A New Technique for Thermal Resistance Measurement in Power Electron Devices

Fabio Filicori, Paola Rinaldi, Giorgio Vannini, *Member, IEEE*, and Alberto Santarelli, *Member, IEEE*

***Abstract*—A simple technique is proposed for the thermal resistance measurement of electron devices. The new approach is based on the standard measurements which are normally carried out for the electrical characterization of power devices, without requiring special-purpose instrumentation and/or physics-based temperature-dependent electrical device models. Experimental results, which confirm the validity of the new method, are provided.**

***Index Terms*—Device characterization, electron device, thermal resistance.**

I. Introduction

THE THERMAL characterization of semiconductor power devices is extremely important for ensuring the safe operating area (SOA) conditions since, for thermally safe design, the junction temperature must not exceed its maximum allowable value.

The thermal resistance R_θ is normally used by device manufacturers and circuit designers for calculating the junction temperature of power devices. The concept is based on the analogy between the electrical and thermal properties of materials with temperature, heat flow, and thermal resistance analogous to voltage, current, and electrical resistance. In particular, $R_\theta = \partial\theta/\partial P_d$ should be determined by measuring the change in junction temperature θ caused by a corresponding variation of the dissipated power P_d.

Beside infrared or other invasive techniques such as those based on liquid crystals, purely electrical approaches may be adopted for thermal resistance measurement. In this case, an electrical temperature sensitive parameter (TSP) of the device, such as the on-voltage or the inverse current of a diode, is typically used as a built-in temperature sensor [1]. Two types of TSP-based methods may be defined: continuous and switched (or pulsed) methods. In the continuous methods, the TSP is used for measuring the junction temperature under two different known dissipative operating conditions, so that the thermal resistance can be obtained by simply computing the ratio between the temperature variation and the corresponding difference in power dissipation. Obviously, this procedure can be correctly applied only if the TSP is independent of the electrical operating conditions. Unfortunately, the most convenient TSPs, like on- or off-state voltages/currents, are current/voltage dependent, respectively. A physics-based electrical model for the TSP could then be adopted, in order to take into account TSP variations due to different operating conditions, but it requires other special-purpose identification procedures. As an alternative, switched (or pulsed) methods may be used. They are based on fast electrical transients, involving both high and low power dissipation states, and exploit the relatively long thermal time constant of electron devices. For instance, in switched methods, the TSP is first characterized versus temperature (thermometer calibration) while the device dissipates zero or near zero power, so that the junction and case (e.g., package) temperatures practically coincide. Then, the TSP is measured immediately after a fast switching transition between a highly dissipative ($P_{\text{diss}} = \widehat{P}$, with junction temperature $\theta_j = \widehat{\theta}$ unknown) and the same very low dissipative condition used in the TSP calibration (where a known constant case temperature θ_C is maintained). If the power switching and the TSP measurement times are sufficiently close to each other (typically $< 1\ \mu$s), the junction temperature does not change, and its value is identified on the basis of the TSP measurement under the same device operating condition used for calibration. Then, the thermal resistance is simply obtained [1] as $R_\theta = (\widehat{\theta} - \theta_C)/\widehat{P}$.

Slightly different procedures are also possible by considering pulsed operating conditions instead of switched ones [1], [2]. Clearly, both these kinds of measurements require relatively simple but special-purpose instrumentation with switched or pulsed measurement capabilities. Other commonly adopted electrical approaches, which do not need a pulsed/transient measurement setup, are somehow invasive, since they use special layout structures, where a "sensing" diode (whose voltage is used as TSP) is integrated close to the power device [3].

In this paper, a new noninvasive method for the accurate measurement of the thermal resistance is proposed, which has the advantage of being based on the standard measurements (i.e., small signal parameters and case-temperature-dependent dc characteristics) that are normally performed for the electrical characterization of a semiconductor power device. With respect to other approaches [4], [5] based on small-signal electrical measurements, the method proposed does not require the identification of a physics-based model of the TSP, which may be difficult to extract and usually requires a detailed knowledge of the device physical parameters.

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II. Measurement Technique

The new approach for thermal resistance measurement will be described by considering a single-port electron device (i.e., a diode). The extension to two-port devices (i.e., transistors) is straightforward and will be discussed later.

In the following, a constant external case (e.g., package) temperature θ_C will be considered, while the internal junction temperature[1] θ can be time-varying under very-low-frequency operating conditions. Therefore, in the low-frequency range where reactive effects due to charge storage are negligible, the device voltage/current characteristics can be expressed in the form

$$V_d = F[I_d, \theta] \tag{1}$$

where F is an algebraic function of both the instantaneous current I_d and the junction temperature θ.

