# VOLTAGE SHARING SCHEME FOR SERIES-CONNECTED POWER SEMICONDUCTORS

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## Abstract

Dynamic voltage-sharing schemes have been investigated which allow high-voltage power-semiconductor devices, such as thyristors, IGCTs, IGBTs or power MOSFETs, to be series connected in strings and switched, as simply in high-voltage applications, as when used as single devices. The circuits have many of the advantages of simply using RC or RCD snubbers, including being easily applicable to both low- and high-side switches. However, because the snubber capacitors are not fully discharged their associated reset current and power-losses are minimized. To illustrate the principle of operation experimentally, a string of three series-connected power MOSFETs switching 100 A from 330 V has been used to obtain practical waveforms. The schemes are discussed and illustrated, using SPICE simulation results. The new, relatively simple voltage-sharing schemes are much easier to design and optimize than recently reported active gate-control and regenerative-snubber methods, allow very rapid turn-on and turn-off switching, and give composite-device switches a usable voltage rating similar to the aggregated voltage ratings of the string.

## 1 Introduction

A novel regenerative snubber scheme, cited in a more complex form in [1] but otherwise rarely discussed, has been developed for controlling the switching voltage transients of individual devices to provide dynamic voltage sharing within strings of series-connected power-semiconductors. It is intended for application in pulsed-power, high-voltage chopper, or inverter bridge-legs, in which series-connected devices are used as single switches. The circuit retains many of the advantages of simply using RC or RCD snubbers across each device in a string, and does not have the complexity, the requirement to modify gate drives, slower switching or other disadvantages of recently investigated active voltage-balancing methods.

With the new method, snubber capacitors are not completely discharged during operation, their associated reset current surge and power-losses are considerably reduced. Like RCD snubbers, the proposed scheme may be used with strings of any number of thyristor or transistor power device operated from a high-voltage DC supply and may be designed to accommodate significant differences in switching characteristics. Snubber circuit development is discussed and illustrated using simulation and experimental results. Experimental operation is investigated at 100 A using series-connected power MOSFETs operating from a 330 V supply.

## 2 Previous methods

Previously reported experimental investigations have shown that modern power-semiconductor devices, such as IGCTs, IGBTs and power MOSFETs, may be series connected and operated synchronously as single switches in high-voltage chopper, inverter and pulsed-power applications [2-7], and series-connected devices are now being used in IGCT and IGBT applications [9] What makes possible direct series operation is the use of an effective voltage balancing scheme, which ensures that the composite switch voltage drop is evenly distributed between the devices in a string during blocking (static voltage balancing) and during switching (dynamic voltage balancing). Without enforced voltage balancing, repeated device breakdown within strings would almost certainly occur because of the variability in off-state leakage current and switching characteristics which arise in practical circuits. Breakdown would arise not only due to production spread in device characteristics, but also due to imperfect synchronisation of isolated drive signals, imbalance in common-mode voltage effects, and imperfect matching of the electrical and thermal impedances of device packages and other related hardware [3,4]. By using an effective voltage balancing scheme, composite-device switches have been shown to have a usable voltage rating comparable to the aggregated voltage ratings of the string.

Voltage balancing is most easily provided by connecting voltage sharing resistors and RC snubbers across each device in a string as successfully applied for many decades in the thyristor strings that constitute the rectifier valves of HVDC-power-transmission converter stations. However, in high-voltage inverter, chopper, and pulsed power applications, it is usually important that devices be switched more rapidly, for example, to implement PWM control at carrier frequencies closer to 1 kHz than line-frequency, and to provide effective electronic over-current protection. With faster higher-frequency switching, the relatively high power-loss and reset-current transient of snubbers have a greater impact on converter efficiency and device utilisation than in thyristor circuits. In recent work on series device operation, a number of efficient innovative voltage-balancing schemes have been developed. Although, no single method outperforms all others in ease of implementation and minimising device switching-loss and –stress, there is some convergence towards proposing using active voltage balancing with IGBTs [4-7] and using refined passive voltage balancing with IGCTs [2, 3]. The active voltage balancing methods involve applying collector voltage feedback to the gate drive of each device, and thus prolonging or retriggering partial reconduction of individual devices to limit voltage drop during blocking, or to clamp voltage overshoot during the final stages of switching. Refined passive voltage balancing methods, developed for IGCTs, generally use RCD snubbers in which the capacitor is fully discharged regeneratively, or discharged to approximately $V_{DC}$ $/N$ where $V_{DC}$ is the converter supply voltage and $N$ the device-string length [2, 3].

