Chapter 5
Fundamentals of Thermal Transient Measurements

Gábor Farkas

As discussed deeply in Chap. 2, a change in the power applied on a system results in a transient change of the temperature. The growth of the powering from a lower P_L level to higher P_H initiates a heating transient. When the power diminishes from P_H to P_L, we observe the cooling of the system (Fig. 5.1).

The transient change is influenced by the material properties and geometry of the structural elements in the heat-conducting path. Consequently, thermal transients carry information on the composition of the structure and eventually on its health and location of potential failures. Also, with an appropriate number of recorded transients, thermal models of the system of different complexity can be built.

Thermal system descriptors such as Z_{th} curves and structure functions can be derived, and thermal models of the system can be constructed with a relatively simple mathematical apparatus if the change in powering occurs as sharp switching between two stable power levels.

Looking into the physical background, we obtained that the origin of thermal transients in a system is a lasting imbalance between the energy *generated* or *applied externally* at specific system locations and the energy which leaves the system.

In this book thermal transients of electronic systems are discussed, where the applied energy is of electrical nature, converted to heat in the tested devices. The energy removal occurs mostly in the form of heat flux, with the important exception of solid-state lighting devices where a large part of the energy is emitted in the form of light [7].

G. Farkas (✉)
Siemens Digital Industry Software STS, Budapest, Hungary
e-mail: Gabor.Farkas@siemens.com

M. Rencz et al. (eds.), *Theory and Practice of Thermal Transient Testing of Electronic Components*, https://doi.org/10.1007/978-3-030-86174-2_5

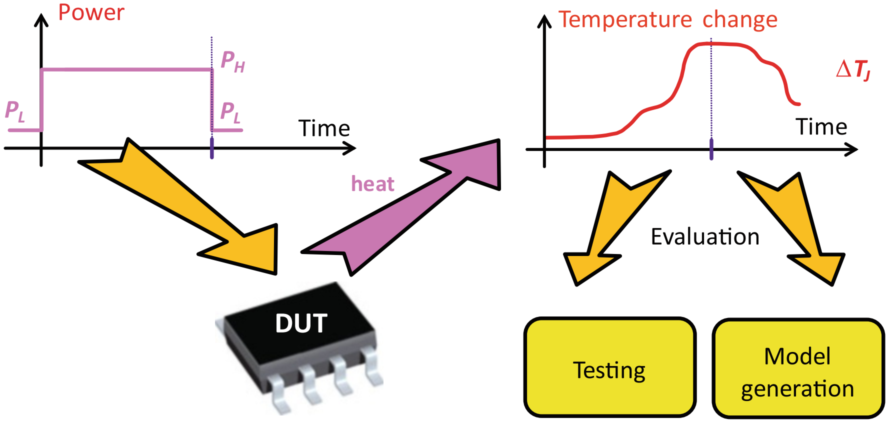

Fig. 5.1 Thermal transients in a system and their use for testing and modeling

In Chap. 2 the energy imbalance was formulated between the momentary power values, the power generated (P_{gen}, P_{in}) and removed (P_{diss}, P_{out}).

The heat removal, more often called dissipation, is governed by the structure in which the powered locations reside. The actual heat flux and the shape of its trajectories depend on the material properties and the geometry of the structure when *conduction* is the mechanism of the heat removal. The heat flow in a continuous, thermally conductive body is determined by a set of partial differential equations, the Fourier law, formulated for an elementary structural detail in Eqs. (2.1), (2.2), (2.3), (2.4), (2.5), (2.6), (2.7) and (2.8). Heat conduction is the most important form of heat transport in the internal parts of electronic systems.

In case when *convective* heat removal is present in some system portions, a different set of equations is to be applied. *Radiation* may play a role at systems operated at high temperature.

The thermal behavior of the system can be investigated based on the *continuous model* of the system if the full geometry and properties of the material components are known [56].

It was presented in Chap. 2 that the conductive heat removal can be modeled by a net of discretized thermal resistance and capacitance elements. An equivalent RC model can be also constructed for convection. In case of radiation, such models are of poor accuracy because of the strong nonlinearity of the effect.

In the practical task of *thermal testing*, it is obvious to partition the above theoretical "structure" into the *device under test* (DUT), the thermal properties of which are to be determined and the *test environment*. The two are separated by a *thermal interface*.

The DUT in the transient test can be a packaged semiconductor device or a larger subsystem in a system-level structural analysis.

Figure 5.1 hints that the thermal transient tests always include a heating and a cooling phase. Placing the obviously non-energized DUT into the test system, these two phases are typically realized as consecutive processes.

In order to provide repeatable testing results, the procedure of thermal testing is regulated in *thermal testing standards* built on the above considerations. Different standards exist for the accomplishment of the task; these will be presented in later chapters in details.

In most standards first the *device categories* to be tested are defined, classified upon their electrical and mechanical construction. The *thermal test equipment* comprises the electronics for powering (excitation), and the electronics for data acquisition. Some standards also specify the conductive or convective thermal test environment around the device.

Today the data acquisition is always accomplished by recording electrical signals. This implicates that thermal sensors (transducers) are needed which convert the transient temperature of the locations of interest into an electrical signal.

The most critical factor in the reliability and lifetime of power applications is the temperature of the semiconductor power devices themselves. As discussed in Chap. 4, practically all parameters of these devices are strongly affected by the device temperature. These temperature-dependent parameters include voltage and current at input and output pins in an actual operating point, the timing of their change, etc. In such a way, semiconductor devices are perfect thermometers; besides they amplify, oscillate, switch current, or emit light. In the literature and in standards, the temperature-dependent electrical parameters that are used for measuring the temperature are referred to as TSP (temperature-sensitive parameter) or TSEP (temperature-sensitive electrical parameter).

The temperature of other accessible points on the device package or in the test environment can be transduced by external *dedicated temperature sensors*, based on the various temperature-dependent effects presented in Chap. 4, e.g., change of resistance or Seebeck effect. In large modules for current switching purposes, it is not uncommon to mount internal resistive temperature sensors on the baseplate hosting the power semiconductors.

5.1 Tester Instrumentation and the Environment

The major elements needed for thermal transient testing are as follows:

- A power switching unit to provide the exact powering, typically switching between two power levels
- A measurement or data acquisition unit to detect the temperature-dependent electric signal
- A temperature-controlled environment to maintain a prescribed ambient into which the DUT is placed (e.g., cold plate, or still-air chamber, etc.)
- A data processing unit, as the measured temperature response has to be further processed

All the above functions have to be controlled by a computer.

The general scheme of a thermal transient tester is shown in Fig. 5.2. The DUT in the figure is a thermal test vehicle with separated powering and sensing, a device in which the functionalities of the thermal testing are least intermingled.

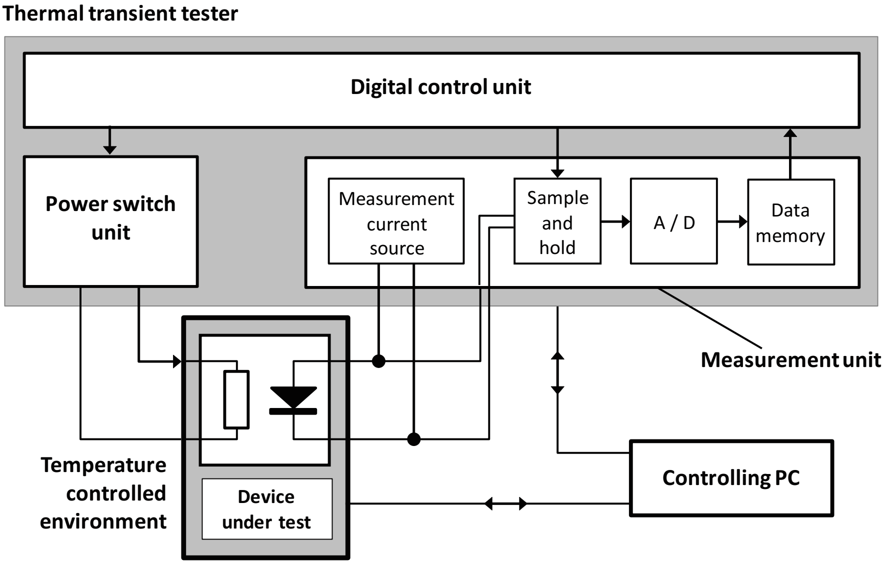

Fig. 5.2 The general scheme of the tester and test environment

The major functionalities of the tester are discussed below.

Powering

In Chap. 2 it was proved that the mathematical procedure of converting the transient temperature data into relevant thermal descriptors such as Z_{th} curves, time constants, and structure functions is the easiest if a power step is applied on the investigated device, and the power is maintained until a steady state is reached again. In the case when a sharp, steplike power change cannot be produced; this conversion is still feasible, but it needs recording of the variation of the power profile and then a complex calculation of convolution.

It has to be noted that the powering units of testers have to serve drastically different needs and have to be of different power and timing specification. Tiny microelectromechanical systems (MEMS) or laser diodes respond in the microsecond range when milliwatts are applied on them; large press-pack IGBT modules stabilize in a few minutes when powered by appropriate kiloamperes. Large street luminaires with their normal air cooling reach steady state in a daylong transient.

In typical testers the diverse demands are served by a base instrument fulfilling the common voltage and current requirements and further accessories, amplifiers, and boosters which extend the voltage and current range.

In many cases applying a long power step until steady state is reached is not feasible.

For example, in a production line, the measurement time is inherently limited. Considering the number of devices to be measured, the targeted throughput restricts the time slot for thermal testing sometimes just to milliseconds [106, 107].

In a type of lifetime testing (active power cycling), pulses at an appropriate duty cycle are applied on the device until its deterioration (Chap. 7, Sect. 7.4). The wear mechanism typically differs when the pulses are short, below a few seconds, and when they are longer, lasting for minutes [21]. Keeping the power for a prolonged time would cause an abrupt breakdown of the device.

However, also in these cases, a sharp swing between the power levels, in the form of a single pulse or multiple pulses, facilitates the calculation of a limited set of the thermal descriptors.

For satisfying all these demands, the sophisticated hardware of transient testers has to incorporate a powering unit with controlled voltage and current sources, which switch between preprogrammed levels fast and reach the accurate "set value" with short settling time.

The settling time of a current, or power in general, is also related to the targeted power magnitude, partly because of limitations in the electric circuitry, partly because of the inductive nature of the cabling towards the DUT.

Regarding the accuracy of power sources, there is a general contradiction in electronics between the expected speed of change and precision. High accuracy of 4–7 digits of the target value can be achieved with long integration times. Typical power supply units in tester appliances have approximately 0.5% accuracy and a typical relaxation (regulation) time of a few milliseconds at the sudden drastic load change associated with applying or revoking power.

As explained later in this chapter, in a properly constructed measurement arrangement, testers typically switch from high power level to low in a very fast process. In this concept, a higher heating current is switched off by a fast switching circuitry, while a constant sense current from a separate source is maintained on the DUT. The relaxation of the powering source continues in the millisecond order of magnitude, but this wobbling occurs behind the switch, separated from the DUT.

This way, a switching time of a few microseconds can be achieved at the start of a *cooling transient*. In high power testers at hundred and thousand amperes, careful design can achieve a settling time below 100 μs for cooling transients. Power edges of this slew rate are in most cases unfeasible for heating transients due to the relaxation time of power supplies.

As an exception, some advanced testers can turn on several amperes in 10 μs. This can be advantageous when beyond thermal transients also transients in other physical domain (optical, magnetic) are to be investigated.

