms of the maximum acceptable self-heating, $\Delta T_{MEAS,\max,M}$ because any self-heating is undesired and it is then essential to keep it below a safe limit; clearly, $\Delta T_{meas,\max}$ is equal to $P_{MEAS,M}R_{TH}$. In the second case, it is necessary that, when the measured temperature is above the desired temperature (which, however, is much higher than the environment temperature) and, therefore, the resistor must cool down, the heat flowing from the (excessively hot) resistor toward the environment be significantly higher than the Joule heat generated inside the resistor during the resistance measurement, so that the resistor can effectively cool down during a time period with disabled heating; this is verified if

$$P_{MEAS,M}R_{TH} \ll (T_{MIN} - T_{\max,environment}) \tag{2}$$

As a result, in both cases the specification corresponds to requiring that $P_{MEAS,M}R_{TH}$ be much smaller than a certain value ($\Delta T_{MEAS,\max,M}$ in the first case, ($T_{MIN} - T_{\max,environment}$) in the second case). Therefore, in general, since we have assumed that R_{TH} is known, this specification can be expressed in terms of the maximum acceptable value for the product of the thermal resistance and the power during measurement or, equivalently, in terms of the maximum acceptable power during measurement. For clarity, in the rest of the paper we will consider heaters whose minimum operating temperature is environment temperature (or very close to environment temperature); there would be no changes in dealing with heaters whose minimum operating temperature is much higher than environment temperature.

3. Auxiliary relations

In this section we determine auxiliary analytical relations which, in next section, will be used for design.

3.1. Measurement voltage drop and equivalent input voltage error of the interface

Let us define the measurement voltage drop, V_{MEAS}, as the voltage drop across the resistor during the temperature measurement sub-cycle; analogously, we can refer to the reference measurement voltage drop, $V_{MEAS,0}$, as to the value of the measurement voltage drop V_{MEAS} at the reference temperature T_0. The reference measurement voltage drop, $V_{MEAS,0}$, is simply equal to R_0I_{MEAS} (the measurement current, I_{MEAS}, ideally, does not depend on temperature).

The temperature measurement error ΔT_{ERR} is related to the error in the resistance measurement, ΔR_{ERR}, by the relation

$$\Delta T_{ERR} = \frac{\Delta R_{ERR}}{dR/dT} = \frac{\Delta R_{ERR}}{R_0\alpha} \tag{3}$$

In general, since $R = V_{MEAS}/I_{MEAS}$, the error in the resistance measurement, ΔR_{ERR}, is given by

$$\frac{\Delta R_{ERR}}{R} = \frac{\Delta V_{MEAS}}{V_{MEAS}} - \frac{\Delta I_{MEAS}}{I_{MEAS}} \tag{4}$$

where, in an hypothetic circuit constituted by an ideal current source I_{MEAS} injecting current into the resistor and by additional devices (e.g. instrumentation amplifier followed by an analog-to-digital converter) for measuring the voltage across the resistor, ΔI_{MEAS} would be the error in setting the measurement current I_{MEAS} and ΔV_{MEAS} would be the error in measuring the measurement voltage drop V_{MEAS}. However, the strategy illustrated in Fig. 2 compares two voltages generated by injecting two nominally identical currents (I_{MEAS}) in two resistors (we assume there is zero error in the reference resistor), so that the error in I_{MEAS} is automatically zeroed, as long as both the nominally identical currents have the same value. Therefore, by using the same current source for both the resistors, the relative uncertainty in the measurement current will give a negligible contribution to $\Delta R_{ERR}/R$; for instance, in a CMOS implementation switched low-voltage cascode current mirrors could be used, with low-voltage cascode techniques for minimizing the errors due to channel length modulation without requiring excessively high supply voltages; clearly, the current switching in the dynamic current mirror must be properly synchronized with the autozeroing process for the instrumentation amplifiers and the comparator. The error in measuring the measurement voltage drop V_{MEAS} is the equivalent input voltage error of the interface, ΔV_{ERR}, which is generally dominated by the input equivalent error of the instrumentation amplifier, so that

$$\frac{\Delta R_{ERR}}{R} \simeq \frac{\Delta V_{MEAS}}{V_{MEAS}} = \frac{\Delta V_{ERR}}{RI_{MEAS}} \tag{5}$$

Intuitively, since the input voltage error ΔV_{ERR} is added to the *measurement voltage drop* V_{MEAS}, in order to guarantee a small temperature measurement error, $V_{MEAS} = RI_{MEAS}$ must be much higher than ΔV_{ERR}, which corresponds to requiring high values for the resistance R_0 and for the measurement current I_{MEAS}.

Taking into account both (1) and (5), the temperature measurement error ΔT_{ERR} can therefore be expressed as

$$\Delta T_{ERR} = \frac{\Delta R_{ERR}}{dR/dT} = \frac{\Delta R_{ERR}}{R_0\alpha} = \frac{\Delta V_{ERR}}{R_0\alpha I_{MEAS}} \tag{6}$$

The reference measurement voltage drop is then

$$V_{MEAS,0} = R_0 I_{MEAS} = \frac{\Delta V_{ERR}}{\alpha \Delta T_{ERR}} \tag{7}$$