An important advantage of these schemes is their improved capability to provide voltage equalisation or clamping only when required, and to thus avoid the continuous power loss of conventional snubbers which are designed and operated as if worst-case conditions of device mismatch and high load-current are continually prevalent. Alternative voltage balancing schemes do, however, lack the ease of application and scalability of simple passive methods. Active voltage balancing, in particular, requires modification of gate-drive circuits to enable devices to respond to anode-cathode voltage feedback; and a greater optimisation and validation effort seems unavoidable if consistent device protection is to be obtained despite variation in device characteristics and operating conditions, perhaps even after the future replacement of a failed device. It seems possible to overcome the disadvantages of active snubbing and still achieve efficient protection of any power semiconductor device using essentially passive voltage-balancing schemes. The new method proposed, essentially, comprises placing biased RCD-snubbers across each device in a series string. Snubber reset losses are minimal when snubber capacitors are biased at $V_{DC}$ $/N$; however, the method allows the bias level to be adjusted to any value between 0V and $V_{DC}$. The novelty of the method lies in the cellular nature of the snubber-capacitor discharge circuit which may be easily expanded to protect relatively long device strings. The proposed form of protection allows series-connected devices to be switched as rapidly as in conventional bridge-legs or single-ended choppers, without significantly increasing switching stress. The development of the scheme is first discussed and illustrated using circuit simulation results. The results of an experimental evaluation of the method are then presented.

## 3 Biased RCD snubber

A single-ended chopper comprising transistor (or gate-turn-off thyristor) $Q$ and diode $D_{FW}$ may be connected to a DC supply and used as a switching regulator to control current in an inductive load, as shown in Figure 1. However, even with careful layout, the parasitic inductance in loop $C_S$-$D_{FW}$-$Q$, the effect of which may be represented by $L_S$ if $D_{FW}$ is very close to $Q$, produces overshoot and high-frequency ringing in the transistor turn-off voltage waveform. A number of methods may be used to clamp the overshoot and damp the resonance between $L_S$ and the parasitic output capacitance of $Q$ [8]. One with the lowest loss is a biased RCD snubber, comprising $D_C$, $C_C$ and $R_R$ in Figure 1. In this, the reset resistor is connected across $D_{FW}$ rather than $D_C$, so that the capacitor remains charged at approximately $V_{DC}$ during transistor conduction. To be effective, $D_C$ and $C_C$ must be placed very close to $Q$ to provide a low inductance path to which transistor current can commutate to at turn off. $L_S$ can then reset more slowly by resonating $C_C$ above $V_{DC}$. $R_R$ then discharges $C_C$ back to $V_{DC}$. Circuit operation may be understood from the simulated turn-off switching waveforms shown in Figure 2. Between $t_1$ and $t_2$, transistor voltage, $v_Q$, rises to $V_{DC}$ and $D_C$ becomes forward biased. Transistor current, $i_Q$, commutates to $D_C$ and $C_C$, $i_{DC}$. During $t_3$ and $t_4$, $L_S$ resets resonantly into $C_C$, increasing the capacitor voltage to a peak of $\Delta V$ above $V_{DC}$. As $L_S$ is reset and its current $i_{LS}$ falls to zero, current rises in the freewheel-diode at the same rate since $i_{LS}$ and $i_{DFW}$ sum to $I_O$ at any instant. Because the bulk of the parasitic inductance, $L_S$, typically lies between $C_S$ and $D_{FW}$-$Q$, the transfer of energy from $L_S$ to $C_C$ is completed before $i_{RR}$ increases from zero and $R_R$ discharges $C_C$ to $V_{DC}$. In practice, a voltage transient occurs in $v_Q$ at $t_4$ due to the reverse-recovery of $D_C$, and is relatively well damped if $R_R$ is a relatively low-inductance low-value resistor.

