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Cascaded Switched-Capacitor dc-dc Converters for a USB Power Delivery Compliant Charger

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Abstract—The improvements in modern circuit technology reduce system size, improve the efficiency and reduce the cost. Utilization of soft-switching resonant power converters give high density high efficiency power converters for narrow output voltage ranges. To overcome this a cascaded switched-capacitor dc-dc converter is used to implement a Universal Serial Bus Power Delivery compliant charger. The converter uses a custom designed application specific integrated circuit designed for switched-capacitor dc-dc converters and wide bandgap technology allowing high frequency, high power design. The converter achieves peak efficiencies up to 98 % at 60 W output power with a corresponding junction temperature of 50 °C. Worst case junction temperatures reach 64 °C and is viable for use in long-term battery charger systems. The cascaded power stages take up 14 mm × 22 mm.

Index Terms—Battery chargers, Switched mode power supplies, ASIC, Analog integrated circuit

I. INTRODUCTION

With the improvements in modern technology the world has seen an ever increasing demand for power supplies and battery chargers. The global battery charger sales have risen from a value of 10.18 billion USD in 2010 to a staggering 21.35 billion USD in 2018. With sales and the increase in global power bank market size at an expected 17.2 billion USD by 2020, the need for chargers are clear [1], [2].

As more OEM companies turn towards unified chargers like Universal Serial Bus (USB) Power Delivery (PD), all smartphones, tablets, laptops etc. now use the same plug which reduces the number of different chargers increasing popularity of USB-C chargers. The PD standard [3] specifies the following output voltages; 5 V, 9 V, 12 V, 15 V and 20 V for output currents up to 5 A. The operating specifications aimed for in this work is 5 V 2 A, 12 V 3 A and 20 V 3 A.

Resonant power converters utilizing a series-resonant tank have shown high efficiencies due to zero voltage switching

(ZVS) and zero current switching (ZCS) even with very high switching frequencies. Increasing switching frequencies reduces implementation size [4], [5]. To further reduce implementation cost, size and complexity self-oscillating gate-drives have become very popular [6]. The primary drawback of resonant converters is the high dependency upon load [7]. This load dependence makes the available output voltage regulation range narrow, reducing the operating efficiencies drastically outside the nominal operating point, rendering resonant power converters non-optimal for use in USB-PD compliant chargers.

To overcome this drawback, this work seeks to utilize downstream switched-capacitor dc-dc converters to implement a USB-PD compliant charger. A front-end resonant ac-dc converter is used to supply an output voltage range of 40 V to 48 V ensuring high efficiency within this output voltage range.

Utilizing switched-capacitor dc-dc converters improves efficiency and power density [8]–[10] and improves system integration [11], [12]. To further increase switching frequencies modern wide bandgap technologies are used. Wide bandgap increases device breakdown thresholds and reduces parasitic capacitances allowing for smaller and cheaper devices [7], [13]–[15].

To generate the USB-PD compliant voltages using switched-capacitor converters, means to control the output voltage must be implemented. Three straightforward solutions exist:

- Vary the input voltage for a 2:1 switched-capacitor dc-dc converter from 10 V to 40 V
- Implement gearbox control by including several flying capacitors
- Cascade two or three switched-capacitor dc-dc converters to implement 2:1, 4:1 and 8:1 conversion ratios

Each suggested implementation carries its own pros and cons. The first solution requires a front-end ac-dc converter with a wide range output voltage, heavily deteriorating the

efficiency of the ac-dc converter. With this required wide output voltage range of the ac-dc converter, the need for the switched-capacitor converter is virtually removed.

The second solution is widely used for implementing large output voltage ranges in integrated switched-capacitor dc-dc converters but is seldomly used for discrete implementations, due to the nature of the many power switches, flying capacitors and floating voltage domains [10].

The third solution investigated in this paper requires an output voltage variation from 40 V to 48 V from the front-end converter utilizing either 2:1, 4:1 or 8:1 conversion ratios from cascaded switched-capacitor converters utilizing wide bandgap gallium nitride field effect transistors (GaNFET) and an application specific integrated circuit (ASIC) for easing the switched-capacitor converter design.

The structure of this paper is divided into an introduction of the custom ASIC in Section II, the basic operation and mathematical derivation of switched-capacitor dc-dc converters in Section III, a study on power-transistors in Section IV, a discussion of the measurement results in Section V and lastly a conclusion in Section VI.

