On $V_{ce}$ method: in-situ temperature estimation and aging detection of high-current IGBT modules used in magnet power supplies for particle accelerators

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Abstract—Magnet power supplies used in particle accelerators are a specialized application of power electronics that requires long lifetime operation, with no unscheduled interruptions and high output current precision. In this paper, the $V_{ce}$ method with sensing current is considered for the estimation of the junction temperature and the aging detection of high-current IGBT modules used in this specialized application. Firstly, sensitivity analysis of the sensing current and the gate-emitter voltage are documented to quantify their impact on the precision of the $V_{ce}$ method for high-current modules independently of the application of interest. Two measuring circuits offering ease of calibration and reduced number of components are proposed but their applicability is not limited to this application. Their performance is tested on a 1.6kA IGBT module that exhibits the highest current ratings documented for the $V_{ce}$ method with sensing current. The test is carried out at a switching frequency of 6.5kHz using an experimental set-up that emulates a phase-leg of the considered power supply. The $V_{ce}$ method with sensing current is evaluated for the aging detection of high-current IGBT modules in general and its effectiveness for the specialized application is verified experimentally.

Index Terms—Insulated Gate Bipolar Transistor (IGBT), temperature monitoring, $V_{ce}$ method, measuring circuits design, aging detection, particle accelerators, magnet power supplies

I. INTRODUCTION

Temperature estimation and aging detection of IGBTs are important to avoid unexpected failures and to accordingly plan the service stops for power electronic converters. Specialised applications have various requirements that need to be taken into account in the selection of the method that performs such estimation and detection, as well as in the design of the circuitry that is used for the method’s implementation.

One such specialised application is the magnet power supply of particle accelerators at CERN, the European Organization for Nuclear Research, which requires high precision for the output current and long life with no unscheduled interruptions. Such a magnet power supply is shown in Fig. 1, where the key parameters of the power circuit are also given. The specialised application in question requires long power cycling with a predefined mix of output current profiles. The output current is of a trapezoidal shape with varying ramp-up, ramp-down and flat-top duration times [1].

Fig. 1. Power circuit of magnet power supply

It is important that the IGBTs in such a specialised application are overrated in order to reduce the thermal stressing and, as a result, to extend their lifetime. According to the extensive work in the field of accelerated testing on IGBT modules [2], [3], [4] and [5], the thermal stressing can, mainly, be determined with the help of the IGBT module junction or case temperature fluctuation combined with the maximum, minimum or average temperature that the IGBT junction or case is exposed to, as well as the on-time of the load profile. Therefore, it is critical to monitor online the junction temperature of the IGBT. However, the pre-defined nature of the current profile for this specialized application does not...
require an online and continuous temperature monitoring. This is not the case with other applications of power electronics, such as the power-electronic converters of a wind turbine, of a solar panel array or of the powertrain of an electric vehicle, where the arbitrary load profile causes unpredictable stressing to the IGBT modules. For the specialized application, it is necessary to estimate the junction temperature for specific operating points of pre-defined load profiles, in order to evaluate the thermal performance of the whole power stack (from IGBT module level to the cooling medium) during its commissioning phase. Moreover, the temperature estimation is needed during planned service stops for the aging detection of the IGBT module.

There are several parameters that can be monitored to assist with the temperature estimation issue of the IGBTs. These include the saturation voltage drop \( V_{ce} \) with sensing current [6] or with load current [7], the gate-emitter threshold voltage \( V_{ce,th} \) [8], the short-circuit current [9], the internal gate resistance related to the peak gate current [10], the voltage dynamics during the turn-off delay [11], the current dynamics during turn-on [12] and others. The \( V_{ce} \) method with sensing current is of interest for the specialised application due to the advantages of relatively simple thermal calibration, of the linearity of the relation between the temperature and \( V_{ce} \) and, finally, of the similar sensitivity of the method for all the Si-based IGBTs [13]. In order to carry out the \( V_{ce} \) measurement with sensing current, it is required to isolate the IGBT switch from the power circuitry and inject a sensing current. Additionally, any change on \( V_{ce} \) detected over time while the semiconductor operates under the same operating conditions could indicate aging of the power modules [14].

The specialized application would benefit from \( V_{ce} \) measuring circuits that offer ease of implementation and of calibration at low cost. These circuits would allow the time-efficient testing of several power-electronic converters. Along with the measuring circuits, the sensitivity analysis of the parameters that may affect the performance of such measuring circuits has yet to be documented and especially for high-current IGBT modules concerning the specialized application. In recent \( V_{ce} \) method reviews such as in [15], the impact of critical parameters, such as the sensing current source precision and the gate-emitter voltage \( V_{ge} \), on the \( V_{ce} \) measurement are not quantified.

