A transient enhanced DVS DC–DC converter based on state trajectory analysis

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Abstract Dynamic voltage scaling (DVS) can effectively reduce energy consumption by dynamically varying the supply voltage of the system accordingly to the clock frequency. A new DVS-enabled DC–DC converter is presented in this paper. State trajectory is employed to analyze the transient features of PWM and PFM Buck converters. A novel transient enhancement circuit is designed to improve the transient response of the DVS-enabled Buck converter. To further expand the output voltage range of the converter, a current-starved voltage controlled delay line is proposed in the controller of DC–DC converter to obtain an ultra low voltage of 0.5 V. When the input voltage is 3.3 V, the output voltage of the converter can be dynamically regulated from 0.5 to 2.0 V. The output voltage tracking speed is less than 7.5 $\mu\text{s/V}$ and the recovery speed is 33 $\mu\text{s/A}$ for a load current step from 0.6 to 0.2 A at output voltage of 0.5 V. The chip area is 1.75 mm $\times$ 1.33 mm in a 0.18 μm standard CMOS process.

Keywords Dynamic voltage scaling · DC–DC converter · Fast response · Transient enhancement · Low voltage

1 Introduction

DC–DC converter has been regarded as the core of the power management unit. Nowadays, dynamic voltage scaling (DVS) is widely used to reduce power consumption by adjusting the supply voltage of the system. Thus it urgently needs to develop the DC–DC converter whose output

voltage can be dynamically regulated with fast response speed. Apart from the high power conversion efficiency, two other considerations should be taken into account in DVS enabled converter. One is the demanding requirement in transient response [1–3], and the other is the range of the output voltage. Power supply voltage becomes lower and lower in system-on-chip (SoC) integration due to the technology shrinking tendency. Thus DC–DC converters with high dynamic response and low minimum output voltage have attracted more and more attentions.

Many techniques have been proposed to improve the transient behaviors of the switching DC–DC converters. To speed up the transient response, a dual-mode control is proposed in [4], which enables hysteretic control during transient and resumes normal voltage-mode control at steady state. In [5, 6], a fast transient recovery circuit is proposed to achieve a high slew rate of load variation in a voltage-mode Buck converter. In [7, 8], a pulse-train control scheme and a PWM controller are reported. Two states can be transited seamlessly depending on the dynamic operating status. All above techniques are aimed at the non-DVS system and they distinguishes the operating mode of the converter by a fixed threshold band. However, in the DVS utilization, different thresholds are required because the output voltage of the converter is dynamically changed. It brings some difficulties on the loop analysis and transient design.

Moreover, for a DVS enabled system, transient behaviors become more important especially on the output voltage tracking features. Actually, state trajectory is a good method to help us know more behavior details about the switching DC–DC converters. In [9], state trajectory is used to analyze the transient response of the switching converter and a free-running control law is developed. Based on the state trajectory predictions, an enhanced free-

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running control law is proposed in a hysteretic Buck converter to improve the transient behaviors [10]. As a result, the output voltage can reach its steady state within two switching periods even under a large transient step. State trajectory is used to study the large signal stability of the switching converter in [11].

In this paper, a DVS-enabled Buck converter with low output voltage and fast transient response is presented. State trajectory is used to analyze the transient behaviors of PWM and PFM Buck converters, respectively. A transient enhancement circuit is proposed to improve the transient and overshooting of the converter. Additionally, a current-starved voltage controlled delay line (VCDL) is designed in the controller to reduce the minimal output voltage of the converter. The proposed DVS-enabled Buck converter is designed in a 0.18 μm CMOS technology. It can dynamically regulate the voltage from 0.5 to 2.0 V with an input voltage of 3.3 V. The output-voltage and load tracking capability are both improved greatly.

The transient features of different converters are discussed by using state trajectory in Sect. 2. The structure and the detailed circuit implementation of the proposed DC–DC converter are described in Sect. 3. The design results and the chip layout are provided in Sect. 4. Finally, conclusion is given in Sect. 5.