3.2. Heating

Since the heater can generally be modeled as first order low pass filter and we assumed the thermal time constant of the heater is sufficiently larger than the period, only the average (over the entire "control period") self-heating power is important (rather than the peak self-heating power, which occurs during the temperature measurement sub-cycle). The average (over the entire "control period") measurement power is equal to

$$P_{MEAS} = RI_{MEAS}^2(1-d) \tag{8}$$

where the duty cycle d is the ratio between the time duration of the heating sub-cycle and the total "control period" (see Section 2.9). The power P_{MEAS} will obviously result in an undesired self-heating equal to

$$\Delta T_{MEAS} = P_{MEAS} R_{TH} = RI_{MEAS}^2(1-d)R_{TH} \tag{9}$$

If, for simplicity, we neglect the temperature dependence of the thermal resistance between the microheater and the environment (if the dependence on temperature of the thermal resistance is known the extension would be obvious), the maximum self-heating will be

$$\Delta T_{MEAS,\max} = P_{MEAS,\max} R_{TH} = R_{\max} I_{MEAS}^2(1-d)R_{TH} \tag{10}$$

where, as above discussed, the measurement current does not depend on the resistor temperature, and $R_{\max}$ is the maximum resistance value over the temperature range of interest. The maximum power during measurement is then

$$P_{MEAS,\max} = \frac{\Delta T_{MEAS,\max}}{R_{TH}} = R_{\max} I_{MEAS}^2(1-d) \tag{11}$$

From (1), $R_{\max}$ can be written as $a_{\max}R_0$, where $a_{\max}$ is a positive real coefficient, equal to $[1+\alpha(T_{MAX}-T_0)]$ for PTC (positive temperature coefficient) resistors or to $[1+\alpha(T_{MIN}-T_0)]$ for NTC (negative temperature coefficient) resistors, where T_{MIN} and T_{MAX} are, respectively, the minimum and maximum operating temperatures; clearly, the modifications for resistances showing an arbitrary temperature dependence would be obvious.

Similarly, the average (over the entire "control period") heating power during the heating sub-cycle is

$$P_{HEAT} = RI_{HEAT}^2 d \tag{12}$$

and will be minimum and maximum in correspondence of the minimum and maximum, respectively, values of the resistance.

4. Design

Here, starting from the "inequality specifications" (see the last five entries in the list of specifications given in Section 1) we find inequalities which can be easily used by designers for determining if a practical solution exists and, in case no solutions are possible, for understanding which specifications must be somehow relaxed and how much. Afterwards, we describe a method for systematic design of the integrated resistors.

4.1. Determination of the design inequalities

Starting from the "inequality specifications" listed in the introduction we now deduce inequalities which can be used for design.

First, taking into account that the heating power must be larger than $P_{HEAT,m}$ and that the current density during heating must be smaller than J_{MAX} we find

$$I_{HEAT}^2 R_{\min} d \geq P_{HEAT,m} \Rightarrow I_{HEAT} \geq \sqrt{\frac{P_{HEAT,m}tW}{d\rho_0 a_{\min} L}} = \sqrt{\frac{P_{HEAT,m}tW^2}{d\rho_0 a_{\min} A}} = k_A\sqrt{t}W \tag{13}$$

where $a_{\min}$ is equal to $[1+\alpha(T_{MIN}-T_0)]$ for PTC resistors or to $[1+\alpha(T_{MAX}-T_0)]$ for NTC resistors, ρ_0 is the resistivity at the temperature T_0, and k_A is a constant univocally set by the specifications; we stress that if resistor with an arbitrary temperature dependence are considered, it is still possible to define the parameter $a_{\min}$ (and $a_{\max}$, see later) as the ratios between the minimum (maximum) resistance values in the temperature range of interest and the resistance at the reference temperature, so the extension to arbitrary temperature dependence is obvious apart the error which would be introduced in (6) (the derivative of the resistance with respect to temperature would not be constant).

Obviously, we also have

$$I_{HEAT} \leq J_{MAX}Wt = k_B Wt \tag{14}$$

where k_B is a constant univocally set by the specifications.

Moreover, from (13), taking into account that the current density must be smaller than J_{MAX} we get

$$J_{MAX}Wt \geq I_{HEAT} \Rightarrow t \geq \frac{I_{HEAT}}{J_{MAX}W} \geq \frac{k_A\sqrt{t}W}{J_{MAX}W} \Rightarrow t \geq \frac{k_A^2}{J_{MAX}^2} \tag{15}$$

which may be re-written as

$$t \geq \frac{P_{HEAT,m}}{d\rho_0 a_{\min} A J_{MAX}^2} = k_C \tag{16}$$

where k_C is a constant univocally set by the specifications.

Then, taking into account that ΔT_{ERR} must be smaller than $\Delta T_{ERR,M}$, we find

$$I_{MEAS} \geq \frac{\Delta V_{ERR}}{\alpha R_0 \Delta T_{ERR,M}} = \frac{\Delta V_{ERR} W^2 t}{\alpha \rho_0 A \Delta T_{ERR,M}} = k_D W^2 t \tag{17}$$

Afterwards, taking into account that $\Delta T_{MEAS,\max}$ must be smaller than $\Delta T_{MEAS,\max,M}$ we find

$$\Delta T_{MEAS,\max,M} \geq R_0 a_{\max} I_{MEAS}^2(1-d)R_{TH} \tag{18}$$

which gives

$$\Delta T_{MEAS,\max,M} \geq R_0 a_{\max} I_{MEAS}^2(1-d)R_{TH}$$
$$I_{MEAS} \leq \sqrt{\frac{\Delta T_{MEAS,\max,M}}{R_0 a_{\max}(1-d)R_{TH}}} = \sqrt{\frac{\Delta T_{MEAS,\max,M}}{\rho_0 A a_{\max}(1-d)R_{TH}}} \times W\sqrt{t} = k_E W\sqrt{t} \tag{19}$$

where k_D is a constant univocally set by the specifications.

