Reducing the Impact of Skin Effect Induced Measurement Errors in M-Shunts by Deliberate Field Coupling

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Keywords

« Device Characterisation » « Sensor » « Measurement » « Noise » « Pulsed Power »

Abstract

The ever-increasing switching speed of semiconductor devices requires a precise measurement of steep current transients. The M-shunt concept offers high signal fidelity, good cooling, and simple manufacturing. Depending on the resistive material used, temperature as well as skin and proximity effects impede static and dynamic measurements to a different degree. A step forward has been derived from the ideal coaxial shunt, so far, a purely theoretical concept, which is hardly producible due to its sophisticated structure. By transferring this concept to the M-shunt structure with its improved PCB manufacturing technologies it can now be realised in practice. Nevertheless, the calibration and the correct degree of delay compensation remain challenging and are investigated more closely within this paper. Furthermore, it will be discussed why the conventional method of bandwidth determination doesn't work for the M-shunt structure. In addition to the low inductance introduced into the load circuit, the high bandwidth of the shunts could be demonstrated, as well as the possibility to extend this by design rules. Supplemented by the advantages of the lower load inductance, the M-shunt will become the tool of choice for characterising switching transients at least up to 200 MHz required bandwidth eventually. Although it is obviously difficult to improve the 3 dB bandwidth with suitable design rules, the range of nearly entirely unaffected measurement frequencies (e.g. < 1 dB) can be significantly extended by limited coupling. For even higher frequencies, measurements of the current M-shunt models, as well as for the coaxial shunts used as reference, should be corrected by post processing to get precise measurement results.

Introduction

Modern power electronics utilise wide band-gap (WBG) semiconductor devices made from SiC or GaN. These devices have properties superior to those of silicon and conquer a growing market share in power applications like automotive, aerospace, microgrid, renewable energy, and many more. One of the devices' main characteristics is their extremely fast switching with switching times down to the nanosecond range. On the one hand, this enables low switching losses and high-frequency operation, but on the other hand, it is a challenge for dielectrics and may cause severe EMI, especially when interacting with parasitics. In other words, a detailed and high-fidelity characterisation of the switching transients is required to avoid SOA or EMI problems. While voltage measurements are basic craftmanship, it is challenging to introduce a current sensor into the power loop without introducing additional parasitics, thereby making measurement errors due to the high dv/dt, di/dt, EMI, or cross-talk, which come in addition to the usual measurement inaccuracies. For high fidelity measurements, the coaxial shunt concept is state of the art and the M-shunt concept was proposed as an alternative, offering good signal fidelity, better cooling, lower inductance introduced into the power loop, and lower cost due to simple manufacturing [1]. Furthermore, the M-shunt allows for a simple and reproducible, though only partial compensation of the skin and proximity effect [2]. The concept was theoretically proposed for the coaxial shunt [3], but has just not been feasible in (mass-) production. However, the choice of resistive materials and the temperature impact significantly influence the actual measurement performance with conflicting optimisation targets.

This paper shows the quantitative extinction of coupled fields and the influences of the skin effect. In particular, the use of temperature-compensated resistive materials is influencing the reduction of inductance in the measuring circuit, such that a non-compensated material with post-processing for temperature compensation could be a better option. Furthermore, a delay-free measurement requires different design rules for the M-shunt and offers additional degrees of freedom, which provide another advantage over the coaxial shunt. [2] The investigation was carried out through repetitive double pulse tests (DPT), comparing the fidelity of the current measurements. Moreover, transfer characteristics of the manufactured shunts were recorded with a network analyser to obtain a better knowledge of the shunt's measurement behaviour and precision.

II. Double pulse measurement and parasitics

The measurements for this investigation were carried out with a double pulse test. Figure 4 shows the DPT board with its integrated PCB-based M-shunt, which facilitates the lowest additional stray inductance due to the shortest connecting PCB traces and due to adhering strictly to the principle of coplanar conductors. Furthermore, particular attention was paid to reducing parasitics, which would otherwise result in oscillations, and current or voltage overshoot during the fast switching of the WBG devices [4]. Such parasitic inductances and capacitances have an adverse influence on the measurement and lead to deviation from the ideal switching waveforms [5]. Figure 6 shows the schematic of the DPT set-up including parasitics. Especially the parasitic capacitance C_L of the load inductor, the parasitics capacitance C_{FWD} of the free-wheeling diode, and the overall stray inductance of the power loop cause significant overshoot and ringing [6]. Because slowing down the switching by increasing R_G or C_{GS} is not an option, the reduction of the overall parasitic inductance in the DPT board, also including the measuring system, was the main design objective.