2 Transient analysis of the converters by state trajectory method

There exist different mathematical methods to analyze the switching DC–DC converters. State trajectory is regarded as one of the good methods to examine the transient behaviors [9]. In this section, state trajectory is employed to analyze the transient response of a voltage-mode PWM Buck converter and a hysteretic PFM Buck converter, respectively.

2.1 State trajectories of the Buck converter

The schematic diagram of a Buck converter is shown in Fig. 1. Its behavior can be represented mathematically by a sequence of differential equations. Then the time-domain state solutions can be obtained. For different time intervals, the corresponding state equations are as follows:

When M_1 is on and M_2 is off, the time-domain differential equations are

$$\begin{cases} L \frac{di_L}{dt} = V_{in} - \left(v_C + C \frac{dv_C}{dt} \times R_C \right) - R_L \times i_L \\ C \frac{dv_C}{dt} = i_L - \frac{1}{R_{load}} \times \left(v_C + C \frac{dv_C}{dt} \times R_C \right) \end{cases} \quad (1)$$

The inductor current i_L and the capacitor voltage v_C are state variables. V_{in} is the input voltage of the Buck converter,

Fig. 1 Schematic diagram of the Buck converter

Fig. 2 Families of M_1 -on trajectories (directional dash lines) and M_1 -off trajectories (directional solid lines) in the state plane of i_L versus v_C with $V_{in} = 3.3$ V, $L = 5$ μH , $C = 1$ μF and $R_{load} = 4$ Ω . Directional closed curve ABA represents one of the steady-state trajectories

R_{load} is the load resistance. R_C and R_L are the parasitic resistance of the capacitor and the inductor, respectively. $v_C(t_0)$ and $i_L(t_0)$ represent the arbitrary initial state variables of the converter, and t_0 is an arbitrary initial time.

Similarly, when M_1 turns off and M_2 turns on, the differential equations are

$$\begin{cases} L \frac{di_L}{dt} = - \left(v_C + C \frac{dv_C}{dt} \times R_C \right) - R_L \times i_L \\ C \frac{dv_C}{dt} = i_L - \frac{1}{R_{load}} \times \left(v_C + C \frac{dv_C}{dt} \times R_C \right) \end{cases} \quad (2)$$

These differential equations, when plotted in a state plane of i_L versus v_C , are defined as the state trajectories. In Fig. 2, the directional dash lines are called M_1 -on trajectories which decided by Eq. (1). The directional solid lines are M_1 -off trajectories which decided by Eq. (2). The operations of the converter can be observed as series trajectory connections of M_1 -on and M_1 -off. The closed curve, which consisting of one M_1 -on trajectory and one M_1 -off trajectory, is one steady-state of the converter.

Fig. 3 State trajectories of Buck converters: **a** PWM converter with output-voltage step change, **b** PWM converter with load-current step change, **c** PFM converter with output-voltage step change and **d** PFM converter with load-current step change

2.2 Transient response comparison based on state trajectory

To show the transient behaviors, a voltage-mode PWM Buck converter and a hysteretic PFM Buck converter are built, where V_{in} is 3.3 V, the filter capacitor C is 1 μ F, the inductor L is 10 μ H, the output load R_{load} is 4 Ω and the parasitic resistance of inductor and capacitor are all 50 m Ω . The switching frequency f_s is 800 kHz. When the above parameters are substituted in Eqs. (1) and (2), the state trajectories corresponding to an output-voltage step changed from 1.08 to 2.2 V and a load-current step changed from 412.5 to 206.25 mA for the voltage-mode PWM Buck converter are shown in Fig. 3(a, b), respectively, where “A” represents the initial state and “B” is the final state. The state of the converter alternately follows M_1 -on and M_1 -off trajectories in the state plane, and eventually, converges to a closed steady-state trajectory curve. Two adjacent segments of state trajectories represent an on/off switching cycle. The state trajectories of the hysteretic PFM Buck converter with an output-voltage step change from 1.08 to 2.2 V and a load-current step change from

412.5 to 206.25 mA are also shown in Fig. 3(c, d), respectively.

