A Lumped, Resonant Switched Capacitor Converter

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Abstract— This paper considers the output short-circuit protection of a Series-Parallel (S-P), switched capacitor converter for interfacing high-power PV farms and/or MV DC micro-grids. A so-called active interrupter can be used to isolate the converter from the fault; however, the inductor associated with this interrupter can be used as the resonant inductor for the circuit, offering potential savings in inductor size/weight/cost and losses. A number of alternative S-P topologies were devised, which were compared against the existing circuit. Results from a scaled hardware prototype are presented, which highlights a number of practical issues related to circuit parasitics.

Keywords—Power Electronics, DC-DC converters, Power Electronics Applications in Power Systems and Utilities

I. INTRODUCTION

Resonant switched capacitor (SC) converters are being proposed for high-power renewable energy applications that require a high DC voltage-conversion ratio. For example, High Voltage DC networks for off-shore wind and step-up, DC-DC converters for MV grid-tied inverters that connect LV batteries or PV panels to MV DC networks [1]–[3].

SC converters have the advantage that they can easily achieve high voltage-conversion ratios with high efficiency and where the voltage stress is inherently shared across the individual cells within the converter. This is achieved without, large bulky inductors making them particularly suitable for space and/or weight critical applications, for example urban sub-stations and offshore platforms [4]. Importantly, with the addition of very small inductors – typically air-cored – the circuit operates with zero-current switching (ZCS) across the complete range of output load.

This paper describes work that has been carried out between Loughborough University in the UK and the Indian Institute of Science (IISc), Bangalore as part of the Joint UK-India Clean Energy initiative (JUICE) [5]. The project has focused on the investigation of direct connection of PV to the MV AC grid or DC network. The requirement is for a 100 kW, non-isolated, unidirectional DC-DC converter, which steps-up the voltage from 600 V to several kV – gains of around 5–30. This can be difficult to achieve in high-power high-voltage applications where voltage insulation and a lack of high frequency transformers are a constraint [6], which is why an SC converter is being considered here. This paper focusses on the protection of the SC converter against short-circuit-to-ground faults on the network side of the converter.

High-power converters that connect energy sources to power networks must achieve high availability. However, they are particularly vulnerable to indiscriminate short-circuit faults within the network to which they are connected. Short-circuit faults are a particular issue for SC converters because (a) the overall stored energy in the cell capacitors can be significant, and (b) the SC converter has no inductor and so cannot operate in current-limit during external faults.

This paper describes a modification to a traditional SC resonant topology that uses the cell ZCS inductors to limit the rate-of-rise of short-circuit fault current. This then allows a blocking transistor to rapidly isolate the converter from the faulted network. Three new converter circuits are compared against the existing circuit and results from a 4-cell, low-power, hardware prototype are presented.

There are numerous papers on the protection of SC converters primarily for use in low-power, on-chip applications such as over temperature, startup-shutdown and even cyber-attacks [7]. But nothing specifically about short-circuits faults with high-power converters. Whilst there appears to be related work on the protection of the flying capacitor multilevel converter (FCMC) [8], the inductor and modulation control used in the flying capacitor circuit allow current-mode control. This is a critical difference between how faults are handled in the FCMC converter compared with the SC converter.

II. FAULT ANALYSIS FOR THE SERIES-PARALLEL SC CONVERTER AND SYNTHESIS OF THE LUMPED VERSION

A SC converter can be realised as the interconnection of the so-called Canonical Cell shown in Fig. 1.

Fig. 1. Switched Capacitor Canonical Cell

The cell consists of three-switches $S_1 \rightarrow S_3$, with gate-drive signals g_T and g , and a cell capacitor C_c [9]. For a resonant converter, which provides soft-switching, a small inductor is connected in series with the capacitor. This ensures the cell current is sinusoidal and synchronised to the gate-signal so that all three switches turn on/off at the zero crossing, resulting in ZCS. The switches can be implemented as either transistors or diodes depending on whether the converter is unidirectional or bi-directional. The gate signals are switched with a constant 50% duty-cycle, the two halves of the cycle corresponding to a cell capacitor charging cycle, followed by a discharge cycle.

