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Simple and Affordable Method for Fast Transient Measurements of SiC Devices

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***Abstract*—The measurement of fast voltage and current transients of SiC devices requires high bandwidth probes. Commercially available voltage and current probes can be expensive and in addition, the delay introduced by them must be compensated to achieve a proper time alignment (de-skew). Determining this de-skewing value is not a trivial task.**

In this paper, a simple and affordable measurement method is presented for the simultaneous measurement of the voltage and current transients of SiC devices. The voltage is measured by means of a high bandwidth RC attenuator while the current is estimated from the voltage drop in the stray inductance of the switching loop. Since both voltages are measured with two equal and matched high bandwidth passive voltage probes there is no need to apply any deskew. This is one of the most important advantages of the method. The presented method is experimentally evaluated and used for the estimation of energy switching losses in the tested SiC - MOSFET.

***Index Terms*—SiC, MOSFET, Measurement, Fast Switching, Voltage probes, Current probes, High Frequency.**

I. Introduction

OVER the last few years, Silicon Carbide (SiC) devices are gaining a strong acceptance in power electronics applications due to their superior thermal and electrical performance compared to their Silicon (Si) counterparts. Probably one of the most attractive characteristic of SiC devices is their ability to operate at fast switching transients. This leads to low switching losses and makes possible the operation at high switching frequencies. In consequence, the volume of passive components is reduced [1].

In spite of the benefits, the operation at high voltage and current slopes in the presence of parasitic capacitances and inductances in the converter makes the design of the converter more challenging in terms of overvoltages, ringings and EMCs. To guarantee the operation of the device within the limits of the Safe Operation Area (SOA), a proper current and voltage measurement is necessary. In addition, to obtain the switching loss data that makes possible the evaluation of the converter power losses, the measurement of the instantaneous current and voltages is mandatory. This data is typically obtained through a Double Pulse Test (DPT) [2], [3]. This DPT should be done in the final layout of the converter to consider the influence of the real stray inductance.

This measurement becomes complicated due to the fast switching nature of SiC devices. On one hand, this requires high bandwidth (BW) voltage and current probes and on the other hand, probes should not add extra parasitic elements to the layout. In addition, the skew of the probes has a big influence on the evaluation of power losses and therefore a sync of voltage and current signals must be applied. Although different de-skewing methods can be found in the literature [4], [5], their application is not trivial and the determination of the proper de-skewing value becomes challenging.

Voltage measurements can be easily performed with high bandwidth active differential probes. It could be said that in the voltage range of 1500 volts there are several affordable voltage probes commercially available with bandwidths exceeding hundreds of MHz. However, as the measured voltage increases, the cost of the probe can become prohibitive if high bandwidth is still required [6]. Thus, while 1500 V probes are good candidates for voltage measurement of 1.2 kV SiC devices, the evaluation of newer 1.7 kV or higher voltage SiC devices (3.3 - 6.5 kV) [7] demands new and expensive voltage probes.

Similarly, commercial Rogowski coils, Pearson current transformers or Coaxial Shunts can measure fast current transients. Although Rogowski coils and Pearson probes are suitable for fast switching transients, their useable rise/fall time must be shorter than the rise time of the measured signal. Thus, a current probe rated at 200 A with a useable rise time of 15 ns, suitable for fast switching transient measurements of power modules (20 ns), is not able to measure fast current transients of discrete devices (30 A - 5 ns). In consequence, different current probes must be used if different power/voltage/current rated devices are evaluated.

Due to their high cost, it is a common practice to buy voltage/current probes that serve as many applications as possible and thus reduce the need of major investments. However, in this scenario, there may be specific applications where the available current and voltage probes are not suitable. If the purchase of the probes cannot be justified or the delivery times are too long it could be complicated to perform the required measurements.

Therefore, in this paper, the suitability of different voltage and current probes is evaluated for fast transient measurements of discrete SiC - MOSFETs. Due to the limitations presented by the available voltage and current probes, this paper presents a simple and affordable method for the simultaneous measurement of fast voltage and current transients. The voltage measurement is performed with a simple high bandwidth RC

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voltage divider and a high bandwidth passive voltage probe. The current measurement is estimated from the voltage drop of a partial stray inductance in the switching loop. This voltage drop is measured with a high bandwidth passive voltage probe.

Since both voltages are measured with equal passive voltage probes, the proposed method simplifies the time alignment problem. The proposed method is experimentally evaluated and finally, a switching loss estimation has been carried out in a Double Pulse Test set up.

II. Double Pulse Test setup and required probe bandwidth

The evaluation of the switching process has been performed through the Double Pulse Test (DPT) on the C3M0120100K SiC - MOSFET (Wolfspeed) [8]. The DC bus voltage is 500 volts and the maximum switched current is 50 A. The expected voltage and current transients with this device are around 5 - 10 ns. The layout has been carefully designed to minimize the stray inductance. This makes possible the operation at high current slopes without excessive overvoltage on the semiconductors. All the paper is focused on the evaluation of transistor Q_2 of Fig. 1.

Basically, the test consists in applying one turn on pulse to transistor Q_2 to get close to the desired current level in the load inductor, L_{LOAD}. During the second pulse the switch on (*ON*) and switch off (*OFF*) transients are evaluated at the desired current levels.

SiC devices can operate at voltage and current slopes in the range of 100 V/ns and 10 A/ns respectively [9], [10]. This means that both, current and voltage, transients finish in few nanoseconds. A proper measurement of these transients demands voltage and current probes capable to react faster than the measured event.