II. ASIC IMPLEMENTATION

Since switched-capacitor dc-dc converters often require several floating power devices and floating voltage domains, a custom ASIC was designed and manufactured in a TSMC 0.18 μm process.

The ASIC features three 75 V set/reset triggered latch-based level-shifters presented in [16], [17]. The level-shifters are optimized for high speed, low power consumption and small chip area. Implemented on the same ASIC is four 5 V gate-drivers intended to drive light loads such as wide bandgap GaN devices or logic-level power transistors. The whole ASIC is designed to optimally operate in the frequency range of 100 kHz – 1 MHz from a 5 V rail voltage.

Fig. II shows the active part of the ASIC that is implemented on a multi project wafer (MPW).

The active circuitry takes up $700\mu\text{m} \times 350\mu\text{m}$ corresponding to 0.245mm^2 . The ASIC enables utilizing three floating voltage domains for floating transistors and is usable for both half-bridge and full-bridge power stages.

III. SWITCHED-CAPACITOR DC-DC CONVERTER

A. Basic operation

A 2:1 switched-capacitor dc-dc converter is depicted in Fig. 2. The circuit consists of four switches, S_1 through S_4 with resistance R_x , which can all be implemented with metal-oxide-semiconductor field-effect transistor (MOSFET) or GaNFET technology. The basic principle is as follows: During the first of two equally long states, S_1 and S_3 conduct charging the flying capacitor C_{fly} to $V_{in} - V_{out}$. During the second states S_2 and S_4 conduct, discharging C_{fly} to V_{out} , here disregarding the ripple voltage of the flying capacitor.

Fig. 1. Die photography. Three level-shifters and four gate-drivers in TSMC 180 nm process.

Fig. 2. Switched-capacitor dc-dc converter with a step-down ratio of 2:1.

This behaviour gives rise to a conversion ratio $M = \frac{1}{2}$ thus reducing the input voltage. The corresponding ripple voltage seen across C_{fly} is given by (1) [18].

$$\Delta V_C = (V_{in} - 2V_{out}) \frac{1 - e^{-1/(2f_{sw}R_{tot}C_{fly})}}{1 + e^{-1/(2f_{sw}R_{tot}C_{fly})}} \quad (1)$$

where f_{sw} is the switching frequency and $R_{tot} = R_{esr} + 2R_{ds,on}$ for which R_{esr} is the equivalent series resistance (ESR) of the flying capacitor and $R_{ds,on}$ the switches' ($S_1 - S_4$) on-resistance.

This charging and discharging behavior of C_{fly} results in the voltage depicted in Fig. 3. What is important to note from the figure is, how the specific frequency and capacitance are chosen gives rise to different waveforms across C_{fly} . If the switching frequency is comparably low, the voltage will be much less linear. Choosing a comparably higher frequency will make the voltage more triangular.

The switched-capacitor dc-dc converter can in short be approximated by an equivalent output impedance R_{eq} and a voltage conversion ratio denoted M [19]. The simplification is depicted in Fig. 4.

B. Mathematical derivation

As the switched-capacitor converter in Fig. 2 is a 2:1 converter the conversion ratio is $M = 1/2$. The corresponding

Fig. 3. Normalized flying capacitor voltage for a switched-capacitor dc-dc converter.

Fig. 4. Simplified model for a switched-capacitor converter with a voltage conversion ratio M and output impedance R_{eq} .

R_{eq} of the converter is a function of the flying capacitor, switches and switching frequency. To derive R_{eq} the circuit will be analyzed utilizing the flow of charge as presented by [19].

A generic charge flow vector is shown in (2).

$$\mathbf{a}^{(1/2)} = [q_{out}^{(1/2)} \quad q_1^{(1/2)} \quad \dots \quad q_n^{(1/2)} \quad q_{in}^{(1/2)}] / q_{out} \quad (2)$$

where q_{in} is the input charge and q_{out} is the output charge, of the given state (1/2). Correspondingly q_n is the charge through device n . This could either be the charge flow through a resistor or a capacitor.

The charges through the corresponding resistances R_1 - R_4 (which constitute the device S_1 - S_4 on-resistances) and R_{ESR} are analyzed by the charge flow vector in (3).

$$\mathbf{a}_r = [q_{out} \quad q_1 \quad q_2 \quad q_{esr} \quad q_3 \quad q_4 \quad q_{in}] / q_{out} \quad (3)$$

Correspondingly the charge through the flying capacitor is given in (4).

$$\mathbf{a}_c = [q_{out} \quad q_{Cfly} \quad q_{in}] / q_{out} \quad (4)$$

To find the correct terms for the charge flow vectors, each individual state is investigated. Fig. 5 shows both of the individual states, charging ϕ_1 and discharging ϕ_2 of the flying capacitor.