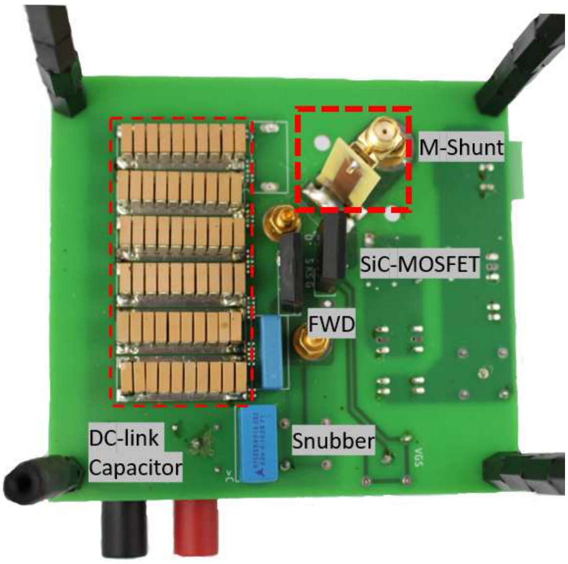

Figure 1 M-shunt integrated into the measurement board of the DPT set-up (red box top right)

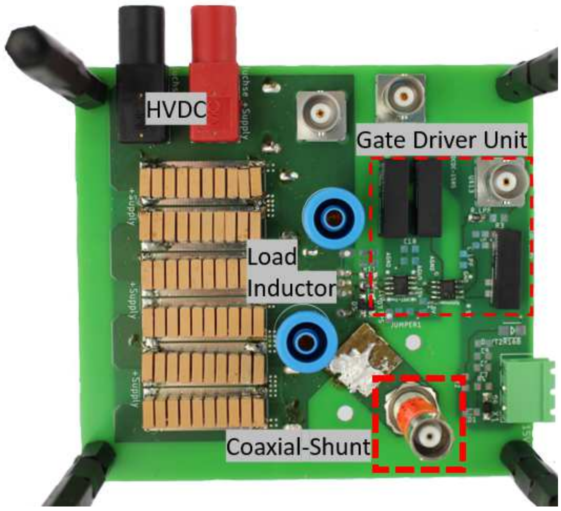

Figure 2 Bottom side of the measurement board shown in Figure 1 with a coaxial shunt measurement in series (red box bottom right)

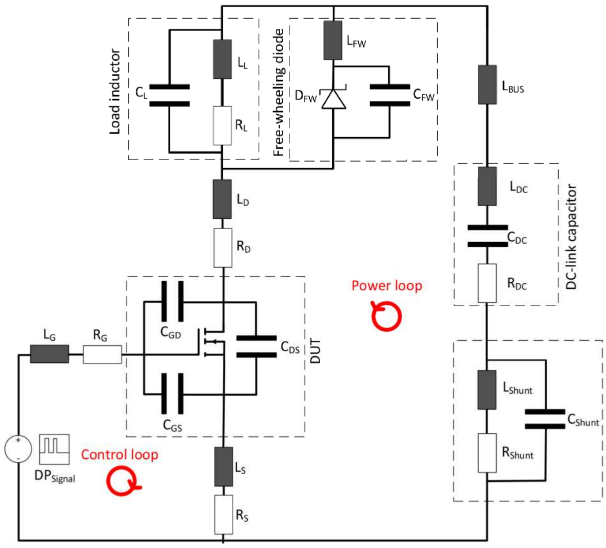

Figure 3 Circuit diagram of the DPT set-up including parasitics

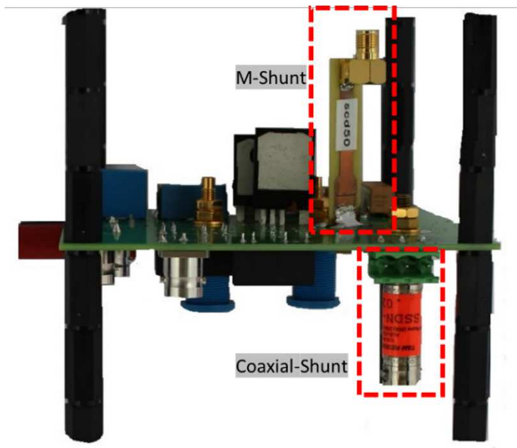

Figure 4 Side view of the measurement board for shunt comparison with the M-shunt sticking out upwards and the coaxial shunt sticking out downwards