The rise time of a signal is the time it takes to rise from 10% to 90% of its final value. The bandwidth of this signal is given by Eq. 1 [11]:

$$f_{3dB} = \frac{0.35}{t_{rise}} \tag{1}$$

Therefore, for example, a signal with a rise time of 8 ns has 44 MHz of bandwidth. The voltage and current probes should have a bandwidth at least 3 times higher in order to properly measure the signal [12]. Thus, the required probe should have at least a bandwidth of 132 MHZ.

In the case of Rogowski and Pearson current probes it is usual to provide the usable rise time of the probe. The usable rise time is the minimum measurable rise time to get a measurement error of 10% [13]. The measured rise time should be larger than the useable rise time of the probe. It is necessary that the bandwidth of the oscilloscope exceed that of the probes themselves, so it does not affect the measurement. For that porpose 1 GHz oscilloscope (Tektronix MDO3104) has been used for all tests.

III. Voltage Measurement

The voltage probe reduces the measured voltage to the input voltage range of the oscilloscope. There are two types of voltage probes: active-differential and passive probes. Therefore, the proper probe for a given application should satisfy at least bandwidth, dynamic voltage range and accuracy requirements.

Active differential probes can perform the measurement between two points, neither of which is at ground potential and where the common reference may be elevated (floating) to hundreds or thousands of volts from ground. Till nowadays, to measure Si device transients, active differential probes have been the preferred choice due to their voltage range, isolation capability and common mode voltage rejection ratio (CMRR). Medium cost active differential probes can be found in the range of 1500 volts with bandwidths up to 200 MHz (≈1.7 k€) [6], [14]. However, exceeding those limits increases the cost of probes becoming them prohibitive (>3 k€) [15] for many users. As shown by [7], 3.3 kV to 6.5 kV MOSFET modules are expected in the near future and in consequence, high bandwidth high voltage probes will be required.

On the other hand, passive probes have no isolation and no CMRR rating, as they are single ended probes. This means that the oscilloscope ground and the probe reference ground are at the same potential. But unlike active differential probes, passive probes do offer high bandwidths, up to 1 GHz, at moderate prices (≈1 k€) [16]. As a drawback, they have a low voltage range (300 V), so an extra attenuation is needed to measure higher voltages. Nevertheless, passive probes are good candidates for the voltage measurement of SiC devices up to 300 volts and short rise times (5 -10 ns). Although galvanic isolation is not provided by passive probes, depending on the setup of the test bench and/or the position of the DUT (Device Under Test), the galvanic isolation requirement may be avoided.

A. Description of available voltage probes

In this section, the available voltage probes in the laboratory have been evaluated for the required measurement, Table I.

TABLE I: Evaluated voltage probes.

Device	Vmax (kV)	Type	BW (MHz)	t_{rise} (ns)	Input Z	Cost* (k€)
*DP*25 (Ch.−Arn.)	1	Diff.	25	14	4 MΩ; 1.2 pF	≈0.25
*HZ*115 (Hameg)	1.4	Diff.	35	17/12	60 MΩ; 1.5 pF	≈0.5
*P*5205 (Tektronix)	1.3	Diff.	100	3.5	8 MΩ; 3.5 pF	≈1.7
*P*2220 (Tektronix)	0.3	Pass.	200	<2.2	1 MΩ; 17 pF	≈0.25
*TPP*1000** (Tektronix)	0.3	Pass.	1000	<0.45	10 MΩ; <4 pF	≈1

*Average cost from different sellers in 2018.
**Not used in the first comparison.

As some low voltage passive probes are included in the comparison the bus voltage has been set to 300 V for the first test.

In this first test, a relatively slow voltage transient, (t_{rise}=20 ns), has been measured (15 V/ns). Thus, the signal has a bandwidth of 17.5 MHz and the required sensor bandwidth

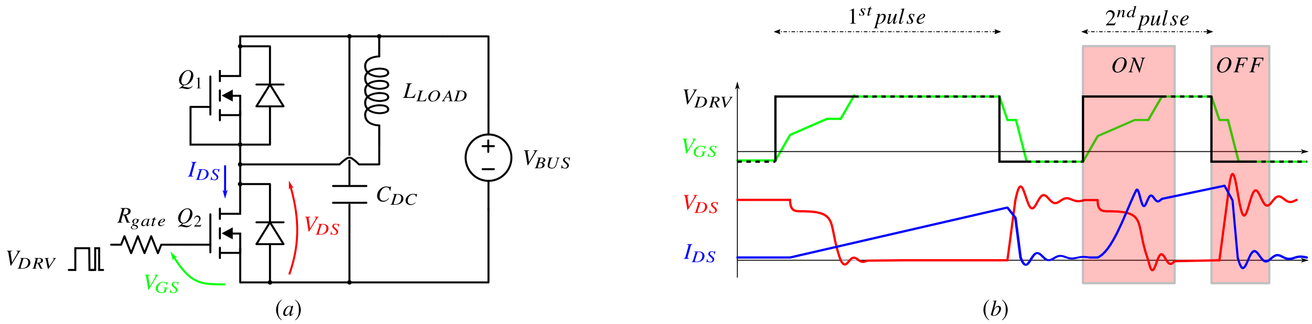

Fig. 1: Double pulse test (a) circuit and (b) waveforms.

is at least of 52.5 MHz [12]. Consequently, only the *P*5205, *P*2220 and *TPP*1000 (not used here) probes should be able to measure the signal. As expected, the *DP*25 and *HZ*115 probes distort the measurement, Fig. 2. In addition, the *P*5205 and the passive probe *P*2220 have enough bandwidth to measure the signal. It can be observed that if a proper de-skew (-8.8 ns) is applied both probes have a very similar response. This time delay between both signals is mainly caused by the voltage differential amplifier. It should be noted that this delay is in the range of the rise time of the measured voltage. In consequence, if this delay time is not properly compensated, the energy loss estimation will have a low accuracy. Thus, it would be desirable to perform the voltage transient measurement without any intermediate electronic amplifier. In this scenario, apart from their lower cost, passive probes are better suited for fast transient measurements. However, their low voltage range makes compulsory the use of some high bandwidth attenuator.