Typically, aging detection relies upon the increment of \( V_{ce} \) indicating bond-wire lift-off or solder delamination at the chip or substrate level. The lack of experimental confirmation on how the \( V_{ce} \) method with sensing current performs with the detection of aging in the context of the specialised application with high-current IGBT modules is, also, yet to be reported. For the specialized application, the IGBT modules operate to a large extent at their Negative Temperature Coefficient (NTC) operating region (low-current region) that is below 400A [16], where \( V_{ce} \) is decreasing with temperature, and to a smaller extent at the Positive Temperature Coefficient (PTC) operating region (high-current region), where \( V_{ce} \) is increasing with the temperature. Between the aforementioned operating regions, there is the narrow in terms of current range Zero Temperature Coefficient (ZTC) operating region, where \( V_{ce} \) has no dependence on temperature. Based on the results in [17], \( V_{ce} \) could for the specialized application be reduced due to the temperature increase, if solder delamination occurred before bond-wire lift-offs. This effect is followed by bond-wire lift-offs that would, initially, be covered by the \( V_{ce} \) decrease, if a continuous \( V_{ce} \) monitoring is not applied to separate the two aging mechanisms.

Regarding IGBT modules with low-current ratings [18], it is reported that the \( V_{ce} \) method with sensing current is, probably, not capable of detecting the very first stages of aging, such as a single bond-wire lift-off. A high current at the PTC operating region is needed to amplify the change in \( V_{ce} \) for the aging detection at an early stage. For the specialized application, the 1.6kA module contains many parallel bond-wires and the loss of a single bond-wire causes a very small change to the \( V_{ce} \) value under a sensing current. The detection with the \( V_{ce} \) method can be more challenging for the 1.6kA IGBT module, but the IGBT module overrating for the specialized application provides redundancy and allows the operation with a reduced number of bond-wires. Therefore, the earliest aging detection stage of a single bond-wire lift-off is not required.

Another method that can detect an early stage of bond-wire lift-offs for low-current IGBTs is the ZTC operating region measurement of \( V_{ce} \) [19]. The reason is that the temperature impact on \( V_{ce} \) due to the solder delamination could be avoided. Specifically, for the IGBT module of the specialized application, the anti-parallel diode has the ZTC operating region at a current level higher than the maximum output current of the magnet power supply [20]. Therefore, the temperature impact of the NTC operating region to the aging detection measurement with high current cannot be avoided.

The first objective of this paper is to document the impact of sensing current and gate-emitter voltage on the \( V_{ce} \) measurement using a sensitivity analysis. These findings are crucial for the utilization of the method in any application. Secondly, two measuring circuits that offer convenient implementation and calibration at low cost are proposed for the specialised application in question. A detailed discussion on the selection of the components of the measuring circuit is included. The utilization of the circuits is not limited to the application of interest. Finally, to fulfil the contribution of this article, experimental results taken from a test bed operating at 6.5kHz and utilising a 1.6kA IGBT module confirm the performance of the measuring circuits along with the evaluation of the \( V_{ce} \) method for the aging detection of high-current IGBTs with the specialised application as an example.

The remainder of the paper is organised as follows. Section II gives a brief introduction of the \( V_{ce} \) method and of the trends of the \( V_{ce} \) measuring circuits. More important is that it presents the main design principles for a \( V_{ce} \) measuring circuit and quantifies the impact of the sensing current and of \( V_{ge} \) to the measurement precision with experimental results for the 1.6kA IGBT module. In section III two circuits developed for swift implementation and calibration are proposed for the \( V_{ce} \) measurement. Section IV presents the experimental results from the performance tests of the two measuring circuits. Apart from the targeted switching frequency of 6.5kHz, a second switching frequency of 3kHz is chosen as an intermediate switching frequency level for the circuits testing to analytically explain the measuring procedure. In the same section, the
detection of aging in a sample IGBT is demonstrated and the applicability of the method in the state-of-the-art detection of a high-current module is discussed. Conclusions are summarized in section V.