III. Current measurement methods and their limitations

There are several well-established current measurement techniques based on shunt resistors, current transformers, Rogowski coils, the Hall effect with open and closed-loop sensing, and fluxgate [7]. All of these techniques are well-suited for low frequency or slow transient measurements. In contrast, fast switching applications utilising WBG semiconductors are still a challenge due to the high di/dt and dv/dt rates. On the one hand, the bandwidth has to be very high to precisely measure the steep current slopes, but on the other hand, inductance inserted into the power loop must be negligibly small to avoid disturbance of the circuitry to be measured. Here a trade-off between bandwidth, accuracy, power limitation, interference with the circuitry, and of course cost occurs. Commercial Rogowski coils are not suited for steep transients, though, Rogowski coils with a higher bandwidth have already been reported [8] [9]. However, these are still limited due to inductive or capacitive couplings from the power loop, inserted resistance, and saturation or filter effects. State-of-the-art for fast switching devices are shunt resistors, especially the coaxial shunt with claimed bandwidths up to 2 GHz [10]. However, the coaxial shunt shows important limitations because it introduces a high additional inductance into the power circuits, e.g. 2 to 8 nH for the 100 mΩ version [11], and is difficult to cool, thus, it is quite limited in the energy of a single measurement pulse and power dissipation during continuous operation. An alternative is the M-shunt, which introduces only little inductance into the power circuit. Mostly less than 1 nH for the versions used. [11] [2] Additionally, it is much easier to cool. For comparison: The 25 mΩ coaxial shunt models are specified with a maximum energy of 3 J. [10] Depending on the specific model, the M-shunts used here were classified with maximum energies ranging from 100 J to significantly more than 200 J, respectively, while not requiring a larger installation space on the circuit board. [12] A comparison of the pros and cons of the main current measurement methods is listed in Table 1.

Table 1 Pros and Cons of selected current measurement methods, c.f. [1].

	Rogowski coil	Coaxial shunt	M-shunt
Advantages	• Galvanic insulation • Almost no additional loop inductance • High current measurement	• High bandwidth (claimed up to 2 GHz) • High accuracy • No auxiliary power required	• High bandwidth • High accuracy • Easy integration • Easy cooling • No auxiliary power required • Low additional inductance
Disadvantages	• Low bandwidth • Low accuracy, dependent on coil position and dv/dt • Not suited for DC measurement • Auxiliary power required • Signal delay caused by integrator	• No galvanic insulation • Additional loop inductance • Difficult cooling • Pulse energy is quite limited • Large mechanical size	• No galvanic insulation • Additional loop inductance

Within the scope of this paper, the M-shunts manufactured, will be compared to commercial coaxial shunts as the reference. The reference chosen is one of the most common versions. For the DPT the 25 mΩ was used for proper comparability. However, the overall inductance in the loop will be increased by two current sensors in series leading to overshoots due to the introduced inductance. The comparison will show differences in the output of the measurements and allows a dedicated comparison of the shunt models.

IV. Shunt Measurement of high fidelity

The general cross-sections of the coaxial shunt and the M-shunt are shown in Figure 5 and Figure 6, respectively. The most important advantage of the coaxial shunt is the ohmic current measurement in a field-free region, which is ensured by the radial symmetry of the resistive cylinder and which allows for direct measurement without the necessity of a compensation circuitry. However, parasitic inductance is indeed introduced into the circuit to be measured. This inductance can be reduced by large radii (leading to low H–Field), and small radius differences (resulting in small flux), which makes the design more complicated and the device larger. The M-shunt is a planar construction with the same cross-section. It can approach the field-free condition if designed according to the usual rules of coplanar conductors, i.e. mainly the dielectric between the conductors needs to be as thin as possible to avoid stray inductance. A pivotal point of the M-shunt design is the current splitting part “current distributor”, which was already introduced in [2] and is shown in the box marked red on the left side of Figure 7 and which ensures an even current flow through the top and bottom traces of the shunt. Another challenge of the M-shunt design is avoiding edge effects – or turning them into a feature. The M-shunts characterised within this paper are based on the manufacturing and design concepts of [12], although some of the newer variants are equipped with SMA sockets, as can be seen in Figure 8.

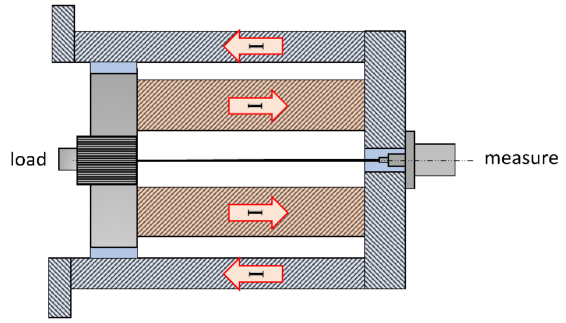

Figure 5: Schematic cross-section of the rotational symmetric coaxial shunt [3]

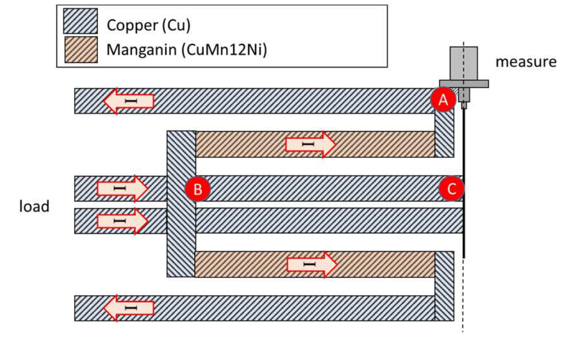

Figure 6: Schematic cross-section of planar 6-layer PCB M-shunt [1]