The cells can be connected in numerous ways to form a variety of circuits. For protection purposes the Series-Parallel converter (S-P) is a good choice because the overall Volt-Ampere rating of the capacitors is a minimum and hence stored energy is low, which helps reduce the level of short-circuit fault currents.

A modified version of the traditional resonant Series-Parallel resonant converter, with input voltage V_{in} is shown as the N-cell circuit in Fig. 2. The cell capacitors and resonant

inductors, which are denoted C_c and L_c respectively, have the same value for each cell.

Fig. 2. Resonant Series-parallel SC converter where S_1 and S_2 are LV rated

The final cell shown in Fig. 2 is known as a sample and-hold cell (S/H), which is used to capture the discharge cycle of the cells; this cell provides the converter DC output voltage and current V_{out} and I_{out} respectively.

The step-up ratio between the input and output voltage is given by $V_{out}/V_{in} = N+1$. The source will be inductive, especially if it is part of a DC network, therefore an input decoupling capacitor is required to ensure the resonant frequency of each cell is unaffected by the supply impedance.

This variation of the traditional S-P topology is more suitable for high-voltage applications because cell switches S_1 and S_2 are only rated to the low-voltage input of the converter V_{in} , whereas with the traditional converter only S_2 is LV rated. An important disadvantage with this circuit is that parasitic circuit resistance significantly affects the quality factor of the downstream cell resonant circuits. This limits the maximum number of cells and hence voltage gain < 10 .

Since the application considered here is for a unidirectional, step-up converter, this means switch S_1 can be implemented as a diode as shown in Fig. 3. For high power applications, the switches S_2 and S_3 are shown as IGBTs.

Fig. 3. High-power, unidirectional converter utilising diodes and IGBTs

If a short-circuit fault occurs on the network side of the converter – Fig. 4, the converter switches will be unable to block the fault. Therefore, the input source V_{in} , input decoupling capacitor C_{in} , cell capacitors and S/H capacitor C_{out} will all contribute to the fault.

Fig. 4. Fault current paths during a network-side fault

The resonant inductor L_c , within each cell limits the magnitude of the resonant fault currents due to cell capacitors C_c , to a peak value of $V_{in}/\sqrt{L_c/C_c}$ (A). However, a rapid disconnection of the converter from the fault is needed to protect the converter against faults from the input supply V_{in} , and the output S/H capacitor C_{out} . This can be achieved using a so-called Active Current Interrupter shown in Fig. 5, which consists of an inductor, transistor and freewheel diode [3]. The inductor is used to limit the rate-of-rise of fault current so that the transistor can open in a few micro-seconds. The diode provides a free-wheel path for the inductor when the transistor opens. In this way, over-currents well within the capability of the converter components can be accommodated.

Fig. 5. Active Interrupter used as a high-speed electronic circuit breaker

The Active Interrupter can be placed either at the output or input of the converter as shown in Fig. 6(a) and (b) respectively – note: one of the cells in the previous figures has been removed for clarity. The circuit in Fig. 6(b), requires an additional capacitor between the interrupter and the first cell to isolate the interrupter inductor from the resonance occurring within the cells.

Fig. 6. Series-Parallel converter with interrupter at (a) output and (b) input

With the first topology Fig. 6(a), the interrupter protects all the switches and capacitors. However, the interrupter switch voltage rating must be at the high-voltage side of the converter. Such a switch, for example using series-connected transistors, is difficult to realise in practice.

The interrupter switch for the second converter Fig. 6 (b), is low-voltage, which can be easily implemented using a single high-current IGBT module. This is a very significant benefit compared to the first circuit. However, unlike the first topology, the interrupter only protects the capacitor C_{in} and isolates the input source voltage from the fault. Fortunately, the cell capacitors themselves are protected from excessive peak currents by their own resonant inductors L_c , but this still leaves the additional decoupling capacitors and the S/H capacitor unprotected. This latter point is considered below.