Let us now consider the small-signal behavior of the device under test (DUT) around bias conditions involving relevant power dissipation, where self-heating effects due to thermal phenomena are not negligible. Due to the relatively slow dynamics of thermal phenomena, the power device exhibits a very different small-signal behavior when considering frequencies well below (i.e., nearly dc) and well above (e.g., 1 kHz) the cutoff frequency of thermal phenomena [6]. Since the proposed approach is based on this important difference, the small-signal resistances at the two outlined frequencies are now computed according to (1). These two different resistances will be referred to in the following as r_{dc} and r_{ac}, respectively.

In small-signal operation at frequencies which are well above the thermal cutoff (e.g., the typical frequency of 1 kHz, normally used for the measurement of differential resistances/conductances), the small-signal resistance r_{ac} is not affected by time-varying thermal effects, since $\theta(t) \simeq \theta^* \doteq P_d^* \cdot R_\theta + \theta_C$, where P_d^* is the power dissipation corresponding to the quiescent condition $I_d = I_d^*$ (and $V_d = V_d^* = F[I_d^*, \theta^*]$). Thus, the small-signal resistance r_{ac} can be simply expressed as

$$r_{\mathrm{ac}} = \left(\frac{dV_d}{dI_d}\right)_{\mathrm{equithermal}} = \frac{\partial}{\partial I_d} F[I_d, \theta^*]. \tag{2}$$

On the other hand, under very low-frequency (i.e., almost dc) operation, the slow time-varying currents and voltages also involve internal temperature variations $\theta(t) = P(t)R_\theta + \theta_C$, with $P(t) = V_d(t)I_d(t)$. Thus, the low-frequency small-signal resistance r_{dc} is different from r_{ac} and is given by

$$\begin{aligned} r_{\mathrm{dc}} &= \left(\frac{dV_d}{dI_d}\right)_{\mathrm{nonequithermal}} \\ &= \frac{\partial}{\partial I_d} F[I_d, \theta^*] + \frac{\partial}{\partial \theta} F[I_d^*, \theta] \cdot \frac{d}{dI_d} \theta[I_d]. \end{aligned} \tag{3}$$

The first partial derivative in (3) clearly coincides with r_{ac}, while

$$\frac{d}{dI_d}\theta[I_d] = \frac{d}{dI_d}(R_\theta V_d I_d + \theta_C) = R_\theta(r_{\mathrm{dc}} I_d^* + V_d^*). \tag{4}$$

[1]Although the internal temperature distribution is not uniform, an equivalent average junction temperature θ will be considered here, according to conventional thermal-resistance-based approaches.

Finally, by substituting (4) into (3), we obtain

$$r_{\mathrm{dc}} = r_{\mathrm{ac}} + K_\theta R_\theta (r_{\mathrm{dc}} I_d^* + V_d^*) \tag{5}$$

where: $K_\theta = \partial F[I_d^*, \theta]/\partial \theta$.

The identification of the thermal resistance R_θ from (5) is now considered. Both r_{dc} and r_{ac} can easily be measured with conventional instrumentation for electrical device characterization provided that a constant case temperature is maintained coherently with the above definitions. The details on the corresponding measurement setup will be provided in Section III. However, to obtain the thermal resistance from (5), the thermal sensitivity K_θ to junction temperature θ is also needed. Although it is theoretically possible to use a physics-based thermal model [5] to derive the temperature sensitivity of F, this approach requires a detailed knowledge of device physics and parameters; this is not always possible especially for commercially available or new, advanced electron devices. Moreover, the identification of the thermal model may be quite a time-consuming task. As a convenient alternative, K_θ can also be indirectly measured. In practice, since the internal temperature θ is not directly controllable, it is easier to express K_θ as a function of the thermal sensitivity $K_{\theta_C} = dV_d/d\theta_C$ to variations of the case temperature θ_C around the nominal value

$$\begin{aligned} K_{\theta_C} &= \frac{dV_d}{d\theta_C} = \frac{\partial}{\partial \theta} F[I_d^*, \theta] \cdot \frac{d}{d\theta_C}(R_\theta V_d I_d + \theta_C) \\ &= K_\theta \cdot (R_\theta K_{\theta_C} I_d^* + 1) \end{aligned} \tag{6}$$

which leads to

$$K_\theta = \frac{K_{\theta_C}}{1 + R_\theta K_{\theta_C} I_d^*}. \tag{7}$$

According to its definition, the thermal sensitivity K_{θ_C} can easily be obtained by measuring the voltage variation due to a small variation of the case temperature for a constant diode current. This measurement is usually carried out as a basic step for the electrical characterization of power electron devices, since the static dc characteristics are normally measured at different, constant case temperatures.