II. CRITICAL FACTORS FOR A $V_{ce}$ MEASUREMENT AND ITS MEASURING CIRCUIT

A. Typical $V_{ce}$ measuring circuits

The main concern for $V_{ce}$ measuring circuits is the possibility to block a high voltage at the kV level and to precisely monitor voltage levels up to a few V. The high voltage clamping can be achieved either with active elements such as MOSFETs [14], [21] and [22], as it is demonstrated in Fig. 2a, depletion MOS [23] and analog switches [24] or with passive elements such as PIN diodes [25] in Fig. 2b, Zener diodes [18], series connection of PIN diodes and voltage sources [26], or even resistive voltage dividers [27]. Therefore, the measuring circuits can be divided in two categories based on the devices that are used for the voltage clamping; circuits using active devices and circuits using passive devices. The required thermal and electrical characterization of the active devices and their driving circuit development contribute to the complexity and uncertainty of the measurement. Among the solutions with passive elements, the resistive voltage divider requires high precision power resistors with low thermal drift and high precision operational amplifier for the measurement adaptation. The PIN diodes necessitate electrical and thermal characterization. The Zener diode is a promising candidate with reduced characterization effort and a detailed analysis and development, continuing from the analysis in [18], follows in section III. Figure 2 illustrates two examples of the two categories of measuring circuits. The circuit in Fig. 2a creates the sensing current by inserting a resistance in series with the DC-link only during the measurement interval. This solution is only targeted to stiff DC-link with no voltage variation, because the resistance value is selected for a specific voltage value. The high-voltage blocking is achieved with MOSFETs that require extra driving circuitry and calibration. The circuit of Fig. 2b, is, initially, developed for the $V_{ce}$ measurement online with load and not with sensing current. This method does not require the IGBT switch isolation, but it requires more effort for the characterization of the switch for several temperature and current levels [7]. It is a method that requires no intervention in the power circuit in contrast to the sensing current method that requires the Device Under Test (DUT) isolation from the power circuit. A precise current source of a few mA is needed to guarantee the full conduction of the PIN diodes and it is challenging to thermally couple the PIN diodes. This measuring circuit could be used for the $V_{ce}$ method with sensing current, if an extra sensing current source injecting current directly to the IGBT was added. A simplified circuit based on a single PIN diode is proposed in section III to reduce the characterization procedure. The requirements for the elements of each of the existing circuits are summarized in Table I. As mentioned earlier, the specialized application necessitates measuring circuits that require even less effort of calibration and simple operation.

Fig. 2. Typical $V_{ce}$ measuring circuits: (a) Measuring circuit with active components (MOSFETs) for high-voltage blocking [14], (b) Measuring circuit with passive components (PIN diodes) for high-voltage blocking and with simplified operational amplifier circuit for the measurement adaptation [25].

B. Protection of measuring circuit

The preferred voltage-blocking elements for the specialized application that is a low voltage application are the passive elements because they offer easy calibration and avoid the complexity of the additional driving circuits of the active components.

C. Precision

The precision of the measuring circuit has to be in the mV scale, due to the sensitivity of the $V_{ce}$ parameter that is, usually, 2mV to 2.3mV/°C for Si-based IGBT modules. The result of the thermal characterization of the complete IGBT power module of the application is illustrated in Fig. 3c. The sensing current source precision and output ripple are crucial for the measurement. For a measuring circuit with a sensing current of 1A injected to the 1.6kA IGBT module, the corresponding loss of accuracy in the voltage measurement and, as a result, in the temperature estimation is illustrated in Fig. 3a.

Fig. 3. (a) Measured sensitivity of $V_{ce}$ with sensing current for the ABB 55NA 1600N170100 IGBT module [20] for $V_{ge}$ equal to 15V and a temperature of 22°C used as a sample in this work, (b) Measured sensitivity of $V_{ce}$ with $V_{ge}$ at 1A for a temperature of 30°C used as an example for the ABB 55NA 1600N170100 IGBT module, (c) Measured relation of $V_{ce}$ with temperature under a sensing current of 1A for the ABB 55NA 1600N170100 IGBT module.

It is observed that for a sensing current of 1A a deviation of 20mA can cause an error of 2mV or 1°C. The required precision for the sensing current source is relatively low at 2% and can be achieved with the current sources available in the market. The characterization is done on a power module at 22°C that is the room temperature.

A second factor that influences the precision is the gate-emitter voltage $V_{ge}$. For the 1.6kA considered module, the collector-emitter voltage at sensing current is not dependent on $V_{ge}$ for values between 10V and 15V. Measurements were performed on the sample IGBT for the $V_{ge}$ - $V_{ce}$ characterization.
TABLE I
OVERVIEW OF THE MAIN DESIGN CONSIDERATIONS FOR $V_{ce}$ MEASURING CIRCUIT

<table>
<thead>
<tr>
<th>Aspect</th>
<th>Requirements</th>
<th>Main concerns</th>
</tr>
</thead>
<tbody>
<tr>
<td>Protection</td>
<td>Withstand in voltage fluctuation from millivolt to kilovolt level</td>
<td>-Resistive voltage divider selection and gain of amplifier for signal adaptation -Blocking diode characterization -Zener diode characterization -Extra circuitry for additional switch for voltage clamping -Additional switch calibration</td>
</tr>
<tr>
<td>Precision</td>
<td>Maximum $2mV$ corresponding to $1^\circ$C</td>
<td>-Current source ripple and accuracy not better than $2%$ -Independent of $V_{ge} &gt; 10V$ -The concerns mentioned for the protection</td>
</tr>
<tr>
<td>Speed</td>
<td>For online measurements: meas. window &lt; Duty cycle</td>
<td>-Current source rise time -Conditioning circuit settling time -Resistive voltage divider dynamics -Zener and measuring resistor matching -$V_{ge}$ level stabilization -Electron recombination</td>
</tr>
</tbody>
</table>