It also has to be noted that the excitation for thermal testing is not necessarily provided by a dedicated powering unit. In many cases the inherent power belonging to the normal operation of the investigated system can be used for powering. In order to meet the sharp timing constraints between the applied power and the captured temperature response, either the recording starts on a trigger event generated from the power change or the operation of the tested system is to be synchronized with trigger signals from the tester.

This external powering approach also opens the way for thermal investigation of live functional systems during their normal operation.

Sensor Biasing

As discussed in Chap. 4, the temperature-to-voltage transducers are dedicated sensors or semiconductor devices operating at some bias. In case of resistive or diode sensors, a separate source of programmable but constant value, referred to as "measurement current" or "sensor current" maintains a low power level on the sensor element.

Measuring thermal transfer effects needs multiple sensors at various locations. For example, when the thermal coupling between different chips in a power module is investigated, each chip has to be biased. Accordingly, testers typically offer multiple independent sensor sources.

The good quality of sensor biasing sources is indispensable in a thermal tester; it highly affects the usability of the recorded thermal signal. Keeping their noise low is essential for identifying tiny variations in the temperature change. They are exposed to sudden harsh jumps in voltage and current levels on the DUT during the switching process; the distortion of the temperature related signal is influenced by the time and amplitude of their relaxation after the perturbation. The contradicting requirements of low noise, precise value, and short relaxation time have to be fulfilled at the same time.

In some practical realization of thermal transient testers, some bias sources are integrated into the powering units in order to simplify the wiring of transient measurements at the driving point, where power is applied.

Several standards define requirements on the expected ratio of the power level in the heating phase of the transient and in the cooling phase when normally only this sensor bias is present. The validity of these expectations is investigated in Chap. 6.

Data Acquisition

The data acquisition part of a thermal tester is also demanding. A thermal tester equipment seems to be an "inefficient" construction in terms of electronics, as hundred watts or kilowatts are fed into the device under test as excitation, while the resulting electrical signals delivered by semiconductor devices and dedicated sensors are in the order of just a few millivolts. The tiny signal has to be formed by amplifiers, which have to fulfill contradictory requirements again: low noise, high speed, and high accuracy. Overdrive conditions frequently occur at switching; in this case fast recovery to normal operation is indispensable.

A tester typically comprises a number of identical data acquisition functionalities, commonly referred to as measurement channels.

As the subsequent sections confirm, a minimum data acquisition rate of 1 megasample/second and a voltage resolution finer than 100 μV are needed to adequately separate the electric and thermal constituents of the recorded signal and to capture the smallest thermal time constants.

Measurement channels designed with this specification can ensure proper suppressing of the inherent noise of the device and such eliminate the need for complex noise attenuation procedures, e.g., capturing the signal multiple times and averaging it.

Sample and hold circuits and high-resolution analog-digital converters in the channels accomplish the transformation of the analog signal into digital data which are stored in the tester and transferred to the external measurement control for further processing.

Temperature-Controlled Environment
An obvious precondition for repeatable measurements is a proper definition of *repeatable boundary conditions*. The related standards [24, 29, 30], define *convective* and *conductive* environments.

Typical convective boundaries are still-air chambers, wind tunnels, and liquid baths. Conductive environment is realized by different cold plate structures.

Plates and baths can be programmed to force different temperature values to the device under test during the transient measurement. In a similar way, they can be used in the calibration process, in which the devices are kept at preprogrammed temperatures for prolonged time and their temperature-sensitive parameter (TSP) value is recorded.

The temperature range and the heat sinking capability of the test environment have to be selected carefully.

The calibration of the TSP typically occurs on a cold plate or in a liquid bath at low power but in a broader range than the expected temperature transient.

During the heating phase of the transient test, the environment may need to keep its own temperature stable while balancing several kilowatts of dissipated power. However, the actual "set temperature" range can be smaller than that of the DUT; it has to match the intended temperature range on the *accessible package surface* only.

As an example, a typical automotive power module built on silicon chips is often calibrated between 25 °C and 150 °C maximum temperature. Then, transient tests for providing thermal data of the module for data sheet are carried out at room temperature. After finishing these tests, for lifetime analysis, the power cycling occurs at an elevated 80 °C baseplate temperature to mimic the typical operating temperature under the hood. In such cases the selected thermostat has to be able to work in wider temperature range and needs to have a high enough heat sinking capability to balance the power dissipation of the component.

Calibration thermostats are often based on thermoelectric coolers (TEC). These fast thermostats may use a combination of resistive heaters for heating and Peltier elements for cooling with a few watts of heat sinking capability at room temperature.

The performance of cold plates is determined by the programmable liquid circulator equipment and the properties of the coolant material. These thermostats also have asymmetric characteristics, with high heating and lower cooling capability, this latter being limited by the cooling aggregate.

In all thermostat constructions, the heat sinking capability diminishes at lower set temperature; the minimum temperature of a construction is defined at zero applied power.

In many branches of electronics, critical subassemblies rely on air cooling (e.g., automotive applications), or complete appliances are cooled by air exclusively (mobile phones). In these constructions the integrated circuits and discrete

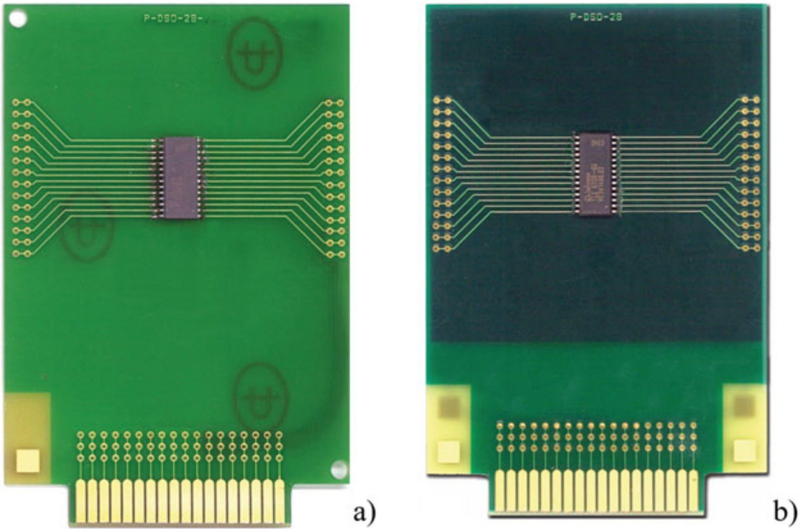

Fig. 5.3 Integrated circuit in DIP package on standard printed boards as defined in the JEDEC JESD51 set of standards, (**a**) worst-case cooling with one patterned copper layer (**b**) with two solid internal copper layers for enhanced cooling. Separated force and sense traces can be observed

components dissipate towards the printed boards hosting them, mostly towards the copper layers.

In order to qualify the thermal performance of packaged semiconductors designed for heat removal towards the board, the whole thermal environment around them must be highly standardized. This uniformity facilitates clear distinction of the share of heat flow in the packaged device and in the environment and helps gaining a valid comparison between different package constructions.

The first elements in the heat-conducting path which belong to the standardized thermal environment are the printed boards. For comparability of measurement results, these have to follow the related standards regarding their geometry and layer structure. Figure 5.3 demonstrates a pair of standard constructions, a worst-case solution, in which cooling occurs towards a single patterned copper layer on board surface with thin traces and a version where two solid internal copper layers ensure enhanced cooling. The JEDEC JESD51 set of thermal measurement standards [29] defines several standard printed board constructions, adjusted to the style of package pins or solder balls, and the presence of cooling tabs or fillers, etc. The two formations in Fig. 5.3 host leaded surface mount packages. An example with their comparison is shown in Chap. 3, Sect. 3.1.2.

The standardization concept is also extended to the farther environment; the JEDEC JESD51 set of standards includes a worst-case still-air conductive environment (Fig. 5.4, JESD51-2A standard) and wind tunnels (JESD51-6 standard).

The construction of the conductive or convective test environments also determines the time needed for a complete thermal test. At aggressive cooling on cold

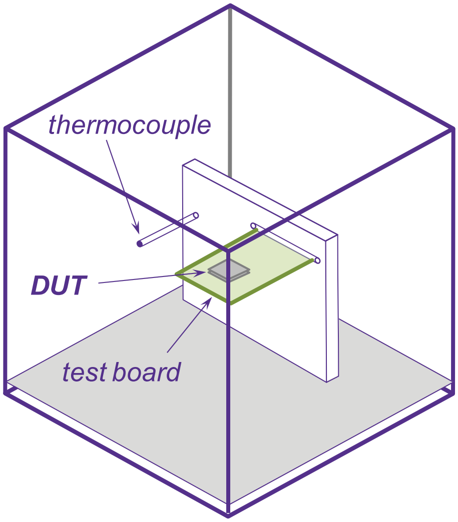

Fig. 5.4 One cubic feet still-air chamber and the location of a standard printed board as defined in the JEDEC JESD51-2A standard

plate, steady state is reached within a couple of minutes; in still air tens of minutes or hours are needed for equalization time.

In case of solid-state lighting devices, additional radiometric and photometric standards (listed in Chap. 6) prescribe an appropriate temperature-stabilized environment for LED measurements, e.g., an integrating sphere with temperature-stabilized fixture. In Sect. 6.10 we present a combined methodology for radiometric/photometric and thermal testing.

5.2 The Interaction of the Components in a Complete Test System

At macroscopic scale the measurement instrument and the entity to be measured have typically minor interaction in a measurement. For example, a tape measure has almost no influence on the size of a measured object; neither do scales impact the weight of it. On the contrary, at atomic magnitudes a measurement severely influences the observation of the phenomena which were discussed in Chap. 4.

In case of thermal testing, the components of the test system and the DUT have much stronger influence on each other. Moreover, this interaction of tester, test environment, and DUT takes place in the electric and the thermal domain in a simultaneous and strongly coupled way. In the testing of power LED devices, this interdependence is extended to the optical domain as well [8] in [7].

In the electric domain, the testing scheme can be partitioned into subsystems such as the tester, the DUT, and the cabling between the two, but all these form a single electric circuitry.

One might think that with the proper construction of the tester, the effect of the attached cabling and DUT can be minimized; the internal sources can be considered as ideal voltage and current generators at a stable operating point. To some extent this can be achieved with careful design. However, the external circuitry can be considered as a load on the sources in the tester, not necessarily an exclusively resistive and time-invariant one. Conversely, the tester constitutes the load on the complex external circuitry composed of the DUT and the cabling. All these together may combine into a high gain amplifier with feedback through the cabling and the parasitic effects of the internal circuitry in the tester. Such a feedback can be present also through the temperature-triggered parameter shift of the DUT and can cause oscillations and runaway.

The testing scheme allows a bunch of distinguished operating points, belonging to the targeted powering of the DUT. Circuit theory offers stability criteria for the operating points in general, but these criteria are always to be interpreted on the complete set of interacting subsystems in the scheme.

The transit between dedicated operating points occurs through transients of electric and thermal nature. After applying a certain power on the DUT, a transient starts, which ends, in best case, in another stable operating point of the DUT and also of the internal circuit elements in the tester.

As discussed throughout this book, the aim of thermal transient testing is to restrict all observed and recorded transients in the complex system to those electric changes, which can be unambiguously tied to the shift of temperature-sensitive parameters of the DUT. Time-variant changes of other root cause are classified as "electric transients" and are suppressed by various means.

Several time-variant effects are inseparable from the physics of the tested device itself. These can be the recombination of charge carriers in *pn* junctions, modulation of the base length of bipolar transistors, or channel length modulation in field effect transistors. These two latter are classified as "backlash effects."