Since all the output charge runs through R_1 , R_{ESR} and R_3 the corresponding charge vector is given in (5).

$$\mathbf{a}_r^{(1)} = [1 \quad 1 \quad 0 \quad 1 \quad 1 \quad 0 \quad 1] / 2 \quad (5)$$

Fig. 5. Switched-capacitor operation states ϕ_1 charging C_{fly} and ϕ_2 discharging C_{fly} .

Looking at the capacitor charge, all the charge flows from input to output (positive) sign, and thus the corresponding charge vector for the capacitors is given in (6).

$$\mathbf{a}_c^{(1)} = [1 \quad 1 \quad 1] / 2 \quad (6)$$

For ϕ_2 all the output charge runs through R_2 , R_{ESR} and R_4 yielding the charge vector shown in (7).

$$\mathbf{a}_r^{(2)} = [1 \quad 0 \quad 1 \quad -1 \quad 0 \quad 1 \quad 1] / 2 \quad (7)$$

For the flying capacitor, all the charge now runs from ground to the output and thus has a negative sign as per (8).

$$\mathbf{a}_c^{(2)} = [1 \quad -1 \quad 1] / 2 \quad (8)$$

Knowing these charge flow vectors, the equivalent output resistance R_{eq} can be approximated by (9) [19].

$$R_{eq} = \sqrt{R_{FSL}^2 + R_{SSL}^2} \quad (9)$$

Where R_{FSL} corresponds to the fast switching limit and R_{SSL} correspond to the slow switching limit [20]. The switching limits are given in (10) and (11) [20], [21].

$$R_{FSL} = 2 \sum_i R_i a_{r,i}^2 \quad (10)$$

$$R_{SSL} = \sum_{i=1}^n \frac{a_{c,i}^2}{f_{sw} C_{fly,i}} \quad (11)$$

Where $a_{c,i}$ and $a_{r,i}$ correspond to the i^{th} element of the capacitor or resistor charge vector respectively. These equations for R_{FSL} and R_{SSL} assume no dead-time and thus a 50% duty-cycle and an infinitely large output capacitor. Studies show better precision can be obtained taking dead-time and duty-cycle into consideration [21], [22].

With the equations (10) and (11), the fast switching limit and slow switching limit can be described in (12) and (13)

$$R_{FSL} = 2R_{ds,on} + R_{ESR} \quad (12)$$

$$R_{SSL} = \frac{1}{4C_{fly}f_{sw}} \quad (13)$$

The fast switching limit ultimately sets a limit for the absolute minimum output impedance and is determined by the choice of semiconductors S_n and ESR of C_{fly} . The slow switching limit correspond to the output impedance due to the non-linear charging and discharging of C_{fly} .

Fig. 6. Conceptual plot of output impedance R_{eq} of a switched-capacitor dc-dc converter vs. log-frequency.

TABLE I

USB-C VOLTAGES AND WHICH SWITCHED-CAPACITOR STAGE IS ENABLED

Enabled stage(s)	V_{in}	V_{out}	M_{eq}
3	40 V	20 V	1/2
3 & 2	48 V	12 V	1/4
3, 2 & 1	40 V	5 V	1/8

How to balance between the two was presented in [21]. For switched-capacitor converters where there is negligible parasitic capacitive loading on the two nodes of C_{fly} to ground, the optimal relation between R_{SSL} and R_{FSL} is given in (14).

$$R_{FSL} = R_{SSL} \quad (14)$$

Since this application utilizes off-chip switching devices S , and off-chip flying capacitor for high power application, C_{fly} will inevitably end up in the range of several μF . Since any parasitic capacitive loading of the flying capacitor's nodes will be in pF, the parasitic capacitances will be negligible and the optimal balance between R_{FSL} and R_{SSL} given by (14) holds true.

Fig. 6 shows the conceptual output impedance vs. log-frequency. As the figure depicts, by increasing the converter switching frequency the output impedance drops, until it reaches the fast switching limit R_{FSL} which is dominated by the device resistances. Correspondingly for low frequencies, the output impedance is dominated by the non-linear charging and discharging of C_{fly} .