Comparing the transient response behaviors in Fig. 3, we can see that the hysteretic PFM converter exhibits faster transient features than the voltage-mode PWM converter either in the output-voltage step change or the load-current step change because fewer switching cycles are needed to reach the final steady state. For the DVS applications, in order to speed up the transient behaviors the hysteretic control is selected in this work.

3 System design and circuit implementation of the proposed converter

3.1 DVS operation and system topology

A DVS system block diagram is shown in Fig. 4 [12]. It consists of a DVS-enabled DC–DC converter, a micro-processor and a DVS algorithm block. The DVS-enabled DC–DC converter provides the dynamically regulated voltage V_{DD} to the microprocessor. The software algorithm

Fig. 4 Block diagram of a DVS system with hysteretic Buck converter

is employed in the microprocessor to calculate the optimal supply voltage for maintaining the minimum power dissipation of the microprocessor. The power stage block, hysteretic controller and VCDL constitute a basic DVS-enabled DC–DC converter. Frequency signal f_{ref} is sent to the VCDL. By detecting the phase difference between f_{ref} and the output signal f_{delay} of the VCDL, the hysteretic controller generates different duty ratio signals to control the on/off of the power transistors providing the required optimal supply to the microprocessor. In addition, sometimes, in order to improve the voltage tracking capability, the transient enhancement circuit is added for DVS system. Note that the output voltage of converter V_{DD} is also the control voltage V_{ctrl} of VCDL, which determines the time delay of the VCDL.

3.2 Design of the transient enhancement circuit

For DVS application, switching DC–DC converters should have fast voltage tracking speed. Many techniques have been proposed to improve the transient behaviors. In [13], a fast transient double Buck converter is proposed. Two control mode are used to achieve linear and nonlinear control strategies during the steady and transient period, respectively. It speeds up the transient speed greatly. In this paper, we give the state trajectory analysis of the double Buck converter and then proposed a novel transient enhanced double Buck converter.

3.2.1 Transient response of the previous double Buck technique

Figure 5 shows a voltage-mode double Buck converter architecture [13]. A main converter and an auxiliary converter are connected in parallel. They play different roles in the converter. The main converter focuses on the loop stability and the low output voltage ripple, while the auxiliary converter provides or takes out current from the output when the output voltage exceeds the voltage thresholds.

Fig. 5 Block diagram of the double Buck converter

We can also apply the hysteretic control in the double Buck converter. The corresponding parameters are: $V_{in} = 3.3$ V, the inductance of main inductor L_1 is 10 μ H, the auxiliary inductance $L_2 = 2$ μ H, the filter capacitance $C = 1$ μ F and R_{load} is 4 Ω . The parasitic resistance of L_1 , L_2 and C are all 50 m Ω . The threshold band of hysteretic control is defined as 0.1 V. The state trajectories are therefore obtained, shown in Fig. 6.

Figure 6(a) illustrates the state trajectories for an output-voltage step from 1.08 to 2.2 V. After several switching periods, the converter state moves from the initial steady state “A” into the threshold band and gradually arrives at the final steady state “B”. For double Buck, if the output voltage is within the threshold band, only the main switching converter operates. It likes a typical Buck converter. If the output voltage goes out of the threshold band, the auxiliary converter works. In Fig. 6(a), when the state trajectory enters the threshold band, the auxiliary converter should stop operating. However, because a large quantity of current have stored in L_2 during the up-step transient, the current continues to charge the filter capacitor through the parasitic diode D_1 of $Mn2$. It forces the state trajectory run out of the threshold band again. In order to make the output voltage go back into the threshold band, $Mn2$ should be turned on to discharge the filter capacitor C . But the discharging behavior could not take effect until the current stored in L_2 exhausts and begins to reverse. As a result, an unexpected inductor current oscillation occurs, which resulting the state trajectory path of NOPQ, slows down the voltage tracking speed of the converter.