As it will be presented in subsequent chapters, these effects cannot be completely eliminated, but they can be significantly restrained with diminishing the difference of the high P_{H} and low P_{L} power levels applied.

However, other electric changes are consequences of the transit of the of tester-cabling-DUT compound circuitry between operating points. For damping the transient or suppressing eventual oscillations, the cabling has to be arranged properly, and capacitive and inductive add-ons may be needed as illustrated in some examples in this chapter and in Chap. 6.

In standard electronics design, the decay of a transient is generally considered finished after a period of a length of four or five times longer than its characteristic time constant.

In thermal testing the electric transients have much higher amplitude (in the range of volts) than the thermally induced change (in the range of millivolts, or just a fraction of them). Still, the electric transients fall off in microseconds, while the thermally induced change is to be recorded from microseconds to minutes or hours continuously.

In thermal transient testing, the data acquisition magnifies the tiny change around the aimed operating point. This way, the falloff of the transient is also recorded in a "zoomed in" way; it perturbs the thermal signal until ten times longer than its characteristic time constant.

At higher currents also the resistive and inductive properties of the cabling may add significant distortion to the thermal signal.

Even with a perfect tester construction and proper routing of the cables, a serious disturbance can be superposed on the temperature-induced voltage signal due to the coupling of the powering and the data acquisition within the device under test.

Inductive and capacitive coupling through the cabling and the DUT may distort the initial, early section of the thermal signal. This effect can be partially eliminated with differential measurement techniques.

The voltage drop on the ill-separated, common resistive sections of the external cabling or the internal wiring of a power module adds a constant error to the measurement of the power level in the heating phase. During the cooling transient, this error typically vanishes; the voltage drop at low measurement current becomes negligible.

Considering the thermal domain, the device under test and the temperature-controlled environment constitute a coupled thermal system. The thermal transient test characterizes the whole heat-conducting path, first its sections within the device and then the thermal interface and the cooling mount. The structure function methodology helps in separating the DUT and the environment, either with repeated tests using different TIM layers or based on the characteristic capacitance of structural elements in the DUT.

The structure functions also portray the TIM layer and the cooling apparatus in the test bench and identify their properties and potential failures.

5.3 Device Under Test Categories and the Related Electrical Arrangement

Thermal testing of power devices is an intricate topic which can be pictured from the perspective of the test system, considering how current or voltage are applied on the DUT and what is the appropriate measurement range in data acquisition channels. Devices are represented by a generic model in this treatment.

In this chapter the test procedure is examined according to this concept.

A deeper analysis of the behavior of particular devices in a test system is given in Chap. 6, with a focus on the temperature-related change of the device characteristics, thermal coefficients, and stability of the operating point.

5.3.1 Devices with Separate Heaters and Sensors

The way of powering and sensing is quite different for different device categories. The highest flexibility in powering can be achieved in devices with fully separated heaters and sensors.

Such devices of monolithic realization and smaller size are called *test chips*. These are used to verify packaging concepts in the design phase when the actual semiconductor device is still not available, but many parameters as the chip size, wafer thickness, and an approximate surface power pattern are already known. Test chips are broadly used for comparing the quality of die attach technologies.

Larger and more sophisticated devices with separate heaters and sensors are denoted as *thermal test vehicles*, TTVs. Using these the thermal management of a complex assembly can be optimized in a mock-up which includes the packaged device and its near environment, thermal interfaces, and cooling appliances.

Test chips and TTVs are typically realized with resistive heaters. The sensors can be resistors or diodes (Fig. 5.5); in TTVs sometimes thermocouples are also used.

In Fig. 5.5 the power step is applied by switching the I_{drive} source on or off, or between I_{H} and I_{M} levels in a more general case. A constant I_{sense} source provides the required bias for the sensor diode (or a resistive sensor).

In TTV devices the power step does not severely interfere with the thermal response transduced by the sensor. A short initial electric transient can appear on the thermal signal because of capacitive or inductive cross talk within the device, or in the cabling, the influence of this latter can be reduced with appropriate routing.

It can be observed that the data acquisition occurs on two separate channels. The voltage on the heater (V_{H} and V_{L} at the two levels of the power step) has to be measured in order to ascertain the $\Delta P = V_{\mathrm{H}} \cdot I_{\mathrm{F}} - V_{\mathrm{L}} \cdot I_{\mathrm{M}}$ power difference which is used during the postprocessing of results for Z_{th} and structure function calculations. V_{H} and V_{L} voltages are typically of several volts.

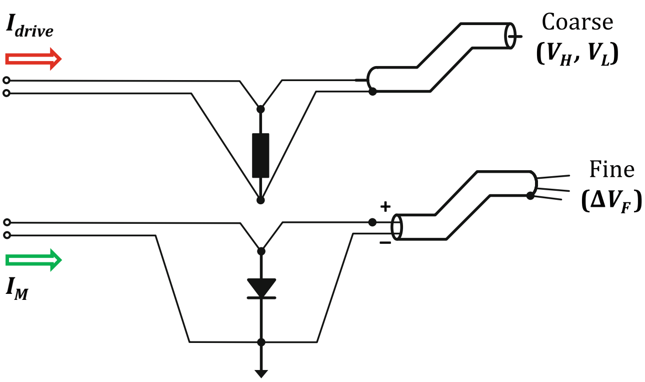

Fig. 5.5 Thermal transient measurement of a thermal test vehicle with separated powering and sensing

Another acquisition channel serves the capture of the thermally induced voltage change on the sensor. For both resistive and diode-type sensors, this change is small; a ΔV_F shift of a few millivolts occurs on the top of a larger V_F bias of many hundred millivolts or volts.

A separated force and sense wiring (four-wire method, Kelvin contact) ensures that the voltage drop caused by I_{drive} or I_{sense} on the wires of the force side is not added to the true voltage on the heater and the sensor element.

It has to be noted that the shielding on the cables in Fig. 5.5 is only needed when the heater or sensor is of resistive type, in the range above a few hundred ohms. At higher impedance the external perturbation on the measurement occurs through electric, capacitive coupling, and shielding can effectively keep off the disturbance.

At low source impedance of the "signal source," which is in this case a low resistance or a diode in forward operation, the source of perturbation is of inductive, magnetic nature. Shielding of wires does not improve the signal quality; tight routing or twisting of the wires is the recommended measure.

The disturbance through common wiring sections and the magnetic interference through grounding loops can be minimized by an appropriate grounding scheme. According to general construction principles in electronics, a single grounding point (star point) minimizes both effects.

Although this clear scheme of powering and sensing cannot be achieved in other classes of thermal systems, the principles of separation, perturbation suppression, grounding, etc. can be used for improving the quality of their thermal testing.

5.3.2 Two-Pin Devices, Diodes

In the general testing scheme of Fig. 5.2 and also in the previous scheme of Fig. 5.5, it was supposed that the I_{drive} current which ensures the powering, the I_{sense} current which provides the bias for the sensor, and the measurement channels are well separated. In electronic systems where the heating source is a device with fully or partly separated input and output ports, this separation can be reached with more or less compromise.

In many cases the powered device is a diode, or another device with just two pins. Moreover, many tutorials and thermal measurement standards recommend a "simplified" test scheme for devices having three or more pins, in which some pins are shorted so that the scheme can be traced back to a two-pin measurement.

This reduction to two pins has some merits and some drawbacks. For example, when a large integrated circuit is to be tested with a high number of pins and limited knowledge on the functionality is available, then a transient measurement on the inherent reverse substrate diode of the chip (Fig. 4.15) can serve as a suitable backup methodology. Also, this approach is the only one which enables a thermal transient measurement in a test system equipped with a single driving current source and some sensor sources only. Up-to-date testers frequently offer more sophisticated powering

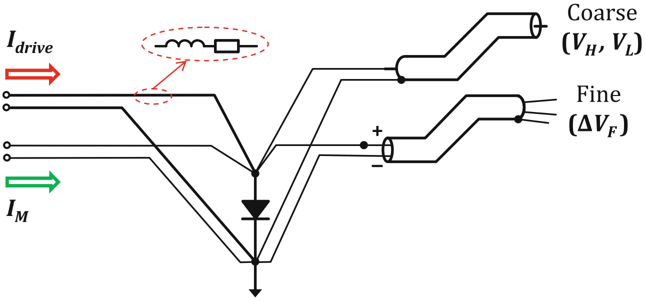

Fig. 5.6 Thermal transient measurement of a diode, powering, and sensing arrangement shown

by a variety of voltage and current sources, but these are not necessarily available at high power levels, for example, in the kiloampere range.

In Fig. 5.6 the typical connections for a thermal transient measurement of a diode are shown. One can recognize again the two current sources and the two data acquisition channels of the previous schemes, but now all these are inevitably connected at the two pins of the device.

The power step is accomplished again by applying and revoking a high I_{drive} current while keeping a constant I_{sense} current on the device. Although the clear separation between the functionalities is now lost, still, a huge change in the power at the input side (left side in the figure) does not rule out a fine detection of a temperature change-induced signal at the output (right side in the figure).

First, as proved in Chap. 4, a diode or other similar device under test has a logarithmic $I_{\mathrm{F}}(V_{\mathrm{F}})$ characteristic, at least for its internal *pn* junction.

As many conclusions in this chapter will be built on this characteristic, it is worthwhile to summarize in Table 5.1 the basic equation and some of its consequences.

All equations referred to in this chapter can be looked up either at the place of their original definition or in the table.

The Shockley equation (4.15) implies that if the ratio of I_{drive} and I_{sense} is 10:1, then the V_{Fpn} forward voltage change on the internal junction is just 60 mV at room temperature (300 K) or 80 mV at 127 °C (400 K). Even at 1000:1 ratio, this elevation is just 180 mV on the top of a typical 0.5 V–0.8 V forward voltage of a silicon diode. The voltage growth on the series R_{S} resistance is higher, but R_{S} is kept low in power devices. The logarithmic nature of the characteristics ensures a strong *decoupling* between the input and the output sides.

The same statement can be formulated for the external disturbance, caused by the inherent noise from the tester, electromagnetic interference, or other effects. These effects can be modeled as an additional current source at the input side, superposing a further excitation on the device at various frequencies and amplitudes. The measured noise voltage amplitude will be the product of the noise current amplitude and the differential (electric) resistance of the diode.

Table 5.1 The diode equation and some affiliated relationships

Equation	Reference in Chap. 4	Notes, typical values
Diode equation, *Shockley equation* for V_{Fpn}; *voltage drop* on the series resistance		
$V_F = V_{Fpn} + V_{FRs} = mV_T \ln \frac{I_F}{I_0} + I_F R_S$	(4.15)	V_{Fpn} typical value 0.5 V to 0.8 V
Differential (electrical) resistance, $dV_F(I_F,T)/\, dI_F$		
$R_D = R_{Dpn} + R_S = m \cdot V_T \,/\, I_F + R_S$	(4.16)	V_T/I_F value at 300 K 28 Ω @ 1 mA, 28 mΩ @ 1A, etc.
Temperature sensitivity, often used as TSP, $dV_F(I_F,T)/\, dT$		
$S_{VF} = \frac{dV_F}{dT} = \frac{V_F - 3V_T - W_g/q}{T}$	(4.17)	S_{VF} typical value –1 mV/K to –2.5 mV/K
Change in forward voltage at constant temperature		
$V_{F1} - V_{F2} = V_T{\cdot}\ln(I_1/I_2) + R_S{\cdot}\,(I_2 - I_1)$	(4.20)	Difference of $V_T \cdot \ln(I_1/I_2)$ at ratio 10:1, 60 mV; at ratio 100:1, 120 mV; at ratio 1000: 1, 180 mV; etc.