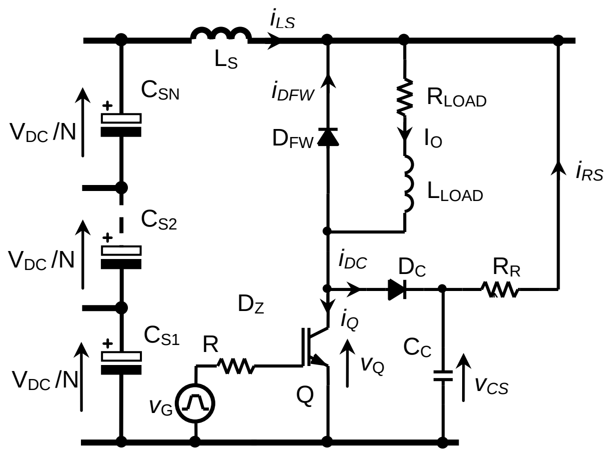


Figure 1: Chopper with biased RCD snubber.

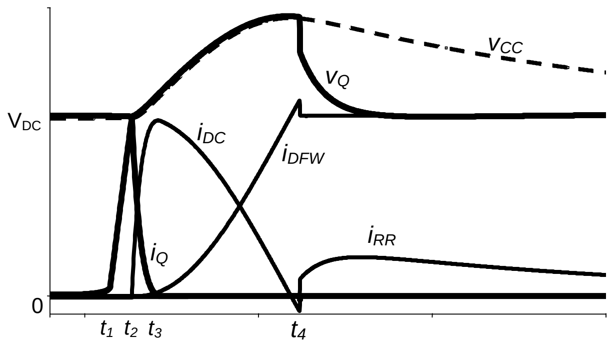


Figure 2: Simulated turn-off switching waveforms.

$$\frac{1}{2}L_S I_{O,MAX}{}^2 = \frac{1}{2}C_C(\Delta V)^2 = W \quad (1)$$

If idealised device switching and clamp diode operation are assumed, such that all the energy trapped in $L_S$ is transferred to $C_C$ and then all dissipated in $R_R$ at $Q$ turn-off, then Equation 1 may be used to estimate worst-case overshoot, $\Delta V$, and resistor dissipation, $W$. Additional energy is dissipated for a brief period in $R_R$ when $Q$ is switched on; however, snubber efficiency may be compared using turn-off loss alone. Energies associated with resetting $L_S$ using a conventional, fully discharging RCD snubber, and using active voltage clamping whereby the transistor is controlled to clamp at $V_{DC}$ + $\Delta V$, as if having repetitive avalanche capability, may be approximated in terms of the original trapped energy as in Table 1 [8]. From the normalised values, $W_N$, determined by assuming constant-current operation and 25% voltage overshoot, it is apparent that the biased RCD snubber is potentially a very efficient method of controlling the reset of unclamped series switch inductance. It should be noted that, in the case of the conventional snubber, the $V_{DC}/\Delta V$ term equates to ½ $C_C$ $V_{DC}{}^2$, which unlike the others does not reduce with current and results in an even lower efficiency [8].

| | Biased RCD | Conventional RCD Snubber | Active Voltage Clamp |
|---|---|---|---|
| $W$ | $\frac{1}{2}L_S I_O{}^2$ | $\left[1+\left(\frac{V_{DC}}{\Delta V}\right)^2\right]\frac{1}{2}L_S I_O{}^2$ | $\left[1+\frac{V_{DC}}{\Delta V}\right]\frac{1}{2}L_S I_O{}^2$ |
| $W_N$ | 1 | 17 | 15 |

Table 1: Normalised energy loss for $\Delta V$ at 25% of $V_{DC}$

# 4 Series-connected RCD snubbers

### 4.1 Basic requirements

With series connected devices, equal voltage clamping and thus dynamic voltage sharing is dependent upon the capacitors being discharged to approximately equal voltage, $V_{DC}/N$ (or as required), prior to synchronised transistor switching. Figure 3 shows, in principle, what is required. However, the circuit is impractical because of the need for multiple, matched, mostly floating, discharge sources. A method of resetting snubber capacitors into one voltage rail is required to make the snubber scheme practical.