During the transient period of a load-current change of 412.5–206.25 mA, as shown in Fig. 6(b), the current stored in L_2 is not large enough to generate the oscillation

Fig. 6 State trajectories of the double Buck converter during transient periods: **a** output-voltage step change and **b** load-current step change

Fig. 7 New transient enhanced double Buck converter

explained above. Thus a relatively faster recovering time and lower overshooting voltage are obtained.

Compared to the state trajectories of the single Buck converter shown in Fig. 3(c, d), the double Buck converter has less voltage overshooting and exhibits better transient behaviors.

3.2.2 New transient enhanced double Buck converter

Previous double Buck converter suffers from the inductor current oscillation. Moreover, when there is a large output-voltage variation, such as the converter start-up period, it suffers from large voltage overshooting. To solve this problem, a novel transient enhancement circuit is proposed in this work.

The structure of transient enhancement circuit is shown in Fig. 7. It is composed of a main power converter loop and two auxiliary current paths. Mn2 and L2 constitute one auxiliary path while Mp2 and L3 constitute the other one. D2 and D3 are two current bypasses. Two current bypasses

Table 1 Operation mode of the proposed converter

Voltage range	Operation mode	On/off state of Mp1/Mn1	On/off state of Mp2/Mn2
$[0, V_o - \Delta V_o]$	Low-voltage transient mode	On/off	On/off
$[V_o - \Delta V_o, V_o]$	Steady-state mode	On/off	Off/off
$[V_o, V_o + \Delta V_o]$	Steady-state mode	Off/on	Off/off
$[V_o + \Delta V_o, V_{dd}]$	High-voltage transient mode	Off/on	Off/on

degrade the inductor's current in terms of the RL loop. Because they are totally independent of the main converter loop, it will not induce the non-stability issues.

Table 1 lists the operation mode of the proposed converter, where V_o is the output voltage and $2\Delta V_o$ represents the threshold band. For the DVS application, V_o is dynamically changed. If the output voltage is within the threshold range, only the main converter loop works. Difference appears when the state trajectory moves into the threshold band. For the pervious double Buck converter, the auxiliary converter is forced to shut down, the current stored in L_2 is charged or discharged through the parasitic diodes of Mp2 or Mn2. Since the charge and discharge actions have to be completed through one current path and the direction of the auxiliary inductor current could not be changed instantaneously, it needs longer transient recovery time. However, for the proposed structure, although the power switches are disabled during the transient period, the current bypass diode D2 and D3 provide the auxiliary current path to avoid the same charge and discharge path described above.

The state trajectories are drawn to illustrate the transient response of the converter. L_2 and L_3 are both 2 μ H and the threshold band $2\Delta V_o$ is 0.1 V. The state trajectories of the

Fig. 8 State trajectories of the proposed converter in transient periods: **a** output-voltage step change and **b** load-current step change

converter with an output-voltage step change and a load-current step change are presented in Fig. 8. As shown in Fig. 8(a), after passing point “M”, the state trajectory staggers along the low threshold boundary and moves into the threshold band. Consequently, oscillation appears in Fig. 6(a) is avoided. The load transient behavior in Fig. 8(b) is similar with that of Fig. 6(b). The proposed transient enhancement circuit maintains excellent current-step transient characteristic and exhibits superior voltage-step transient characteristics, which shows more attractive for the DVS applications.

3.3 Current-starved VCDL for lowering the minimum output voltage of the converter

3.3.1 Operational principle of VCDL based Buck converter

VCDL is used to construct a hysteretic controller, as shown in Fig. 9(a). It consists of quantities of delay cells. When a frequency signal of f_{ref} is applied to the input of VCDL, it passes through delay cells and being sampled by f_{ref} again at the output. The propagation delay t_{delay} is determined by the control voltage V_{ctrl} , which is equals to the output voltage of converter. When V_{ctrl} decreases, t_{delay} increases. The falling edge of the delay signal passes the sampling edge, as shown in Fig. 9(b). The output of D flip-flop becomes “1” and power transistor Mp turns on. The inductor is charged and the output voltage increases. On the contrary, if V_{ctrl} increases, t_{delay} decreases. The falling edge of delay signal does not reach the sampling edge, as shown in Fig. 9(c). The output of D flip-flop keeps its initial value of “0”. In this case, the inductor is discharged and the output voltage of the converter is reduced. On the other hand, if f_{ref} decreases, the falling edge of the delay signal will not reach the sampling edge, as shown in

Fig. 9(d). The output of D flip-flop becomes “0”, and the inductor discharges current from output capacitor, resulting decrease of the output voltage. Similarly, for a increased f_{ref} , as shown in Fig. 9(e), the output voltage is increased.