From (18) we also have

$$\Delta T_{MEAS,\max,M} \geq R_0 a_{\max} I_{MEAS}^2(1-d)R_{TH} = V_{MEAS,0} a_{\max} I_{MEAS}(1-d)R_{TH} \tag{20}$$

Since ΔT_{ERR} must be smaller than $\Delta T_{ERR,M}$ or, equivalently (see (7)),

$$V_{MEAS,0} \geq \frac{\Delta V_{ERR}}{\alpha \Delta T_{ERR,M}} \tag{21}$$

we can rewrite (18) as

$$I_{MEAS} \leq \frac{\alpha \Delta T_{MEAS,\max,M} \Delta T_{ERR,M}}{a_{\max}(1-d) R_{TH} \Delta V_{ERR}} = k_F \tag{22}$$

where k_F is a constant univocally set by the specifications.

Finally, the voltage drop across the resistor supply voltage must be smaller than the supply voltage V_{DD} (in reality, the voltage which can be applied to the resistor must be somewhat inferior to the supply voltage V_{DD} but, with proper low-voltage techniques this difference can be rather small, e.g. around 0.5 V for a cascode current mirror, eventually with low-voltage gain boosting for further increasing the output resistance to extremely high values); this condition is satisfied if V_{DD} is larger than the maximum voltage drop across the resistor, that is (the heating current is the maximum current)

$$R_{\max} I_{HEAT} \leq V_{DD,R} \Rightarrow \frac{\rho_0 a_{\max} L}{Wt} I_{HEAT} \leq V_{DD,R} \tag{23}$$

which can be re-written as

$$I_{HEAT} \leq \frac{V_{DD} W^2 t}{\rho_0 a_{\max} A} = k_G W^2 t \tag{24}$$

where k_G is a constant univocally set by the specifications.

4.2. Identification of the realistic domain (W,t)

Obviously, there may be technological limits (strongly dependent on the available process) on both the minimum and maximum values of both the thickness and the width of the resistor. For instance, the minimum thickness required for achieving an electrically continuous thin Pt film depends on the type of deposition as well as on the deposition parameters [38]; moreover, at low thickness, the electrical resistivity may be strongly affected by grain size and roughness [38].Besides the technological limits, the design inequalities further restrict the range of possible values for W and t. In fact, since

$$k_F \geq I_{MEAS} \geq k_D W^2 t \tag{25}$$

and

$$t \geq k_C \tag{26}$$

we have

$$k_F \geq I_{MEAS} \geq k_D W^2 t \geq k_D W^2 k_C \Rightarrow W \leq \sqrt{\frac{k_F}{k_C k_D}} \tag{27}$$

Moreover, since the resistivities of typical heater materials are very low the length is generally larger than the width of the resistor, or equivalently

$$W \leq L \Leftrightarrow W^2 \leq LW = A \Leftrightarrow W \leq \sqrt{A} \tag{28}$$

Therefore, taking into account the technological limits, (26)–(28) we find the range of possible values for (W,t), which will be referred to as the realistic domain (W,t) and is

$$\begin{cases} W_m \leq W \leq \min\left(W_M, \sqrt{\dfrac{k_F}{k_C k_D}}, \sqrt{A}\right) \\ \max(k_C, t_m) \leq t \leq t_M \end{cases} \tag{29}$$

where A, k_C, k_D and k_F are constants univocally determined by the specifications, as evident from (16), (17), and (22). If one between the inequalities in (29) is not satisfied, there is no solution and another process must be considered or the specifications must be somehow relaxed by taking into account (29) and the definitions of k_C, k_D, and k_F (see Section 5 for a practical example).

4.3. Determination of the measurement current

Clearly, each point in the realistic domain (W,t) univocally determines both the length L (because of the specification $A = WL$) and the resistance at the reference temperature (because $R_0 = \rho_0(L/tW)$), so that only the measurement current and the heating current must still be determined.

Taking advantage of standard numerical computing software, after discretization of the realistic domain (W,t), for each point of this domain, we may compute the upper and lower limit for the measurement current using

$$\begin{cases} I_{MEAS} \geq \dfrac{\Delta V_{ERR} W^2 t}{\alpha \rho_0 A \Delta T_{ERR,M}} = k_D W^2 t \\ I_{MEAS} \leq \sqrt{\dfrac{\Delta T_{MEAS,\max,M}}{\rho_0 A a_{\max}(1-d) R_{TH}}} \times W\sqrt{t} = k_E W \sqrt{t} \\ I_{MEAS} \leq \dfrac{\alpha \Delta T_{MEAS,\max,M} \Delta T_{ERR,M}}{a_{\max}(1-d) R_{TH} \Delta V_{ERR}} = k_F \end{cases} \tag{30}$$

By inspection of (30), we also see that a value for the measurement current which satisfies all the specifications can be found if and only if

$$\begin{cases} k_F \geq k_D W^2 t \Leftrightarrow W^2 t \leq \dfrac{k_F}{k_D} \Leftrightarrow W^2 t \leq \dfrac{\Delta T_{MEAS,\max,M} \rho_0 A}{a_{\max}(1-d) R_{TH}} \left(\dfrac{\alpha \Delta T_{ERR,M}}{\Delta V_{ERR}}\right)^2 \\ k_E W \sqrt{t} \geq k_D W^2 t \Leftrightarrow W\sqrt{t} \leq \dfrac{k_E}{k_D} \Leftrightarrow W^2 t \leq \left(\dfrac{k_E}{k_D}\right)^2 = \dfrac{\Delta T_{MEAS,\max,M} \rho_0 A}{a_{\max}(1-d) R_{TH}} \left(\dfrac{\alpha \Delta T_{ERR,M}}{\Delta V_{ERR}}\right)^2 \end{cases} \tag{31}$$