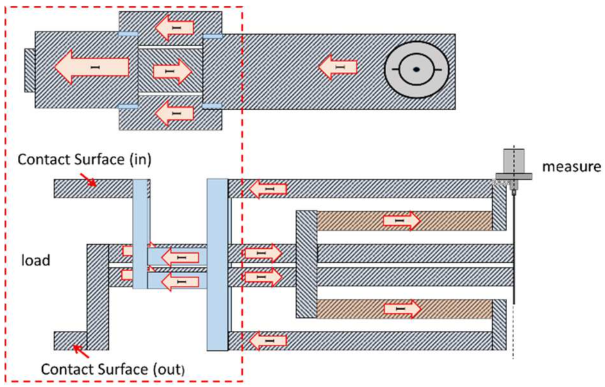

Figure 7: Shunt concept including current distributor

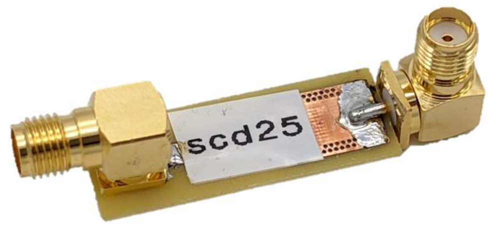

Figure 8: Picture of an M-shunt reduced to 25% size with an SMA socket for the measurement (left) and for load connection (right)

Fundamentally, there is a severe frequency limit for the coaxial- and M-shunt if the thickness of the resistive layer is larger than the skin depth. The current density and ohmic field will not reach the inner region of the resistive layer. For coaxial shunts, this is a known phenomenon [3] and maybe best understood using the concept of “field diffusion”. As shown in Figure 9, the skin and proximity effects cause a current displacement towards the inner edge of the individual top and bottom power traces that make up the M-shunt, or to the centre of the tubes of the coaxial shunt, respectively. This uneven current distribution across the conductors leads to a delay in the measurement signal picked up by the taps at the outer edges. The delay of the measured signal concerning the actual current signal is shown in Figure 10 and comes on top of the inaccuracy caused by the higher effective resistance, i.e. measurement signal is larger than it should be. Even worse, the temperature-dependent increase of the resistivity in the shunt would cause an elusive change of skin depth. This additional portion is difficult to determine and could be avoided if the resistive layer was made from a temperature compensated material like Manganin ® (registered trademark of Isabellenhütte Heusler GmbH & Co. KG). The higher resistivity of Manganin (copper-mangenese-nickel alloy), compared with pure copper generates a larger skin depth per se and accordingly less delay at steep transients. To achieve the identical resistance value at room temperature, the Manganin must be about 25 times thicker than copper, which substantially increases the influence of the skin effect, as shown in Figure 10. At the same time, the skin depth improves only by a factor of less than 16. Depending on the ringing frequency and the overall layer thickness, the influence of the skin effect can increase significantly.

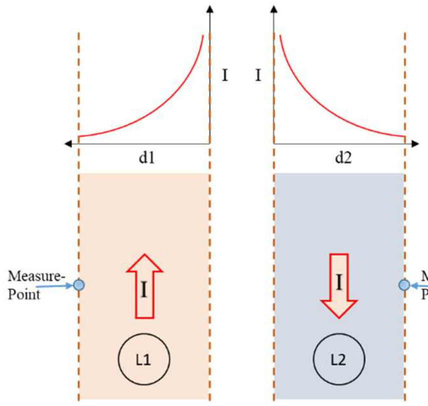

Figure 9: Principle diagram of current distribution in the conductor (L1) and resistor (L2) in coaxial- and M-shunts [2]

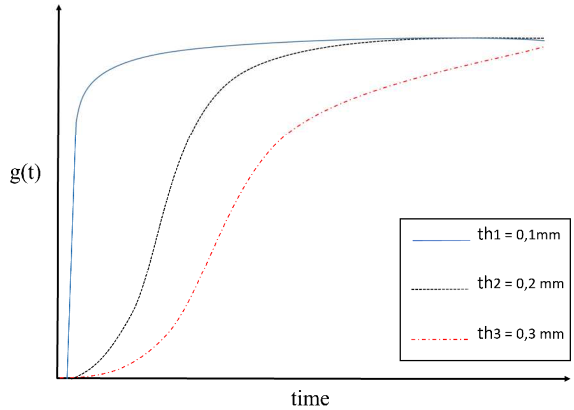

Figure 10: Sketch of the step response of coaxial shunts at high frequencies depending on the wall thickness (th_1=0.1 mm; th_2 = 0.2 mm th_3 = 0.3 mm) [3]