### 4.2 Practical scheme of biased series snubbers

Two practical cellular snubber schemes and simulated operating waveforms are shown in Figures 4 to 7, in which $C_1$ to $C_N$ are effectively discharged in parallel to the same source, which may be set to $V_{DC}/N$ or a higher or lower value. In the Figure 4 scheme, auxiliary switchable devices are required (represented with thyristor symbols), which have the same forward voltage-blocking capability as $Q_1$ to $Q_N$, but the reset-source voltage polarity required is likely to be available as part of the high-voltage DC supply system.

The need for auxiliary switches is avoided in the Figure 6 self-commutating scheme; only auxiliary reset diodes are required. However, a purpose-built auxiliary reset source is required because of its polarity. In both schemes, the reset of snubber capacitors is performed when the composite switch turns on.

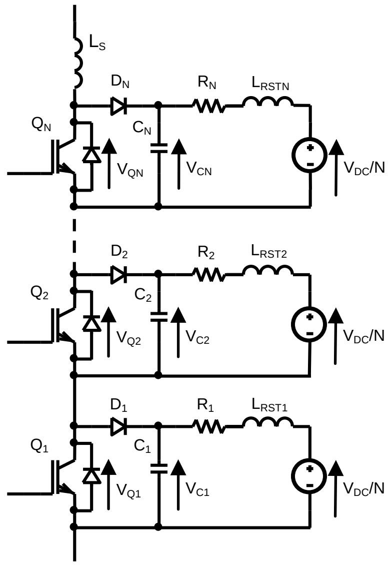


Figure 3: Impractical RCD scheme for series devices.

Simulated turn-off waveforms are shown in Figures 5 and 7, for a three-device composite switch and freewheel diode, applied in a single-ended chopper, such as Figure 1, and protected by Figures 4 and 6 snubber schemes. The chopper is assumed to be delivering a constant current to an inductive load. A relatively low DC supply voltage of 300 V is used and allows voltage overshoot features to be clearly seen. The waveforms are for synchronous, balanced operation of $Q_1$ to $Q_3$. At turn off, just after 0µs, individual transistor voltages, $v_{Q1}$ to $v_{Q3}$ rise above $V_{DC}/3$ and transistor currents commutate to $D_{C1}$ to $D_{C3}$, and $L_S$ resonates up the series-connected snubber capacitors, $C_1$ to $C_3$, which were previously discharged to $V_{DC}/3$. As the reset of $L_S$ completes $i_{LS}$ falls to zero and $v_{C1}$ to $v_{C3}$ reach a peak overshoot value of $\Delta V/3$. Although the snubber capacitors remain charged above $V_{DC}/3$ until the transistors are next turned on, the transistor voltages fall back to $V_{DC}/3$ due to discharge of junction capacitance and clamp-diode reverse-recovery into the DC supply and static voltage sharing resistors which are used but omitted from the circuit for clarity.

When $Q_1$ to $Q_3$ turn back on at 108µs, $v_{Q1}$ to $v_{Q3}$ fall, and load current commutates to them from the composite freewheel diode at a rate governed by $L_S$. Freewheel diode reverse-recovery is allowed to complete before $S_1$ to $S_3$ are switched (or $D_{R1}$ to $D_{R3}$ forward biased in Figure6) to discharge $C_1$ to $C_3$ into $V_{RST}$, and back to $V_{DC}/3$, in readiness for the next composite-switch turn off. $L_{RST}$ is added in series with the discharge path to control the rate-of-rise and peak value of reset current in $S_1$ to $S_3$. The snubber reset currents in $R_{R1}$ to

$R_{R3}$ are all approximately equal. However, reset currents accumulate as they cascade down through the reset switches, as evident from the reset current waveforms in Figures 5 and 7. The duration of the reset period may be reduced at the expense of increased peak current by reducing $R_{R1}$ to $R_{R3}$ and $L_{RST}$ values.

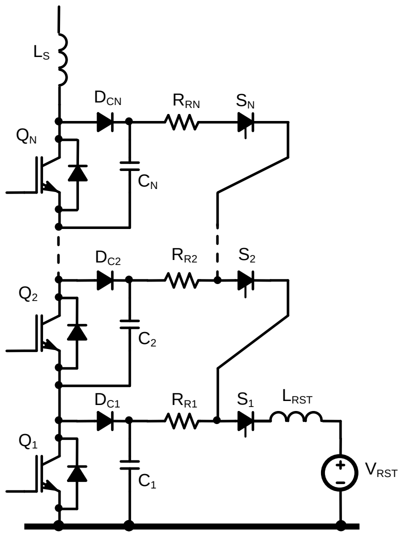


Figure 4: Series RCD scheme with reset switches.

