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Voltage Balancing of Series IGBTs in Short-Circuit Conditions

Mostafa Zarghani, Saeed Peyghami, *Member, IEEE*, Francesco Iannuzzo, *Senior Member, IEEE*, Frede Blaabjerg, *Fellow, IEEE*, Shahriyar Kaboli

***Abstract*—This paper proposes a scheme for balancing the voltage of series-connected Insulated Gate Bipolar Transistors (IGBTs), which is also very effective under short-circuit conditions. An optimized clamp-mode snubber is proposed, including an active-driver to balance the currents of the IGBTs during short-circuit, which in turn allows a considerable reduction of the snubber capacitance. The approach is proven to be effective under short-circuit conditions by a maximum 10% voltage imbalance and a negligible difference in the current values. The effectiveness of the proposed approach is demonstrated through simulations in PSPICE software and experimental tests performed at 2000 V.**

***Index Terms*—Series IGBTs, Short-circuit Fault, Voltage Balancing, Clamp Mode Snubber.**

I. Introduction

One of the most popular switches used in high-power electronic converters is the IGBT. High voltage, high current, a simple gate-driver circuit, medium switching losses, low conduction losses, high short-circuit current capability, high reliability, and advantageous economic parameters such as cost and size per rated power are the main merits of IGBTs over other power electronic switches [1]. However, they face some limitations, at high-voltage applications, such as limited maximum current density and low switching speed[2]. Some remarkable approaches have been presented to improve the IGBT characteristics [3]–[7]. But operations beyond 3 kV - 4 kV per device still present a big hurdle. To overcome these issues in applications like high-voltage motor drivers [8], [9], solid-state circuit breakers, medium/high-voltage DC transmission systems [10]–[13], and solid-state modulators[14]–[16], the only option is still the use of a series stack of IGBTs. However, the unbalanced voltage sharing and the galvanic isolation, which are needed to connect the gate drivers in the string, are the most critical issues. Various solutions have been presented in the literature to address these issues under normal operating conditions [8]–[16].

Short-circuit faults are likely to occur in the operating life of power converters. IGBTs can tolerate short-circuit currents for a limited time as far as the operating voltage and the current remain inside the safe operating area (SOA) [17]. Various techniques have been presented to protect the IGBTs in short circuit situations. For instance, de-saturation detection [20], current sensing [19], gate voltage sensing [20], and *di/dt* feedback [21] are common schemes to detect the short circuit condition. After detecting a short-circuit fault, the IGBT is turned off slowly according to the chosen strategy and opportune system considerations [22]. However, the mentioned techniques for protecting IGBTs under short-circuit faults are mainly suitable for a single unit. since IGBT units do not have identical I-V characteristics, these techniques may not succeed on IGBT strings. The intrinsic differences of IGBTs and their drivers stem from 1) parametric variation in the manufacturing process and 2) the aging process during operation.

The facts mentioned above will induce unbalance voltage sharing over series IGBTs in the short-circuit fault duration. Two IGBTs in series are assumed to clarify this issue, as shown in Fig. 1 (a). If a short circuit fault occurs, the short circuit current will pass through the IGBTs. Since IGBTs have different I-V characteristics, as shown in Fig. 1 (b), they will experience different voltages, and the IGBT with a lower short-circuit current will have a higher voltage. In this condition, even if the IGBT can tolerate the short circuit energy shock, its immediate failure is obvious due to the extra overvoltage [23]. In [17] short-circuit safe operating area for a single IGBT is investigated. When a short-circuit fault occurs for an IGBT and its voltage is low, the IGBT can tolerate its nominal short-circuit current for a determined time. But for medium or high voltage ratios, the tolerable short-circuit current decrease as the IGBT voltage increase. This means that for a specified short circuit current, the maximum tolerable voltage in short circuit fault conditions is lower than the maximum blocking voltage in the turn-off state.

Notably, the schemes that are reported in [8]–[16] for balancing the voltage of IGBTs in the normal operation condition cannot guarantee that the voltages of the IGBTs are lower than the maximum tolerable voltage in the short-circuit fault occurrence. To solve this issue, several approaches, such as active clamp mode [8], [24], or passive clamp mode [16], [23] can be used to limit the voltage of IGBTs in the short-circuit duration. Using this approach, in overvoltage conditions, the gate voltage of IGBTs will increase by the devised high voltage zener diode. Hence, its current will increase, and the voltage of the series IGBTs will be clamped [24]. The two main drawbacks of this method are as follows; (1) this approach has some delay, and (2) in the series structure, the short-circuit current of all IGBTs will increase to the short-circuit current of the IGBT which has the maximum short-circuit current.