V. On the Difficulties of Bandwidth

Bandwidth determination is straightforward for the coaxial reference shunt. In most cases the +3 dB limit is determined for the shunt's transfer function. For the coaxial reference the magnitude decreases steadily and would run out of the range, shown in Figure 11, at its +3 dB limit. For the 50 mΩ version characterised within this work this point lies around 400 MHz. This determination method is based on the equivalent circuit model of a resistive current sensor connected to an oscilloscope, shown in Figure 12, while the resistance of the oscilloscope Z_l is set to 50 Ω for all measurements within this paper.
This two-port model consists out of the total parasitic inductance L_{ins} and the mutual inductance L_m caused by the magnetic coupling between load and measuring circuit. [13] The L_s seen in the Figure 12 represents the inductance component introduced exclusively into the load circuit, consisting essentially of the current distributor, but not the resistor, so the route from mark A to B in Figure 6 is excluded from the total inserted inductance L_{ins}.
In this case the transfer function is basically described by:

$$G = \frac{V}{I} = R + j\omega \mathrm{L}$$

The situation is more difficult in the case of the determination of the M-shunt's bandwidth. Considering the curves for shunts of the basic design, only changed in their overall size, a completely different behaviour is obtained. The additional design freedom, which will be used for the dedicated coupling described, works here as a capacitive component, so the circuit diagram has to be changed and consists of further capacitive and inductive components but is not fully explained so far.
To determine the limiting frequency, the +3 dB gain limit seems to be method of choice and is given for the reference model used at 400 MHz by the datasheet [10] and measured as 45 MHz, shown in Figure 11. Figure 13 shows a picture of the versions compared, which are all designed as 25mΩ resistances, but changed in the overall area, through simultaneous reducing of the length and the width to 75 %, 50 % and 25 % of original 100 % length and width, respectively.
This option, tagged as f1 in Table 1, seems to be too optimistic for the M-shunt because the negative trend upfront already exceeds comparable deviations, but in the opposite direction (damping). To transfer the 3dB limit in the negative direction on the other hand seems to be too pessimistic because the amplification factor of 0.7 (-3 dB) in other terms would be much lower than the 1.41 of the +3dB boundary. Permitting the same deviation from the 0 dB mark, the resulting frequency would be even higher but still arbitrary. In any case, it can be ensured that the most pessimistic value (±3 dB) allows equal or smaller deviations than the positive gain limit of the coaxial shunt. Based on this pessimistic variant, a useful bandwidth of values between 141.6 and 214.74 MHz results for the considered shunts. In any case, this is still lower than the values given in the datasheet for the coaxial

reference (400 MHz at 25 mΩ), whereby the reference shunts used in this study, as in other studies, fall behind the datasheet promise by far. [13].

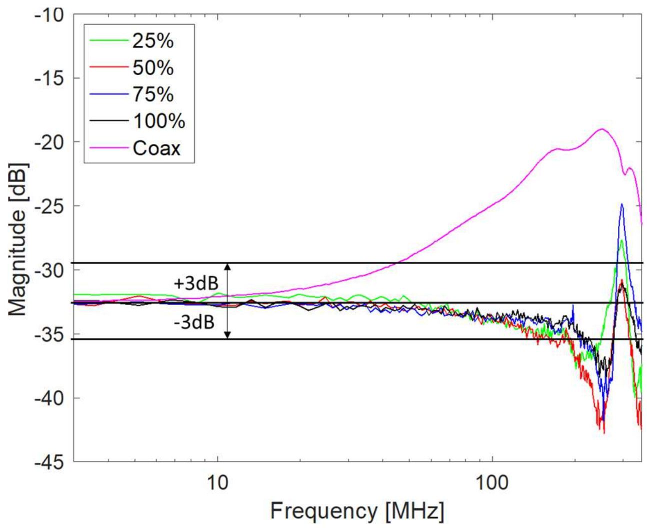

Figure 11 S-Parameter measurement for M-shunts of different size but with nearly same resistance (also shown in Figure 13) and reference coaxial shunt (-1dB Offset for better comparative view)

Area [%]	f1 (+3dB) [MHz]	f2 (-3dB) [MHz]	R [mΩ]
25	286	141	25.3
50	570	146	23.6
75	284	209	22.6
100	478	214	23.4

Table 1 Summary of the positive and negative 3 dB limits for shunts of different area