As evident the inequalities (31) are identical, a result which is not surprising because, starting from the five "inequality specifications", we have found the seven inequalities which define the constant k_A, k_B, k_C, k_D, k_E, k_F, and k_G, respectively. Though, therefore, the seven inequalities are not independent and the same information could, of course, be more concisely expressed by the five "inequality specifications" only, increasing the number of inequalities is instrumental to gain insight for design and, in particular, will allow: to restrict the domain (W,t) to possible values only; to identify, if a solution does not exist, which specifications must be relaxed or, if a solution exists, to compute possible values of the measurement and heating current; to optimize the design.

Afterwards, taking into account the definition of the realistic domain (29), the last two inequalities define the portion, if any, of the realistic domain (W,t) which allows to find measurement current values meeting all the specifications.

4.4. Determination of the heating current

For each point of the realistic domain (W,t) we may compute the upper and lower limit for the heating current using

$$\begin{cases} I_{HEAT} \geq \sqrt{\dfrac{P_{HEAT,m}tW^2}{d\rho_0 a_{\min}A}} = k_A\sqrt{t}W \\ I_{HEAT} \leq J_{MAX}Wt = k_BWt \\ I_{HEAT} \leq \dfrac{V_{DD}W^2t}{\rho_0 a_{\max}A} = k_GW^2t \end{cases} \tag{32}$$

By inspection of these inequalities we see that a value for the heating current which satisfies all the specifications can be found if and only if

$$\begin{cases} k_BWt \geq k_A\sqrt{t}W \Leftrightarrow t \geq \left(\dfrac{k_A}{k_B}\right)^2 = \dfrac{P_{HEAT,m}}{d\rho_0 a_{\min}AJ^2_{MAX}} \\ k_GW^2t \geq k_A\sqrt{t}W \Leftrightarrow W^2t \geq \left(\dfrac{k_A}{k_G}\right)^2 = \dfrac{P_{HEAT,m}\rho_0 Aa^2_{\max}}{V^2_{DD}da_{\min}} \end{cases} \tag{33}$$

where only the second inequality is important and defines the portion, if any, of the realistic domain (W,t) which allows to find heating current values which meet all the specifications (the first inequality is equivalent to (16) and has therefore already been taken into account when defining the realistic domain (W,t)).

As a result, by considering (31) and (33) we find the portion of the realistic domain (W,t) which allow to find both measurement current values and heating current values which meet all the specifications if and only if

$$\left(\frac{k_A}{k_G}\right)^2 \leq \frac{k_F}{k_D} \Leftrightarrow \frac{P_{HEAT,m}a^3_{\max}}{V^2_{DD}a_{\min}} \leq \frac{\Delta T_{MEAS,\max,M}d}{R_{TH}(1-d)}\left(\frac{\alpha\Delta T_{ERR,M}}{\Delta V_{ERR}}\right)^2 \tag{34}$$

Provided that the realistic domain (W,t) above defined (29) is not empty, the last inequality is the necessary and sufficient condition for the existence of possible solutions. Moreover, if there is no possible solution, from (34) we may also identify which specification can be relaxed and how much. For instance if the second term of the inequality (34) is smaller than the first term, no solution is possible; however, if the supply voltage is sufficiently increased a solution may become possible.

For clarity, we will refer to the portion of the realistic domain (W,t) which satisfy (29), (31), and (33) as to the design-domain (W,t). For each point of the design-domain (W,t) the designer may compute the inferior and superior limit of both the measurement current (from (30)) and the heating current (from (32)). Within such limits, the design values of the measurement and heating currents can then be chosen arbitrarily and this degree of freedom can be used for a more aggressive or optimized design. As an example, for a given point of the design-domain (W,t), which correspond to a well-defined value of the resistance (because the resistor top area and the resistivity are specifications), increasing the measurement current results, ceteris paribus, in a reduction of the temperature measurement error because the ratio between the measurement voltage drop and ΔV_{ERR} is increased; similarly, if the heating current is increased, the heating will be faster.

5. Design examples

In this section we deliberately start by choosing such severe specifications that no solution exists. Then we show that our approach, beside easily finding that solutions are impossible, also allows to identify which changes in the specifications may make the problem solvable; in fact, as an illustration of the power of the method, we gradually modify the specifications so that, after eliminating all the obstacles, the problem becomes barely solvable (i.e. a hypothetical design by trial and error would be extremely difficult), thus showing that the proposed approach easily allows to identify a tight solution domain. Finally, we discuss how an optimum (i.e. more aggressive than the initial specifications) may be found by simply iterating the methodology after refining, at every step, one or more specifications.