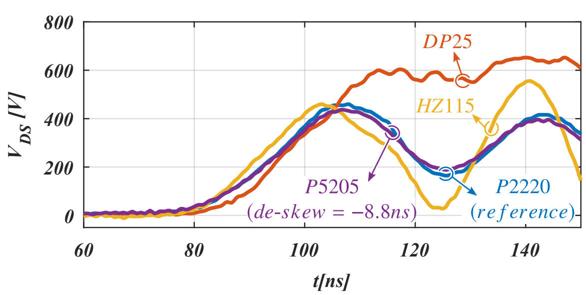

Fig. 2: Dynamic response of the evaluated voltage probes.

In the next section, a single, affordable and high bandwidth voltage divider is proposed that makes possible the use of passive probes at voltages higher than their rated voltage.

B. Proposed voltage divider

The proposed measurement method uses non isolated passive probes. Therefore, for safety reasons, it is only suitable for controlled test environments. One of the first problems that arise when using passive probes is the need to attenuate the signal without losing bandwidth. Due to parasitic capacitances and inductances in the body of a resistor, a highly resistive voltage divider would present a bad frequency response, Fig. 5. A resistive-capacitive voltage divider, Fig. 3, is a better solution since the desired voltage attenuation can be achieve with a highly resistive voltage divider (low standby power losses) and in addition, during the voltage transient, the current injected by the parallel capacitors makes possible the fast response of the network.

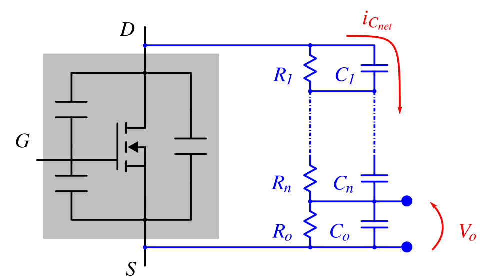

Fig. 3: Proposed RC voltage divider for the V_{DS} measurement.

This measurement method is widely used to perform high voltage transient measurements, [17] and [18]. Assuming that the time constants of all the resistor-capacitor pairs are the same (Eq. 2),

$$R_1 \cdot C_1 = \ldots = R_n \cdot C_n = R_o \cdot C_o \tag{2}$$

the RC network behaves as a pure resistive voltage divider (Eq. 3):

$$H(s) = \frac{R_o}{R_o + \sum_{i=1}^{n} R_i} \tag{3}$$

The RC network should satisfy three design criteria:

1) It should have a high enough bandwidth for the measured signal.
2) It should attenuate the measured maximum voltage down to the dynamic range of the voltage probe.
3) It should present a low equivalent capacitance and high equivalent resistance in order to not affect the test circuit.

Thus, the first step to design the RC network is the definition of the voltage attenuation. Passive voltage probes with an attenuation of 10 (10x) have a dynamic voltage range of 300 V_{RMS}. As a first approach, to have a good voltage sensitivity range, the DC bus voltage is attenuated at R_o to $3/4$ of the probe dynamic range (Eq. 4). In this way, the ground lead noise and ground loop noise effect can be considered despicable. In addition, during the overvoltage of the switch off transient the probe dynamic range is not exceeded.

$$V_o\left(V_{DC}\right) = V_{DC} \cdot \frac{R_o}{R_o + \sum_{i=1}^{n} R_i} \approx \frac{3}{4} V_{probe\ dynamic\ range} \tag{4}$$

In addition, the maximum power dissipated must be defined (Eq. 5). Thus the number of resistors and their resistive values can be estimated.

$$P_{measurement} = \frac{V_{DC}^2}{R_o + \sum_{i=1}^{n} R_i} \tag{5}$$

Once the number of RC pairs has been defined, the value of the equivalent capacitance must be calculated in order to not affect the switching behavior of the device under test (DUT). The current deviated to the RC network during the switch off transient ($i_{C_{net}}$), Fig. 3, is dependent on the drain-source voltage slope as can be seen in Eq. 6.

$$i_{C_{net}} = C_{net} \cdot \frac{dv_{DS}}{dt} = \left(\frac{1}{\frac{1}{C_o} + \sum_{i=1}^{n} \frac{1}{C_i}} \right) \cdot \frac{dv_{DS}}{dt} \tag{6}$$

On the other hand, the switch off voltage slope (dv_{DS}/dt), Eq. 7, of the MOSFET depends on the drain current (i_{DS}), Fig. 1.

$$\left| \frac{dv_{DS}}{dt} \right| = \left| \frac{v_{GS}\left(i_{DS}\right) - v_{DRV}}{R_{gate} \cdot C_{GD}} \right| \tag{7}$$

In consequence, to reduce the influence of the RC network on the switching transient the deviated current ($i_{C_{net}}$) should be reduced as much as possible. In this work, the maximum deviated current ($i_{C_{net}}$) has been limited to the 10% of the drain current (i_{DS}).

For AC signals, high frequency probe suppliers provide a simplified model of the input impedance of the probe, R_{probe} and C_{probe} of Fig. 4. This is enough to analyze how the probe itself affects the measurement. As shown in Table I, the used Tektronix *TPP*1000 has an input impedance of 10 MΩ and less than 4 pF.