for a temperature of $30^\circ$C and they are illustrated in Fig. 3b. The $V_{ge}$ - $V_{ce}$ graph has the same form and is valid at any temperature level, as it was observed during the measurements. The $V_{ce}$ method does not need $V_{ge}$ monitoring or characterization, considering that $V_{ge}$ rises to $10V$ in a negligible time period that is faster than the bandwidth of the measuring circuits for this application. In contrast with the $V_{ce}$ method with sensing current, the $V_{ce}$ method at load current is sensitive to $V_{ge}$, as it has been previously described [7]. This delay may hinder the measurement, in the case of a high switching frequency where the measuring window is narrow. Moreover, this parameter affects the measurement even at the ZTC point that for this module is at approximately $400A$. Therefore, $V_{ce}$ at load current and at ZTC is not directly temperature dependent, however, indirectly it is, due to $V_{ge}$.

D. Speed of measurement

The main limiting factors in terms of speed for the measurement within the duty cycle are the current source rise time, the signal conditioning circuit settling time and desaturation, the $V_{ge}$ settling time and the high-voltage blocking elements such as resistive voltage dividers or measuring resistors causing slower dynamics. For the active voltage-blocking elements, there are no concerns regarding the speed of measurement. For an interval equal to the duty cycle $D$ available for measurement the following relation holds

$$t_{setting} + k t_{sampling,DAQ} < D \quad (1)$$

where $t_{sampling,DAQ}$ is the sampling time of the Data Acquisition system (DAQ) and $k$ defined by the Nyquist criterion. Equation (1) is valid only if the sensing current is injected in the device directly after the device is turned on. If the sensing current follows a full current conduction, then a time interval at the level of hundreds of microseconds has to pass, in order to start the measurement and to avoid the electron recombination effect that could influence the measurement

[28]. Table I summarizes the requirements and considerations for the three main design aspects of the measuring circuit.

III. DESIGN OF PROPOSED MEASURING CIRCUITS

A. Zener-based measuring circuit design

The circuit and the working principle of the measuring circuit for the $V_{ce}$ method are presented in Fig. 4 and it is based on a concept similar to the one in [18]. It comprises a measuring resistance in series with a Zener diode, as it is depicted Fig. 4a. The measuring circuit is connected in parallel to one of the two single-switch IGBTs of a phase leg. The voltage across the Zener diode is the $V_{ce}$ voltage, if the DUT is conducting, according to Fig. 4c. If the DUT is blocking, the voltage across the Zener diode is the Zener voltage and the rest of the DC-link voltage is applied across the measuring resistor, as it is shown in Fig. 4b.

The current source is protected from the high voltage with two diodes in series. In parallel to the source, there is a MOSFET and a resistor; this parallel branch is optional, in the case that the current source is not fast enough to raise the current within the measuring window. The current would circulate in advance, during the off-time of the DUT. The sensing current that is injected to the DUT is measured with a Direct-Current Current Transformer (DCCT) [16], [29] to ensure that the current reaches the expected value within the measuring time window.

B. Components selection

The Zener voltage $V_z$ has to, initially, fulfill two requirements

$$V_{ce} < V_z < \min(V_{\text{max,op-amp}}, V_{DAQ}) \quad (2a)$$

$$V_{ce} \ll V_z \quad (2b)$$

where $V_{\text{max,op-amp}}$ is the maximum allowable voltage of the operational amplifier and $V_{DAQ}$ of the DAQ. The saturation of the amplifier can delay its response during the desaturation time and has to be avoided, if the measurement requires fast dynamics in a small measuring time window. The second requirement results from the leakage current of the Zener diode, when the DUT is conducting. The voltage $V_{ce}$ has to be significantly lower than $V_z$ to ensure that the minimum leakage current will flow in the measuring resistor. The Zener diode characterization in terms of applied voltage and leakage current is presented later in this paper.
The measuring resistor is selected based on four criteria. The first one is the measurement dynamics due to the RC circuit created by the Zener diode capacitance and the measuring resistor. The second criterion is the maximum allowable Zener current flowing in the diode, when the DUT is blocking the voltage. The increase of the diode temperature increases this current. The thermal resistance junction-to-ambient \( R_{th,j-a} \) of the diode package limits the current value according to

\[
R_{th,j-a} P_z = R_{th} I_z,AV V_z = \Delta T_{jc} \text{ close to zero} \quad (3)
\]

where \( P_z \) is the conduction power losses of the Zener diode, \( I_z \) is the average Zener current during the switching of the DUT and \( T_{jc} \) is the estimated junction temperature of the diode that has to remain as close as possible to the nominal conditions of 25°C in the datasheet. Thirdly, it is the leakage current of the diode with its voltage-current relation,