First, let us assume that in the available MEMS process, due to technological limits, the width W and the thickness t of the integrated resistors must satisfy

$$\begin{aligned} 10\ \text{nm} \leq t \leq 1\ \mu\text{m} \\ 1\ \mu\text{m} \leq W \end{aligned} \tag{35}$$

Furthermore, let us consider the following (deliberately severe) specifications:

- duty cycle d equal to 0.9,
- reference temperature $T_0 = 20\,°\text{C} = 293.15\,\text{K}$,
- minimum temperature $T_{MIN} = 300\,°\text{C} = 573.15\,\text{K}$,
- maximum temperature $T_{MAX} = 500\,°\text{C} = 773.15\,\text{K}$,
- platinum as the resistor material, with $\alpha = 3920\,\text{ppm K}^{-1}$, electrical resistivity at the reference temperature $\rho_0 = 1.06 \times 10^{-7}\,\Omega\,\text{m}$, maximum current density $J_{MAX} = 10^{10}\,\text{A/m}^2 = 10^6\,\text{A/cm}^2$,
- input equivalent voltage error of the electronic interface $\Delta V_{ERR} = 50\,\mu\text{V}$,
- maximum acceptable temperature measurement error $\Delta T_{ERR,M} = 0.25\,\text{K}$,
- maximum acceptable self-heating during temperature measurement $\Delta T_{MEAS,\max,M} = 0.25\,\text{K}$,
- minimum acceptable value for the heating power $P_{HEAT,m} = 0.5\,\text{W}$.

With these values, if we consider a resistor area $A_1 = 100\,\mu\text{m} \times 100\,\mu\text{m}$ and a thermal resistance between the heater and the environment $R_{\text{TH}} = 200{,}000K/W$, when checking if there is a realistic domain (29) we find $t \geq k_C \simeq 2.5\,\mu\text{m}$ which is not compatible with (35), so that a solution is impossible.

However, it is easy to identify which changes in the specifications can make a solution possible. For instance, by inspection of (16) we see that k_C is inversely proportional to the top area of the resistor and therefore can be reduced by increasing the top area; for instance, if we can consider $A_2 = 1000\,\mu\text{m} \times 1000\,\mu\text{m} = 1\,\text{mm}^2$ and a thermal resistance $R_{\text{TH}} = 2000K/W$ (the thermal resistance can be approximately considered as inversely proportional to the heater area), we find a non-empty realistic domain (in particular, we find $t \geq k_C \simeq 25\,\text{nm}$ which is now compatible with (35)). With these values, if we now consider a supply voltage V_{DD} equal to 3.5 V we find that (34) is not verified because

$$\frac{k_F/k_D}{(k_A/k_G)^2} \simeq \frac{1.76}{1.9} < 1 \tag{36}$$

so that no portion of the domain (W,t) allows to find values for the measurement and heating currents which meet all the specifications.

However, by considering the relations above found for k_A, k_D, k_F, and k_G we see that increasing the supply voltage may allow to satisfy (34) (k_G will be proportionally increased, whereas k_A, k_D, and k_F will stay constant); in fact, if we consider $V_{DD} = 5\,\text{V}$ we find

$$\frac{k_F/k_D}{(k_A/k_G)^2} \simeq \frac{1.76}{0.93} > 1 \tag{37}$$

which is now correct.

Therefore, we can now compute, for each point of the design domain, the minimum and maximum values of both the measurement and heating current. For clarity we consider as the domain for

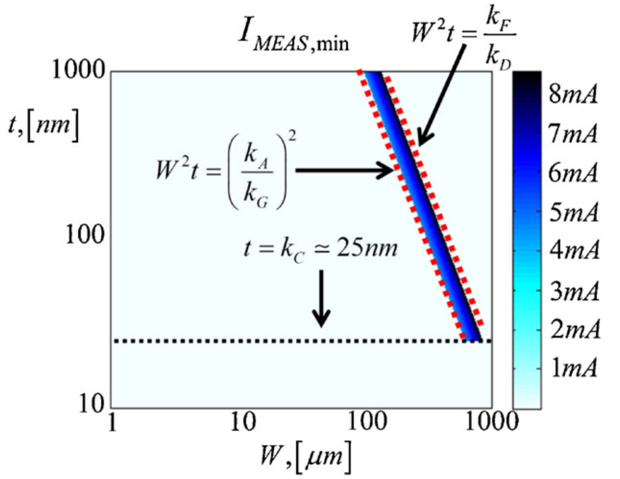

Fig. 3. Minimum possible value of the measurement current for the example discussed in Section 5.

the plot (taking into account the process limitations (35) and that W is smaller than the square root of the resistor area (28)).

$$\begin{aligned} &10\,\text{nm} \le t \le 1\,\mu\text{m} \\ &1\,\mu\text{m} \le W \le 1\,\text{mm} \end{aligned} \tag{38}$$

Then, we restrict the domain to the design domain by excluding the portions of (W,t) which do not satisfy (29).

Then, we compute, for each point of the design domain the minimum and the maximum values of the measurement current, which, according to (30), are

$$\begin{cases} I_{MEAS,\min} = \dfrac{\Delta V_{ERR} W^2 t}{\alpha \rho_0 A \Delta T_{ERR,M}} = k_D W^2 t \\ I_{MEAS,\max} = \min(k_E W \sqrt{t},\, k_F) \end{cases} \tag{39}$$

Fig. 3 shows the minimum measurement current $I_{MEAS,\min}$ for each point of the design domain and clearly shows the limits to the design domain.

Fig. 4 shows the maximum measurement current $I_{MEAS,\max}$.

Figs. 3 and 4 are very similar, which is a clear indication that the inequalities (39) are very tight (due to the severe specifications; by gradually relaxing the specifications, the values of the minimum and maximum measurement currents become more and more different).