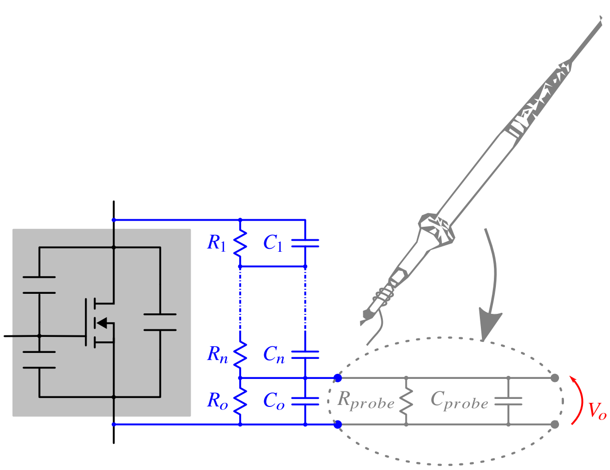

Fig. 4: Influence of the voltage probe in the RC divider.

To avoid any loading on the voltage divider, the value of R_o must be considerably lower than the input resistor of the passive probe. For values of some tens of kΩ the loading error can be considered negligible. Similarly, the capacitive loading should be negligible to avoid any distortion in the transient measurement.

It should be highlighted that all the RC pairs must have an equal time constant. This means, that the tolerance of the used components is of high relevance. Different time constants lead to a distorted voltage measurement.

For comparison purposes, the frequency response of three different voltage dividers have been evaluated and compared. The first one is a pure through hole resistor voltage divider, the second one is a Through Hole (TH) RC voltage divider and the third one is a Surface Mount Device (SMD) RC voltage divider (less stray inductance). As expected, the pure voltage divider has a poor frequency behavior Fig. 5 with a bandwidth lower than 500 kHz. The through hole RC network has a considerably better frequency response than the pure resistor voltage divider. It shows a flat gain response up to 300 MHz while the phase is flat up to 60 MHz. Despite of the resonance at 100 MHz, the SMD RC voltage divider has a quasi-flat gain response up to 500 MHz and the phase is flat up to 400 MHz. Due to their superior frequency response, SMD components have been chosen for the voltage divider.

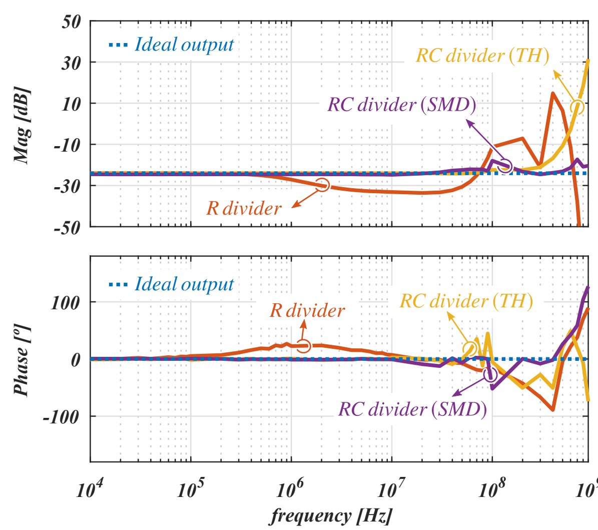

Fig. 5: Frequency response of evaluated voltage dividers.

IV. Current Measurement

The current probe provides an image of the circulating current in the tested device. There are three main types of high bandwidth current probes: Rogowski coils, Pearson current probes and Coaxial Shunt resistors. It must be taken into account that current probes add insertion impedance into the circuit. This should be carefully considered especially when the device operates at fast switching speeds since the additional inductance inserted by the measurement circuit increases the overvoltages and ringings. Up to now, Pearson probes and Rogowski coils have been the preferred choices to test

Si IGBT/Diode modules [19]. There is a wide current range for both sensors with enough bandwidth for fast transients on Si IGBT/diode modules. Both probes can surround screws or terminals and therefore the circuit is not broken to introduce the probe. In small designs, with discrete packages, it is usual to break the circuit to introduce the current probe. This procedure is also required by shunt resistors.

A. Description of available current probes

As it has been done with voltage probes, in this section a comparison of the available current probes have been performed. Table II summarizes the compared three current probes. It can be observed that the three current probes are able to measure very short rise times. However, it can be expected that the available Pearson probe and the Rogowski coil are not suitable to measure the fast current transients (t_{rise}≈5-10 ns) of the tested SiC-MOSFET.

TABLE II: Evaluated current probes.

Model	Current type	Galvanic isolation	BW (MHz)	t_{rise} (ns)	I_{max} (A)	Cost* (€)
Rogowski CWT MiniHF 6	ac	Yes	30	12	1200	≈820
Pearson 410 Current Transformer	ac	Yes	20	20	5000	≈650
T&M SDN-015 Coax. shunt	ac-dc	No	1200	0.3	E_{max}**	≈350

*Cost from different sellers in 2018.
**Pulse width dependant.

The Rogowski coil is a flexible and thin coil that can be inserted easily into the circuit without altering the layout. This current probe provides galvanic isolation. The Rogowski current transducer is composed of a Rogowski coil and an electronic integrator. The operating principle is that the change on the measured current induces a voltage in the coil (Eq. 8).

$$V_{signal} = L_M \cdot \frac{di_{signal}}{dt} \tag{8}$$

where L_M is the mutual inductance between the coil and the signal conductor. The integration of the induced voltage is proportional to the measured current (Eq. 9) [20], [21].

$$i_{signal} = \frac{1}{L_M} \int V_{signal}\, dt \tag{9}$$

One of the main drawbacks of this current probe is that it suffers of capacitive coupling (Eq. 10) when the probe is placed close to a high dv/dt [22].

$$i_{C_{coupling}} = C_{coupling} \cdot \frac{dV_{DS}}{dt} \tag{10}$$

This coupling current distorts the measurement and must be corrected. [23] proposes a two-step measurement procedure to reject this perturbation (Fig. 6 (*a*)).