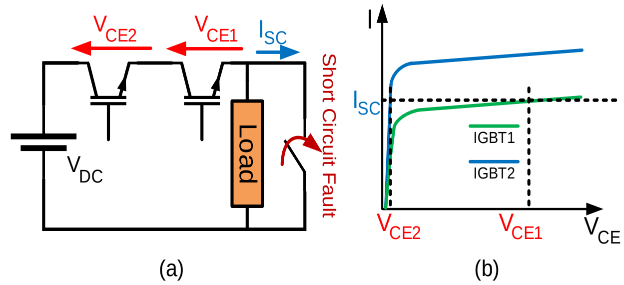

Fig. 1. Series connection of IGBTs a) Two series IGBTs in a circuit with short-circuit fault occurrence b) I-V characteristics of tow series connected IGBTs.

a clamp mode passive snubber for voltage limiting of IGBTs in the short-circuit duration is used in [23]. Furthermore, using a series resistor to limit the short-circuit current and decrease the differences in the current of the IGBTs is introduced. The critical limitation of this approach is the resistor used in the current path. For applications where the duty cycle is not low, the power loss of this resistor is not negligible. The differences of the IGBTs short-circuit current in the series structure with a clamp mode snubber pass through the corresponding snubber. If these differences are high, then the snubber voltage will increase considerably unless the clamp mode snubber capacitor has a large value. Therefore, the power loss of the series resistors in the emitter and using a large capacitor for limiting the voltage of IGBTs are the main drawbacks of this method.

The maximum tolerable short-circuit time is achieved when the voltage and the current of IGBTs are identical. Furthermore, the desired voltage-limiting technique should have a minimum number of components, minimum power losses and be simple to implement. The previously discussed voltage-limiting techniques are unable to achieve all the aforementioned goals simultaneously without significant compromises. In this paper, in order to minimize the differences of the short-circuit currents of the IGBTs, a feedback of the short-circuit current difference of each IGBT with the last IGBT before itself is provided. This feedback is used in the IGBT gate driver circuit to minimize the differences in the short-circuit current from the previous IGBT in respect to itself. Thus, in the proposed scheme, the current that will be inject into the clamp mode snubber will be minimized, and this results in reducing the snubber capacitor. Moreover, there is no need for a series resistor in the emitter in the proposed scheme giving a higher efficiency than the previous techniques.

This paper is organized as follows. In section II, the behavior of the IGBTs in short-circuit conditions and the parameters that affect the short-circuit current of IGBTs are explored. The proposed method for providing the safe condition for the series IGBTs in the short-circuit fault duration is presented in section III. The design procedure and the simulation of a case study with the proposed scheme are provided in section IV. Moreover, the experimental demonstrations are presented in Section V. Finally, the outcomes are summarized in Section VI.

II. SHORT-CIRCUIT CURRENT CHARACTERISTICS

Due to uncertainties in the manufacturing process of IGBTs and their gate driver circuit components, the short circuit-current of IGBTs with similar part-number might not be identical. Furthermore, the characteristics of an IGBT changes due to aging, and for two IGBTs with the same aging process, these changes may be different. In the series structure of IGBTs, the current that passes through the IGBTs are equal, but the turn-off collector-emitter voltage $V_{CE}(off)$, saturation voltage $V_{CE}(on)$, and turn on/off times of IGBTs are different. Thus, the switching loss $P_{Loss-Switching}$ and the conduction loss $P_{Loss-Cunduction}$ will be different. On the other hand, the thermal resistance R_{Th} of the IGBTs and their heat sinks are not identical. Therefore, the junction temperature T_J of the IGBTs will be even. All of these parameters in addition to the differences in the voltage amplitude of the gate driver and the IGBT transconductance, affect the IGBT short-circuit current. A flow diagram of the parameters that affect the IGBT short circuit current is presented in Fig. 2

In the short circuit duration, the linear equation of IGBT short circuit current is as [23].

$$I_{sc} = g_{fs}.\left(V_{ge} - V_{th}\right) \tag{1}$$

In this equation I_{sc} is the IGBT short-circuit current, V_{ge} is the gate-emitter voltage, V_{th} is the linearized gate-emitter threshold voltage, and g_{fs} is the IGBT transconductance. As shown in Fig. 2, g_{fs} and V_{th} are dependent on the IGBT junction temperature. Moreover, the V_{ge}, g_{fs} and V_{th} for each IGBT and its driver are different. These differences are caused by intrinsic uncertainties in manufacturing and physical material deformation due to aging during operation. The IGBT used in this work is IKW40N120H3. In Table I, for each parameter of this IGBT, two coefficients are considered with the normal probability distribution function modeling variation due to aging and intrinsic uncertainties. The mean values of the coefficients are equal to one, and the standard deviation for aging and intrinsic uncertainties are equal to σ_A and σ_I respectively, as summarized in Table I.

Based on the parameters in Table 1 and their impact on the IGBTs short-circuit current (see Fig. 2), the probability density function of the IGBT short-circuit current can be obtained. Using the Monte Carlo simulation, the probability distribution of the short-circuit current of an IGBT will be as depicted in Fig. 3.

The mean value μ, and the standard deviation σ of short-circuit current amplitude are 85.5 A and 8.25 A respectively. For a parameter that has a normal distribution, 95.45% of the data are between μ-2σ and μ+2σ. In other word, it can be assumed that a parameter just varies between μ-2σ and μ+2σ with the confidence of 95.45%. As shown in Fig. 3, the short-circuit current of each IGBT has a value between 109 A and 142 A. The junction temperatures of IGBTs obtained from the simulation are between 85 ̊C and 111 ̊C.