It is now proposed that under normal operation the function of the cell resonant inductors in Fig. 6 (b), can be subsumed by the interrupter inductor. The resonant inductors and additional decoupling capacitor can be removed from the circuit. Whilst an additional inductor is needed at the converter output to protect C_{out} , the overall benefits of only needing an LV interrupter switch and merging the cell inductors into one component are very desirable. The investigation of this converter, which has been termed a Lumped Series-Parallel (S-P) converter is described in the following sections of this paper. The circuit is shown Fig. 7, where the inductor L_{out} has been added to protect the S/H capacitor. The active interrupter inductor now takes on the additional function of a resonant inductor for the cell capacitors and is denoted L_{in} .

Fig. 7. Proposed Lumped Series-Parallel (S-P) converter circuit

III. LUMPED S-P CONVERTER VARIANTS

An initial analysis of the new Lumped converter during normal operation assumes that the resonant period during capacitor charging is the same as that for the traditional S-P converter – the latter will herein be referred to as a Distributed converter.

A. Traditional Distributed Converter

All cells have the same value for the cell capacitor C_c and also the inductor L_c , other than the sample and hold cell, which has $C_{out} \gg C_c$. The cell capacitor current waveform i_{C_c} , over one switching period T_s (s), is the same for all N cells and is shown in Fig. 8:

Fig. 8. Distributed converter cell capacitor current waveform.

By design, the resonant circuit should have a high-quality factor Q , such that the currents in the charge / discharge periods are sinusoidal. Expressions for the peak currents in terms of the converter output current I_{out} , can be obtained by considering the average power through a cell in terms of the converter output power. This analysis then gives amplitudes of $\pm \pi I_{out}$ as shown in Fig. 8.

The cell resonant components are related to the switching period by,

$$T_s = 2\pi\sqrt{L_c C_c} \quad (1)$$

The capacitor C_c is chosen based on a per-unit, peak-peak voltage ripple specification for the cell, denoted r , typically having a value equivalent to 5 %,

$$C_c = \frac{I_{out} T_s}{r V_{in}} \quad (2)$$

B. Lumped Converter

During charging, the converter must achieve the same resonant period as the Distributed converter (1). Since the lumped inductor L_{in} now resonates with all the parallel connection of the cell capacitors, then,

$$L_{in} = \frac{L_c}{N} \quad (3)$$

During the subsequent discharge period, L_{in} resonates with the series connection of the capacitors and hence unlike the distributed converter, the resonant period will be shorter than the charging period. This is shown by the cell capacitor current waveform in Fig. 9. The cell capacitor current waveform terminates early during the discharge period, such that the switching cycle has a period of zero-current. Generally, this mode of operation should be avoided as it leads to higher RMS current and peak current in the components. Two other variants of the circuit are therefore proposed for comparison. In the first case, the transistor duty-cycle and converter switching frequency are adjusted as shown in Fig.

10, which shortens the switching cycle and avoids the period of zero current.

Fig. 9. Lumped converter current waveform for the input inductor L_{in} .

Fig. 10. Lumped converter L_{in} current waveform, avoiding zero current

The new switching frequency period T_A , is given by $T_A = (N+1)T_s/2N$. The duty-cycle D , which is defined as the duration of the charging period over the switching period T_A is given by $D = N/(N+1)$. Since the converter maintains ZCS, the only penalty in increasing the switching frequency is that the AC resistance of the wound components will be higher.

No increase in the switching frequency or duty-cycle is needed for the second circuit variant, but it does require the additional retuning inductor L_{res} at the output of the N^{th} cell as shown in Fig. 11,

Fig. 11. Lumped converter L_{in} current waveform, & retuning inductor L_{res}

The function of L_{res} is to retune the resonant frequency during the discharge period of the converter to remove the period of zero current, so that,

$$L_{res} = (N^2 - 1)L_{in} \quad (5)$$

The current waveform for the cell capacitors i_{C_c} , is then identical to that of the Distributed converter shown in Fig. 8.