As it was formulated in (4.16), the R_{Dpn} differential resistance of the internal junction is around 30 Ω at 1 mA sensor current and in the mΩ range at 1 A driving current (at room temperature). This low differential value represents the source resistance of the temperature-induced voltage signal towards the data acquisition part of the tester, and ensures the strong decoupling of possible perturbations from the input side. Routing and shielding rules declared in the previous section apply again.

A clear separation of the "force" wiring at the input and the "sense" wiring at the output side prevents adding the voltage drop on the cables to the measured voltage on the device. Due to the inductivity of the cabling, at switching off a high I_{drive} current also an additional voltage spike is superposed on the "force" wiring. These effects are indicated by the resistor and inductance symbols above the corresponding cable section in Fig. 5.6.

In many testers the two measurements at the "sense" side are integrated into the same unit; the scheme of Fig. 5.6 is reduced to six wires, four for applying the driving current and the sensor current and two towards the data acquisition. Separation of the wires of I_{drive} and I_{sense} is essential at higher currents. The resistive and inductive voltage jumps at switching I_{drive} may cause wobbling effects in the sensor current and may cause initial artifacts in the measured signal.

Even in an integrated data acquisition unit, two interrelated measurement functions are to be performed, still of different required accuracy.

The measurement of the full V_F forward voltage is performed as a "coarse" functionality; the V_H voltage on the device at I_{drive} and the V_L voltage at I_{sense} have to be measured on the diode for calculating the power step. In case of a standard silicon diode, the V_H and V_L values are in the 0.5 V–1 V range, and it is sufficient to measure these at the 0.5% accuracy mentioned previously. Transient testing of a complex subassembly in a two-pin arrangement may involve a high voltage measurement, e.g., the driving current may induce a cumulated forward voltage above 100 V on a screen backlight unit built of a chain of LED diodes.

The other function of the data acquisition unit is to record the tiny ΔV_F change of the forward voltage induced by the temperature change, typically in the range of a few millivolts or tens of millivolts.

5.3.3 Discrete Devices with Three or More Pins

Three-pin devices are typically discrete with a single active component. The three accessible leads offer a wide variety of powering and sensing options. Several pins can be assigned as "power ports" where the powering occurs, and one pair of pins can be used as "sense port" where the temperature-sensitive parameter is measured.

Connecting two of their pins, three-pin devices can be "degraded"; in this case the measurement falls back to one of the methodologies used for two-pin devices.

The general thermal transient testing methods of three-pin devices will be presented in Sect. 5.5; specific methods for different device types will be shown in Chap. 6 in a detailed way.

In some designs the powering and sensing of the active element are separated; such devices have four or more pins.

Certain amplifiers and voltage stabilizers are also encapsulated in a package with three (or a few) pins. It has to be noted that the active operation of such devices may impede the fast switching between power levels as it is targeted in thermal transient testing. For relevant transient measurements, the internal complex circuitry is to be thwarted, by controlling it out of its normal operation range. More details on the measurements of amplifiers and voltage stabilizers are presented in Sect. 6.11.2.

5.3.4 Modules with Multiple Active Devices

In modules and in other subassemblies containing more devices, the powering and sensing schemes may became very intricate.

The devices are always coupled in the thermal domain, various thermal transient measurements can be carried out, and different effects of self-heating and transfer heating between devices can be examined.

In the electric domain, the voltage and current constraints are determined by the circuit topology. Independent voltage and current values cannot always be maintained on electrically connected devices.

Figure 5.7 illustrates an arrangement in which the self-heating and transfer heating between transistors in a half bridge module can be measured. The I_{drive} current is always applied on one transistor only. Several sense currents of I_{sense} value enable the transient temperature measurement on both devices.

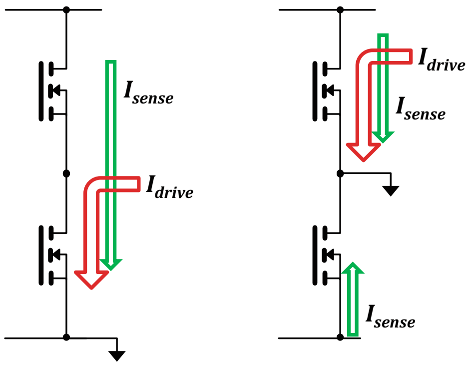

Fig. 5.7 Thermal transient measurement in a half bridge module, (**a**) low-side transistor heated, transient measured on both conducting channels, (**b**) high-side transistor heated, transient measured on the conducting channel of the high-side transistor and on the reverse diode of the low-side transistor

5.4 Thermal Transient Measurement of Two-Pin Devices

As discussed above, several device classes can be measured in the two-pin measurement scheme. In this subsection a simplified diode model will be used to outline the generic rules of this type of testing. A detailed examination of the behavior of real diodes in a thermal transient test will be given in Chap. 6, Sect. 6.1. Other examples of testing more complex semiconductor structures (bipolar transistors, MOSFETs, resistive structures) as two-pin devices will be also shown in that chapter.

The test techniques treated in this section differ in their powering and timing concept. Still, in all methodologies heating and cooling transients follow each other, caused by switching between different current levels. Some aspects on the selection of the current or power levels, especially its relation to reliability testing, will be expanded in Chap. 7, Sect. 7.4.

5.4.1 Continuous Cooling Measurements

Continuous cooling is the most frequently used thermal transient measurement technique. In this methodology the change between power levels occurs only once, as shown in Fig. 5.8.

Instead of using the notation I_{drive} and I_{sense} used throughout this book, several standards define the switching between an I_{H} *heating current* and an I_{M} *measurement current*.

Figure 5.9 shows the theoretical measurement arrangement for diodes as recommended by the classical technical literature [112] and the JEDEC thermal testing standard [30].

The scheme is clearly just of symbolic nature; the I_{H} or I_{M} current sources cannot be left open in the transient phases when they are not used.

In Fig. 5.10a, b, the practical realization of the switching is shown.

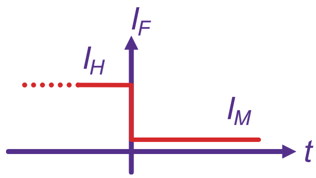

Fig. 5.8 Stepwise change of the forward current on the diode under test

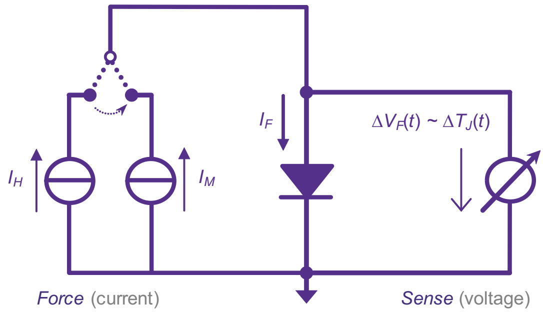

Fig. 5.9 Diode measurement scheme as defined by the JESD51-1 standard

The I_H heating current is provided as the sum of the currents of the I_{drive} and I_{sense} current sources; thus, in the heating period, both currents are applied on the device for an equalization time of appropriate duration (Fig. 5.10a). When the device reaches the "hot" steady state, the drive current is switched off.

At the moment of the switching, the transient recording of the cooling starts at I_M current bias, provided by the I_{sense} current source (Fig. 5.10b).

In this book we shall use the I_{drive} and I_{sense} notation when we refer to the instrumentation of a test, and the I_H and I_M notation when the operating point of the DUT is considered. In most cases the heating current is simply $I_H = I_{drive} + I_{sense}$ and the measurement current is $I_M = I_{sense}$.

In some modules with complex current paths, the heating and sensing currents are composed as sums and differences of several currents from different sources. In modules with simple serial elements like in Fig. 5.7, the sum of I_{drive} and I_{sense} provides the powering on the heated device; heat transfer can be measured on the other devices biased by I_{sense}.

Equation (4.17) indicates that diodes have negative temperature sensitivity at low current level ($S_{VF} \approx -1$ to -2.5 mV/K). The V_F forward voltage at high I_F forward current may have positive or negative thermal coefficient depending on the extent of the effects cumulated into the R_S series resistance. Considering moderate heating currents, the negative thermal coefficient dominates; consequently during the heating period the temperature increases and the forward voltage decreases by a ΔV_{FH} value.

At constant I_H the decrease of ΔV_{FH} results in a proportional decrease in P_H during the heating (Fig. 5.11), $\Delta P_H = \Delta V_{FH} \cdot I_H$.

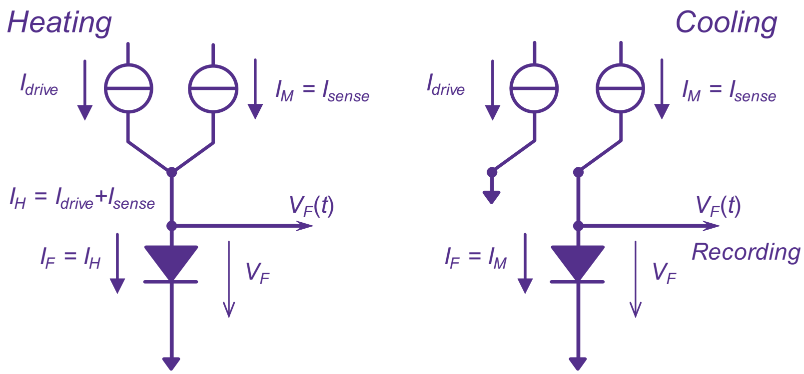

Fig. 5.10 Practical realization of the diode measurement scheme: (**a**) applying heating current; (**b**) cooling at constant measurement current

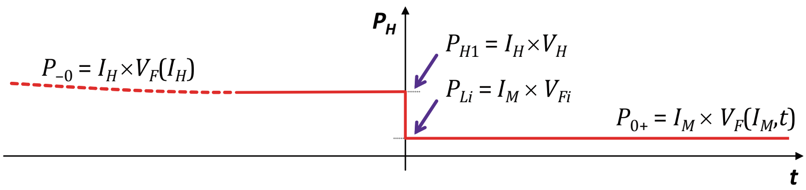

Fig. 5.11 Power change before and during the cooling measurement on a diode

The change of the heating power in time can be mathematically handled with the convolution apparatus, even if the change to lower power occurs before steady state is reached. Nevertheless, in thermal transient testing, the common practice is that I_H is maintained until a "hot" steady state is reached and P_H is considered to be P_{H1}, the value measured just before switching off.

In case of in-line measurements on the production lines, the high-powered steady state cannot be reached, because the production throughput allows only a short time heating and recording only a partial transient. The related problems are exposed in details in [106, 107].

When the test is carried out between two steady-state situations, the power levels can be calculated as

$$P_H = I_H \cdot V_H, \quad P_{Li} = I_M \cdot V_{Fi}, \tag{5.1}$$

and the power step is

$$\Delta P_H = P_H - P_{Li} = I_H \cdot V_H - I_M \cdot V_{Fi}. \tag{5.2}$$

where V_H is the forward voltage of the hot diode biased with the I_H heating current at the time instance of switching and V_{Fi} is the initial value of the forward voltage at the beginning of the cooling transient, when only the small I_M measurement current flows through the diode.

During the cooling period, the V_F forward voltage will grow; it nearly repeats the change that took place during the equalization, but now in the opposite direction. The slightly changing power during cooling can be expressed as

$$P_L = I_M \cdot V_{Fi} + I_M \cdot \Delta V_F(t) \tag{5.3}$$

If the I_M measurement current is significantly lower than the I_H heating current, then the *power change* during cooling can be neglected.