III. PROPOSED VOLTAGE BALANCING SCHEME FOR SERIES IGBTS IN SHORT-CIRCUIT CONDITION

The amplitude of the short circuit current in the series IGBTs must be lower than its maximum value. Meanwhile, as shown in Fig. 1, the discrepancies in the short-circuit currents of the IGBTs cause unbalanced voltage sharing among the IGBTs. Thus, the IGBTs with a lower short-circuit current will experience higher voltage for the same short-circuit current.

In order to limit the voltage of the IGBTs, in short-circuit periods, clamp-mode snubbers can be employed. Limiting voltages of IGBTs causes differences in the short-circuit current of the IGBTs, and these differences inject into the snubbers. Therefore, the clamp-mode snubbers should tolerate these currents, or the short-circuit currents of the IGBTs should be controlled to minimize their differences.

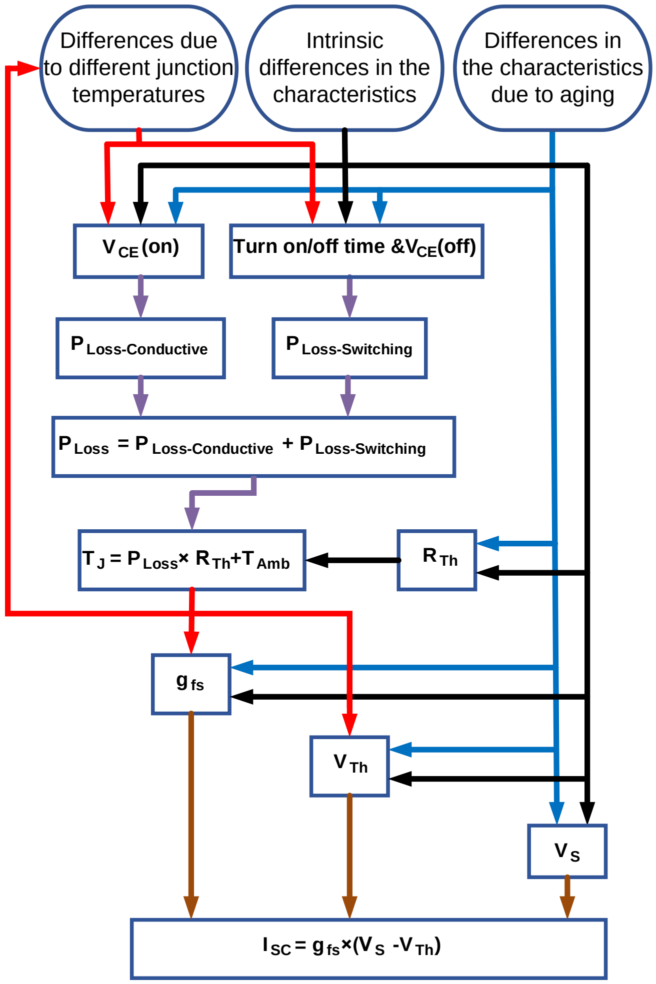

Fig. 2. Flow diagram of the parameters that affect the IGBT short-circuit current.

Table I: IGBT parameters and their standard deviations.

Parameter	Nominal value	σ_I	σ_A
R_{Th} (Thermal resistance of IGBT)	0.75 K/W	0.03	0.05
E_{Sw} (Turn on & off switching energy)	4 mJ	0.03	0.05
V_{CE}(off) (Turn off voltage of IGBTs)	٦00	0.03	0.02
V_{CE}(on) (Turn on saturation voltage)	2.5 V	0.05	0.05
V_{th} (The threshold voltage of IGBT obtained from the linearized curve)	8 V	0.03	0.02
V_s (gate driver voltage)	13 V	0.02	0.01
g_{fs} (IGBT transconductance)	20 S	0.03	0.05
I (Normal operation current)	25 A	0	0

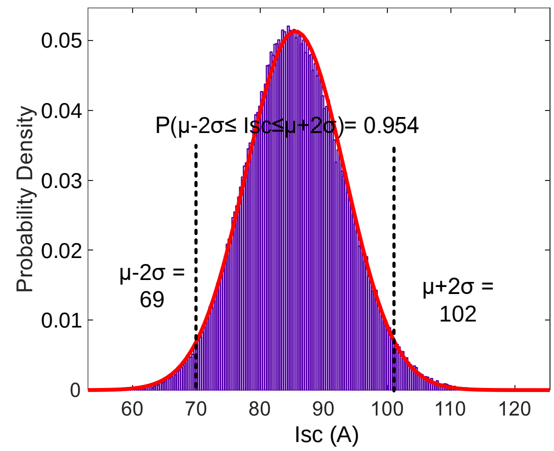

Fig. 3. Probability density function of the IGBT short-circuit current.

In [23], in addition to the clamp mode snubber, a series resistor is used in the emitters of the IGBTs to control the short circuit current of the IGBTs. For the proper performance of this method, the value of this resistor should be high. However, in high current applications, due to the power loss of this resistor, it cannot be high enough, and thus, the short circuit currents of IGBTs may not be balanced. Also, this resistor cannot reduce the effect of differences in the gate-emitter threshold voltage and gate driver voltage on the short-circuit current. In this paper, instead of controlling the short-circuit current of each IGBT to have a determined short-circuit current, their short-circuit current mismatch is controlled to have a minimum value. Therefore, the current that injects into the snubber will have a low value. The snubber capacitance, that limits the voltage, will be reduced. The schematics of the proposed scheme is shown in Fig. 4 with the sampling resistor of R_s connected to the snubbers. Therefore, in the normal operation of the switch, the power loss over this resistor is not remarkable.