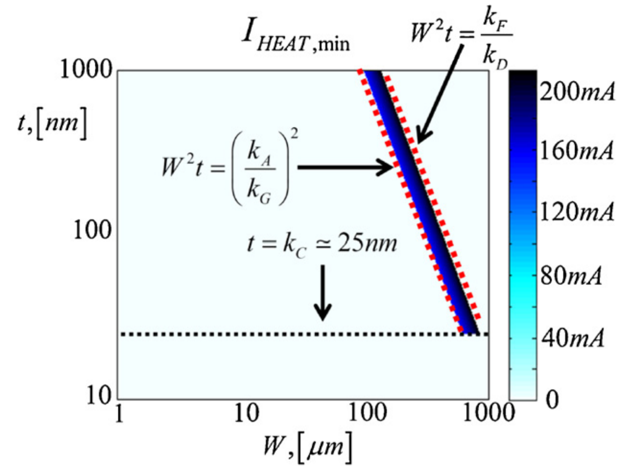

Fig. 5. Minimum possible value of the heating current for the example discussed in Section 5.

Afterwards, we compute, for each point of the design domain, the minimum and the maximum values of the heating current, which, according to (32), are

$$\begin{cases} I_{HEAT,\min} = \sqrt{\dfrac{P_{HEAT,m} t W^2}{d \rho_0 a_{\min} A}} = k_A \sqrt{t} W \\ I_{HEAT,\max} = \min\left(J_{MAX} W t,\, \dfrac{V_{DD} W^2 t}{\rho_0 a_{\max} A}\right) = \min(k_B W t,\, k_G W^2 t) \end{cases} \tag{40}$$

Fig. 5 shows the minimum heating current $I_{HEAT,\min}$.

Fig. 6 shows the maximum heating current $I_{HEAT,\max}$. Similar to Figs. 3–6 are also very similar due to the severe specifications.

If we now choose a point of the design domain, for instance $W = 350\,\mu\text{m}$, $t = 100\,\text{nm}$, we can arbitrarily choose both the measurement and heating current

$$\begin{cases} 5.96\,\text{mA} \le I_{MEAS} \le 7.11\,\text{mA} \\ 176\,\text{mA} \le I_{HEAT} \le 202\,\text{mA} \end{cases} \tag{41}$$

We stress that, though the problem is still barely solvable (the inequalities (41) are found by taking into account the nominal values of many parameters which, however, may have relatively large spread), our approach quickly allows to determine the possible solutions. We also observe that an extended solution domain can be

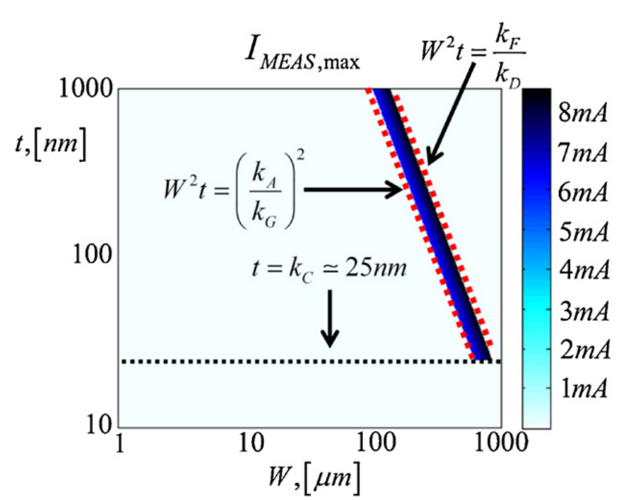

Fig. 4. Maximum possible value of the measurement current for the example discussed in Section 5.

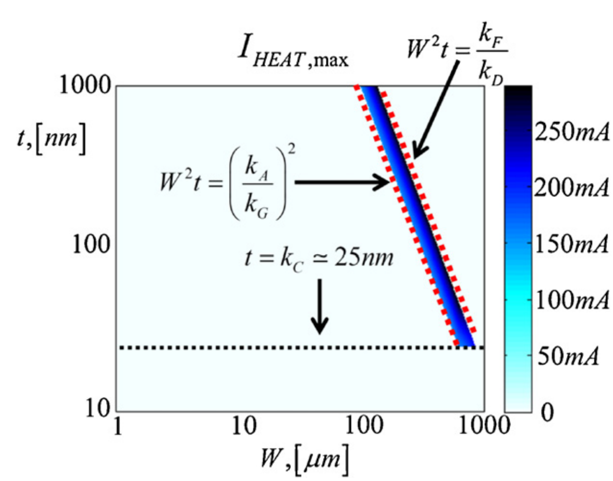

Fig. 6. Maximum possible value of the heating current for the example discussed in Section 5.

obtained by increasing the distance, in the (W,t) domain between the borders of the design domain.

As a validation test, we may verify that if we choose values of the measurement and heating currents which satisfy the inequalities (41), all the specifications can be met. If, for instance, we choose

$$I_{MEAS} = 7\,\text{mA}, \qquad I_{HEAT} = 200\,\text{mA} \tag{42}$$

we find

$$\begin{aligned} & L \simeq 2.85\,\text{mm}, \quad R_0 = 8\,\Omega, \\ & P_{HEAT} \geq 0.65\,\text{W} > P_{HEAT,m} = 0.5\,\text{W} \\ & \Delta T_{ERR} \leq 0.22\,^{\circ}\text{C} < 0.25\,^{\circ}\text{C} \\ & \Delta T_{MEAS} \leq 0.24\,^{\circ}\text{C} < 0.25\,^{\circ}\text{C} \\ & J \leq 5.7 \times 10^{9}\,\text{A/m}^2 < 10^{10}\,\text{A/m}^2 \\ & V_{DROP} \leq 4.93\,\text{V} < 5\,\text{V} \end{aligned} \tag{43}$$

which satisfy all the specifications (for simplicity we neglect the minimum voltage headroom requirements of the current sources, which can obviously be easily taken into account).