When the VCDL based Buck converter is operating in its steady state, the propagation delay t_{delay} and f_{ref} satisfy the following relationship:

$$t_{\text{delay}} = \frac{1}{2f_{\text{ref}}} \quad (3)$$

For the DVS-enabled converter, an excellent linearity between the output voltage and the reference frequency f_{ref} is required. An inverse proportion between the control voltage V_{ctrl} and the delay time t_{delay} is required to compensate the nonlinearity between t_{delay} and f_{ref} according to Eq. (3).

3.3.2 Current-starved VCDL

In order to regulate the output voltage as low as 0.5 V, VCDL should effectively operate in sub-threshold region. Traditional CMOS-inverter based VCDL shows approximate inverse-proportion feature between V_{ctrl} and t_{delay} when V_{ctrl} is much higher than the threshold voltage of the MOSFETs. But when V_{ctrl} is lower or approximate to the threshold voltage, CMOS-inverter based VCDL lost the good inverse-proportion feature between V_{ctrl} and t_{delay} . In this work, a current-starved VCDL is proposed to improve the non-linearity at low voltage under the threshold voltage of the MOSFETs. The current-starved VCDL is shown in Fig. 10, which consists of quantities of current-starved inverters and a V–I converter. By controlling the charge or discharge current of the output parasitic capacitor of each current-starved inverter, the propagation delay of the delay line can be regulated. Time delay of one current-starved inverter is as follows [14]:

Fig. 9 Operational principle of VCDL based Buck converter: **a** block diagram of the VCDL based Buck converter and the waveforms when **b** V_{out} decreases, **c** V_{out} increases, **d** f_{ref} decreases and **e** f_{ref} increases

Fig. 10 Schematic of the proposed current-starved VCDL

$$t_{delay_inv} = \frac{C_{tot} \times V_{dd}}{I_D} \quad (4)$$

where C_{tot} is the equivalent parasitic capacitance of the inverter. It is a constant. V_{dd} is the power supply voltage of VCDL and I_D is the bias current of the inverter.

A V–I converter is designed to achieve a linear variation of t_{delay_inv} . The structure of V–I converter is shown in Fig. 10. $Mp1$ is controlled by V_{ctrl} . The symmetrical bias voltages V_{bp} and V_{bn} provide DC operating conditions of the transistors. They also determine the charging or discharging current I_D of C_{tot} . $Mn1$, $Mn2$ and $Mn4$ have the same dimensions, while $Mp2$ and $Mp3$ have the same dimensions. $Mp1$ operates in the linear-region. A bias voltage of V_{bias} , which is a little higher than V_{dd} , is added to the bulk of $Mp2$ to lower the threshold voltage of $Mp2$. Following equations can be obtained:

$$\begin{cases} (V_{GS})_{Mp2} = (V_{GS})_{Mp3} + (V_{DS})_{Mp1} \\ (V_{TH})_{Mp2} = V_{THp0} + \gamma(\sqrt{|2\phi_F + V_{dd} - V_{bias}|} - \sqrt{|2\phi_F|}) \\ (V_{DS})_{Mp1} \approx -\frac{(I_{drain})_{Mp1}}{(k)_{Mp1}[V_{ctrl} - V_{dd} - (V_{TH})_{Mp1}]} \end{cases} \quad (5)$$

where V_{GS} is the gate voltage, V_{DS} is the voltage between drain and source, V_{TH} is the threshold voltage and ϕ_F is the Fermi potential. That is,