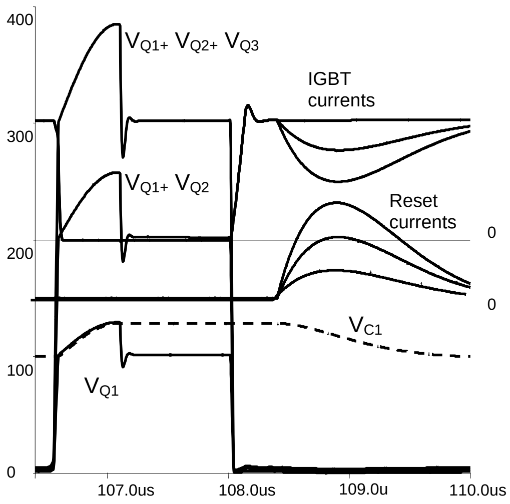


Figure 5: Operating waveforms for Figure 4.

In practice, differences in device characteristics, component values, discharge paths etc will cause asymmetry in the switching behaviour of devices within a string. However, device voltages are effectively clamped by the energy absorption capacity of the RCD snubber effect. For example, Figure 8 shows that the peak voltage arising across $Q_2$ and $Q_3$ (or $Q_1$) when the turn off (turn on) of $Q_1$ is delayed by 200 ns in the string of three devices switching 100 A.

## 5 Experimental results

The viability of both Figure 4 and 6 snubber schemes was proved experimentally using a string of three series connected power MOSFET modules switching 100 A from a 330 V supply. Static voltage-sharing resistors and small RC snubbers were added across MOSFETs in the practical circuits. In the oscilloscope traces shown in Figures 9 to 14, turn on of the devices occurs first followed by turn off and the following scaling factors apply: voltage scale 25V/div, current scale 20A/div, time base 500ns/div. In Figures 5, 7 and 8, turn off precedes turn on. A single current waveform corresponding to $i_{LS}$ is shown, which first rises to 120 A due to diode reverse recovery, because of the difficulty of measuring individual device currents within a circuit which has been highly integrated to minimise stray inductance. Figures 10 and 11 show that Figure 6, the simpler self-commutating scheme, provides satisfactory uniform voltage clamping effect, just as the more complex Figure 4 scheme. In both, the initial voltage transient at device turn off, of a duration approximating to the power MOSFET current fall time, arises when device current commutates to $D_{C1}$ and $C_1$ due to $D_{C1}$ forward recovery and $D_{C1}$-$C_1$ stray inductance.

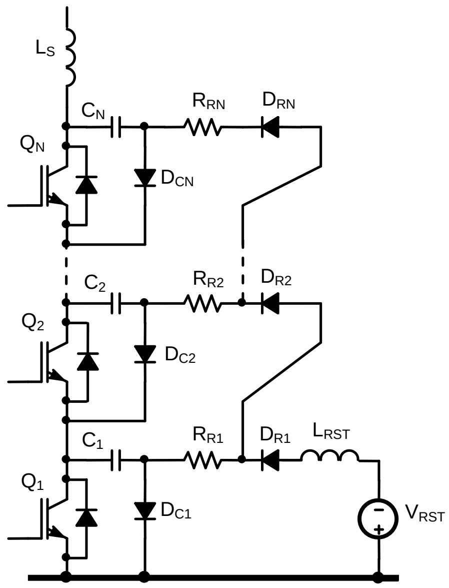


Figure 6: Series RCD scheme without switches.

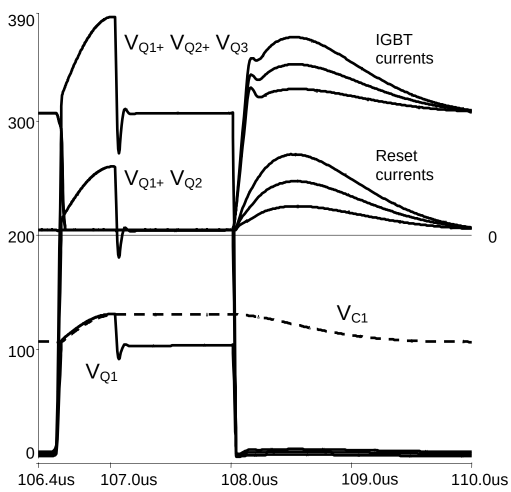


Figure 7: Operating waveforms for Figure 6.