Example 5.1 Thermal Transient Testing of a Medium Power Rectifier Diode

In order to follow the course of a typical thermal transient measurement on a medium power rectifier diode, let us consider powering parameters as $I_H = 20$ A and $I_M = 100$ mA. Assuming a forward voltage of about 0.8 V at I_H and 0.6 V at I_M, the total power step will be (20 A × 0.8 V) – (100 mA × 0.6 V) = 15.94 W.

Suppose that the temperature sensitivity of the diode is −2 mV/°C at the measurement current and the temperature change during the transient is 50 °C in the actual cooling environment. Accordingly, the temperature-induced forward voltage change is 100 mV. The error term (power instability during the cooling) will be $I_M \cdot \Delta V_F(t) = 0.1$ A × 0.1 V = 10 mW which is about 0.06% of the total power change. Thus, at low ΔV_F and low I_F, the change of the power on the device is a secondary order effect only. In such a way, the cooling starts with a nearly perfect *power step*.

After switching off first a sudden large voltage jump occurs on the diode – the V_F forward voltage sinks from the V_H value belonging to the "hot" diode characteristics at I_H to the lower V_{Fi} value belonging to the "hot" diode characteristics at the smaller I_M (Fig. 5.12).

This change, referred to as electrical transient, can be many hundred millivolt for silicon diodes, and for diodes with higher series resistance even more than one volt. The voltage change is usually in the 10 μs–100 μs range, because a large amount of stored diffusion charge has to be removed from the forward-biased *pn* junction (see, e.g., [10]).

After this electric transient, the pure thermal part of the transient can be recorded. The V_F forward voltage slowly increases from its V_{Fi} initial value as the operating point moves from the "hot" to the "cold" diode characteristics, reaching its final, V_{Ff} value, always at I_M bias (Fig. 5.13).

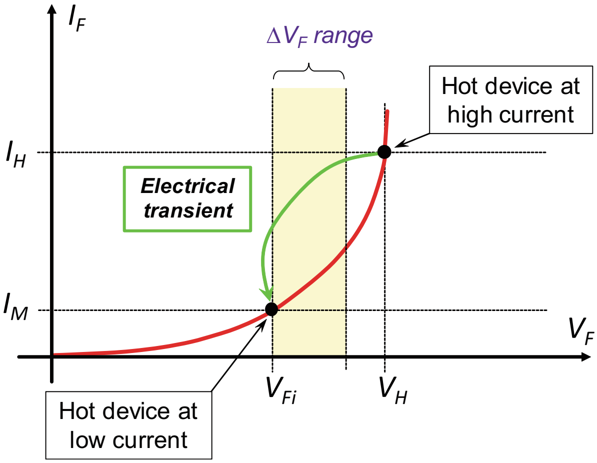

Fig. 5.12 Electrical transient of the hot diode shown in its *I–V* characteristics, switching from the heating current to the measurement current. This chart is an enlarged portion of Fig. 6.1

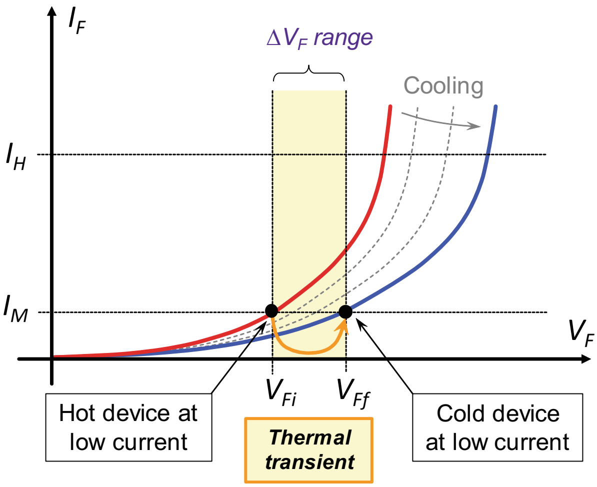

Fig. 5.13 Thermal (cooling) transient of the diode at constant measurement current shown in its *I–V* characteristics. This chart is an enlarged portion of Fig. 6.1

The measurement channel has to be capable to record the complete transit of the V_F voltage between its hot and cold value, in an appropriate "fine" measurement range covering ΔV_F. Still, the fast travel of the signal during the electric transient is not recorded always in its entirety; the signal may jump into the "fine" measurement range from an "out of range" condition. The necessary ΔV_F range of recording is highlighted in Figs. 5.12 and 5.13.

The V_H and V_{Fi} endpoints are to be recorded in the "coarse" range of the data acquisition unit.

For achieving proper signal to noise ratio in the record, it is often proposed to select the smallest "fine" measurement range in which the temperature-induced signal still fits. Selecting larger range enhances quantization noise on the measured signal.

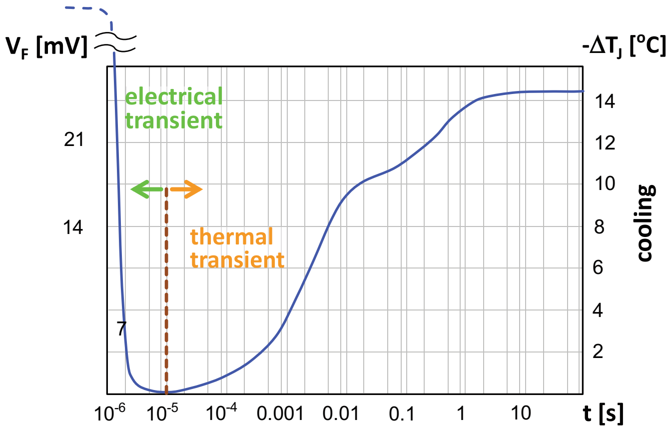

Fig. 5.14 Recorded transient of an actual diode, scaled in voltage (on the left) and in temperature (on the right)

In Fig. 5.14 a recorded transient of a diode is shown; the electric transient finishes at approximately 10 μs.

Various techniques are available to reconstruct the early part of the measured transient which is covered by the electric distortion. Some of these are based on the considerations formulated in Example 2.6 of Chap. 2. A detailed analysis on the initial transient correction techniques is given in Chap. 6 Sect. 6.1.4.

So far in the examples, the I_H and I_M values used in the measurements have been taken for granted. Even in the daily engineering practice, their selection is often based on data sheet values and local traditions.

As thermal testers are typically sensitive, for analyzing the structure integrity, a relatively low power level is sufficient to ensure a few centigrade temperature elevation. However, in case of nonlinearities in the material parameters of the devices, it is safer to accomplish the measurements with the power levels that are foreseen in the normal operation. For reliability testing always the normal operational power has to be ensured. For accelerated reliability testing, higher than normal power has to be applied unto the device. A comparison of results measured at various heating currents is presented in Chap. 6.

A common misconception broadly publicized in the literature is that the measurement current should be kept low in order to avoid the self-heating of the component. In reality, self-heating has only some minor influence in an *absolute* measurement technique, and can be nearly neglected when using a *differential* technique. That means that sensor current values can be freely selected in a very wide range.

One of the advantages of selecting higher measurement currents is that the initial electric transients are smaller and faster. Moreover, at higher measurement current, the noise on the measured signal is lower and external disturbances are better suppressed. The physical background of these effects is explained in depth in Sect.

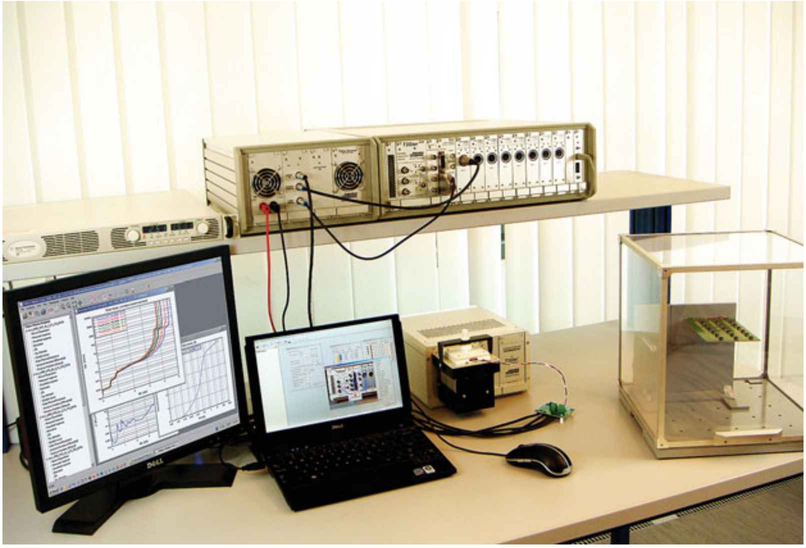

Fig. 5.15 Typical thermal transient measurement arrangement: test equipment, thermostat for TSP calibration or to be used as a cold plate, natural convection test environment, and PC with measurement control and data processing software

5.7. A comparison of results measured at various measurement currents is presented Chap. 6, Sect. 6.1.4.

An actual thermal measurement arrangement with the tester and the equipment realizing the thermal boundary is shown in Fig. 5.15.

5.4.2 Continuous Heating Measurements

In these measurements, only the sensor current is applied to the device for an equalization time of appropriate duration.[1] When the device reaches the "cold" steady state (the equilibrium at $P_{\mathrm{H1}} = I_{\mathrm{M}} \cdot V_{\mathrm{F1}}$) the drive current will be switched on in addition to the sensor current, and the transient recording of the heating will start.

[1] In a strict sense, heating transients can be carried out without an applied measurement current. Still, because of the difficulties of calibration at high heating current, in practical cases heating measurements always are carried out with interlaced cooling transients. The calibration of the heating occurs with fitting towards calibrated cooling transients. In many tester solutions, "single power pulse" and "multiple power pulse" test modes are provided for various test purposes. These test modes play a primary role at thermal quality test applications in production where the measurement time is inherently limited.

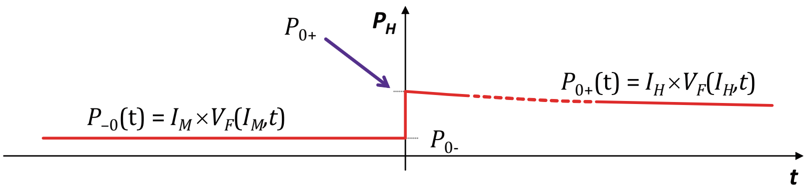

Fig. 5.16 Power change before and during the heating measurement on a diode

Most considerations introduced for the cooling operate in the same way. However, here we experience a larger power change on the hot device during the recording time; approximately the opposite forward voltage change of what we saw at cooling measurements occurs now at high current. The powering is far from an ideal step function, as shown in Fig. 5.16.

This problem can be again mathematically handled, but it is easier to use the cooling method for the thermal measurements of two-pin devices.

Both the cooling and the heating measurements described up to this point correspond to the transient extension of the JEDEC JESD51-1 “static” test method, defined in [30].

5.4.3 Pulsed Measurements

Older tester types use the so-called “dynamic” method, also defined in [30]. The measurement principle is illustrated in Fig. 5.17. In this method the measurement is based on a series of high current pulses for heating and switching back to low current for temperature recording.

The temperature value T_1 corresponding to t_1 time instance is measured such that a power pulse with a duration of t_1 is applied to the device under test. When t_1 time is elapsed, the power pulse is switched off (switching from I_H heating current to the I_M measurement current), and with a short delay t_{MD} (called the measurement delay), the value of the TSP (forward voltage) is measured and through the K-factor is converted to junction temperature. Then, the device under test needs to cool down; the cooling time must be at least as long as t_1. Then, the process is repeated for a longer pulse width t_2, etc.