C. Cascading switched-capacitor stages

The structure of cascading switched-capacitor dc-dc converters is depicted in Fig. 7.

To implement the correct voltages according to Section I, the cascaded switched-capacitor stages must be controlled according to Table I.

Thus to get the correct voltages, there must be a way for the output of each switched-capacitor stage to be fed to the corresponding USB output, which enables three discrete output voltages. The straightforward approach is to use low-resistance

TABLE II
CASCADED SWITCHED-CAPACITOR CONVERTER PERFORMANCE WITH DIFFERENT CONTROL SCHEMES FOR A 10 W 5 V OUTPUT.

Control Scheme	ϕ	$R_{eq,casc}$	η
FD		88 m Ω	94.28 %
PS	$\phi_1 = 0, \phi_2 = 90, \phi_3 = 180$	88 m Ω	92.9 %
PS	$\phi_1 = 180, \phi_2 = 90, \phi_3 = 0$	88 m Ω	93.2 %
PS	$\phi_1 = 0, \phi_2 = 90, \phi_3 = 0$	88 m Ω	93.35 %

TABLE III

CASCADED SWITCHED-CAPACITOR CONVERTER DESIGN PARAMETERS.

Parameter	Stage 1	Stage 2	Stage 3
f_{sw}	1 MHz	500 kHz	250 kHz
C_{fly}	5.6 μF	11 μF	22.3 μF
R_{ESR}	2.8 m Ω	2.8 m Ω	2.8 m Ω
$R_{ds,on}$	16 m Ω	16 m Ω	13.5 m Ω

switches such as back-to-back PMOS/NMOS devices. The output impedance for each case can then be described as:

$$R_{eq,20V} = R_{eq,3}$$

$$R_{eq,12V} = R_{eq,2} + M^2 R_{eq,3}$$

$$R_{eq,5V} = R_{eq,1} + M^2 R_{eq,2} + M^4 R_{eq,3}$$

where M is the step-down ratio of each stage. This in turn yields (15) for N cascaded switched-capacitor stages with output impedance R_{eq} of stage i .

$$R_{eq,casc} = \sum_{i=0}^{N-1} R_{eq,i} M^{2i} \quad (15)$$

These assumptions only hold true for infinitely high input impedances, which is naturally not achieved, but the implications of non-infinite input impedances is not significant.

When it comes to controlling each individual switched-capacitor stage in Fig. 7, there is a variety of control methods. The two most obvious methods consist of frequency doubling (FD) between each stage or phase shift (PS) between each stage. The primary concern is to ensure a current-path from input to output in the case where multiple cascaded stages are used.

To test the two control methods LTSpice was used. Table II shows the results for the simulations. Here FD denotes frequency doubling for each subsequent stage, meaning stage 2 is the double frequency of stage 3, and stage 1 is the double frequency of stage 2. The chosen base frequency of stage 3 is 250 kHz. The parameter ϕ denotes the phase shift between each subsequent power stage. The chosen frequency utilizes the power transistors chosen in the Section IV.

From which it is clear that doubling the frequency for each subsequent stage yields the best performance. Utilizing the FD control method for the cascaded switched-capacitor stages the chosen system parameters are shown in Table III.

Since the design of the switched-capacitor converters is an iterative process, the $R_{ds,on}$ of the switch devices found in following Section IV is included in Table III.

Fig. 7. Equivalent model of three cascaded switched-capacitor dc-dc converter stages. All i converters have output impedance $R_{eq,i}$ and conversion ratio M .

TABLE IV
POWER TRANSISTOR REQUIREMENTS FOR EACH POWER-STAGE

SC Stage	Voltage [V]	Average Current [A]
1	12	2
2	30	3
3	60	3

IV. POWER TRANSISTOR STUDY

For any power converter a key component is often the power transistor. Much of the converter stresses and limits are based on the power transistor, and the past years present interesting findings within new technologies like wide bandgap devices.

Choosing the right power transistor is not always straightforward, and many aspects must be kept in mind. A conservative approach is to find transistors that comply to voltage and current specifications and then looking at the figure-of-merits (FOM), that is most often defined by (16) and (17) [23], [24]

$$FOM_1 = R_{ds,on} Q_g \quad (16)$$

$$FOM_2 = R_{ds,on} Q_{oss} \quad (17)$$

$$FOM = FOM_1 + FOM_2 \quad (18)$$

Where $R_{ds,on}$ is the on-resistance of the power transistor, Q_g the total gate charge and Q_{oss} the total output charge.