By substituting (7) into (5), the expression of the thermal resistance as a function of the measured parameters r_{dc}, r_{ac} and K_{θ_C} is finally obtained

$$R_\theta = \frac{r_{\mathrm{dc}} - r_{\mathrm{ac}}}{K_{\theta_C}(V_d^* + r_{\mathrm{ac}} I_d^*)}. \tag{8}$$

The approach described above for single-port devices can be directly extended to power transistors. To this aim, typical on-state bias conditions of power transistors in switching applications may be conveniently considered. For instance, we consider here a MOSFET biased in the triode region (i.e., linear), where the output differential resistance is still quite small yet important self-heating phenomena occur. In such a case, the differences between dc and ac differential parameters are significant and can still be accurately measured by means of conventional instrumentation. Moreover, in this bias condition the power dissipation may be evaluated by only considering the electrical variables at the output (*power*) port, since the contribution associated to the input (*control*) port is normally negligible. Thus,

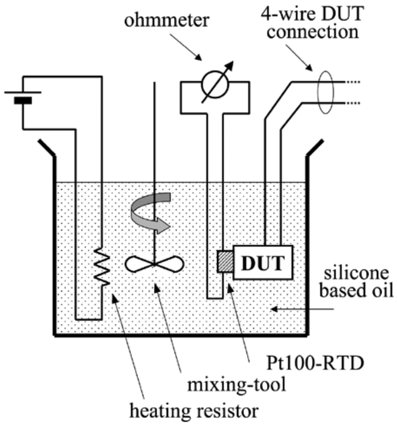

Fig. 1. DUT measurement setup.

the output characteristics at a given gate voltage bias V_{GS}^* may be written as

$$V_{DS} = F_T[I_{DS}, \theta; V_{GS}^*] \tag{9}$$

where F_T is an algebraic function of two variables with parametric dependence on the given applied voltage at the input transistor port. In this case, θ can be interpreted as the average temperature of the MOSFET channel. Clearly, (9) may be treated analogously to (1), and all the theoretical developments for the single port device may be repeated for the two-port device by simply replacing V_d, I_d with the voltage and current at the output port of the power transistor. Thus, (8) can be directly used for the evaluation of the thermal resistance of power transistors. Experimental validation of the procedure is provided in the following section.

III. Experimental Results

The experimental validation of the new approach for thermal resistance measurement is initially carried out using a Si-based power diode STPR520F ($P_{\max} = 12$ W) manufactured by STM. To accurately set the case temperature, the diode under test is immersed in silicone-based oil, acting as a large heat sink, whose temperature is controlled by a conventional thermostatic system. In particular, the diode temperature is measured by monitoring the variation of a platinum-based PT100 resistance temperature detector ($R = 100\ \Omega$ at 0 °C, with temperature coefficient equal to $0.385\%/°\mathrm{C}$) closely connected to the case, as shown in Fig. 1.

The proposed procedure based on (8) is used here to measure the thermal resistance of the diode with reference to a given operating condition. In particular, a bias current is chosen equal to $I_d^* = 5\ \mathrm{A} \pm 0.001\ \mathrm{A}$ ($V_d^* = 0.86\ \mathrm{mV} \pm 0.03\ \mathrm{mV}$ at 60 °C). Relevant self-heating phenomena around this bias condition cause important differences in the dc and ac differential resistances, as will be evident in the following.

The different steps of the proposed method are now outlined. First, the diode differential resistance r_{dc} is obtained by differentiation of the device static characteristics voltage versus current measured at the arbitrary constant case temperature $\theta_C = 60$ °C (the same nominal case temperature is assumed in all the examples presented). Thus, the diode is excited with two different current values I_{d1} and I_{d2} ($I_{d1} = 4.5$ A, $I_{d2} = 5.5$ A),

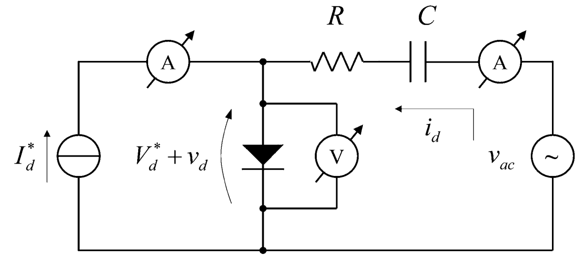

Fig. 2. Electrical circuit schematic used for the diode characterization.