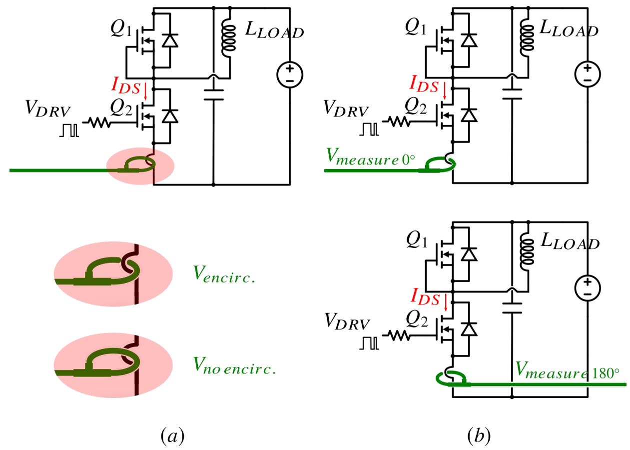

Fig. 6: Noise cancelling methods for Rogowski probes (*a*) noise subtracting and (*b*) angle shift.

After a first measurement (Eq. 11), a second measurement is performed with the Rogowski coil located close to the initial measurement but without encircling the conductor. Thus the second measurement records only the dv/dt perturbation (Eq. 12).

$$V_{encirc.} = V_{signal} + V_{coupling} \tag{11}$$

$$V_{no\,encirc.} = V_{coupling} \tag{12}$$

This perturbation is subtracted to the first measurement and only the current measurement is obtained (Eq. 13).

$$V_{meas.final} = V_{encirc.} - V_{no\,encirc.} = V_{signal} \tag{13}$$

Similarly, another two step procedure to subtract the coupling noise is presented in [24], (Fig. 6 (*b*)).

In the first step the current is measured (Eq. 14) and in the second step the same current is measured placing the coil at 180° (Eq. 15). Thus the coupling effect is the same in both measurements but the sign of the measured current is the opposite.

$$V_{measure\,0°} = V_{signal} + V_{coupling} \tag{14}$$

$$V_{measure\,180°} = -V_{signal} + V_{coupling} \tag{15}$$

Finally both measurements are subtracted to obtain the desired signal (Eq. 16):

$$V_{meas.final} = V_{measure\,0°} - V_{measure\,180°} = 2 \cdot V_{signal} \tag{16}$$

Both methods have been evaluated to remove the coupled noise from a measured current transient, Fig. 7. The fall time of the measured current is higher than the usable rise time of the probe (12 ns). As it can be observed, the same drain current has been measured with a shunt resistor and the Rogowski coil. It can be noted that the Rogowski measurements present a high distortion level due to common mode noise. With the first method, the resultant current has a slower fall rate than the shunt resistor, so the method or at least the performed measurements should be put in doubt.

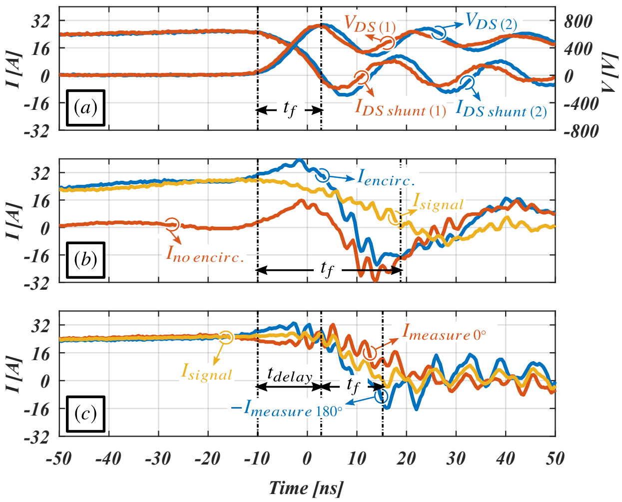

Fig. 7: Measurement of SIC - MOSFET turn-off process: (a) Voltage and shunt current of both tests, (b) Rogowski measurements applying noise subtracting method and (c) Rogowski measurements applying angle shift method.

With the second method, the obtained current has a current derivative similar to the one measured with the shunt resistor (2 A/ns). However, compared with the shunt measurement and the drain-source voltage the resultant current still has a high noise level. In addition, as expected, the integrator amplifier of the Rogowski coil introduces a large delay in the measurement. Due to the difficulties for obtaining a clear measurement with the Rogowski coil, this current probe has been discarded.

Pearson current transformers are often used by many power device manufacturers to characterize their devices [19], [5]. These probes consist of a magnetic core, a secondary winding, a terminating resistor and the electromagnetic shielding. It has a large bandwidth, large linearity and low insertion losses [25]. The operating principle is similar to the Rogowski coil, but instead of an air core, it has magnetic material core where the field is contained. One of the biggest problems with this transducers is that the core is rigid and bulky so it can be difficult its insertion on compact layouts without modifying it.

The coaxial Shunt resistor or Current-Viewing resistor (CVR) is a high frequency current transducer [26]. This coaxial shunt acts as a simple low value resistor. Ideally, the measured current is estimated as Eq. 17:

$$I_{signal} = \frac{V_{shunt}}{R_{shunt}} \tag{17}$$

However, if a high current derivative is measured, the measurement can be distorted by the influence of the stray inductance of the shunt resistor. In addition, the shunt resistor introduces a relatively high impedance into the test circuit. To reduce this insertion impedance, a low shunt resistance can be chosen, however, the signal to noise ratio (SNR) can be compromised and too noisy signals can be obtained. On the contrary, if the resistance value is too high, the high voltage drop in the shunt can influence the estimation of switching losses. Additionally, a shunt resistor does not provide galvanic isolation. This can be a big drawback on field measurements but on test bench environments it is not necessarily a big drawback.