For this application there is a variety of voltage and current specifications set for each switched-capacitor converter, denoted SC Stage x , presented in Fig. 7. The specifications are given in Table IV with an additional safety margin of 20%.

With these device specifications, a list of various devices can be investigated. Since the implemented ASIC is implemented in a 5 V process, the only devices of interest are GaNFETs and logic-level power MOSFETs.

Fig. 8 shows the combined FOM vs. area for some power switching devices that comply to the specs in Table IV. The figure shows some MOS devices beat the GaN devices on their total FOM, which is due to a very low FOM2 compared to GaNFETs, but have reasonably higher FOM1.

This means in general, the GaNFETs are much easier to drive, since their gate capacitance is smaller compared to the MOSFET counterparts. This has the benefit that they are better suited for high switching frequencies compared to the MOSFETs.

Fig. 8. Figure of merits for various logic level power-transistor MOSFETs and GaNFETs.

Since this application will utilize switching frequencies up to around 1 MHz, to achieve high power densities, small switching devices are desired and low gate-capacitances are required because of the ASIC's performance constraints, thus all the MOSFETs are ruled out leaving only GaN devices.

As there are various voltage requirements for each switched-capacitor stage shown in Table IV, different transistors can be chosen if it is more beneficial to the overall design. The first GaNFET of interest is the EPC2014C which has the smallest package size and a total FOM around 100. This device tolerates up to 40 V and 10 A continuous, which fulfills the requirements for stage 1 and 2 but not 3, which requires 60 V devices. However, choosing EPC2014C for the two viable stages will ensure a very small footprint size with the best performance.

The device chosen for stage 3 is the EPC2052, which is a 100 V 8.2 A rated device, which albeit is somewhat overrated but performs much better than the other devices and has a very small form factor similar to the EPC2014.

The switching devices chosen for each power-stage, the technology and their corresponding on-resistance are summarized in Table V.

Knowing the specific devices used for the switches enables us to utilize the equations derived in Section III-B to calculate the remaining parameters such as f_{sw} and C_{fly} for each stage.

The parameters used for f_{sw} and C_{fly} were previously listed in Table III.

TABLE V
CHOSEN POWER TRANSISTORS FOR EACH SWITCHED-CAPACITOR STAGE.

SC Stage	Device	Technology	$R_{ds,on}$
1	EPC2014C	GaN	16 m Ω
2	EPC2014C	GaN	16 m Ω
3	EPC2052	GaN	13.5 m Ω

Fig. 9. Single stage switched-capacitor prototype. Power-stage (red), flying capacitor (yellow), floating voltage supplies (magenta), low voltage power supply (blue) and set/reset pulse generation (teal).

V. MEASUREMENT RESULTS

Two different prototypes have been implemented. One prototype solely for measuring the output impedance to verify the mathematical model and one for implementing the dc-dc converter for generating USB-PD compliant voltages.

A. Single Stage Prototype

The first prototype addresses the modelling of the converter's output impedance and efficiency. It utilizes the designed custom ASIC, four GaN FET GS61004B switches and a 22 μ F flying capacitor.

The prototype is shown in Fig. 9. The prototype includes two different power supplies, one for generating floating voltages for the ASIC gate-drivers and one for low-voltage logic circuitry implementing non-overlapping clock pulses for the logic signals. The single stage prototype is designed to operate at 250 kHz.

To assess the output impedance the output voltage and current were measured using digital multimeters over a broad range of frequency. Fig. 10 shows the measured, simulated and calculated output impedance of the single stage prototype.

The measured output impedance corresponds to both the simulated and calculated output impedance for low frequencies. However, the measurements quickly deviate from the other graphs primarily due to the effects of the duty-cycle, caused by the implemented dead-time and partially due to

Fig. 10. Output impedance of a single stage switched-capacitor dc-dc converter prototype. The graph shows measured (red), simulated (blue) and calculated (green) output impedances.

Fig. 11. Measured and simulated efficiency of a single stage switched-capacitor converter for 40 V to 20 V conversion. Shows measured (red) and simulated (blue) efficiency across load current.

increased junction temperatures. The simulated and calculated output impedances are almost identical and proves the validity of the equations.