\[
R_{meas} I_{leak} = V_{i, error} \quad (4)
\]

where \( I_{leak} \) is the diode leakage current and \( V_{i, error} \) the measurement error in the input voltage. The leakage current can create a notable voltage drop across the resistor that has to be compensated at the circuit output. Finally, it is the impedance matching with the operational amplifier input impedance. The measuring resistor has to be much smaller than the amplifier input impedance, in order to eliminate the voltage divider effect of the measuring path and to maintain a precision of 2mV. For the \( V_{ce} \) method with sensing current and for a known maximum measured voltage \( V_{ce,max,1A} \) the relation of the two resistors are

\[
V_{R_{meas}} = \frac{R_{meas}}{R_{meas} + R_{op-amp}} V_{op-amp}
\]

with

\[
R_{meas} + R_{op-amp} V_{R_{meas}} = V_{ce,max,1A} \text{ and } R_{meas} \leq 2mV
\]

\[
\frac{R_{meas}}{R_{op-amp}} = \frac{V_{R_{meas}}}{V_{ce,max,1A} - 2V_{R_{meas}}} \quad (5)
\]

where \( V_{ce,max,1A} \) corresponds to the minimum operation temperature of the module. This value increases if there are \( V_{ce} \) measurements after bond-wire lift-off incidents. The thermal drift of \( R_{meas} \) does not have a significant impact neither at the speed nor at the precision of the measurement.

For this design \( R_{meas} \) should be at least 80kΩ, in order to maintain a low temperature increase for the Zener diode due to its leakage current. The Zener diode BZX55C6V8 with a Zener voltage of 6.8V is selected for the low leakage current at the measuring range of the application. According to the datasheet [30], for a Zener voltage of up to 2V, it limits the leakage current to below 100nA. For the characterization of the Zener diode, the set-up of Fig. 5a is used and the measurement error is obtained along the \( V_{ce} \) measuring range for the \( V_{ce} \) method with the sensing and with load current in Fig. 5b.

For commonly available Zener diodes with a leakage current at the level of 1μA and at the worst case in terms of leakage current that is for 125°C, the voltage drop across the measuring resistance may reach 80mV or more that corresponds to almost 40°C error at the temperature estimation with the \( V_{ce} \) method with sensing current.

The implemented measuring circuit is illustrated in Fig. 6a.

![Fig. 5](image)

**Fig. 5.** (a) Set-up for the measurement of the error due to the leakage current of the Zener diode, (b) Measured voltage drop across measuring resistor as a function of the total voltage applied across the series-connected measuring resistor and Zener diode

The leakage current increases, as the measuring voltage approaches the Zener voltage. The Zener voltage should be much lower than the measured \( V_{ce} \) and the input amplifier saturation voltage. For \( V_{ce} \) with load current, a Zener diode with a Zener voltage over 10V is recommended.

From the thermal response graphs at [30], the expected temperature increase for the diode BZX55C6V8 is approximately 6°C according to (3). Due to the RC circuit that is formed by the diode and \( R_{meas} \), the Zener voltage has a slow rise time to reach its nominal value that prohibits the Zener voltage from rising when the DC-link voltage is applied. Therefore, the calculated temperature rise is the worst case in terms of thermal stressing.

A linear regulator with adequate precision and a current rise time of 800ns is used as the sensing current source, in order to avoid noise from switching sources. The operational amplifiers circuit consists of a precision amplifier with a high input impedance that, directly, receives the \( V_{ce} \) value and the output of this amplifier is connected to the isolation operational amplifier ISO124 from Texas Instruments. The amplifier before the isolation amplifier is introduced because the input impedance of the latter is only 200kΩ that is comparable to \( R_{meas} \) causing a voltage divider effect. The settling time for the \( V_{ce} \) level is approximately 30μs.

The set of batteries of Fig. 6b is used for the power supply of the card that makes the measuring device portable and eliminates any noise introduced by the supply. The device is enclosed in a metallic and grounded box. It can be, directly, connected in parallel to the DUT at the magnet supply in the field. The advantage of this design is the simple circuitry with low number of components and the fast and easy calibration of the Zener diode.
As an advancement to the method for the measuring circuit of [18], the effect of the Zener diode selection and its leakage current, especially for $V_{ce}$ under high current, is highlighted and it is considered for the accuracy improvement. A complete design procedure for the measuring circuit is provided. The Zener diode-based circuit that is utilized for the method with sensing current does not require an electrical and thermal characterization like the PIN diode [7] and the depletion MOS [23]. Furthermore, the MOS requires a driving circuit that adds extra complexity. Finally, a switch in parallel with the sensing current source is proposed in case the current source has a rising time that is slow compared to the measuring window inside the switching period of the DUT.