In the circuit shown in Fig. 4, the first IGBT has a short-circuit current that is determined by its own characteristics and the voltage amplitude of the related gate driver circuit. The other IGBTs follow the short-circuit current of the last IGBT in respect to itself. The following relations can be written for the i^{th} IGBT.

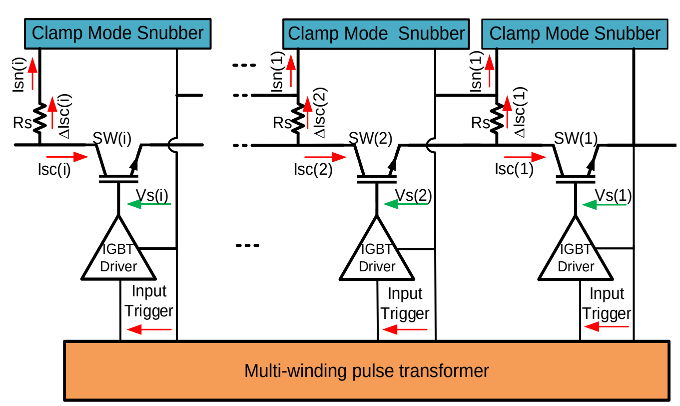

Fig. 4. Proposed scheme for voltage balancing of IGBTs in short circuit conditions, including clamp mode snubbers.

$$V_{ge}(i) = V_s(i) - R_s.\Delta I_{sc}(i-1) \tag{2}$$

$$I_{sc}(i) = g_{fs}(i).(V_{ge}(i) - V_{th}(i)) \tag{3}$$

$$\Delta I_{sc}(i-1) = I_{sc}(i) - I_{sc}(i-1) \tag{4}$$

$$I_{sc}(i) = \frac{g_{fs}(i).(V_s(i) - V_{th}(i) + R_s.I_{sc}(i-1))}{1 + g_{fs}(i).R_s} \tag{5}$$

In these equations, V_{ge} is the gate-emitter voltage, I_{sc} is the short-circuit current, V_s is the gate-driver voltage, and R_s is the sampling resistor. The relationship between the short-circuit current of the IGBT and sampling resistor (R_s) is given (5). Considering:

$$\begin{cases} R_s.I_{sc}(i-1) \gg V_s(i) - V_{th}(i) \\ g_{fs}(i).R_s \gg 1 \end{cases} \tag{6}$$

then,

$$I_{sc}(i) \simeq I_{sc}(i-1) \tag{7}$$

which means that the differences of $I_{sc}(i)$ and $I_{sc}(i-1)$ will be minimized. R_s can have a large value to satisfy the conditions in (6). The main limitation of selecting this resistor with a very large value is its effect on the switching time of the IGBTs. This resistor is in series with the gate driver circuit. Thus, if it has a large value, it will affect the switching times. In the gate driver circuit, usually, there is a resistor; adding this resistor (which is about a few ohms) or replacing it with the gate driver resistor has no significant impact on the switching times. When the proposed method is used, the maximum injected current to the snubbers (I_{sn-Max}) is equal to (8). In (8), $I_{sn}(i)$ is the injected current to the snubber of the i$^{\text{th}}$ IGBT.

$$\begin{aligned} I_{sn-Max} &= Max(I_{sn}(1), I_{sn}(2), \ldots, I_{sn}(n)) \\ &= Max(I_{sc}(1), I_{sc}(2), \ldots, I_{sc}(n)) \\ &\quad -Min(I_{sc}(1), I_{sc}(2), \ldots, I_{sc}(n)) \end{aligned} \tag{8}$$

In order to investigate the effectiveness of the proposed method, a case study with 4 IGBTs is assumed. The IGBTs used in this study are IKW40N120H3. The short-circuit current for a single IGBT considering the probabilistic values for its characteristics is investigated. The maximum injected current to the IGBTs snubbers can be obtained when $R_s = 5\Omega$ as shown in Fig. 5(a) for the conventional method and in Fig. 5(b) for the proposed method. For obtaining the curves in Fig. 5, it is assumed that the effect of voltage changes of the clamp mode snubbers in the short-circuit duration is negligible. As shown in Fig. 5, if it is assumed that the maximum of maximum injected current to the snubber is $\mu(I_{sn-Max}) + 2\sigma(I_{sn-Max})$, the maximum of maximum injected current to the snubbers in the conventional scheme is 30.3 A, while this value in the proposed scheme is 0.4 A. These results show that the current injected into the snubber using the proposed scheme is considerably lower than the conventional structure. Since IGBTs follow the short-circuit current of the first IGBT with a negligible difference in the short-circuit interval, if the short-circuit current of the first IGBT has been controlled, then the short-circuit current of the equivalent series switch is also controlled. The effect of the R_s value is further explored for this case study and is shown in Fig. 6. The graph in Fig. 6 indicates that the maximum short circuit current mismatch of IGBTs decreases by increasing the value of the sampling resistor.