Similarly the efficiency of the single stage switched-capacitor converter at 250 kHz is measured. Fig. 11 shows the measured and simulated efficiency of the single stage converter. The measured efficiency includes any power loss in the on-board power supplies, which is indicated by the deviation between the two curves at low output currents. At higher output currents the two curves converge and correspond well to each other.

With the use of a single stage switched-capacitor dc-dc converter the rough equations for the output impedance was tested and verified.

B. Triple cascaded prototype

The triple cascaded switched-capacitor dc-dc converter, which is the heart of this project, is the prototype that implements the switchable USB-PD compliant voltages. The prototype does not include any USB-C controller but only the required power-stage for generating the compliant voltages and currents.

Fig. 12. USB-C PD compliant cascaded switched-capacitor dc-dc converters. Switched-capacitor stages (white), clock generation (blue), set/reset pulse generators (yellow), bias power supply (magenta) and output control logic (red).

Fig. 12 shows the prototype of the cascaded switched-capacitor dc-dc converters. The cascaded converters are shown in the white square and measure a total $14 \text{ mm} \times 22 \text{ mm}$ with two-layer component mounting. Support circuitry including clock generation and set/reset pulse generators also depicted on the prototype could for future iterations be included on the designed ASIC, further reducing the PCB area for implementing USB-PD compliant chargers. The output control logic shown in red is implemented with back-to-back MOSFETs ensuring the correct rail of one of the three cascaded switched-capacitor converters are connected to the corresponding USB-C output.

The cascaded prototype's efficiency is measured for the ranges given in Section I which is a part of the USB-PD compliance range.

Fig. 13 shows the measured and simulated efficiencies (including supply circuitry) for the three output voltages.

For a 20 V output where only stage 3 is running, the peak efficiency is 98 %. It reaches 95 % for 12 V output and 86 % for 5 V output. Both the 20 V and 12 V outputs achieve high efficiencies and are comparable to the expectations shown in simulations. However, the 5 V output does not achieve the expected 94 % efficiency as achieved in the simulations. This deviation is likely caused by poor optimization of the duty-

Fig. 13. Measured (solid lines) and simulated (dashed lines) efficiency of cascaded switched-capacitor dc-dc converters for 40 V input and 5 V (red), 12 V (blue) and 20 V (green) output voltages.

cycles of phase 1 and phase 2 of switched-capacitor stage 1. If both states are not enabled equally long, the prior assumptions and formula are no longer true.

For all three output voltages the temperatures of the corresponding stages were measured. Fig. 14, 15 and 16 show the operating temperature of the corresponding switched-capacitor stages.

The peak temperature reached is 64°C of the second switched-capacitor stage utilized for a 4:1 step-down ratio for 12 V, 36 W out. As the figures depict, the GaNFETs are the devices that reach the highest temperatures. A peak operating temperature of 64°C is well below the temperature recommendations for the GaNFETs and also below any significant long term derating effects. This in turn means a long-term reliable USB-PD charger can be implemented with GaN technology, given all external circuitry provide long-term stability as well.

VI. CONCLUSION

The improvement in modern integrated circuit technologies allow small and efficient systems for high power systems including power supplies and battery chargers. Resonant power converters is one topology offering high efficiencies for very narrow output voltages.

As the Universal Serial Bus (USB) Power Delivery (PD) require a wide output voltage range, this paper presents the use of gallium nitride field effect transistors and an application specific integrated circuit level-shifter and gate-driver for implementation of a down-stream dc-dc switched-capacitor converter

The first prototype illustrates the overall design procedure which yields great performance and a low output impedance. Additionally the application specific integrated circuit proved useful in implementing a high performance switched-capacitor dc-dc converters with little circuit complexity.

The second prototype implemented a charger utilizing a cascaded setup of switched-capacitor dc-dc convertes with a 2:1 step-down ratio and series pass switches to implement

Fig. 14. 5 V, 10 W output, max temperature of stage 1 of 50 °C.

Fig. 15. 12 V, 36 W output, max temperature of stage 2 of 64 °C.

Fig. 16. 20 V, 60 W output, max temperature of stage 3 of 50 °C.

USB-PD compliant output voltages of 5 V, 12 V and 20 V. The charger reached an efficiency of 98 % at 60 W 20 V output and 86 % at 10 W 5 V, including external control and logic circuitry.

Active power-stage circuitry with dimensions of 14 mm × 22 mm reached worst-case peak temperatures of 64 °C and best-case full power temperatures of 50 °C which is well within any temperature limits of modern wide bandgap devices.

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