C. Proposed measuring circuit based on desaturation protection circuit

A starting point for the typical desaturation detection circuit is shown Fig. 7a [31]. The blocking diode $D_f$ is used for protection. Resistor $R_1$ and capacitor $C_1$ form a low-pass filter. For the $V_{ce}$ method a measurement throughout a range is necessary and not a threshold voltage detection. To avoid the current source that is proposed in [25], a characterization of $R_1$ and $D_f$ is necessary for the expected variation of $V_{ce}$. For this work, the proposed alternative of the desaturation protection circuit, with limited calibration effort is the circuit in Fig. 7b. It is based again on a PIN diode as a passive voltage-blocking element.

The design principle is to keep the diode forward voltage $V_f$ at a constant value. This enables to subtract a constant $V_f$ from the measurement and obtain $V_{ce}$. A constant $V_f$ requires a constant supply voltage $V_s$ and a $V_{ce}$ that ranges within values that do not practically change $I_{df}$. The resistor $R$ defines the current $I_{df}$. The range of $V_{ce}$ change is from 0.35 to 0.5V, according to Fig. 3c, for the whole module with a small extra margin due to the aging effect, as it will be discussed later, for the IGBT module of the application. For demonstration purposes for the test of the circuit, one substrate is utilized that changes $V_{ce}$ to the level of 0.6V, as it is going to be observed in the results. The current $I_{df}$ remains constant at 2mA throughout the range of $V_{ce}$ change. Therefore, no essential change in $V_f$ is caused within the defined $V_{ce}$ range. The voltage $V_f$ is stabilized between 1.089V and 1.09V. The voltage supply $V_s$ can be formed with a voltage divider from the supply of the operational amplifier connected across $D_f$ and the IGBT emitter. It has to be a fraction of the amplifier supply in order not to drive the amplifier to saturation. The amplifier network is the same as in the Zener diode-based circuit.

The advantage of the proposed circuit comparing to the similar circuit of Fig. 2b is that no precision current source is needed, in order to define a constant $V_f$. For the proposed method, there is no need to use any additional source except for the sensing current source.

Moreover, it requires only one PIN diode instead of two and it avoids the thermal coupling of two PIN diodes in [7] increasing the stability and the linearity of the method. It is a measuring circuit that, only, requires a PIN diode and a precision resistor for the adjustment of the PIN diode current.

D. Test set-up for verifying the $V_{ce}$ measuring circuits performance

The main test set-up of the measuring circuit is shown in Fig. 8a. The measuring card is connected in parallel to the DUT IGBT that is the upper switch at a phase-leg configuration. The DUT is intentionally uncovered to enable the independent control of the four substrates, as it is illustrated in Fig. 8b. The phase-leg is connected to a high-voltage DC source that emulates the DC-link of the magnet supply, in order to test the voltage blocking capability and the dynamic performance of the measuring circuit. The two switches are operated in a complementary manner with $f_{ow}$ reaching up to 6.5kHz that is the upper target of this application for a measurement within the switching period. The duty cycle for each switch is 0.5. This measurement can be useful to estimate the temperature at specified operating points during the thermal performance evaluation phase or at planned service stops.

For ease, one substrate of the DUT is used as the upper switch. The open module is not necessary for this test, in contrast to the substrate failure test that is going to be presented later in this paper. The sensing current is measured with the DCCT to ensure that 1A is injected to the DUT. The $V_{ce}$ measurement is collected into LabVIEW using the National Instruments USB-6251 Data Acquisition System (DAQ). The DAQ sampling rate is 800ksamples/s. A voltmeter is connected across the DUT to measure $V_{ce}$ when there is no switching. This value is taken as a reference for the measurements. The DUT is thermally characterized per substrate and as a complete module by obtaining $V_{ce}$ as a function of the junction temperature for a current of low value (usually about 1/1000 of the device nominal current or 100mA/cm² [28]) within the range of temperatures that are expected during operation, as it is illustrated in Fig. 3c. Since the DUT is not heated, a
thermocouple is placed at the test bench to obtain the ambient temperature and compare the voltage measurement with the thermal characterization curve of Fig. 3c. This method can be more accurate for a first-stage evaluation of the measuring circuit performance than the active heating of the semiconductor with high current and the monitoring with a thermal camera within specific intervals. Due to the thermal camera low sampling rate, the temperature of the DUT can be measured only under a load with low switching frequency that does not exhibit the measuring method’s bandwidth performance. The thermal camera measurement can introduce measurement errors due to the black paint application and the emissivity definition, whereas the proposed test is realized under a fixed and known temperature. Therefore, the proposed evaluation method is, easily, applied and tests the measuring method’s bandwidth too. The tests are implemented in the field environment with several power converters operating at the same time.