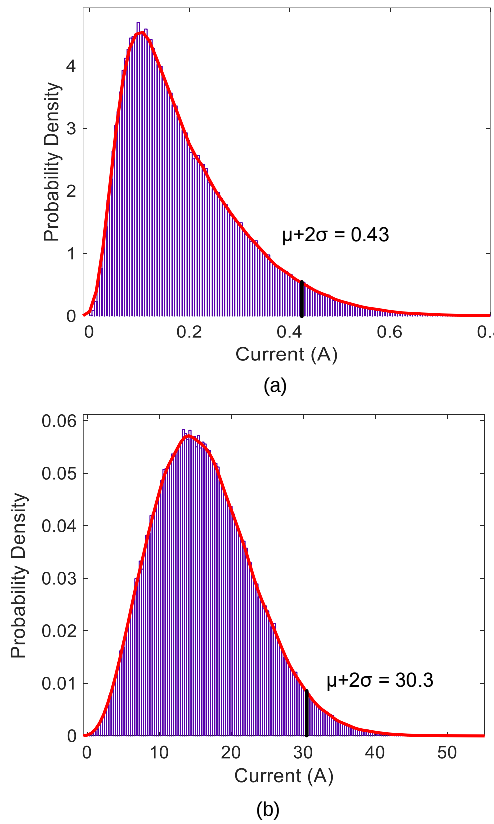

Fig. 5. Distribution of the maximum injected current to the IGBTs snubbers a) the proposed scheme with $R_s = 5\Omega$ b) the conventional scheme.

A clamp mode snubber can be modeled as an ideal zener diode (or transient voltage suppressor) [25]. This zener diode has an unwanted series resistance in the real condition, as shown in Fig. 7. Thus, using a parallel capacitance is suggested in the literature to limit the voltage changes of the snubber in transients. Therefore, after the transient, the extra electrical charge of the capacitor will be discharged in the zener diode.

Using a zener diode just limits the voltage of the clamp mode snubber, and it does not balance the voltage of the series IGBTs. In [25], using a parallel resistor R_p to improve the voltage balance of the clamp mode snubbers is presented. The minimum value of this capacitor for voltage limiting or voltage balancing of the series IGBTs in the normal operating condition is discussed in [13] and [18]. Parasitic capacitance, nonsynchronous operation of IGBTs, and different tail current of IGBTs are the main causes of injecting current into the clamp mode snubber in normal operating conditions [14], [25].

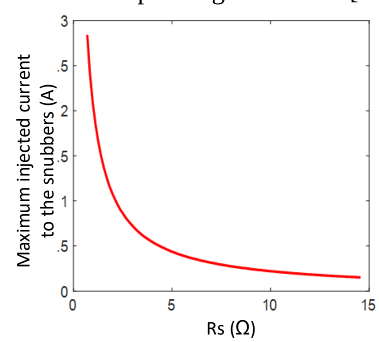

Fig. 6. Maximum injected current in the snubbers as a function of R_s.

The maximum electrical charge that injects into the snubber capacitor in the short circuit duration (t_{sc}) is obtained by using (9).

$$Q_{sc-Max} = I_{sn-Max} \times t_{sc} \tag{9}$$

Therefore, considering a short-circuit fault occurrence, the value of the injected electrical charge during the short-circuit interval and the electrical charges that inject to the snubber due to the nonsynchronous operation of IGBTs and the parasitic capacitances should be considered to calculate the snubber capacitor. When the maximum electrical charge Q_{sc-Max} that injects to the snubber capacitor is divided by the maximum acceptable voltage changes ΔV_{sc-Max} , the snubber capacitor can be defined as (10).

$$C = \frac{Q_{sc-Max}}{\Delta V_{sc-Max}} \tag{10}$$

The switching time of an IGBT is a function of the equivalent resistance in the gate driver circuit. Thus, the equivalent resistance of the gate driver circuit has a specific value for a determined switching time. In the proposed scheme, R_s is in series with the gate driver circuit in the normal operation condition. In this study, the value of R_s is selected to be 5Ω based on proper switching time and internal gate resistances of IGBTs gate driver ICs. As shown in Fig. 5 $\mu(I_{sn-Max})+2\sigma(I_{sn-Max})$ for $R_s = 5\,\Omega$ is equal to 0.43 A. Thus, with the assumptions in Table III, and assuming that the maximum injected current to the snubber is 0.43 A, a proper value for the snubber capacitor is 30 nF. It should be considered that in the conventional scheme, the maximum injected current to the snubber is 30.3, and a 2500 nF capacitor is needed for voltage limiting. This result shows that the calculated snubber in the proposed is about 1/80 of the conventional scheme.

IV. SIMULATION OF PROPOSED SCHEME

A prototype consisting of four series-connected IGBTs using the proposed scheme is simulated by use of PSPICE software.