The available Pearson probe has a relatively slow useable rise time and the shunt resistor introduces a high impedance on the measured circuit. In order to estimate the circulating drain current of the tested SiC - MOSFET a current estimation method is proposed in the following section.

B. Proposed current estimation method

The proposed current estimation method basically consists in the measurement of the voltage drop in the stray inductance from the power source terminal to the ground of the circuit. This voltage drop is proportional to the current derivative and the leakage inductance. This voltage drop must be integrated to obtain the drain current (Eq. 18).

$$I_{DS} = \frac{1}{L_{\sigma\,S}} \int V_{L_{\sigma\,S}}\,dt \tag{18}$$

As shown in Eq. 18, the precise value of the stray inductance ($L_{\sigma\,S}$) must be known to solve this integral and estimate the amplitude of the current. The accurate estimation of the stray inductance is a challenging procedure since the current derivate is not constant during the switching transient. Thus, accurate voltage and current measurements and a numerical post processing are necessary to estimate its value. In order to avoid this complex estimation process, in this paper an alternative and easier method is proposed for the estimation of the stray inductance. To do that, during the switching transient, the $V_{L_{\sigma\,S}}$ voltage and the load current (I_{LOAD}) are measured. Since the load current does not present fast transients, a standard low bandwidth current probe can be used for the load current measurement. Therefore, the measurement process of the drain current transient and the estimation of the $L_{\sigma\,S}$ is:

1) The measured voltage in the stray inductance is integrated in the transient time interval (t_1 to t_2 and t_3 to t_4 in Fig. 8). This gives an image of the transient current waveform but with a wrong amplitude ($I^*_{DS_{L_{\sigma\,S}}}$). since at this point the value of the $L_{\sigma\,S}$ is unknown and can not be applied as shown by Eq. 18.
2) In the turn on transient, the final value of Eq. 19 (at t_2) must be equal to the measured load current (I_{LOAD}) at this instant.

$$I^*_{DS_{L_{\sigma\,S}}} = \int_{t_1}^{t_2} V_{L_{\sigma\,S}}\,dt \tag{19}$$

Therefore $I^*_{DS_{L_{\sigma\,S}}}$ must be multiplied to match the amplitudes of the load current and the estimated drain current in the steady state ($t2$). The multiplication factor is shown in Eq. 20 and in consequence, the value of L_{ON} can be easily estimated.

$$\frac{1}{L_{\sigma\,S_{ON}}} = \frac{I_{LOAD}\,(t_2)}{I^*_{DS_{L_{\sigma\,S}}}\,(t_2)} \tag{20}$$

3) In the turn off transient, the estimated current image has a negative value due to the measured negative voltage drop. In a first step, an offset (I^*_{OFFSET}) is applied to

the estimated image to guarantee that the final transient value of the current is zero ($I^*_{DS_{L_{\sigma S}}}(t_4) = 0$ in Eq. 21).

$$I^*_{DS_{L_{\sigma S}}} = \int_{t_3}^{t_4} V_{L_{\sigma S}}\, dt + I^*_{OFFSET} \tag{21}$$

The value of Eq. 21 at t_3 must be equal to the measured load current at this instant. Therefore, in a second step, $I^*_{DS_{L_{\sigma S}}}$ must be multiplied to match the amplitude of the load current in the steady state, the beginning of the current transient (t_3). The multiplication factor is $1/L_{OFF}$ and in consequence, the value of $1/L_{OFF}$ can be easily estimated.

$$\frac{1}{L_{\sigma\, S_{OFF}}} = \frac{I_{LOAD}\,(t_3)}{I^*_{DS_{L_{\sigma S}}}\,(t_3)} \tag{22}$$

4) If both, L_{ON} and L_{OFF}, have the same value, the current estimations can be considered valid. Different values of L_{ON} and L_{OFF} indicate a bad current estimation. High signal/noise ratio in the voltage measurement in the stray inductance has been found to be the most common cause of invalid current estimations. In this case, a higher stray inductance should be considered to increment this signal/noise ratio.

In this application, both, the drain to source voltage and the voltage drop in the leakage inductance have been measured with two matched high bandwidth passive voltage probes. Therefore, measured voltage and estimated currents do not need any de-skew.

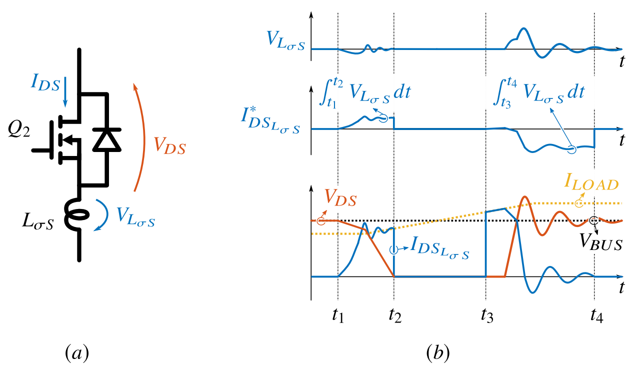

Fig. 8: Ids estimation procedure (a) $V_{DS_{L_{\sigma S}}}$ measurement and (b) fitting method for the estimated current.

V. Results

In order to validate the proposed methods different measurements have been carried out. In a first step the voltage divider has been tested. Fig. 9 shows the design of the voltage divider. All the resistors are 20 kΩ with a tolerance of 0.1% and capacitors are 1 nF with a tolerance of 1%. The layout of the two layer PCB is simetrically designed to minimize the stray inductance in the divider.