IV. COMPARISON OF S-P CONVERTER VARIANTS

Before commencing it is worth summarising the four different converters that are to be compared, which are: (i) the traditional Distributed converter Fig. 6(a)/Fig. 8 and the Lumped converter and its variants, (ii) with $D = 0.5$ duty-cycle Fig. 9, (iii) with an asymmetric duty-cycle $D = N/(N+1)$ Fig. 10 and (iv) with $D = 0.5$ duty-cycle but having an additional re-tuning inductor L_{res} Fig. 11/Fig. 8.

The following comparison is made assuming the cell capacitors have the same value for the Distributed converter and its three variants. Note the voltage rating requirements for the capacitors and switches are identical for all the converters and are therefore ignored. Expressions for the average and RMS currents for the four converter variants have been obtained to compare switch, inductor and capacitor conduction losses. The switch average currents are identical for all four converters and equal to i_{out} for S_2 and S_3 , and $j \times i_{out}$ for S_1 , where $j = N \rightarrow 1$ as the cell number $1 \rightarrow N$. Equations for the RMS currents are shown in TABLE I.

TABLE I. NORMALISED RMS CURRENTS FOR THE SC VARIANTS

	RMS currents normalised to $\pi I_{out}/2$		
	<i>Distributed/ retuned Lumped</i>	<i>Lumped D = 0.5</i>	<i>Lumped D = N/(N + 1)</i>
C_{in}	$\sqrt{N^2 + 1 - \frac{4}{\pi^2}(N+1)^2}$	$\sqrt{N^2 + N - \frac{4}{\pi^2}(N+1)^2}$	$(N+1) \sqrt{\frac{1}{2} - \frac{4}{\pi^2}}$
L_{in}	$\sqrt{N^2 + 1}$ (retuned only)	$\sqrt{N(N+1)}$	$(N+1)/\sqrt{2}$
C_c	$\sqrt{2}$	$\sqrt{N+1}$	$(N+1)/\sqrt{2N}$
$j^{th} \text{ Cell } S_1$	j	j	$j\sqrt{(N+1)/2N}$
S_2	1	$\sqrt{N}$	$\sqrt{(N+1)/2}$
S_3	1	1	$\sqrt{(N+1)/2N}$
L_{res}	1 (retuned only)	-	-
$S/H S_1$	1	$\sqrt{N}$	$\sqrt{(N+1)/2}$
$S/H C_{out}$	$\sqrt{1 - 4/\pi^2}$	$\sqrt{N - 4/\pi^2}$	$\sqrt{\left(\frac{N+1}{2}\right) - \frac{4}{\pi^2}}$
L_{out}	$2/\pi$	$2/\pi$	$2/\pi$

Another metric for comparison purposes is the ripple voltage on the capacitors. Expressions have been derived for each converter and are shown in TABLE II. The value of the input capacitor C_{in} , is determined by grid decoupling requirements and its value will be the same for all the converters. It is therefore not included in TABLE II.

TABLE II. NORMALISED CAPACITOR RIPPLE FOR THE SC VARIANTS

For capacitor:	Capacitor ripple voltage normalised to $I_{out}T_s$		
	<i>Distributed & Retuned Lumped</i>	<i>Lumped D = 0.5</i>	<i>Lumped D = N/(N + 1)</i>
C_c	$1/C_c$	$1/C_c$	$(N+1)/2NC_c$
C_{out}	$\frac{1}{k C_{out}} \left(\frac{\sin^{-1}\left(\frac{1}{k\pi}\right)}{\pi} - \frac{1}{2} + \sqrt{k^2 - \frac{1}{\pi^2}} \right)$		
	$k = 1$	$k = N$	$k = (N+1)/2$
Note: T_s = switching period of the Distributed converter			

The first initial observation from these two tables is that since the switch and inductor components for the two Lumped converters: $D = 0.5$ and $D = N/(N+1)$, are the same, then because the RMS current terms in TABLE I, and the capacitor

voltage ripple, TABLE II, are greater for the $D = 0.5$ case, this circuit can be disregarded from further consideration.