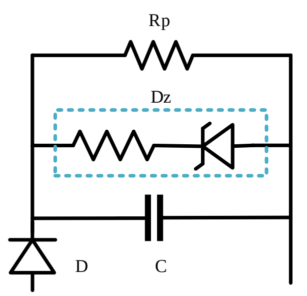

Fig. 7. Schematic of a clamp mode snubber.

Table II. Assumptions for calculating the value of the snubber capacitor

Parameter	Value
R_s	5 Ω
Maximum injected current to the snubber	0.43A
Maximum short circuit current	110 A
Maximum voltage changes in the short circuit duration	60 V
t_{sc} (short circuit time)	5 μs

These IGBTs and their gate drivers are respectively IKW40N120H3 and TC4427. The important specification of the IGBT, driver, and other assumptions for the simulations are summarized in Table III. In this study, different junction temperatures and voltages for the gate drivers are assumed.

The voltages of the IGBTs are presented in Fig. 8. As shown in Fig. 8, the maximum voltage change of IGBTs is 70 V. Therefore, the voltages of the IGBTs remains in their safe operation region. The short circuit current of the series-connected IGBTs is 113 A that is equal to the short circuit current of the first IGBT. Notably, a slight change of IGBTs voltage in the turn-off transient is due to the nonsynchronous turn-off of IGBTs that come from the differences of gate drivers voltages and the discrepancies of IGBTs junction temperatures. Fig. 9 illustrates the injected currents to the snubbers. These currents are the difference of short-circuit current of the IGBTs that inject into the snubber capacitor and lead to voltage increment of the capacitor. A comparison of the injected current to the clamp mode snubbers using the conventional scheme is illustrated in Fig. 10. As shown in Fig. 10, the maximum of the current differences of the IGBTs that are injected into the snubber is 42 A. It is noted that for the simulation of the conventional scheme, the snubber capacitor is 2.5 μF. Compared to the results shown in Fig. 9 and Fig. 10, the injected current to the snubber using the proposed scheme is about 0.01 of the conventional one.

Table III. The assumptions for simulation of the series IGBTs

Parameter	Value
Junction temperature	105±15̊C
Gate driver voltage	13±1 V
Dc link voltage	2400 V
R_s	5 Ω
Snubber capacitor for the proposed scheme	30 nF
Nominal current	40 A
Gate driver resistance	10 Ω
Snubber capacitor for conventional scheme	2.5 μF
Short circuit time	5 μs

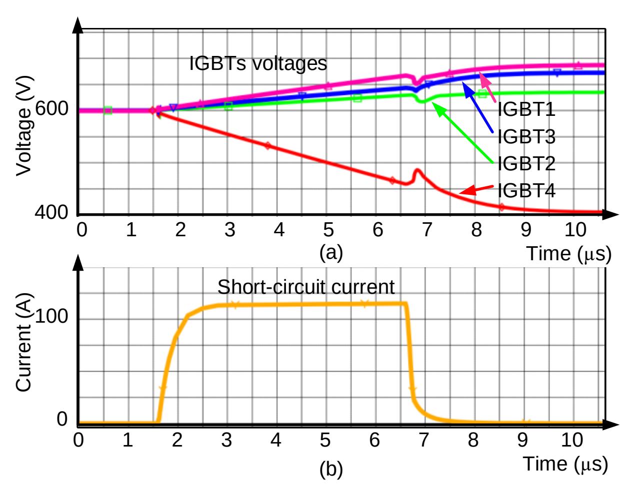

Fig. 8. Collector-emitter voltage of IGBTs in short circuit duration using the proposed scheme.

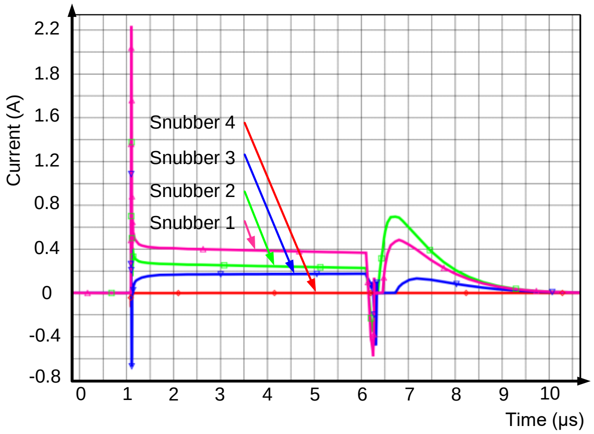

Fig. 9. Injected current in the snubbers of IGBTs in the short-circuit duration using the proposed scheme.

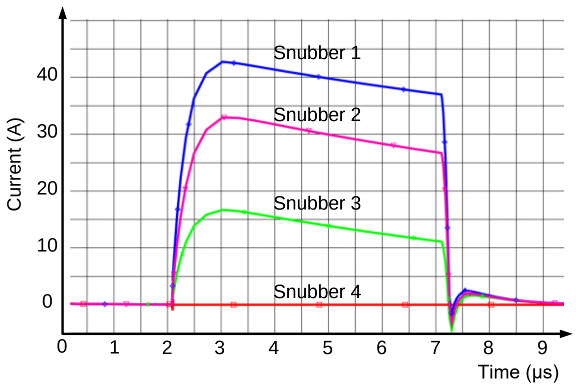

Fig. 10. Injected current in the snubber using the conventional scheme.