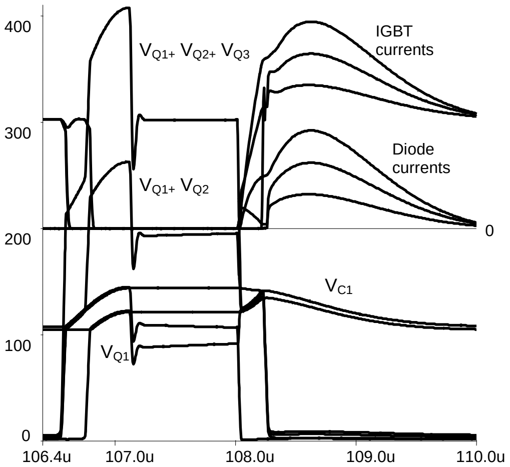


Figure 8: Asymmetric switching waveforms for Figure 6.

The following voltage rise, lasting less that 1μs, is controlled by the snubber capacitors. It is worth noting that the magnitude of voltage overshoot may be reduced by increasing the value of capacitor. However, because the capacitor charges and discharges to $V_{DC}/N$ as shown in Figure 12, turn-off-associated snubber-reset loss remains relatively constant at about $1/2L_SI^2$. Snubber capacitor and $L_S$ value determine the $L_S$ reset time at turn-off and snubber capacitor reset time at turn-on. The composite capacitor-reset-current is shown in Figure 12, reaching 60 A and lasting 3μs.

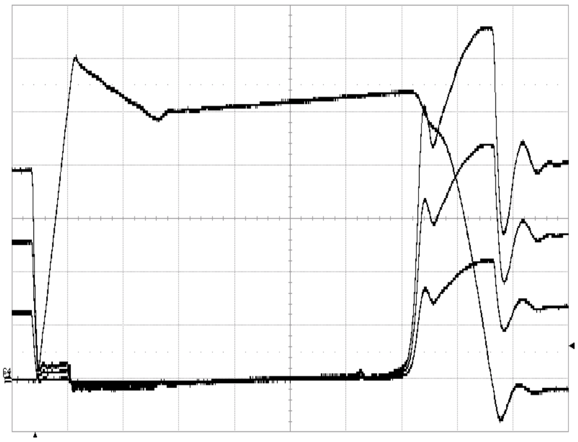

Figure 9: Current and $v_{Q1}$, $v_{Q1}+v_{Q2}$ and $v_{Q1}+v_{Q2}+v_{Q3}$ waveforms for a practical implementation of Figure 4.

To confirm the tolerance of the snubber scheme to asymmetry in gate–drive and power-semiconductor switching transients, $Q_1$ was switched off 200ns before $Q_2$ and $Q_3$ in the circuit in Figure 6. The resulting device voltage waveforms are shown in Figure 13. During the interval between $Q_1$ and $Q_2$, $Q_3$ switching off, $D_{C1}$ and $C_1$ conduct the full load current and limit the voltage across $Q_1$ to below its rated value. It should be noted that 200ns has previously been used as a safe worst-case difference due to differences in the drive circuits and gate-drive circuits of 2.5 kV, 1.8 kA flat-packaged IGBTs [9]. Snubber capacitor value must therefore be chosen by anticipating the worst case difference in switching times. As previously noted, a conservative design results in longer reset time, but not excessive reset power loss. Turn off conditions were the same as for Figure 8; however, cumulative device voltages are given there.

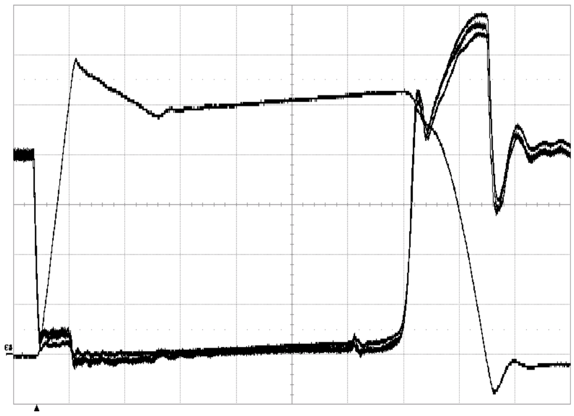

Figure 10: Current and individual $v_{Q1}$, $v_{Q2}$ and $v_{Q3}$ waveforms for a practical implementation of Figure 4.