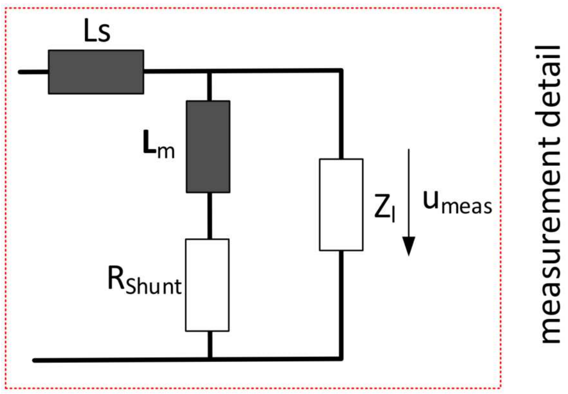

Figure 12 Equivalent circuit for resistive current measurement with termination resistor [13]

Figure 13 Picture of the shunts of different area

VI. Concept of adjustable field coupling for delay compensation

The basic idea of the designs analysed pursues influencing design-related inconsistencies and measurement errors through targeted field coupling as introduced for the ideal coax-shunt. (Figure 14) [3] For this purpose, the basic design, presented in [12] was modified in three different ways. Figure 15 shows the basic concept of the delayed measurement signal, and the intended field coupling. The central goal is to achieve the ideal step response from the superposition of the delayed measurement signal and the signal components generated by coupling [2], at least in the range of the usual measuring frequencies.

In the first step, only the outer lead of the M-shunt's cross-section (conduction layer) was widened, while the measurement tap and resistor layer remained the same width. Without changes at the measurement tap nearly no changes could be detected for the measured signal (design "CL"). In the second step, in addition to the outer lead, also the measurement taps were widened equally (Figure 16 of design "CLML"). The number written in the identification of the "CL" or "CLML" curves is representing the proportional width extension, compared to the width of the resistance layer, as illustrated in Figure 16. In this case, a clear influence on the measured output signal can be noticed (Figure 18). A wider overhanging surface (tested up to 50 % width) results in a steeper transient and a larger overshoot. By comparing to coaxial shunt measurements performed in series with the M-shunt measurements shown, it can be concluded, that the coupling is (as expected) only part of the measurement and is not influencing the load circuit itself. (Figure 19) The reference measurements of all curves, performed with the coaxial shunts in series, show almost a perfect match. Therefore, one may conclude that the M-shunt model is suitable to transfer the concept of the ideal coaxial shunt into practice to remain capable of measuring at high frequencies. In principle, these improved frequency responses should be recognisable through an improved transfer function, which will assist to determine the right amount of overhang in order not to overcompensate for the delay.

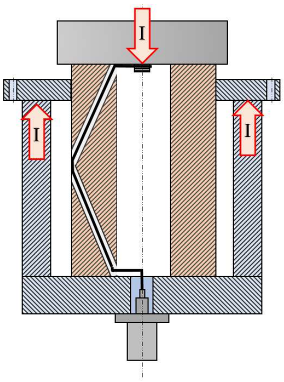

Figure 14 Cross section of a coaxial shunt with improved measuring tap [3]

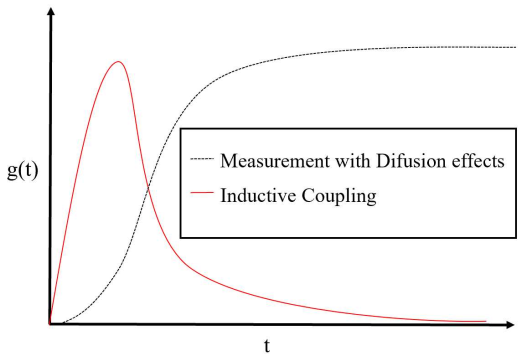

Figure 15 Visualisation of the signal components [2]

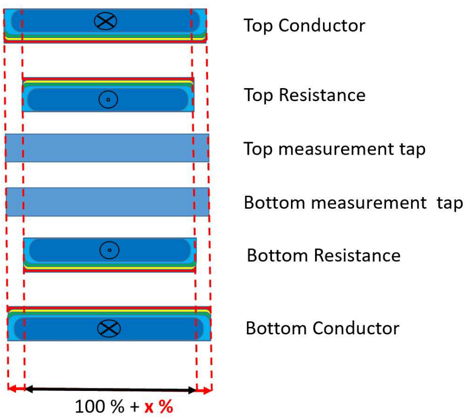

Figure 16 Design Cross section drawing for shunts with dedicated overhanging coupling areas in conductor only (CL) and conductor plus measurement Layer (CLML) [2]