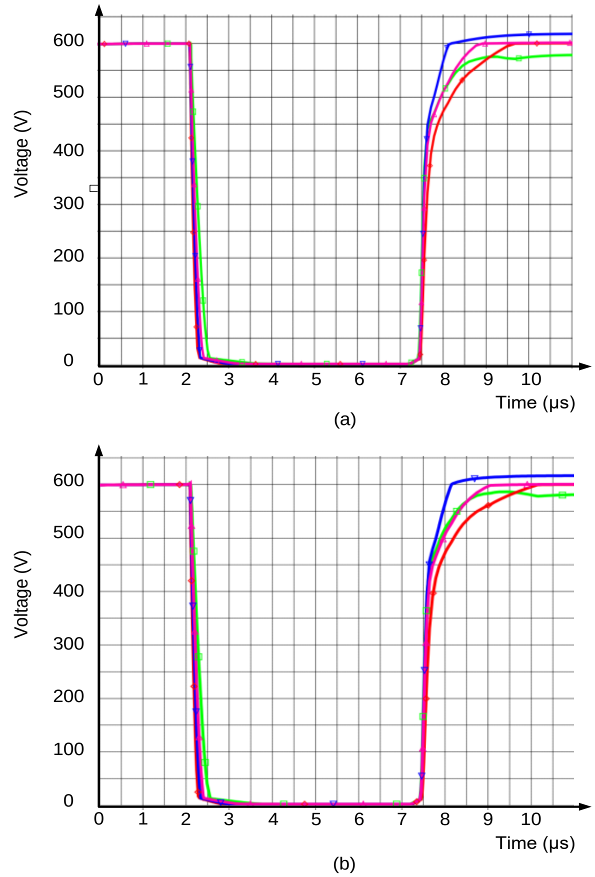

Fig. 11. The collector-emitter voltage of IGBTs in normal operation condition a) without Rs and b) with Rs.

The effect of adding the resistor to the IGBTs gate drivers and snubber circuits on switching times of IGBTs is shown in Fig. 11(a) for normal operating conditions. Also, for comparison, the voltage waveforms of the IGBTs without this resistor are illustrated in Fig. 11(b). As shown in Fig. 11, the voltage waveforms of IGBTs after adding this resistor have negligible changes in normal operating conditions.

V. EXPERIMENTAL RESULTS

In order to verify the proposed scheme, a prototype consisting of four series IGBTs has been constructed and tested. The block diagram of the test setup is presented in Fig. 12, and a photograph of the constructed series IGBTs board is presented in Fig. 13. The parameters of the constructed board are the same as the parameters used in the simulations. Although the IGBTs and their characteristics are different in this test, different gate driver voltages are used to make more differences in the short circuit currents of the IGBTs. The gate driver voltages are presented in Fig. 14. First, each IGBT block is tested separately with a 500 V collector-emitter voltage, and the short-circuit current is measured using a current probe (T3RC0300 Rogowski coil current probe). In Fig. 15, the short-circuit currents of each of the IGBTs are presented. As shown in this figure, the maximum difference in the IGBTs short-circuit current is equal to 40 A.

The series-connected IGBTs are tested with a 2000 V power supply. As shown in Fig. 15 the short-circuit current of the first IGBT is 132 A, and the other IGBTs follow this current. Therefore, their currents are almost equal. Fig. 16 presents the short-circuit current of the IGBTs by using the proposed scheme. As shown in Fig. 16 the differences in the short-circuit currents of the IGBTs decreased considerably, and the short-circuit currents of all the IGBTs are almost 132 A. This means that the first IGBT controls the short-circuit currents of the rest of the IGBTs.

Since differences in the short-circuit currents of IGBTs inject into the snubber, therefore, balancing the short-circuit currents of the IGBTs will reduce the injected currents to the snubbers; therefore, the voltage changes of the snubber will be limited.

In order to measure collector-emitter voltages, the voltage of each node to the ground is recorded, and the collector-emitter voltage of each IGBT is obtained by subtracting the recorded voltages of the IGBT block nodes. The collector-emitter voltage of the IGBTs during short-circuit time is shown in Fig. 17. As shown in Fig. 17, the maximum voltage increment of the IGBTs during the short-circuit interval is about 70 V, which is consistent with the analyses and simulations. Therefore, using the proposed scheme, the collector-emitter voltages of IGBTs are limited to 570 V, which is lower than the permitted short-circuit voltage presented in the datasheet of the used IGBTs.

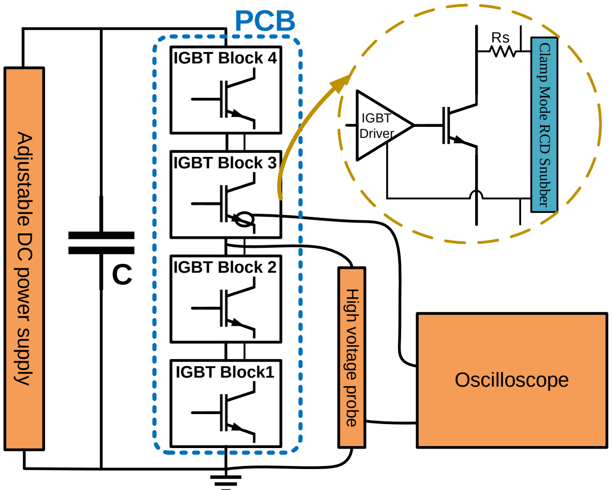

Fig. 12. Block diagram of the test setup.