$$\begin{aligned} -\sqrt{\frac{2|(I_{drain})_{Mp2}|}{(k)_{Mp2}}} + (V_{TH})_{Mp2} &= -\sqrt{\frac{2|(I_{drain})_{Mp3}|}{(k)_{Mp3}}} \\ &+ (V_{TH})_{Mp3} - \frac{(I_{drain})_{Mp1}}{(k)_{Mp1}[V_{ctrl} - V_{dd} - (V_{TH})_{Mp1}]} \end{aligned} \quad (6)$$

where I_{drain} is the drain current of the MOSFET and k is the transconductance coefficient. Considering that,

$$(I_{\text{drain}})_{Mp1} = (I_{\text{drain}})_{Mp2} = (I_{\text{drain}})_{Mp3} = -I_D \quad (7)$$

$$(V_{TH})_{Mp2} = (V_{TH})_{Mp3} = V_{THp0} \quad (8)$$

$$(k)_{Mp1} = (k)_{Mp2} = (k)_{Mp3} \quad (9)$$

From equations from (6) to (9), we can obtain

$$I_D = K \times (V_{\text{ctrl}} - V_{dd} - V_{THp0}) \quad (10)$$

where

$K = \gamma \times (\sqrt{|2\phi_F + V_{dd} - V_{bias}|} - \sqrt{|2\phi_F|}) \times (k)_{Mp1}$. V_{THp0} and K are both process-related constants. γ is the substrate bias coefficient, $\gamma < 0$.

Assuming there are N stage current-starved inverters in the VCDL, the total delay t_{delay} of the VCDL is obtained by combining Eqs. (4) and (10).

$$t_{\text{delay}} = \frac{N \times C_{\text{tot}} \times V_{dd}}{K \times (V_{\text{ctrl}} - V_{dd} - V_{THp0})} \quad (11)$$

Therefore, combining Eqs. (3) and (11), the linear relationship between V_{ctrl} and f_{ref} is obtained as follows:

$$f_{\text{ref}} = \frac{K \times (V_{\text{ctrl}} - V_{dd} - V_{THp0})}{2N \times C_{\text{tot}} \times V_{dd}} \quad (12)$$

The current-starved VCDL can work at an ultra low voltage, even the voltage a little higher than 0 V is acceptable, which makes it possible for the converter to regulate the ultra low output voltage of 0.5 V.

4 Circuit design and result discussion

A DVS-enabled fast transient double Buck converter with a current-starved VCDL is designed in a 0.18 μm CMOS process. The circuit configuration is shown in Fig. 11. The Clock Generator generates the periodic pulses for the Boost Circuit and the enable signals for the converter. A bias voltage V_{bias} being a little higher than the supply voltage V_{in} is generated by the Boost Circuit module to provide a N-well bias of the VCDL. The Control Block generates the duty variation PFM control signals, which are sent to the Dead Time and Driving Block to drive the power switches. The converter operates in three modes: the high-voltage transient mode when V_{out} is higher than $V_o + \Delta V$, the low-voltage transient mode when V_{out} is lower than $V_o - \Delta V$, and the steady-state mode when V_{out} is between $V_o - \Delta V$ and $V_o + \Delta V$. The voltage and current references are generated by Bandgap and the Protection Circuit provides the protections of the chip.

The chip layout is shown in Fig. 12. The die area is 1.75 mm $\times$ 1.33 mm, including pads and ESD protections. The post-layout simulation results of the converter with

Fig. 11 Topology of the proposed DVS-enabled Buck converter

Fig. 12 Layout of the proposed converter

Process, Voltage and Temperature (PVT) variations are shown in Fig. 13. The same parameters as those of the hysteretic converter are used. Three process corners are simulated and the waveforms are divided into three groups, which depict the converter's transient characteristics. Figure 13(a–d) are the results under Typical NMOS and Typical PMOS (TNTP) corner, Fig. 13(e–h) are the results under Slow NMOS and Slow PMOS (SNSP) corner and Fig. 13(i–l) are the results under Fast NMOS and Fast PMOS (FNFP) corner. Take the TNTP corner as an example, Fig. 13(a) shows the waveforms of the converter with the output voltage step between 0.5 V and 1.8 V. When the reference frequency f_{ref} is 106 MHz, the output voltage of the converter V_{out} is 0.5 V. When the