The requirements for the inductive components are used as a final means of comparison. To simplify this appraisal, it is assumed from (3) that L_{in} for the Lumped converter, is fabricated from $N \times$ cell inductors L_c , connected in parallel. This means the cost - including size/weight - is the same for both the Distributed and Lumped converters. However, in practice there will be advantages/disadvantages in constructing L_{in} as a single component, for example less copper, but a reduced surface area for cooling. A discussion of inductor requirements now follows:

A. Resonant inductors L_c , L_{in} and L_{res}

As stated above, the size/weight of the N cell inductors for the Distributed converter and L_{in} for the Lumped converter are the same. Therefore, only a comparison of conduction losses is needed using the 2nd and 3rd rows of TABLE I.

The Retuned Lumped converter also has the additional resonant inductor L_{res} . Using (5) and making the simplifying assumption that $N^2 - 1 \approx N^2$, then L_{res} can be referred to the input side of the converter as L_{in} . The Retuned converter therefore appears to have an input inductor $2L_{in}$.

The inductor conduction losses for L_c , L_{in} and L_{res} as a function of the number of cells N , are shown in TABLE III.

TABLE III. TOTAL RESONANT INDUCTOR CONDUCTION LOSS NORMALISED TO THE DISTRIBUTED CONVERTER, FOR $N=2 \rightarrow \infty$

	<i>Distributed</i>	<i>Lumped D = N/(N + 1)</i>	<i>Retuned Lumped</i>
Inductor cond. loss, L_c & L_{in}	1	$\frac{9}{16} \rightarrow \frac{1}{4}$	1 (Includes L_{res})

From (5), for large values of N , inductor L_{res} will have a relatively high value of inductance and therefore require a ferro-magnetic core. In which case, core-losses should also be factored into the entry for the Retuned Lumped converter in TABLE III. Assuming core and conduction losses are equal, this would effectively raise the table entry from 1 to 1.5.

B. Output inductor L_{out} , and switch interruptor inductance

These two inductors are used solely for protection. The Distributed converter only has an interrupter inductor Fig. 6 (a), and inductor L_{out} is not required, whereas, the opposite is true for the Lumped circuits where the role of the interrupter inductor is subsumed by L_{in} . Note, the retuning inductor L_{res} also appears to assist these two inductors in terms of protection. However, since it is fabricated on a ferromagnetic core it would saturate under short-circuit conditions and can be ignored.

The appraisal therefore reduces to a comparison between the interrupter inductor for the Distributed converter - here termed L_{int} - and L_{out} for the Lumped converter. During normal operation, both inductors conduct the same current i_{out} so this mode is not relevant. Under faulted conditions, the output short-circuit current waveform will initially be dominated by the discharge of the output capacitor C_{out} since the upstream cell capacitors and V_{in} are initially blocked by the S_1 diode. This discharge current will be sinusoidal through either the interrupter inductor L_{int} for the Distributed converter, or L_{out} for the Lumped converter. The peak fault current for the Lumped converter will then be

$V_{out}/\sqrt{L_{out}/C_{out}}$. For the Distributed converter, the peak fault current will be the over-current trip level set for the interrupter:

$$\text{overcurrent trip level} = \frac{V_{out}}{\sqrt{L_{int}/C_{out}}} \sin\left(\frac{T_{SC}}{\sqrt{L_{int}C_{out}}}\right) \quad (6)$$

where T_{SC} is the trip time of the interrupter, having a typical value for IGBTs of $10 \mu\text{s}$. The ratio of the peak fault currents for the two converters termed m , is then,

$$m = \sqrt{\frac{L_{int}}{L_{out}}} \frac{1}{\sin\left(\frac{T_{SC}}{\sqrt{L_{int}C_{out}}}\right)} \quad (7)$$