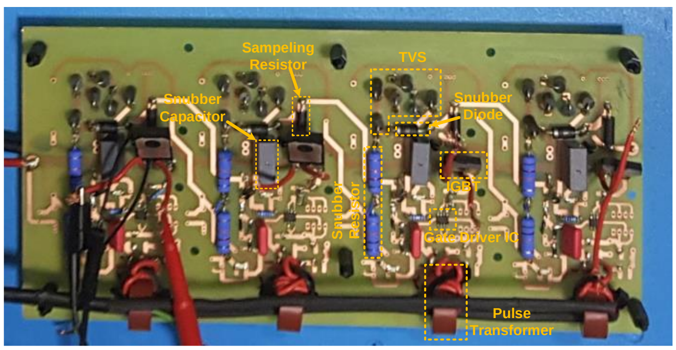

Fig. 13. Photograph of the constructed series IGBTs board.

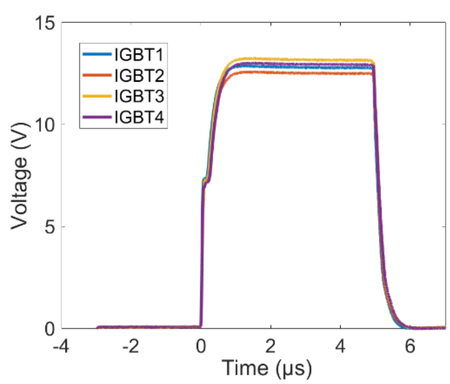

Fig. 14. Output voltages of the gate drivers.

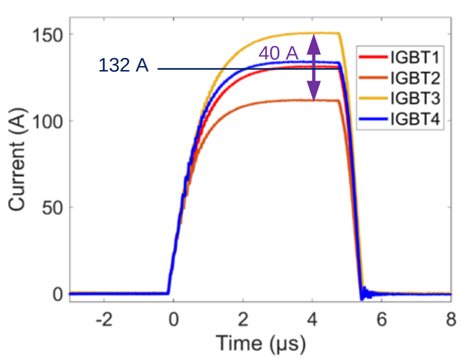

Fig. 15. Short-circuit current of the individual IGBTs ($V_{CE} = 500V$).

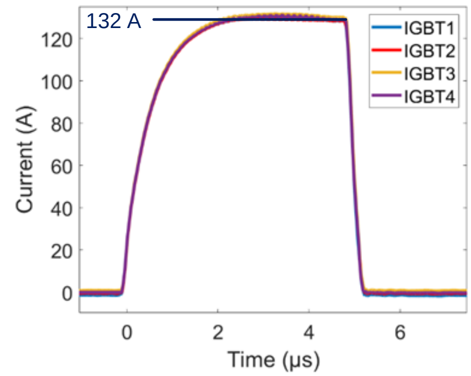

Fig. 16. Short-circuit currents of the IGBTs in the series structure using the proposed scheme.

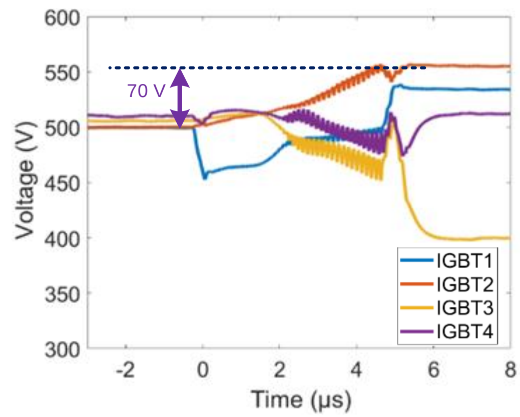

Fig. 17. Collector-emitter voltage of series IGBTs using the proposed scheme in short-circuit condition.

VI. CONCLUSION

In this paper, the short-circuit fault control and behavior in the series configuration of IGBTs is discussed. The impact of uncertainties of IGBTs parameters, due to aging and intrinsic differences, on the short-circuit characteristics of IGBTs is explored. The drawbacks of the conventional schemes for voltage limiting of IGBTs in the series configuration are explained, and a new scheme for voltage balancing of IGBTs in the short-circuit fault condition is proposed. Employing the proposed scheme achieves proper voltage sharing on IGBTs, by equalizing the short-circuit current of the IGBTs and using a clamp mode snubber. Furthermore, unlike the conventional approaches, there is no resistance in the current path of the switch in the proposed scheme, thereby implying a higher efficiency. The injected current into the snubber in the short-circuit fault duration is significantly smaller than the conventional scheme. Moreover, the clamp mode snubber components are considerably smaller than the conventional structure, which reduces the cost and size of the converter. The simulation and experimental results validate the effectiveness of the proposed scheme on the voltage and current balancing in short-circuit faults without disrupting the switching behavior of IGBTs.

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