Finally, we observe that if the specifications are less severe, much larger solution domains will be found; in these cases a designer can iterate the approach after changing the specifications for a more aggressive design (e.g. reduce the temperature measurement error for higher accuracy or increase $P_{HEAT,m}$ for faster heating), thus allowing to optimize the design toward a well-defined direction, e.g. reducing the maximum temperature measurement error.

6. Conclusions

Temperature control is crucial for microelectronic devices, pressure sensors, implantable systems, chemical sensors, bio-MEMS, memories, material science and more. For temperature control, two distinct devices can be used for, respectively, temperature sensing and heating, but it is also possible to quasi-simultaneously use a single resistor for both measuring and heating. Though the "single-resistor" strategy requires less wires and is ideally suited for temperature control of very small regions, the design becomes more difficult due to many design parameters and conflicting specifications, so that until now it was impossible to predict if, for a given set of specifications, a solution exists and no systematic method for design has been reported.

Here we determine a complete set of analytical relations which allows to easily find if, for a given set of specifications, a solution is possible and, if not, to identify which specifications must be relaxed and how much. Moreover, we demonstrate that these relations, even with very severe specifications, provide the insight required for optimized design of integrated resistors suitable for temperature control by quasi-simultaneous heating and temperature sensing.

Acknowledgments

This research has been supported by the Italian Institute of Technology (IIT - Project Seed API NANE, "Arrays of Piezoelectric Nanowires for Energy Harvesting"). Thanks to Francesco Gatta and Ivan Pini for useful discussions.

I wish to express my deepest gratitude to Professor Arnaldo D'Amico; he gave me the opportunity to conduct research on analog electronics, sensors, and micro-nano-systems; his passion for multi-disciplinarity is an inspiration and his enthusiasm for research is an example.

References

[1] R. McGowen, C. Poirier, C. Bostak, Power and temperature control on a 90-nm Itanium family processor, IEEE Journal of Solid-State Circuits 41 (2006) 229–237.
[2] C.H. Mastrangelo, R.S. Muller, Microfabricated thermal absolute-pressure sensor with on-chip digital front-end processor, IEEE Journal of Solid-State Circuits 26 (1991) 1998–2007.
[3] J. Mitchell, G. Lahiji, K. Najafi, An improved performance poly-Si Pirani vacuum gauge using heat-distributing structural supports, IEEE Journal of Microelectromechanical Systems 17 (2008) 93–102.
[4] Y. Kakubari, F. Sato, H. Matsuki, T. Sato, Temperature control of SMA artificial anal sphincter, IEEE Transactions on Magnetics 39 (2003) 3384–3386.
[5] S. Semancik, R.E. Cavicchi, M.C. Wheeler, J.E. Tiffany, G.E. Poirier, R.M. Walton, J.S. Suehle, B. Panchapakesan, D.L. De Voe, Microhotplate platforms for chemical sensor research, Sensors and Actuators B: Chemical 77 (2001) 579–591.
[6] M. Afridi, J. Suehle, M. Zaghloul, A monolithic CMOS microhotplate-based gas sensor system, Sensors Journal 2 (2002) 644–655.
[7] T.a. Kunt, T.J. McAvoy, R.E. Cavicchi, S. Semancik, Optimization of temperature programmed sensing for gas identification using micro-hotplate sensors, Sensors and Actuators B: Chemical 53 (1998) 24–43.
[8] V. Guidi, G. Carlo Cardinali, L. Dori, G. Faglia, M. Ferroni, G. Martinelli, P. Nelli, G. Sberveglieri, Thin-film gas sensor implemented on a low-power-consumption micromachined silicon structure, Sensors and Actuators B: Chemical 49 (1998) 88–92.
[9] D. Barrettino, M. Graf, W. Song, Hotplate-based monolithic CMOS microsystems for gas detection and material characterization for operating temperatures up to 500 C, IEEE Journal of Solid-State Circuits 39 (2004) 1202–1207.
[10] E. Lauwers, J. Suls, W. Gumbrecht, A CMOS multiparameter biochemical microsensor with temperature control and signal interfacing, IEEE Journal of Solid-State Circuits 36 (2001) 2030–2038.
[11] D. Sadler, R. Changrani, P. Roberts, Thermal management of BioMEMS: temperature control for ceramic-based PCR and DNA detection devices, IEEE Transactions on Components and Packaging Technologies 26 (2003) 309–316.
[12] W. King, T. Kenny, K. Goodson, Design of atomic force microscope cantilevers for combined thermomechanical writing and thermal reading in array operation, IEEE Journal of Microelectromechanical Systems 11 (2002) 765–774.
[13] C.J. Taylor, S. Semancik, The use of microhotplate arrays as microdeposition substrates for materials exploration, Chemistry of Materials 14 (2002) 1671–1677.
[14] A. Ludwig, J. Cao, J. Brugger, I. Takeuchi, MEMS tools for combinatorial materials processing and high-throughput characterization, Measurement Science and Technology 16 (2005) 111–118.
[15] H. Baltes, O. Paul, Micromachined thermally based CMOS microsensors, Proceedings of the IEEE 86 (1998) 1660–1678.
[16] C. Falconi, C. Di Natale, E. Martinelli, A. D'Amico, E. Zampetti, J.W. Gardner, C.M. Van Vliet, 1/f noise and its unusual high-frequency deactivation at high biasing currents in carbon black polymers with residual 1/f̂(((= 2.2) noise and a preliminary estimation of the average trap energy, Sensors and Actuators B: Chemical 174 (2012) 577–585.
[17] G. Ferri, V. Stornelli, A high precision temperature control system for cmos integrated wide range resistive gas sensors, Analog Integrated Circuits and Signal Processing. 47 (2006) 293–301.
[18] Z.L. Wang, J. Song, Piezoelectric nanogenerators based on zinc oxide nanowire arrays, Science (New York, N.Y.) 312 (2006) 242–246.
[19] G. Romano, G. Mantini, A. Di Carlo, A. D'Amico, C. Falconi, Z.L. Wang, Piezoelectric potential in vertically aligned nanowires for high output nanogenerators, Nanotechnology 22 (2011) 465401.
[20] R. Araneo, G. Lovat, P. Burghignoli, C. Falconi, Piezo-semiconductive quasi-1d nanodevices with or without anti-symmetry, Advanced Materials 24 (2012) 4719–4724.
[21] G.C. Cardinali, L. Dori, M. Fiorini, I. Sayago, G. Faglia, C. Perego, A smart sensor system for carbon monoxide detection, Analog Integrated Circuits and Signal Processing 14 (1997) 275–296.
[22] C. Falconi, M. Fratini, CMOS microsystems temperature control, Sensors and Actuators B: Chemical 129 (2008) 59–66.
[23] G. Araujo, R. Freire, J. da Silva, DC-amplifier-input-offset-voltage control in a constant-temperature thermoresistive-sensor-measurement instrument, IEEE Transactions on Instrumentation and Measurement 56 (2007) 778–783.
[24] C. Falconi, E. Zampetti, S. Pantalei, E. Martinelli, C. di Natale, A. D'Amico, et al., Temperature and flow velocity control for quartz crystal microbalances, in: 2006 IEEE International Symposium on Circuits and Systems, 2006, pp. 4399–4402.
[25] J. Lee, T. Beechem, T.L. Wright, B.A. Nelson, S. Graham, W.P. King, Electrical, thermal, and mechanical characterization of silicon microcantilever heaters, Journal of Microelectromechanical Systems 15 (2006) 1644–1655.
[26] A.I.K. Lao, T.M.H. Lee, I.-M. Hsing, N.Y. Ip, Precise temperature control of microfluidic chamber for gas and liquid phase reactions, Sensors and Actuators A: Physical 84 (2000) 11–17.
[27] N.R. Swart, A. Nathan, Design optimisation of integrated microhotplates, Sensors and Actuators A: Physical 43 (1994) 3–10.
[28] A.E. Perry, G.L. Morrison, A study of the constant-temperature hot-wire anemometer, Journal of Fluid Mechanics 47 (1971) 577–599.
[29] P. Freymuth, Feedback control theory for constant-temperature hot-wire anemometers, Review of Scientific Instruments 38 (1967) 677.