Capacitors have a higher surge current rating than IGBTs, so that typically m has a value of at least 10. Re-arranging (7) as an implicit equation in terms of L_{int} ,

$$L_{int} = \frac{1}{C_{out}} \left(\frac{1}{T_{SC} \sin^{-1}\left(\frac{\sqrt{L_{int}}}{\sqrt{L_{out}}} \frac{1}{m}\right)} \right)^2 \quad (8)$$

Typical values for the output capacitor for an SC converter with a switching frequency of 10 kHz are $C_{out} = 1 \rightarrow 10 \mu\text{F}$. Therefore, in order to get practical values for L_{int} from equation (8), which would typically be in the range of 10s of μH for an air-core inductor, the equation shows the term L_{int}/L_{out} in the denominator gives $L_{int} > 50L_{out}$. The inductor L_{out} therefore adds no significant additional cost to the Lumped converter when compared with L_{int} for the Distributed converter.

A summary of the comparison of conduction losses and inductive requirements against the Distributed circuit are shown in TABLE IV.

TABLE IV. SUMMARY OF CONVERTER REQUIREMENTS, FOR $N=2 \rightarrow \infty$

Criterion	Converter	
	Lumped $D = N/(N+1)$	Retuned Lumped
Cond. loss : L_{in}	$\approx 45\text{-}50\%$ less	No change
Cond. loss : S_1 & S_3	$\approx 25\text{-}50\%$ less	"
Cond. loss : S_2	At least 50% higher, increasing with N	"
Cond. loss : C_{in}	$\approx 37\text{-}85\%$ less	"
Cond. loss : C_C	At least 10% higher, increasing with N	"
Cond. loss : C_{out}	At least 85% higher, increasing with N	"
Cond. loss : L_{int}	N/A (part of L_{in} above)	"
Additional inductors	L_{out} : no significant cost	L_{res} : costly L_{out} : no significant cost
Interrupter switch	High-current, low-voltage rated	High-current, low-voltage rated

From TABLE IV the Retuned Lumped converter has almost the same performance as the Distributed circuit, other than it has the disadvantage of needing the additional inductor L_{res} . On the other hand, it has the significant advantage that the interrupter switch is rated at the LV side of the converter.

Likewise, the Lumped converter with asymmetric duty-cycle $D = N/(N+1)$, has the advantage of requiring a low-voltage interrupter switch. In addition, unlike the Lumped Retuned converter it does not require the inductor L_{res} . More importantly, the conduction losses in the more lossy components – inductors and switches – are lower than both the other two converters. However, this is not the case for IGBT S_3 , where losses are higher. Whilst the losses in the cell and output capacitors are also higher, these components generally have much lower equivalent series resistance. Therefore, the Lumped converter with asymmetric converter may be a good candidate for low values of N , and for higher values of N if the overall losses in IGBT S_3 are offset by the savings in the other two switches and inductors.

V. HARDWARE PROTOTYPE AND RESULTS

Scaled-hardware prototypes of the Distributed, Lumped asymmetric duty-cycle and Lumped Retuned converters were built and tested in the laboratory. The converter had three cells and a S/H cell giving a voltage gain of four. The nominal input/output voltage was 30 V : 120 V and the output power 100 W. Each cell consisted of two Infineon IRF100P218, 100V, 218 A, MOSFETs for S_2 and S_3 , and a Wolfspeed CVFD20065A, 200 V, 65 A silicon carbide schottky diode for S_1 . The cell-capacitors were 22 μF polypropylene giving a cell-ripple of 5 % for a nominal switching frequency of 10 kHz. The S/H cell capacitor C_{out} was selected as 470 μF , and the converter input capacitor 22 μF . However, the bench DC supply used to emulate V_{in} had a poor high frequency response and required additional 4mF decoupling at its output terminals. The global clock signal, which included a dead-time of 80ns between S_2 and S_3 , was generated by an off-board DSP.

A. Distributed SC converter

Using (1), the cell inductor was chosen as 10 μH . Measured waveforms of the cell 1, S_3 gate signal (yellow), cell 1 input current (blue), and the converter DC input and output voltages V_{in} (green) and V_{out} (red), are shown in Fig. 12.