IV. RESULTS

A. Performance of the proposed circuits

The design of the measuring circuits aims to the measurement within a window of approximately 75μs that corresponds to the DUT on-time for a duty cycle of 0.5 and a switching frequency of 6.5kHz. A second switching frequency of 3kHz is chosen as an intermediate switching frequency level for the performance test of the Zener-based circuit to demonstrate analytically the measuring procedure and the difference in the response of the circuit.

A filter to eliminate the 500kHz ripple produced by the modulator/demodulator for the digital-to-analog conversion of the input signal in the isolated amplifier causes the oscillation before reaching the low state. Figures 9a and 9b show the $V_{ce}$ voltage measurement with the Zener-based circuit and for a switching frequency of 3 kHz. In Fig. 9a the $V_{ce}$ measurement fluctuates from approximately 4V, when the DUT is blocking to the value corresponding to the sensing current of 1A providing 0.618V at 22°C. The saturation voltage $V_{ce}$ at low state is processed for the last 50μs of the DUT on-time.

The measurement for a switching frequency of 6.5kHz is illustrated in Fig. 10. In this case $V_{ce}$ at low state is processed for the last 20μs of the DUT on-time, according to Fig. 10a because of the measuring circuit settling time. For the maximum values of Zener voltage for the 3kHz and the 6.5kHz case, it is noticed that at 3kHz the maximum Zener voltage is higher than at 6.5kHz according to Fig. 9a and Fig. 10a. The RC response of the Zener capacitance and the measuring resistor reduces the $V_{ce}$ maximum value. Therefore, the actual losses and the temperature increase of the diode are smaller than the theoretical calculations.

![Fig. 10. Measurements with the Zener-based circuit for a switching frequency of 6.5 kHz: (a) Voltage across Zener diode: high value for DUT at off-state, low value for DUT at ON-state, (b) measuring interval of $V_{ce}$ in the end of the ON-state](image)

The noise reaches up to 20mV due to few sparse measurements. An additional step of the investigation is to change the isolation stage with a digital isolator as an input to an ADC converter, in order to compare with the isolation amplifier in terms of noise level. A median filter with a band of 3 samples, taking into account the next three values, is utilized to smooth out the measurement [32]. The measurement ripple is reduced to about 15mV peak to peak in the worst case. The principle of the median filter for the spikes elimination is illustrated in Fig. 11.

![Fig. 11. Median filter logic for eliminating the spike value P2](image)

The selection of the next value is within a band of points that are ranked from the smallest to the greatest value. The median value is the output. The filtered signal at 3kHz has a band of approximately 7mV per side around the reference value of 0.618V and its mean value matches with the reference. The same technique is used for a switching frequency of 6.5kHz, where the measuring window is now 20μs due to the reduced duty cycle D. The mean value deviates from the reference by less than 1mV. The level of accuracy remains unchanged over the switching periods, which is satisfactory.

During the performance tests with the set-up of Fig. 8a, the second proposed circuit exhibited comparable performance with the desaturation circuit. The performance of the simplified circuit is similar to the Zener diode circuit for the two switching frequency levels. The median filter is applied again. The performance test is again repeated and the results for 6.5kHz are illustrated in Fig. 12b. The mean value of the median filter output is 3mV higher than the reference for the measurements at 3kHz and 6.5kHz. Similar to the Zener-based circuit, due to capacitance of the antiparallel diodes in Fig. 7b the maximum voltage level does not have enough time to rise to the $V_{CE}$ level but it reaches about 5V, as it is illustrated in Fig. 12a. The noise level is comparable to the Zener-based measuring circuit.
B. Aging detection

The open IGBT sample is used to demonstrate the detection of the failure of one of the substrates by using the $V_{ce}$ method with sensing current. The independent control of each substrate and the switching from four-substrate to three-substrate operation demonstrates the loss of one substrate and the detection by the $V_{ce}$ measuring circuit based on the Zener diode. The voltage step with the loss of one substrate is around 14mV, as it is illustrated in Fig. 13.

Considering the four chips per substrate and the ten bond-wires per chip, the detection of the lift-off of one bond-wire would require a very precise detection system that may exceed the 2mV accuracy of the current $V_{ce}$ measuring circuit. It would be challenging to distinguish the incident from the noise in the field environment.

As it is mentioned before and it is also, observed in [18], at the NTC operating region the junction voltage drop dominates the resistive part contribution in the total $V_{ce}$ value. This effect is even more evident for the sensing current, because the voltage drop at the resistive part due to the sensing current is not significant. On the other hand, a change in the parasitic resistance due to a bond-wire lift-off that is difficult to detect for a sensing current, is amplified in the case of the load current at PTC. Table II provides examples from the literature concerning the bond-wire lift-off detection for modules of different ratings with the $V_{ce}$ method with load current and compares the results with the substrate failure detection with sensing current for the 1.6kA module. The difference in the absolute value of the $V_{ce}$ change due to the bond-wire lift-off is greater for the low-current modules than for the high-current ones.