[30] A. Oliveira, P. Lobo, G. Deep, R. Freire, J. da Rocha Neto, Frequency domain analysis of an electrical substitution radiometer, Transactions of the ASME 121 (1999) 110–115.
[31] C. Falconi, E. Martinelli, C. Di Natale, A. D'Amico, F. Maloberti, P. Malcovati, A. Baschirotto, V. Stornelli, G. Ferri, Electronic interfaces, Sensors and Actuators B: Chemical 121 (2007) 295–329.
[32] C. Lauterbach, W. Weber, S. Member, D. Römer, Charge sharing concept and new clocking scheme for power efficiency and electromagnetic emission improvement of boosted charge pumps, IEEE Journal of Solid-state Circuits 35 (2000) 719–723.
[33] C. Falconi, G. Savone, a. D'Amico, High light-load efficiency charge pumps, in: 2005 IEEE International Symposium on Circuits and Systems, 2005, pp. 1887–1890.
[34] J. Laconte, C. Dupont, D. Flandre, J.-P. Raskin, SOI CMOS compatible low-power microheater optimization for the fabrication of smart gas sensors, IEEE Sensors Journal 4 (2004) 670–680.
[35] P.C.D. Jong, G.C.M. Meijer, Using Dynamic Feedback Control 46 (1997) 758–763.
[36] C. Falconi, M. Faccio, A. D'Amico, High-accuracy instrumentation amplifier for low voltage low power CMOS smart sensors, in: Proceedings of the 2003 IEEE International Symposium on Circuits and Systems, Bangkok, Thailand, 2003.
[37] C. Falconi, A. D'Amico, M. Faccio, Design of accurate analog circuits for low voltage low power CMOS systems, in: Proceedings of the 2003 IEEE International Symposium on Circuits and Systems, Bangkok, Thailand, 2003.
[38] J.S. Agustsson, U.B. Arnalds, A.S. Ingason, K.B. Gylfason, K. Johnsen, S. Olafsson, J.T. Gudmundsson, Electrical resistivity and morphology of ultra thin Pt films grown by dc magnetron sputtering on SiO 2, Journal of Physics: Conference Series 100 (2008) 082006.

Biography

Christian Falconi was born in Rome, Italy, 1973. He received the MSc (cum laude) and the PhD degrees in electronic engineering from the University of Tor Vergata, Rome, Italy, in, respectively, 1998 and 2001; as a part of his PhD program he visited the University of Linkoping (1 month), and TU Delft (7 months). In 2002 he has supervised the design of the electronic interface for the ST-Microelectronics DNA chip. Since 2002 Christian Falconi is Assistant Professor at the Department of Electronic Engineering, University of Tor Vergata, where he currently teaches Micro-Nano-Systems and Electronic Interfaces. Since 2003 he has visited the Georgia Institute of Technology (14 months). His research interests include electronic devices, analog circuits, electronic interfaces, sensors, microsystems, and nanosystems.