To validate the voltage divider, a low drain-source voltage has been measured with the passive voltage probe ($TPP1000$) and the proposed voltage divider. As it can be seen in Fig. 10 both measurements match perfectly.

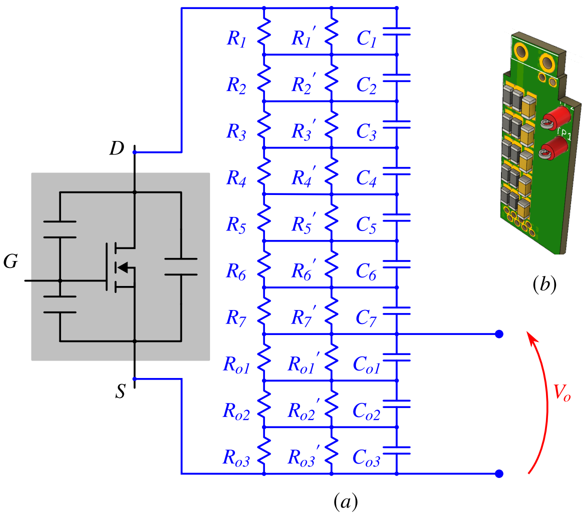

Fig. 9: (a) Configuration of the proposed voltage divider and (b) its PCB design.

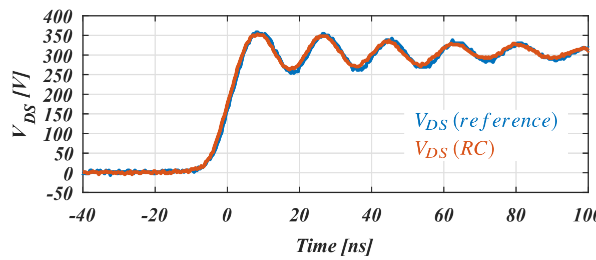

Fig. 10: Validation of the proposed voltage divider.

On a second step, the proposed current estimation method has been validated. In this case, the first measurements have been performed with the coaxial shunt, the Pearson current transformer (Table II) and the proposed current estimation method (Fig. 11). As the Pearson probe has a relatively low bandwidth, this test has been done slowing down the switching process of the MOSFET.

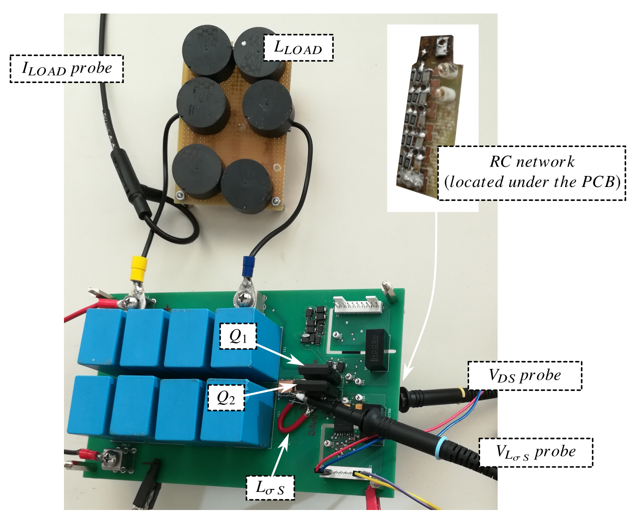

Fig. 11: Test-bench using the proposed voltage and current measuring methods.

It can be observed in (Fig. 12) that the Pearson and the proposed method responses have the same behaviour after applying the required de-skew to the Pearson probe (12 ns). By contrast, the response of the shunt resistor is slightly ahead of the others during the turn on process and considerably ahead during the turn off process. This behaviour is due to the current derivative in the stray inductance of the shunt resistor itself. Thus, a higher distortion is expected if higher current derivatives are measured.

It can be seen, that power losses estimated with the Pearson and the proposed method match perfectly while the losses with the shunt are higher at the turn on transient and lower during the turn off transient due to the influence of the stray inductance.

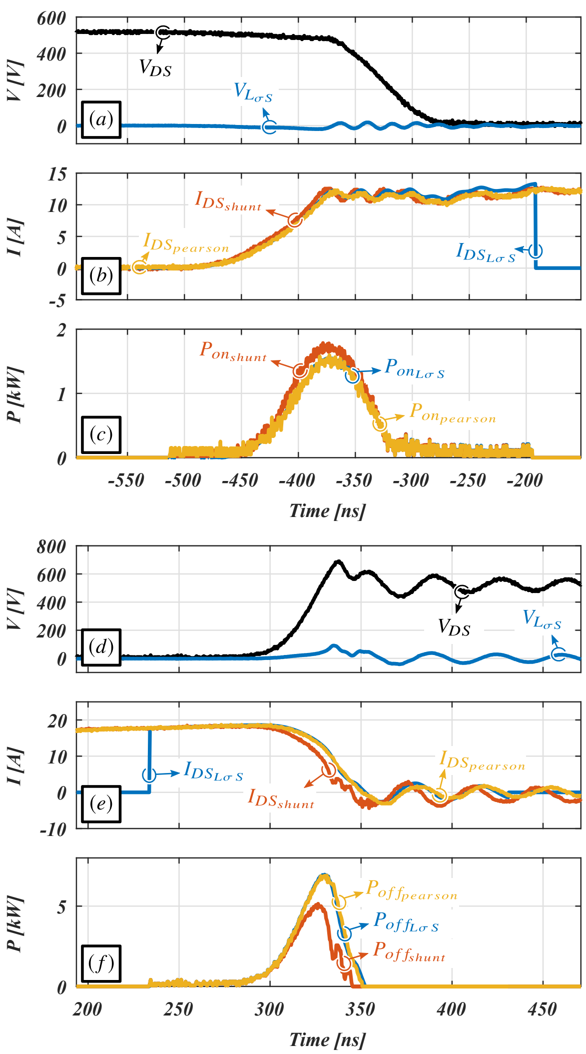

Fig. 12: Switching transient measurement of drain - source voltage (a) turn-on and (d) turn-off drain. Comparison of coaxial shunt, Pearson probe and estimated currents during the (b) turn-on and (e) turn-off process. Calculated switching power losses for the (c) turn-on and (f) turn-off transients.