Fig. 12. Measured cell 1 input current (C4, blue, 2A/div), global gate signal (C1, yellow, 10V/div), V_{in} (C2, green, 20V/div), V_{out} (C3, red, 50V/div); time axis: 20 $\mu\text{s}/\text{div}$

It was initially found that circuit parasitics caused the resonant frequency to be lower than expected so that ZCS was not achieved. This was overcome by reducing the switching frequency to 9.4 kHz. The converter then operated as expected (Fig. 12), having a ripple-free output voltage and ZCS operation, the latter indicated by the current reaching zero as the gate voltage changes state. The step-up ratio was 3.7, which corresponds to a full-load voltage regulation of 7.5 %, which is typical for S/C converters.

Nevertheless, it can be seen from Fig. 12 that the shape of the current was not the expected ideal sinusoidal shape. This was due to excessive damping – low Q. Also, the resonant frequency during the discharge period did not match switching frequency. This was thought to be due to parasitic resistance and inductance. To confirm this, further measurements were used to calculate the parasitics using the non-linear curve fitting technique described in [10]. Subsequent circuit simulation which included these parasitics, then gave a close agreement with the measured current waveform of Fig. 12.

B. Lumped SC

Both the Lumped asymmetric duty-cycle and Retuned circuits were tested using the prototype by shorting the cell inductors of the Distributed converter. Equation (4) was used to calculate $L_{in} = 3.3 \mu\text{H}$. However, it was then found that with this reduced inductance the parasitics were more dominant, making ZCS in all cells difficult to achieve. For example, with the Retuned Lumped converter, this meant that the switching frequency had to be reduced to from 9.4 to 9.0 kHz, and the calculated value of L_{res} from (5), had to be increased from $26.6 \mu\text{H}$ to $37.5 \mu\text{H}$ in order to maintain ZCS throughout the converter cells. The waveforms for the Lumped Retuned converter are shown in Fig. 13.

Fig. 13. Measured cell 1 input current (C4, blue, 4A/div), global gate signal (C1, yellow, 10V/div), V_{in} (C2, green, 20V/div), V_{out} (C3, red, 50V/div); time axis: 20 $\mu\text{s}/\text{div}$

Finally, it was noted that there was excessive ringing on the turn-off voltage of S_3 for cell 1, as shown in Fig. 14(a), where an overshoot of approximately 300% is seen.

Fig. 14. Measured Cell 1, S_3 voltage (C1, yellow, 20V/div) (a) without snubber and (b) with snubber; time axis: 20 $\mu\text{s}/\text{div}$

This was found to be due to resonance between the output parasitic capacitance of the MOSFET S_3 of cell 1 and the input inductor L_{in} . A simple 400 nF, 10 Ω dissipative snubber was used to suppress the overshoot as shown in Fig. 14(b). However, the efficiency of the prototype then dropped by 1.2 % at rated power. This would be overcome in a real full-scale application using lossless clamping techniques.

VI. CONCLUSIONS

This paper has compared a number of new, series-parallel topologies, where lumping the cell resonant inductor into a single component offers a number of advantages. Two Lumped circuits were considered; both have the significant benefit of a low-voltage interrupter switch. The use of the interrupter inductor as the resonant inductor offers further potential savings in conduction losses for the asymmetric duty-cycle converter for moderate step-up ratios. However, there are no such savings for the Retuned converter.

The impact of circuit parasitics is a particular issue with the Lumped converter as the reduced value of the resonant inductor L_{in} means that parasitics then dominate the resonant frequency. Maintaining ZCS in each cell then becomes a problem. Another practical consideration is the resonance between L_{in} and the parasitic output capacitance of the cell 1, transistor S_3 , which requires a relatively large snubber circuit and if dissipative, can reduce the converter efficiency.

The next stage of this work would be to carry out tests on a full-scale circuit to assess the impact of parasitics on a converter rated at several 10s kW and 10s kV output voltage.

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