Fig. 13 Post-simulated waveforms with output-voltage/load-current step changes: **a–d**, **e–h** and **i–l** are the results under the TNTP corner, the SNSP corner and the FNFP corner, respectively. **a**, **e**, **i** are the results when output voltage is step-changed between 0.5 and 1.8 V. **b**, **f**, **j** are the results when output voltage is step-changed between 1.0

and 2.0 V. **c**, **g**, **k** are the results when $V_{out} = 0.5$ V and output current is step-changed between 200 and 600 mA. **d**, **h**, **l** are the results when $V_{out} = 2.0$ V and output current is step-changed between 200 and 600 mA

corresponding f_{ref} changes to 57 MHz, V_{out} is 1.8 V. The voltage up-tracking speed of 4.74 $\mu s/V$ and the down-tracking speed of 7.46 $\mu s/V$ are obtained. Figure 13(b) shows the results when the output-voltage

changes between 1.0 V and 2.0 V. When f_{ref} is 92 MHz, V_{out} is 1.0 V, while when f_{ref} is 48 MHz, V_{out} is 2.0 V. The voltage up-tracking speed is 2.0 $\mu s/V$ and the down-tracking speed is 5.4 $\mu s/V$. Figure 13(c, d) show the

Fig. 13 continued

waveforms when the load-current of the converter changes between 200 mA and 600 mA. When the output voltage is 0.5 V, the load up-tracking speed of 9.13 μ s/A and the down-tracking speed of 33.0 μ s/A are obtained. When the output voltage is 2.0 V, the up-tracking speed is 16.1 μ s/A and the down-tracking speed is 15.25 μ s/A. Similarly, the voltage tracking speed and the load tracking speed under SNSP corner and FNFP corner can be obtained,

respectively, which are summarized in Table 2. This converter exhibits a slower transient response under the conditions of high junction temperature with SNSP corner than it does under the conditions of low junction temperature with FNFP corner. However, it is still a robust design against PVT variations.

Table 3 summarizes the features of the proposed converter. The output-voltage and the load-current tracking

Table 2 Transient response of the proposed converter with PVT variations

Parameters	Values	Values	Values
Process corner	FNFP	TNTP	SNSP
Input voltage	3.6 V	3.3 V	3.0 V
Temperature	−40 °C	25 °C	125 °C
Reference frequency for 0.5 V output	117 MHz	106 MHz	97 MHz
Reference frequency for 2.0 V output	53 MHz	48 MHz	43 MHz
Up-tracking speed of the output voltage	4.5 $\mu\text{s/V}$ @ (from 0.5 to 1.8 V)	4.7 $\mu\text{s/V}$ @ (from 0.5 to 1.8 V)	6.5 $\mu\text{s/V}$ @ (from 0.5 to 1.8 V)
Down-tracking speed of the output voltage	5.1 $\mu\text{s/V}$ @ (from 1.8 to 0.5 V)	7.5 $\mu\text{s/V}$ @ (from 1.8 to 0.5 V)	10.8 $\mu\text{s/V}$ @ (from 1.8 to 0.5 V)
Load up-tracking speed	12.3 $\mu\text{s/A}$ @ (from 200 to 600 mA)	16.1 $\mu\text{s/A}$ @ (from 200 to 600 mA)	20.0 $\mu\text{s/A}$ @ (from 200 to 600 mA)
Load down-tracking speed	23.3 $\mu\text{s/A}$ @ (from 600 to 200 mA)	33.0 $\mu\text{s/A}$ @ (from 600 to 200 mA)	40.0 $\mu\text{s/A}$ @ (from 600 to 200 mA)