Once the proposed current estimation method has been validated, a faster switching transition has been measured. In this case, the rise and fall times are shorter than the usable rise time of the Pearson (t_{rise}=20 ns) and therefore, only the shunt and the proposed method have been applied. For the current estimation, the voltage drop of the stray inductance have been measured in two different paths of the PCB, one with 18 nH and the other 27 nH to check that in both cases the estimation is similar. From the obtained results, Fig. 13, it can be observed how the influence of the stray inductance of the shunt becomes more notorious. Thus, during the turn on transient the shunt resistor estimates the drain current leading the drain-source voltage. This indicates a poor behaviour of the used shunt.

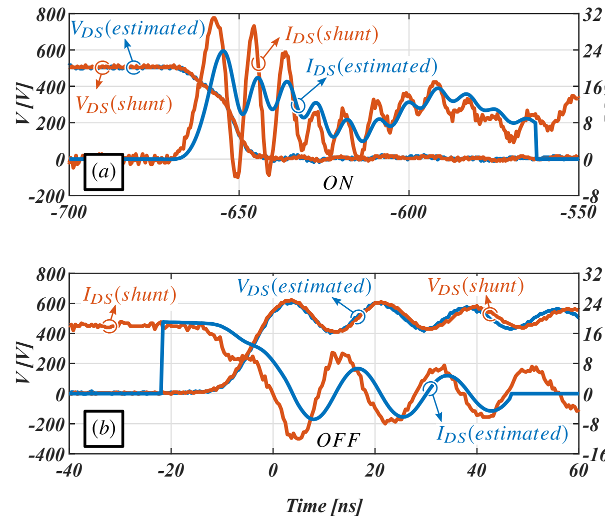

Fig. 13: Comparison of coaxial shunt and estimated currents during the (a) turn-on and (b) turn-off transients.

During the turn off process a similar behaviour has been obtained. A drain-source overvoltage is measured when shunt resistor is measuring a positive current slope. Thus, at least the tested shunt resistor is not considered as a valid device for this kind of measurements. With the proposed current estimation method, both drain current and drain-source voltages are synced. During the off transient the drain-source voltage slope causes the current deviation to the complementary diode. When the DC bus voltage is achieved, the high current derivative leads to an overvoltage. If the current has a zero derivative, no overvoltage is measured. Similarly in the on transient, the current derivative leads to the drain-source voltage drop. Once the complementary PN diode reach the maximum reverse recovery current, the drain-source voltage starts falling to zero.

Thus, with the proposed voltage and current measurement methods, the energy losses of the tested MOSFET have been measured in the real switching circuit. The obtained results have compared with the power losses estimation provided by the manufacturer in their test bench. The same bus voltage and gate resistor values are considered in the comparison.

From Fig. 14 it can be observed that estimated turn on losses are lower than the losses estimated by the manufacturer while turn off losses almost double the losses provided in the datasheet. As there is not any available voltage/current waveforms from manufacturer tests, it is not possible to analyse the nature of the deviation. Different stray inductances, slight non-compensated delays or even errors introduced by the used voltage/current probes could be the reason.

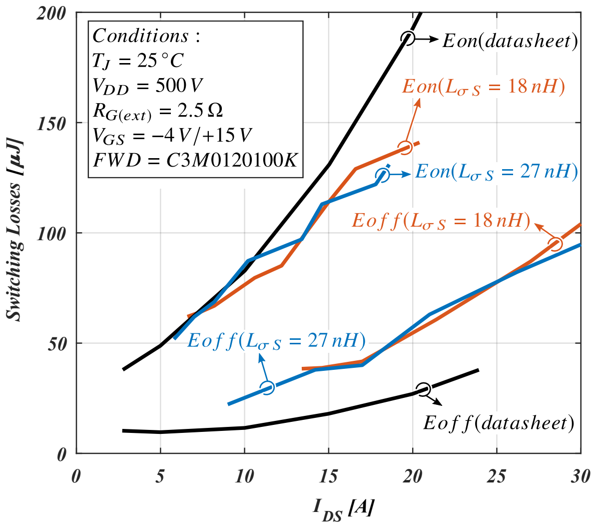

Fig. 14: Switching power losses calculated with the proposed method and switching power losses provided by the manufacturer.

VI. Conclusions

In this paper a simple and affordable method for fast transient measurements of SiC devices has been presented. With this method both, the current and voltage, switching transients are simultaneously measured with by means of two passive voltage probes. Since both measurements rely on matched passive voltage probes, there is no need for any de-skew. This simplifies the transient measurement.

The voltage measurement method uses a high bandwidth voltage divider and a passive voltage probe. The same passive probe can be used even if different drain-source voltages are measured just changing the attenuation of the voltage divider.

The proposed current estimation method relies on the voltage drop in the stray inductance of the current path. If the layout has enough stray inductance, this method provides virtually zero insertion impedance. If a standard current sensor is additionally used for the load current measurement the estimation of the stray inductance is greatly simplified. Thus, the estimation of the transient current is simple to perform.

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