The test result (referred to as the heating curve) is composed from these responses to individual heating pulses of different length. (Such heating curves obtained by two different commercial testers can be found in the technical literature, e.g. in [113] and [114].) This technique distorts each recorded point by an electric transient, and the data correction problem (back extrapolation of the measured i-th junction temperature at $t_i + t_{MD}$ time instance to the t_i time instance) is also present at every data point of the composed Z_{th} curve. Last but not least, the physical time needed for the measurement by the dynamic test method is by orders of magnitude longer than the length of the real cooling transient measured by the transient extension of the static test method, as demonstrated in [115].

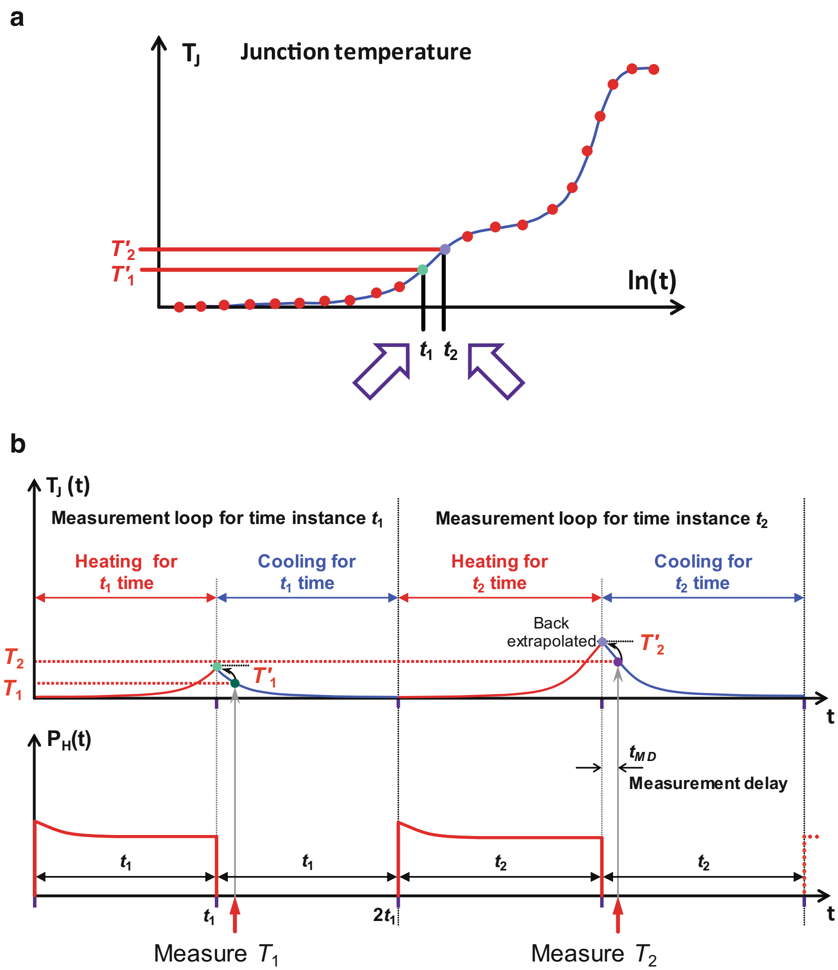

Fig. 5.17 The principle of the JEDEC JESD51-1 dynamic test method: (**b**) the series of heating pulses and corresponding temperature responses, (**a**) heating curve composed of individual temperature values measured at the end of the heating pulses

5.5 Measurement of Three-Pin Devices

Three-pin devices offer certain flexibility, as outlined in Sect. 5.3. Several pins can be assigned as "power ports" where the powering occurs and one pair of pins can be used as "sense port" where the temperature-sensitive parameter is measured.

We shall discuss below the measurement of bipolar junction transistors, but all considerations apply in the same way to MOSFETs, IGBTs, etc.

In the powering and sensing scheme of three-pin devices, one pin is shared between the input and output port. Figure 5.18 presents the common base

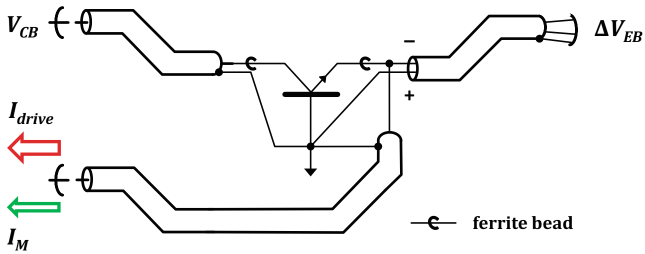

Fig. 5.18 Measurement of a transistor in common base configuration

arrangement, in which the collector-base "port" is of high impedance while the emitter-base "port" is of low impedance. This determines the type of the power sources, which can be used, a V_{CB} voltage is sustained on the collector-base "port" and an I_E current is applied on the emitter-base "port".

Figure 5.18 corresponds to the measurement of an npn bipolar transistor (positive V_{CB} voltage, negative I_E current), but all statements are valid for pnp devices as well, with appropriate signs.

Transistors in the common base setup can be characterized with high gain and high bandwidth; the device with the source units in the tester and the cabling forms a high gain amplifier with feedback, which tends to oscillate at a frequency of many megahertz. This also justifies the extensive use of cooling transients in device characterization. The characteristic amplification parameters of three-pin devices (current gain of bipolar devices, transconductance of MOSFETs) are lower at the operating point of the low measurement current; this makes the DUT-cabling-tester compound less susceptible to oscillations.

Ferrite beads and capacitors are often added to the circuit scheme in order to lower the cutoff frequency.

The measurement scheme in the figure implies that the power on the device can be altered in two obvious ways.

In one mode a fixed V_{CB} voltage is maintained between the collector and emitter, and the I_E emitter current jumps between a higher and a lower level. This mode resembles very much the previous measurement technique introduced for two-pin devices, and is also called *current jump mode*.

Another way for inducing a power step is a sudden change in the collector-emitter voltage, at steady emitter current. This mode is called *voltage jump mode*.

More details about measuring transistors and other three-pin devices will be presented in Sect. 6.1.

5.5.1 Current Jump Measurement of Three-Pin Devices

In *current jump mode*, the power can be calculated as the sum of a $P_{CB} = I_E \cdot V_{CB}$ and a $P_{EB} = I_E \cdot V_{EB}(I_E,T)$ constituent. I_E changes between an I_H and an I_M level.

The measured temperature-induced signal is the $V_{EB}(I_M,T)$ emitter-base voltage at I_M current.

This measurement mode has many advantages. For example, the highest power step can be achieved on a three-pole device in this mode, without special fast high voltage switches in the tester equipment. Due to low power at the low current state, the calibration of the devices can be easily carried out.

On the other hand, the charging or discharging of the diffusion capacitance of the forward-biased emitter-base *pn* junction can be slow; a long electric transient is to be expected at the beginning. Figures 5.12 and 5.13 are valid again for the electric and thermal transient sections.

A significant drawback of the measurements on diode-like devices is the limited power step, constrained by the I_H heating current and the V_F forward voltage. In case of three-pin devices, the high voltage on the collector or drain pin allows a significant increase of the applied power.

In Example 5.1 the diode was powered at $I_H = 20$ A, $I_M = 100$ mA, which resulted in $V_{FH} = 0.8$ V and $V_{FM} = 0.6$ V corresponding forward voltages. The power step was calculated as $\Delta P_D = (20\ \text{A} \times 0.8\ \text{V}) - (100\ \text{mA} \times 0.6\ \text{V}) = 15.94$ W.

Suppose a transistor is measured in the arrangement of Fig. 5.18, in current jump mode with identical I_H and I_M, at $V_{CB} = 20$ V collector-base voltage. The power generated on the collector-base junction also heats the device; further $\Delta P_{CB} = (20\ \text{A} \times 20\ \text{V}) - (100\ \text{mA} \times 20\ \text{V}) = 398$ W adds to the power step.

The error term (power instability during the cooling) calculated for the diode in Example 5.1 was $I_M \cdot \Delta V_F(t) = 0.1\ \text{A} \times 0.1\ \text{V} = 10$ mW which was about 0.06% of the total power change; now this 10 mW is to be related to the 413.94 W total dissipation.

Details of these testing schemes are presented in Sects. 6.2 and 6.3.

5.5.2 *Voltage Jump Measurements of Three-Pin Devices*

In *voltage jump mode*, a steady I_E current has to be maintained, and the voltage is to be switched between a high and a low V_{CB} value for generating a power step (Fig. 5.18). The $V_{EB}(I_M,T)$ voltage change is the temperature-dependent signal again which can be captured. Using different pins, now powering and sensing are well separated.

This measurement mode offers the best resolution when mapping the fine details of the thermal structure, belonging to shortest time constants.[2]

In voltage jump mode, the I_E emitter current does not change; such the diffusion charge profile in the emitter-base *pn* junction remains nearly unchanged. The electric

[2] It was presented in Chap. 4 that in bipolar devices, the major source of the electric transient is the diffusion capacitance.

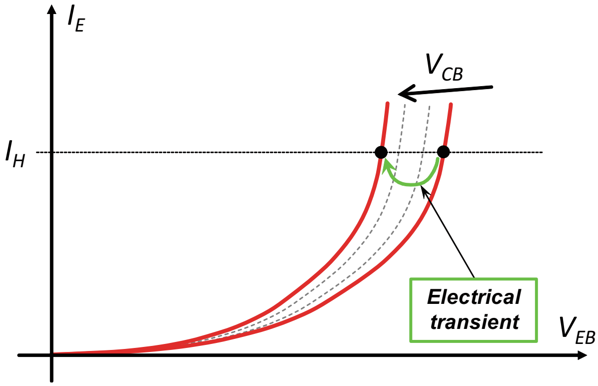

Fig. 5.19 Electric transient of a transistor after switching V_{CB} from high to low voltage

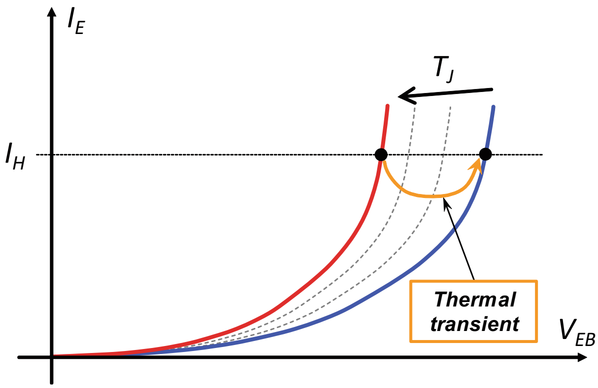

Fig. 5.20 Thermal transient of a transistor after switching V_{CB} from high to low voltage

transient is very fast, just a few μs, which is near to the slew rate of the power sources in the best transient testers.

The electric transient is caused by a backlash effect introduced in Sect. 5.2.

The physical background of backlash is at bipolar transistors the Early effect: at higher V_{CB} the base length of the transistor shortens. This effect is small and very fast.

MOSFET devices have a much stronger backlash, especially at short channel devices. Here the cause of the backlash is the channel length modulation of the device, i.e., the increase of the depletion layer width at the drain as the drain voltage is increased.

At both device categories, an elevating V_{CB} voltage causes some decrease in V_{EB} while I_E stays constant (using the indices for bipolar device). Figure 5.19 illustrates the electric transient at a sudden change from higher to lower V_{CB} voltage.