TABLE I
Diode AC Differential Resistance Measured at Different Frequencies

$Freq.$ $[Hz]$	i_d^{rms} $[mA]$ $(\pm 0.5mA)$	v_d^{rms} $[mV]$ $(\pm 0.03mV)$	r_{ac} $[m\Omega]$ $(\pm 0.4m\Omega)$
20	109.9	3.12	28.4
50	110.2	3.28	29.8
70	110.4	3.31	30.0
100	110.6	3.34	30.2
200	110.6	3.35	30.3
500	110.3	3.34	30.3
1000	110.3	3.34	30.3
5000	109.9	3.33	30.3

which represent small variations with respect to the nominal bias current; the corresponding diode voltages V_{d1}, V_{d2} are measured by means of a precision multimeter ($\pm 0.03 - \mathrm{mV}$ accuracy). After imposing each new current value, it is necessary to wait long enough to guarantee the settlement of the case temperature. In particular, we considered the case temperature as settled when the variations are less than 0.05 °C/min. Then, the dc differential resistance is evaluated as $r_{\mathrm{dc}} = (V_{d2} - V_{d1})/(I_{d2} - I_{d1}) = 23.55\ \mathrm{m}\Omega$, whose associated uncertainty has been estimated as 0.45% (± 0.1 mΩ) according to the formula

$$\frac{\Delta r_{\mathrm{dc}}}{r_{\mathrm{dc}}} = \frac{\Delta V_{d2} + \Delta V_{d1}}{V_{d2} - V_{d1}} + \frac{\Delta I_{d2} + \Delta I_{d1}}{I_{d2} - I_{d1}}. \tag{10}$$

Then, the diode ac differential resistance r_{ac} is measured by superimposing a small-signal sinusoidal excitation ($i_d^{\mathrm{rms}} = 100$ mA) to the nominal bias current. Thus, a sinusoidal voltage source v_{ac} with a high-value series resistor *R* is connected[2] to the diode as shown in Fig. 2. A value of the series resistor ($R = 150\ \Omega$) sufficiently larger than the differential resistance of the diode (< 100 mΩ) guarantees operation of the ac source in Fig. 2 as an almost ideal ac current generator. The series capacitor C prevents the dc current flowing into the ac branch. A suitable large value ($C = 100\ \mu$F) sets the high-pass filter cutoff frequency at nearly 10 Hz. As shown in Fig. 2, voltmeter reading at the diode (v_d^{rms}) provides the differential ac resistance by means of $r_{\mathrm{ac}} = v_d^{\mathrm{rms}}/i_d^{\mathrm{rms}}$. Measurements were carried out at different frequencies and the results obtained in Table I with the corresponding measure uncertainties. Clearly, the electrical impedance magnitude $|Z| = v_d^{\mathrm{rms}}/i_d^{\mathrm{rms}}$ becomes frequency-independent and, consequently, purely resistive (i.e., $r_{\mathrm{ac}} \equiv |Z|$) above 100 Hz. In our validation test of thermal resistance measurement, the value at 1 kHz was chosen, obtaining

[2]A power amplifier, not shown in Fig. 2, was also used in order to deliver sufficiently-high power to the diode.

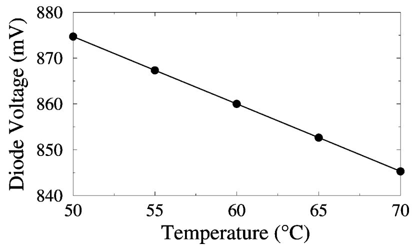

Fig. 3. ST-PR520F diode voltage measured at different case temperatures (constant current $I_d = 5$ A).

$r_{\mathrm{ac}} = 30.3\ \mathrm{m}\Omega \pm 0.4\ \mathrm{m}\Omega$. As it can be seen, an ac value of the differential resistance higher than the corresponding dc one is obtained, coherently with a diode saturation current increasing with temperature.

At this point, the thermal sensitivity K_{θ_C} is evaluated by measuring the voltage variation due to a small perturbation of the case temperature at the constant bias current I_d^*. Again, after changing the current into the heating resistor shown in Fig. 1, it is necessary to wait long enough for the oil temperature to settle to a new stable value before acquisition of the diode voltage. In Fig. 3, the measured diode voltage values are shown at different case temperatures for a constant current of 5 A. Since a linear relationship holds over a large range of temperatures, accurate estimation of the thermal sensitivity ($K_{\theta_C} = -1.47$ mV/°C) can be obtained. Since this value is evaluated by means of a least square fitting of multiple acquisitions, thermal sensitivity is considered known with negligible uncertainty.