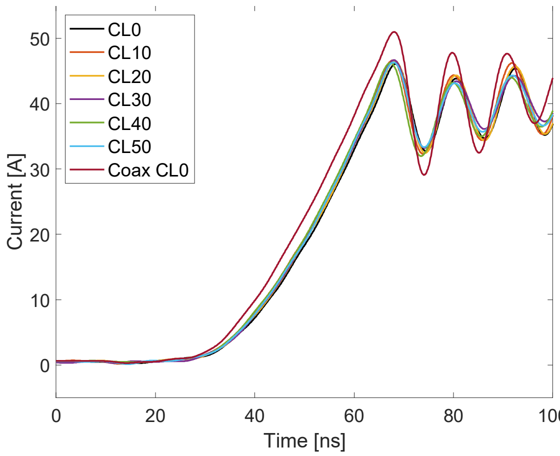

Figure 17 Comparison of turn-on behaviour measured by Shunts with wider conduction layer compared to their coaxial reference shunt measurement in series

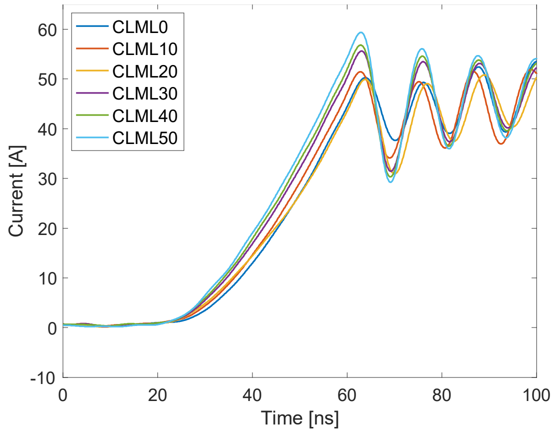

Figure 18 Comparison of turn-on behaviour measured by Shunts with wider conduction and measurement layer from 0 to 50 % overhang (CLML)

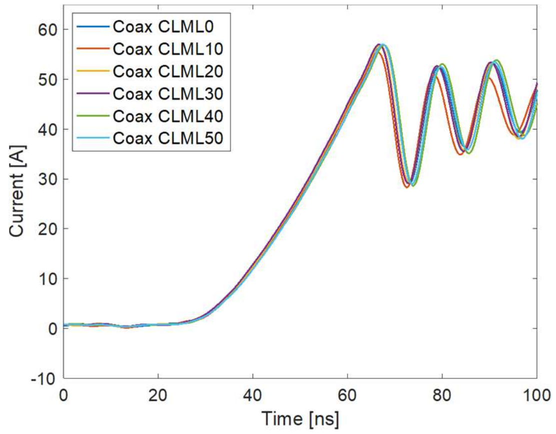

Figure 19 Comparison of coaxial shunt measurements of turn-on behaviour measured in series of the CLML designs. The differences seen in Figure 18 are only measured and not traceable in the circuit itself

VII. Verification of the Transfer Function

In case of the coaxial shunt, the following method, i.e. the calculation of the transfer function from the measurement signals, is used to verify the bandwidth [13]. For the M-shunt, it is rather used to correctly determine the transfer function as a whole, because the bandwidth determination has its difficulties as mentioned. Figure 20 shows the measurement result recorded with the network analyser of the already introduced CLML designs. The change of the gain over the frequency response as a function of the exceeding coupling surface is evident and emphasises the observations described in section VI. Shunts with an overhang of more than 10% obviously overcompensate the measurement delay already as was also shown by the curves of the double pulse test. The longest-lasting bandwidth with almost no deviation (<1dB) seems to be in the range of a coupling area around 5 % overhang (between 0 % and 10 %). Table 2 shows the decreasing bandwidth via the coupling surface set at the +3dB limit. As mentioned the 0% coupling version has the highest bandwidth per definition. On the other hand, due to its previous damping (-3 dB limit), it would basically get a rather bad value, as described in section VI. The calculated measuring inductance is therefore only a mathematical factor and should be treated with caution in a physical context due to the above-mentioned difficulties of the equivalent circuit diagram for the M-shunt.

In the previous section Figure 18 shows the comparison of the measurement curves of the CLML variants. Figure 21 in this section shows the same measurement curves if adjusted by their transfer function to the ‘real current’ considering [7]:

$$i_{meas}(j\omega) = i_{real}(j\omega) \cdot \frac{G(j\omega)}{R}$$

respectively:

$$i_{real}(t) = \frac{1}{R} \cdot \mathcal{F}^{-1}\left\{\frac{\mathcal{F}\{i_{meas}(t)\}}{G(j\omega)}\right\}$$

To verify the concept of the postprocessing Figure 22 shows the measured signals of the “CL0” M-shunt compared to its coaxial reference, like it was already shown in Figure 17. In addition, it is shown how the signal looks after postprocessing of the 44 MHz signal shown. Considering the measurements performed combined by the aid of these calculations, it can be stated that the coaxial shunt clearly falls behind its promise and has a comparable bandwidth to the M-shunt. However, within their 3 dB limits respectively for low frequencies below 150 MHz, the average deviation, i.e. the measurement error at low frequencies, is significantly larger for the coaxial shunt than for the M-shunt. Nevertheless, the comparison shows that not all measurement inaccuracies can be fully prevented by the post-processing, which is the reason for the differences between the final signals.