After this electric transient, the cooling can be observed as growing V_{EB} forward voltage at steady I_E current (Fig. 5.20).

The powering in this mode is not limited by the forward voltage on the emitter-base junction either. In case of, e.g., constant $I_H = 2$ A and a V_{CB} jump from 20 V to

2 V, the power step on the collector-base junction is obviously $\Delta P_{CB} = 36$ W. The power change on the emitter-base junction is caused by the backlash effect and is typically insignificant compared to ΔP_{CB}.

The disadvantage of the method is that the calibration of the devices should occur at higher power.

Assuming $V_{EB} = 0.8$ V forward voltage at 2 A, similarly to the previous section, the calibration is to be carried out at $P_M = 2.8 \text{ V} \cdot 2 \text{ A} = 5.6$ W power. This level already needs a precise calibration arrangement with negligible change in the R_{thJA} junction to ambient thermal resistance (see Sect. 5.6).

Details of measuring various three-pin devices will be presented in Chap. 6.

5.6 The Voltage to Temperature Calibration Process

The calibration process connects a measured temperature-sensitive parameter value to the actual device temperature. Its precision is of high importance because this step influences the overall accuracy of the measurement. All other steps in thermal measurements are practically voltage and current measurements for which instruments of high precision and high time resolution are available. On the contrary, it is easy to perform a bad calibration and undermine the validity of thermal data.

5.6.1 *The Temperature-Sensitive Parameter*

The most often used parameter for temperature sensing is the forward voltage of a diode-like structure. Previously we defined this parameter as $dV_F(I_F,T)/dT$ and deduced an appropriate formula for it in (4.17) and also in Table 5.1.

In the thermal transient testing, other thermally sensitive parameters can be equally used. In this section we denote the TSP as V_F, but substituting it by other electric parameters, all considerations apply in the same way.

Thermostats produce highly repeatable boundary conditions at various temperature levels. So they can be used for *temperature-voltage calibration*, which means *recording* V_F forward *voltage* (or other parameter) values *at different component temperatures*.

We have to make a difference between *absolute* and *relative* calibration. The former means mapping the $V_F(I_F,T)$ function, the latter deriving the S_{VF} sensitivity parameter only.

In case of relative calibration, the temperature-related change of the parameter is approximated by

$$V_F(I_M, T_J) \cong V_{F0}(I_M) + S_{VF} \cdot (T_J - T_{J0}). \tag{5.4}$$

Absolute calibration has a power-dependent error, but is exempt from nonlinearity problems.

For operating a component, some P power needs to be applied on it. Supposing all paths in the heat conduction path arrive at the same temperature-controlled surface named X, one experiences a

$$T_J = P \cdot R_{thJX} + T_X \tag{5.5}$$

junction temperature value, the junction temperature differs from the controlled temperature.

For the differential measurement principle mentioned in Chap. 3 Sect. 3.1.5, only the S_{VF} sensitivity value (or its reciprocal, the *K-factor*) is to be determined in the calibration process; this approach remains valid as long as the temperature to voltage mapping is obviously linear.

The power-dependent error of the absolute calibration procedure expressed in (5.5) has no practical consequences when the calibration occurs at low power and the result is used in cooling measurements, as demonstrated below in Example 5.3. In fact, there is no reason to use the "sensitivity" or "K-factor" quantities in the measurement evaluation; up-to-date testers are able to provide an absolute temperature calibration. The only reason to use a sensitivity value instead is when many semiconductor devices of the same manufacturing batch are to be compared. Typically, the reproducibility of the samples of the same batch is better than the repeatability of the calibration procedure; in such a way, forcing the same sensitivity value to all samples ensures a comparison of higher accuracy.

It has to be noted that during the calibration, the chip surface and the junction are practically at homogeneous T_J temperature due to the low power applied. In the actual test process at the beginning of a cooling transient, the chip surface has a bell-shaped temperature distribution, as illustrated in Fig. 3.17. The recorded T_{VJ} virtual temperature is a weighted average of the temperature distribution over the chip surface. Still, as Eq. (4.21) and Table 4.3 in Sect. 4.3.1 indicate, the hot center is represented with an overwhelmingly high weight in the average; T_{VJ} corresponds to the hottest point. During the cooling transient, the temperature of the chip surface levels out; the measured junction temperature matches the calibrated value.

In this section we focus on the calibration of devices with single energy transport, where all input energy is converted into heat. Further important aspects of the TSP calibration will be expounded in Chap. 6 Sect. 6.10.5.

5.6.2 Calibration on a Cold Plate

Components having a large cooling surface (case, tab) where most heat flows through can be measured and calibrated when mounted on a cold plate (Fig. 5.21). This is actually an easy process:

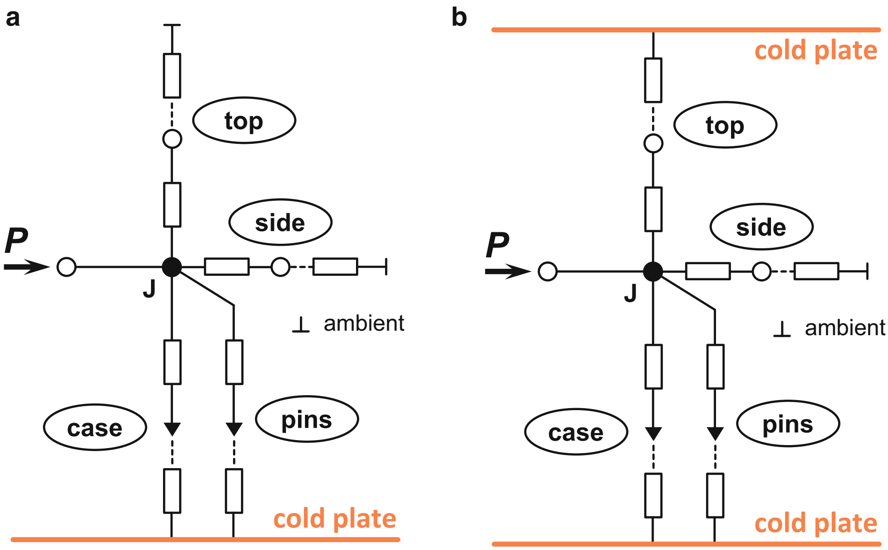

Fig. 5.21 Calibration of a component mounted on single and dual cold plate

- Set the cold plate temperature to several values.
- Record the corresponding V_F forward voltage.

Despite its simplicity we have to be aware of some important issues for doing the calibration correctly:

- As S_{VF} depends on the operating point, always *apply the voltage and/or current* on the component to be calibrated, which *corresponds exactly to the transient measurement* circumstances.
- For cooling measurements, this should be the lower, for heating the higher powering of the two levels used in the differential method.
- For diode-type components, this practically means applying the sensor current only if cooling (or pulsed method) is the selected transient type. The same is true for other devices in current jump mode.
- The power sinking capability of (liquid circulator driven) cold plates is high, even the calibration at high power needed for heating can be carried out easily.

Figure 5.21 shows the simplified model of a packaged device mounted on a single-side and a dual cold plate. The figure reveals that some surfaces of the package are terminated by the ambient (room temperature) rather than by the temperature-controlled plate. The junction temperature is "downscaled" by the appropriate thermal resistances in the thermal circuit; it does not follow exactly the set point of the plate.

Figure 5.22 shows the consequences. The dashed line labeled as "absolute" shows the $R_{thJX} = 0$ case ("chip not packaged, just attached to the cold plate"). The solid SC curve corresponds to the forward voltage of the packaged component

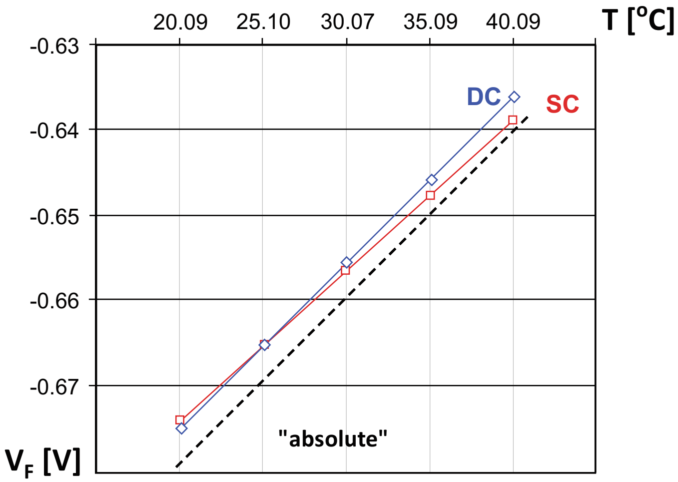

Fig. 5.22 Example on the absolute calibration error

on a single cold plate, while the solid DC curve corresponds to the dual cold plate case. The temperature axis is scaled in the measured *cold plate temperature*. V_F is negative (calibration of an anode-grounded device).

If the cold plate is set to room temperature (25 °C in the figure), then the SC and DC curves coincide. The junction-to-ambient thermal resistance can be derived from the plot, using the $T_J = P{\cdot}R_{thJA} + T_A$ equation. At all other temperatures, the junction is between the ambient and the cold plate temperature; we underestimate the actual S sensitivity.

Figure 5.21 also hints that if a large portion of the heat leaves through the pins (package with small tab and many pins), also a good thermal contact is needed between the wires feeding the package through the pins and the cold plate.

The calibration process can be manual or automatic, using calibration software. In both cases the following steps have to be carried out:

- Select four or five temperature set points, spanning the whole temperature range of the future measurement.
- Apply the appropriate power on the device.
- Program the lowest temperature value, and wait for t_1 time until the cold plate temperature stabilizes.
- Checking the component voltage wait for t_2 time until the voltage stabilizes.

t_1 and t_2 waiting times are needed because the "thermal resistance" elements shown in Fig. 5.21 are complex impedances; their capacitive part expresses heat storage in different material sections. If t_2 is too short, we face the problem shown in Fig. 5.23. The component voltage (dynamic curve labeled "fast" in the figure) follows the cold plate with some delay; at the lowest temperature point, we also see the previous cooling from room temperature.

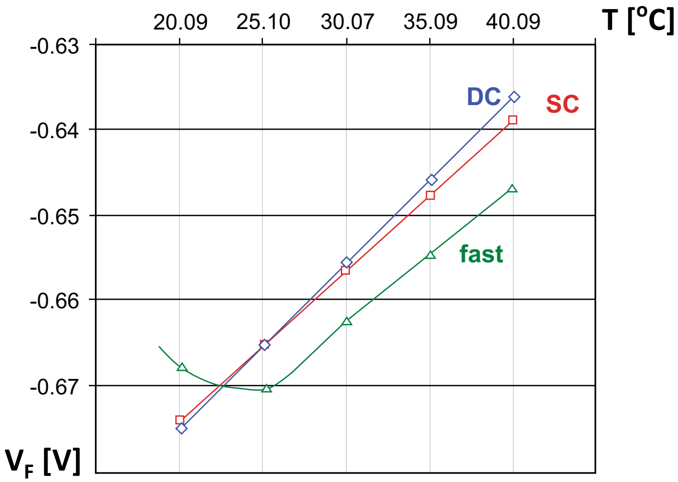

Fig. 5.23 Example on the dynamic calibration error

Even waiting for long t_2 times, we cannot get rid completely of the dynamic effect. On the other hand, even with long t_1 times, we experience small changes also in the cold plate temperature due to the control loop of the liquid circulator. The best practice to minimize these effects is:

- Record the actual temperature of the cold plate after a t_1 stabilization time instead of the programmed set point temperature (as in Fig. 5.23).
- Omit the first voltage reading at lowest temperature.
- Try to read all voltage values at equidistant $t_1 + t_2$ times (in such a way the dynamic curve runs parallel to the SC or DC curve).