Table 3 Key features of the proposed converter

Parameters	Values	Parameters	Values
Input voltage	3.0–3.6 V	Load current	200–600 mA
Output voltage	0.5–2.0 V	Switching frequency	0.74–1.71 MHz
Linear regulation	2.0 %/V @ ($V_{\text{out}} = 0.5$ V)	Load regulation	2.5 %/A @ ($V_{\text{out}} = 1.0$ V)
Output voltage tracking speed	7.5 $\mu\text{s/V}$ @ (V_{out} changes from 1.8 to 0.5 V)	Load current tracking speed	33.0 $\mu\text{s/A}$ @ ($V_{\text{out}} = 0.5$ V, I_{load} changes from 600 to 200 mA)

speeds are compared with those of the previous works in Tables 4 and 5, respectively. This work shows the lowest minimal output voltage and thus a wider output-voltage range. Moreover, it has a faster output-voltage tracking speed comparing with the other DVS-enabled converters. Besides this, our work shows a superior load tracking capability.

5 Conclusion

A fast transient DVS-enabled Buck converter has been designed in this paper. The transient features of the

Table 4 Comparisons of the output-voltage tracking speed

Converter	This work ^a	Work in [15] ^b	Work in [16] ^c
Type	Buck	Buck	Boost
DVS	Enable	Enable	Enable
Input/output voltage range	3.0–3.6/ 0.5–2.0 V	2.7–3.3/ 0.8–1.6 V	1.5/0.9–2.2 V
Up-tracking speed of the output voltage	4.7 $\mu\text{s/V}$ @ (from 0.5 to 1.8 V)	12.5 $\mu\text{s/V}$ @ (from 0.8 to 1.2 V)	93.3 $\mu\text{s/V}$ @ (from 1.2 to 1.8 V)
Down-tracking speed of the output voltage	7.5 $\mu\text{s/V}$ @ (from 1.8 to 0.5 V)	22.5 $\mu\text{s/V}$ @ (from 1.2 to 0.8 V)	26.7 $\mu\text{s/V}$ @ (from 1.8 to 1.2 V)

^a The results in this work are based on simulated data

^b The results in [15] are based on simulated data

^c The results in [16] are based on experimental data

Table 5 Comparisons of the load-current tracking speed

Converter	This work ^a	Work in [17] ^b	Work in [18] ^c
Type	Buck	Buck	Buck
DVS technique	Enable	Disable	Disable
Input/output voltage	3.0–3.6/ 0.5–2.0 V	3.0–5.0/2.0 V	3.3/2.0 V
Load up-tracking speed	16.1 $\mu\text{s/A}$ @ (from 200 to 600 mA)	37.5 $\mu\text{s/A}$ @ (from 100 to 500 mA)	66.7 $\mu\text{s/A}$ @ (from 100 to 400 mA)
Load down-tracking speed	33.0 $\mu\text{s/A}$ @ (from 600 to 200 mA)	20 $\mu\text{s/A}$ @ (from 500 to 100 mA)	83.3 $\mu\text{s/A}$ @ (from 400 to 100 mA)

^a The results in this work are based on simulated data

^b The results in [17] are based on simulated data

^c The results in [18] are based on simulated data

voltage-mode PWM and hysteretic PFM converters are analyzed by using state trajectory. It can be concluded that PFM control shows better load and output voltage tracking capabilities than PWM control. Thus the hysteretic control is used and it also ensure the unconditional stability of the converter. Furthermore, a novel transient enhancement circuit is proposed based on the state trajectory analysis to effectively reduce the voltage oscillation and the overshooting during the mode transition. The transient recovery time of the Buck converter is improved. In hysteretic controller, the current-starved VCDL is designed to enable the output voltage of the converter reaching to a very low value of 0.5 V, which is very suitable for the low supply-voltage applications. The converter is designed in a 0.18 μm CMOS technology. Post-layout simulation results show that the converter's output voltage can be dynamically regulated from 0.5 to 2.0 V with a load current range

from 200 to 600 mA. The tracking speed is less than $7.5 \mu\text{s/V}$ for the output-voltage step from 1.8 to 0.5 V. For a load-current step from 0.6 to 0.2 A, the tracking speed is within $33 \mu\text{s/A}$. The results show a better dynamic features compared with the other designs.

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