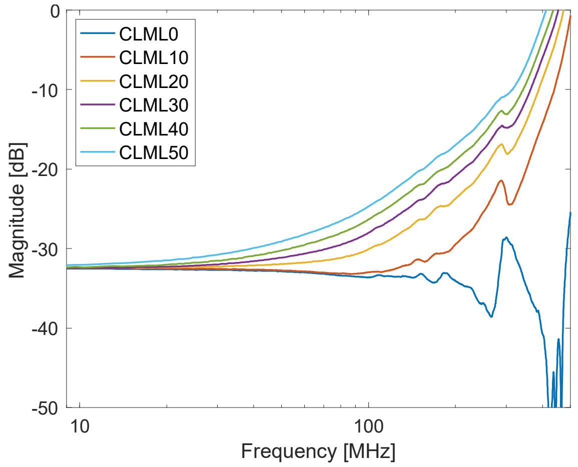

Figure 20 S-Parameter measurement for shunts with wider conduction and measurement layer from 0 to 50 % overhang (CLML)

Area [%]	R [mΩ]	f_{BW} [MHz]	L [pH]
0	24,16	290	13,22
10	24,26	202	19,09
20	24,39	113	34,23
30	24,43	82	46,97
40	24,47	62	62,67
50	24,48	47	83,72

Table 2 Overview on the ±3dB bandwidth for the CLML M-shunts with different amount of overhang.

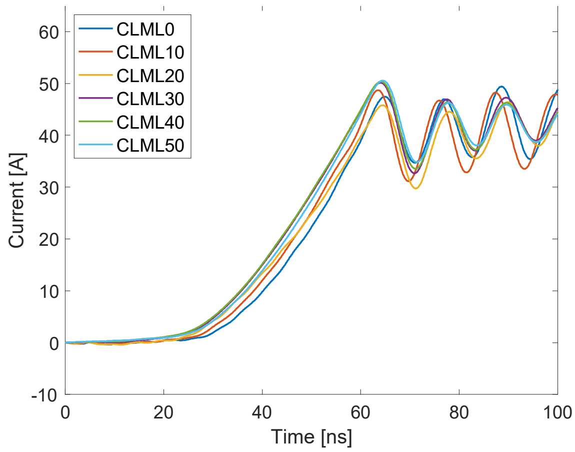

Figure 21 Comparison of turn-on behaviour measured by Shunts with wider conduction and measurement layer from 0 to 50 % overhang (CLML) after post-processing

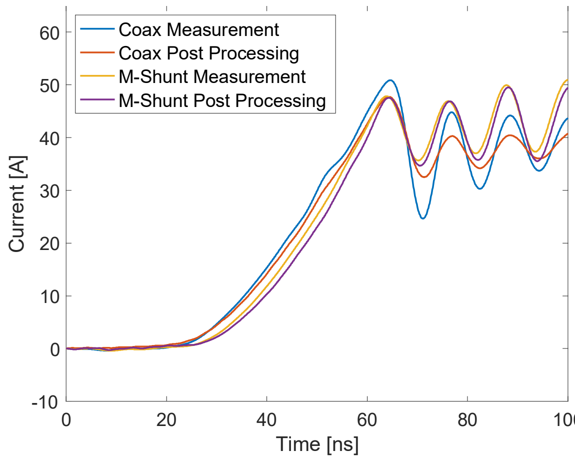

Figure 22 Comparison of turn on characterisation measured with M-shunt and coaxial shunt. Furthermore same measurement results after postprocessing

Conclusion

In summary, the series of experiments presented here prove the possibility to shape the output signal of the M-shunt according to the principle of the ideal coaxial shunt. However, the theoretical idea of a step response is naturally subject to delays and limits the validity of a clean calibration by means of controlled coupling. Nevertheless, the "almost" instantaneous step responses of GaN switching can certainly be used for iterative approximation towards the "ideal M-shunt", although it should be noted that there is always a narrow line between glossing over results and genuine calibration or correction of errors. It must be mentioned that the regular bandwidth definition is not directly transferable for the M-shunt, due to its capacitive components. Nevertheless, at least at frequencies up to 200 MHz, the M-shunt is an equivalent measuring device to the coaxial shunt without any post processing or compensation. Supplemented by the advantages of the lower introduced inductance [11], the M-shunt will become the tool of choice for the characterisation of switching transients at least up to 200 MHz required bandwidth, eventually. Although it is obviously difficult to improve the 3dB bandwidth with suitable design rules, the range of almost entirely unaffected measurement frequencies (e.g. < 1dB) can be significantly extended through defined coupling areas. For even higher frequencies, measurements of the current M-shunt models, as well as for the coaxial shunts, should be corrected by post processing to get precise measurement results.

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