For the specialized application, the first aging stage of a bond-wire lift-off cannot be detected with sensing current for the 1.6kA module. Due to the current overrating of the IGBT module for lifetime extension, the module could, theoretically, operate even with the loss of a substrate. Therefore, the detection of a substrate loss at the service stop may be sufficient as a warning to prevent the complete failure of the module and the unplanned interruption of the power-electronic converter operation. The bond-wire lift-off detection with the sensing current method prevents the heating of the power module that occurs for high currents and could be misinterpreted as solder delamination. The ambient temperature should be taken into consideration for the measurement, because it would add an offset that has to be removed for the comparison of the measurement with the initial measurement at the converter thermal evaluation phase. For this measurement, there is no requirement for the converter to be in operation. Furthermore, there is no need to experimentally find the exact ZTC point of the power module. The solder delamination can be detected with the sensing current method, because it results to a temperature change that is observed as $V_{ce}$ reduction.

<table>
<thead>
<tr>
<th>Module ratings [V/A]</th>
<th>Module structure [chips, bond-wires per chip]</th>
<th>Current at $V_{ce}$ measurement [A]</th>
<th>No. of bond wires lifted off</th>
<th>Change in $V_{ce}$ [mV]</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>1700/1000</td>
<td>12, 8</td>
<td>850</td>
<td>1, 2</td>
<td>approx 50, 10</td>
<td>[33]</td>
</tr>
<tr>
<td>600/70</td>
<td>1, 6</td>
<td>50</td>
<td>1, 2, 3</td>
<td>20, 50, 140</td>
<td>[18]</td>
</tr>
<tr>
<td>1700/1600</td>
<td>16, 10</td>
<td>1</td>
<td>4 chips or 40 bond-wires</td>
<td>14</td>
<td>Measured in this paper</td>
</tr>
</tbody>
</table>

For the specialized application although the sensing current method may be sufficient for the detection of the aging stages that are critical, a solution that could detect earlier aging is proposed. In particle accelerators the load current value is regulated to a precision in the order of 5ppm. This allows using a load current cycle for performing $V_{ce}$ measurements that are, highly, reproducible. As a result, the use of a service cycle is periodically applied and its $V_{ce}$ signature is compared to a reference measurement performed during commissioning of the power converter. This cycle could be used during service stops and factory acceptance tests, it requires no current control of the power electronic converter and it is independent of the actual operation cycles. Compared to the method in [18], there is no need for an extra high current source and extra circuitry for the separation of the high current source from the main power circuit. Compared to the method in [14], no extra power circuitry is needed for the inductance current redirection.

According to Fig. 14b, the engagement of only two diagonal switches and the anti-parallel diodes of the other two diagonal switches can produce a current ramp-up and ramp-down for the inductive load leaving a time interval in which the H-bridge operates in discontinuous current mode. This cycle is repeated for every switching period. The $V_{ce}$ measurement at the high-current region of the current ramp, can provide information.
about the bond-wires condition. When the thermal steady state is reached and during the time interval of zero current in the discontinuous mode, one of the two diagonal switches, that were previously conducting, is turned on. The sensing current can be injected to the switch and the measuring method can provide the temperature estimation that helps with the health-check of the solder layers in the module. Moreover, the temperature estimation gives an indication of the thermal stressing for specified rms current levels and it can be used to verify the thermal model of the power electronic converter. For given power stack ratings, the rms current levels depend on the duty cycle and the value of the inductor that is used as a load. The cycle is illustrated in Fig. 14a.

![Diagram of Power Electronic Converter Test Cycle](image)

**V. CONCLUSION**

In this paper, the $V_{ce}$ method with sensing current for junction temperature estimation and aging detection of high-current IGBT modules has been discussed. In particle accelerator applications, this method can be used during the validation and factory acceptance of power semiconductor stacks, as well as to detect premature aging such as bond wire lift-off or solder delamination during the service stops. The main contributions of this paper can be summarized in the following:

- Sensitivity analysis on the impact of the sensing current and the gate-emitter voltage at the $V_{ce}$ method, in order to define the general requirements for the development of a $V_{ce}$ measuring system.

- Proposal of two $V_{ce}$ measuring systems that are designed to offer fast implementation, minimum calibration effort and versatility for utilization in various applications. Experimental results on a 1.6kA IGBT module confirm the performance of the proposed method and circuitry.

- Experimental evaluation of the applicability of the $V_{ce}$ method for the aging detection of high-current IGBT modules. It was demonstrated that the $V_{ce}$ method meets the requirement for aging detection set by the specialized application.

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**REFERENCES**


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