Example 5.2: Calibration of a Package on Single and Dual Cold Plate
A flat power package with exposed cooling surface (tab) is calibrated in single-side cold plate setup.

Suppose that in the model of Fig. 5.21 the junction-to-top resistance is approximately 10 K/W and the still air cooling on the top surface corresponds to approximately 20 K/W.

The parallel path composed of the junction-to-case and junction-to-pin resistance is 2 K/W, the spreading in the cold plate is below 0.1 K/W. The wires towards the pins are in good thermal contact with the cold plate. The junction-to-side path is approximately 200 K/W.

Without detailed calculations we can see that nearly 10% of the heat leaves towards the ambient; we underestimate the sensitivity by almost 10% in this setup.

In the dual cold plate setup, we have some heat loss towards the sides only. However, as there is a thin air gap between the two temperature-stabilized metal plates, the air is practically of the same temperature as the cold plate. The side resistance is connected to the cold plate temperature rather than to room temperature.

5.6.3 Calibration in a Closed Chamber or Bath

In case all branches of the heat conductance path end at the same temperature, many problems of the previous section are automatically solved. This is the case when using a closed chamber with thermo-electric heating and cooling, or a liquid bath (Fig. 5.24).

Otherwise, the way of programming temperatures and reading voltages is much the same as in case of cold plate calibrations.

The t_1 and t_2 equalization periods are not fixed values. Instead, we should accept the state as thermal equilibrium if the changes of the chamber temperature remain below a given limit in a given time window, and after reaching this, also the measured voltage does not change more than a predefined limit for a similar time window.

For cold plates and chambers, the actual liquid or plate temperature can be different from that of the sensor used for controlling the system. In real life 3–5% repeatability can be expected for TSP calibration.

As mentioned before, all other steps in thermal tests need some sort of voltage and current measurement instruments which are of high accuracy and stability. Recalibration of these is also only very rarely needed. Calibration thermostats need regular recalibration using stable reference devices.

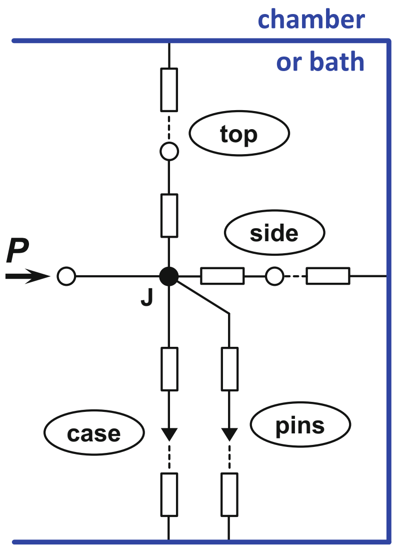

Fig. 5.24 Calibration of a component in closed chamber or bath

Example 5.3: Calibration of Devices Mounted on Printed Board in a Bath

The natural thermal environment for surface mount devices (SMD) is being mounted on a printed board. For this reason, their thermal transient test typically occurs on test boards similar to the design of Fig. 5.3. The boards represent a reproducible thermal boundary, and their edge connectors provide electrical access to the relevant pins of the device for powering and for measurement.

Several guidelines for thermal testing propose the calibration of board-mounted devices in a temperature-stabilized liquid bath filled with dielectric oil. Although this approach seems to be rational because the dielectric oil ensures homogeneous temperature on the whole board and package surface, the handling of the boards contaminated with sticky, greasy material during the test is rather inconvenient.

Many years of practice confirms that good thermal contact towards the liquid and the electric insulation of the board can be realized in an easy way, without compromise.

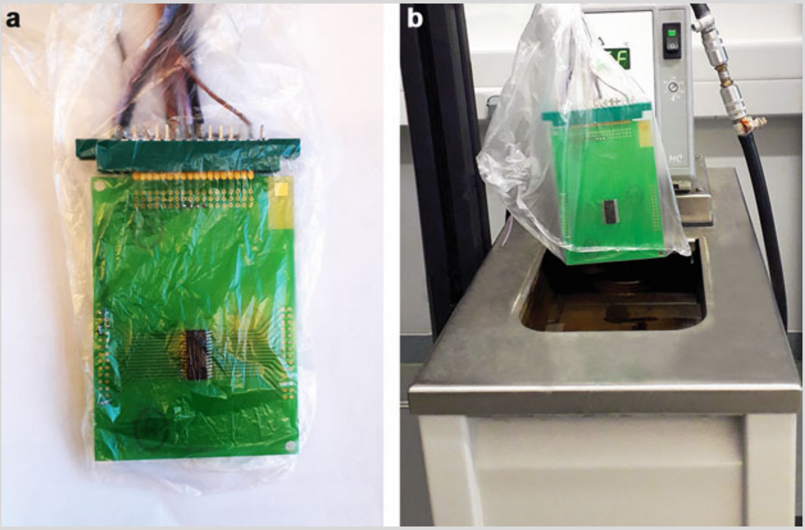

Fig. 5.25 Water-filled liquid circulator with open vessel. Printed board placed in three plastic bags prepared for TSP calibration

A simple test arrangement is presented in Fig. 5.25. After plugging the edge connector of the board into an appropriate socket with the powering and sensing wires, the board is placed into thin plastic bags of appropriate size. Two or three plastic bags layers are recommended in order to avoid leakage problems.

(continued)

Example 5.3 (continued)

Immersing the board in bags into the vessel of a liquid thermostat, an air bubble of continuously changing shape and volume builds around the board. Now, analyzing the influence of this fluctuation on the measured TSP value, we find that it is negligible.

Suppose the junction to ambient thermal resistance from the device to the ambient (which is in case the water in the vessel) changes massively, for example, between 10 K/W and 20 K/W.

If the forward voltage on the device is 0.7 V when it is driven by 10 mA sensor current during the calibration, the resulting power dissipation will be 7 mW. The difference of the junction temperature from the ambient will swing between 70 mK and 140 mK as the size of the air bubble changes. That is, in a calibration temperature sweep of 50 K, the related error is below 0.15% which is negligible compared to other errors of the calibration process.

Still, it has to be emphasized again that the thermal transient measurement has to be carried out in cooling, at the same low measurement current where the calibration occurred.

It is advisable to compose the calibration curve of five or more TSP versus temperature pairs, in order to detect possible nonlinearity. The typical time needed to stabilize the temperature of both the thermostat and the PCB-mounted device is around 20 minutes for each set point in liquid bath. For devices attached to a TEC thermostat plate, 5 minutes is expected. Typical calibration software tools automatically check the stabilization of the thermostat temperature and of the TSP. These values are considered to be stable if their change remains within a predefined limit in a predefined time window, e.g., the change is less than 0.1 °C and 2 mV in 120 seconds.

5.7 Noise and Immunity in the Thermal Measurements

Thermal measurements operate on very low signals; accordingly they are most susceptible to noise and external interference. In a way they are prodigal, investing many watts or kilowatts for heating the thermal response can be just a few millivolts.

The electrical noise and the external interference in thermal measurements have several distinct sources.

One major source of the noise superposed on the actual measured thermally induced signal is the internal noise of the device (shot noise in case of diodes).

The amplitude of the shot noise can be derived from the equations of the thermodynamics. At a certain bandwidth of the measurement, it is

$$i = \sqrt{4kT\Delta f / R_D} \tag{5.6}$$

where k is the Boltzmann constant in joule per kelvin, T is the absolute temperature in kelvin, Δf is the bandwidth of the measurement and R_D is the dynamic electrical resistance of the device under test.

From (4.16), the differential resistance of a diode is inversely proportional to the I_M measurement current, $R_D = mV_T/I_M$, and such

$$i = \sqrt{4kT\Delta f \cdot I_M / mV_T} \tag{5.7}$$

the noise current grows with the square root of the I_M measurement current. This result is physically sound; a larger particle flow involves larger fluctuations.

However, the measurement records the resulting voltage fluctuation, which is

$$v = i \cdot R_D = \sqrt{4kT\Delta f \cdot mV_T / I_M}, \tag{5.8}$$

the internal noise in the forward voltage diminishes with the square root of the measurement current.

Selecting a higher measurement current is also advantageous for lower susceptibility to external interference. The thermal measurements are normally not carried out in a shielding cage; signals from powerful high-frequency sources such as broadcast stations, motors, and similar are demodulated and added to the temperature-related parameter change. The measurement current source of the thermal tester also produces some unavoidable noise current.

The noise components of the internal device noise, the external perturbations, and the tester yield an i_{noise} noise current compound. (The *power* of independent noise generators is to be added up which results in a square root growth of the resulting noise *voltage*.)

However, the testers measure the device voltage, along with the noise voltage. Supposing that the current generators in Fig. 5.10 yield both the necessary bias and the inevitable noise, we can state that the R_D dynamic resistance of the diode shunts the noise current. The resulting noise voltage can be given as

$$v_{noise} = i_{noise} \cdot R_D \tag{5.9}$$

Accordingly, the voltage of all perturbations shows square root decrease with higher measurement current.

Electrical noise calculation plays an important role in tester construction, but it also has importance for a broader audience when selecting the measurement current for thermal transient measurements. A higher measurement current results in lower

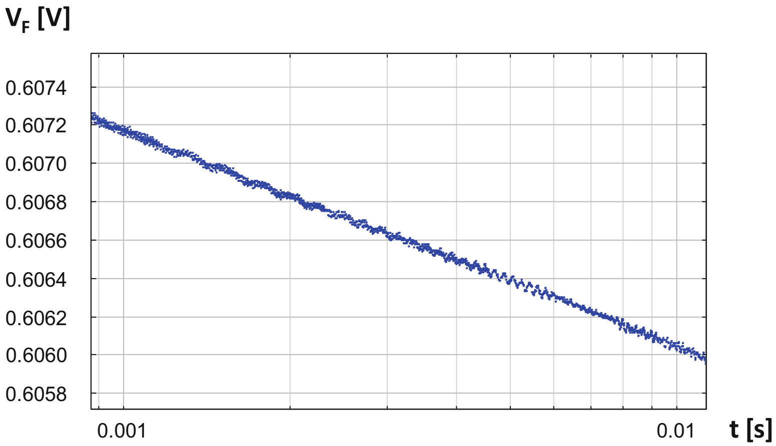

Fig. 5.26 High-frequency interference from an external source on a measured thermal signal

noise accordingly to (5.8). It is also advantageous for shorter electric transients as shown in Fig. 2.68.

A general practice is to select the lowest measurement current with already acceptable noise. For a broad category of silicon devices 1 mA and for LEDs, 10 mA is an acceptable value. Larger modules may need a measurement current of several amperes.

As it can be calculated from (4.16) at $I_F = 1$ mA and $m = 1$ ideality factor, the differential resistance of a pn junction is $R_D = 26\ \Omega$, at $I_F = 10$ mA it is $R_D = 2.6\ \Omega$, and so on. This low source resistance of the thermal measurements effectively shunts not only the noise content from the noise sources discussed so far but also the possible external interference superposed on the wires of the measurement arrangement by capacitive coupling. This makes the shielding of force or sense wiring in Fig. 5.6 actually unnecessary.

On the other hand, this low source resistance makes the loops formed by the wiring of the current source and the device under test (Fig. 5.6) susceptible to electromagnetic induction. External electromagnetic sources can cause observable interference (Fig. 5.26).

This can be prevented by clear separation of the force and sense wiring (Fig. 5.6), and twisting of the wires on the sense side.