After evaluation of the three parameters r_{dc}, r_{ac}, and K_{θ_C}, the diode thermal resistance is obtained through (8), which provides a value $R_\theta = 4.5$ °C/W with an uncertainty of about 8% given by

$$\frac{\Delta R_\theta}{R_\theta} = \frac{\Delta r_{\mathrm{dc}} + \Delta r_{\mathrm{ac}}}{|r_{\mathrm{dc}} - r_{\mathrm{ac}}|} + \frac{\Delta V_d^* + \left(\frac{\Delta r_{\mathrm{ac}}}{r_{\mathrm{ac}}} + \frac{\Delta I_d^*}{I_d^*}\right) r_{\mathrm{ac}} I_d^*}{|V_d^* + r_{\mathrm{ac}} I_d^*|}. \quad (11)$$

In order to validate the results obtained with the new approach, thermal resistance measurements were carried out also with the switched method [1] summarized in Section I, where the diode voltage was used here as TSP. To this end, the diode was excited by a voltage dc source and series-connected with suitable-value switchable resistors. The diode voltage and current were monitored by means of a digital sampling oscilloscope. In particular, after calibration of the TSP under negligible dissipated power, two different power switching conditions were used for thermal resistance measurement. In the first experiment, the diode current was instantaneously switched from 2.2 A to around 50 mA. Instead, in the second case the current switching was between 5 A and 50 mA. Both experiments led to a thermal resistance evaluation of 4.2 °C/W, in excellent agreement with the value obtained with the new method.

The experimental procedure for the evaluation of the thermal resistance was also carried out on a Harris power MOSFET RFP4N06 ($P_{\max} = 25$ W) biased at $V_{GS}^* = 10$ V and $I_{DS}^* = 2$ A $\pm$ 0.001 A ($V_{DS}^* = 2$ V $\pm$ 0.2 mV at 60 °C). The bias condition adopted corresponds to DUT operation in the triode region but it involves a relatively high power dissipation, so that important self-heating effects can still be observed. As previously discussed, (8) is used here by assuming $r_{\mathrm{dc}} = (dI_{DS}/dV_{DS})_{\mathrm{nonequithermal}}$; $r_{\mathrm{ac}} = (dI_{DS}/dV_{DS})_{\mathrm{equithermal}}$, and $K_{\theta_C} = dV_{DS}/d\theta_C$. In particular, the dc differential resistance was evaluated by forward/reverse differentiation of the static current characteristics by measuring the drain source voltage at 1.8 and 2.2 A of drain current ($\theta_C = 60$ °C). An $r_{\mathrm{dc}} = 1.168\ \Omega \pm 0.007\ \Omega$ was obtained.

Moreover, by using the same electrical measurement circuit in Fig. 2 (apart from using an additional voltage source for biasing the gate), the ac differential resistance was measured at different frequencies obtaining $r_{\mathrm{ac}} = 0.806\ \Omega \pm 0.002\ \Omega$ at 1 kHz. Unlike the diode case, an expected ac resistance smaller than the corresponding dc value is obtained. In fact, lower FET drain currents correspond to higher temperature values due to a decrease in carrier mobility.

Finally, thermal sensitivity to case temperature variations in the range between 50 °C and 70°C at $I_{DS}^* = 2$ A was estimated as $K_{\theta_C} = 0.0208$ V/°C and assumed with negligible uncertainty, as in the diode case.

Substitution of r_{dc}, r_{ac}, and K_{θ_C} values into (8) yields a thermal resistance $R_\theta = 5{,}2$ °C/W with an uncertainty of about 3%. The values predicted by means of expression (8) applied at the output port of the device are in excellent agreement with the corresponding values extracted by using the standard switched method [1] and are substantially coherent with the values provided in the manufacturer's data sheet, as can be seen from Table II.

TABLE II
VALUES OF THE THERMAL RESISTANCE (°C/W) OBTAINED WITH DIFFERENT MEASUREMENT APPROACHES AND PROVIDED BY THE MANUFACTURERS FOR A POWER DIODE AND A MOSFET

	Diode	MOS
New approach	4.5	5.2
Standard method	4.2	5.1
Data sheet	6.0 (max)	5.0 (max)

IV. CONCLUSION

The new method proposed for thermal resistance measurements provides accurate results on the basis of conventional electrical characterization without requiring special-purpose instrumentation or specific layout structures of the device.

The same approach can easily be extended to the characterization of the device thermal impedance on the basis of electrical impedance vector measurements.

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