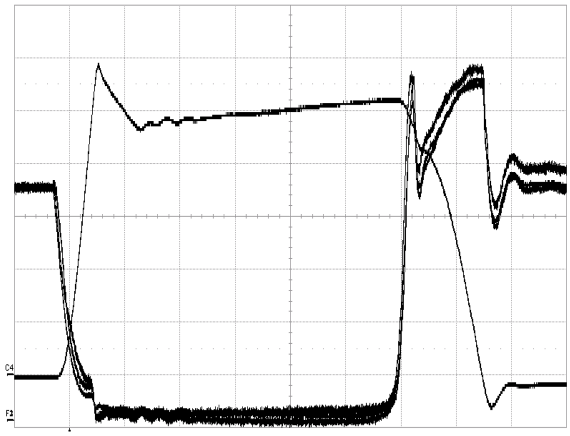

Figure 11: Current and individual $v_{Q1}$, $v_{Q2}$ and $v_{Q3}$ waveforms for a practical implementation of Figure 6.

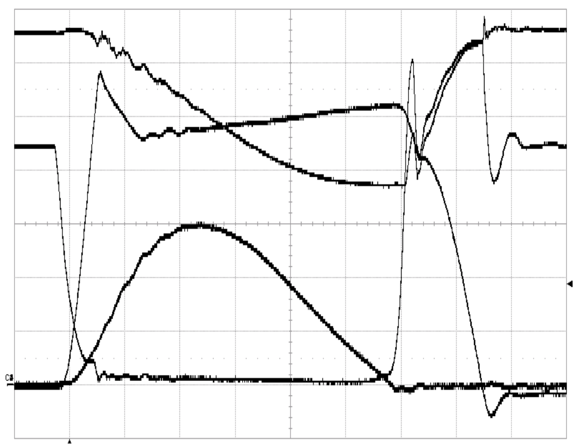

Figure 12: Current $v_{Q1}$, $v_{C1}$ and $i_{D1}$ waveforms for a practical implementation of Figure 6.

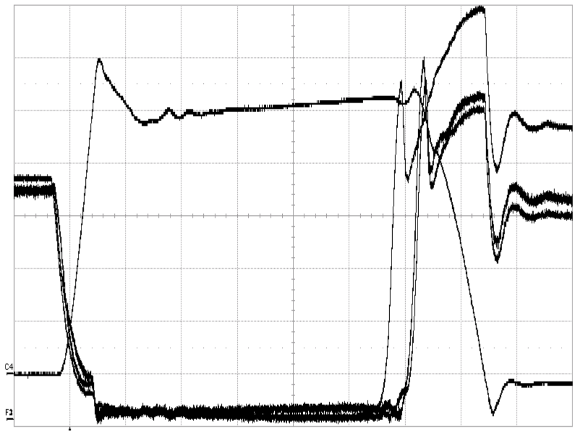

Figure 13: Current and $v_{Q1}$, $v_{Q2}$ and $v_{Q3}$ waveforms for Figure 6 with $Q_1$ turn off 200ns before Q2 and Q3.

In inverter-pole applications, similar voltage sharing action would be produced across the freewheel diodes which are connected in parallel with the switching devices, as shown in Figures 4 and 6, when they turn off at the conclusion of reverse recovery.

## 6 Conclusions

Two dynamic voltage sharing schemes have been investigated which may be used to protect any series connected power-semiconductor device. The schemes have the potential to give significantly less switching loss than active and simpler passive schemes provided that energy returned to the reset sources is regenerated to the main supply or otherwise used. Then, only energy trapped in unclamped inductance at turn off and diode recovery is dissipated. Both schemes may be made fully regenerative by removing resistance in the capacitor reset paths. Circuit operation has been investigated by circuit simulation and an experimental investigation. Results have been shown to be in good agreement.

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