A fully integrated digital LDO with voltage peak detecting and push-pull feedback loop control

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Abstract: A push-pull multi-loop architecture for the digital low drop-out (D-LDO) regulators is presented with small variations of output voltage and 200 mA load capacity. The proposed D-LDO adopts voltage peak detector (VPD) to observe the output voltage ripples. Once undershoot or overshoot on output voltage is detected, the push-pull feedback loop is quickly triggered, which minimizes the voltage shoots even if the load current changes abruptly. Meanwhile, the shift register (S/R) feedback loop regulates the output voltage to desired value with high accuracy. Hence the D-LDO recovers steady state with greatly small voltage spikes. The proposed D-LDO is designed and simulated in SMIC 65 nm CMOS process with a 0.42 mm² active area. The simulated voltage overshoot and undershoot are 27 and 26 mV respectively, with load step of 20 to 200 mA with a 10-ns edge time. The max load current and quiescent current are 200 mA and 400 μ A, respectively, and the peak current efficiency is 99.8%.

Keywords: low drop-out (LDO) regulator, voltage peak detector (VPD), shift register (S/R), push-pull feedback loop

Classification: Integrated circuits

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1 Introduction

Systems-on-Chips (SoCs) integrating many individual blocks (such as memory, digital circuits, analog-to-digital converters among others) becomes more widely used in different applications. Power management circuit likes LDO is becoming critical to provide a clean and regulated power supply for each individual block in SoCs [1]. LDO can be classified into analog LDO (A-LDO) [2] and digital LDO (D-LDO) [3, 4]. Digital LDOs are gaining more attention since they have the advantages of suitable for low operating voltages [3], easy integration, small size, programmability, high stability over a wide range of load current variations and a lower sensitivity to process variations [4]. A baseline discrete-time digital LDO usually consists of a comparator, a serial-in parallel-out bidirectional shift register (S/R), and a P-channel MOSFET (PMOS) array acting as the power transistors [5]. In these discrete-time digital LDOs, S/R shifts the control signal according to the result of voltage comparator to turn on/off PMOS transistors and regulates the

output voltage with a master clock. Thus the generation and propagation of control signal will be synchronous. Due to their synchronous and sequential nature of switching, the baseline digital LDOs suffer from slower transient responses to large load current steps resulting in a large voltage undershoot/overshoot and show a tradeoff between steady-state stability and transient response [6].

Several previous works have been proposed to tackle this issue. A gated voltage controlled oscillator (GVCO) with counters is used in [7] to provide more “continuous” information of the digitalized output voltage level. Based on that information, the D-LDO in [7] controls multiple power transistors concurrently and the transient performance is improved. However, the inherent sensitivity to PVT variations limits its applications. Asynchronous control and adaptive sizing that changes resolution are applied in the D-LDO, targeting at a balance in high speed and fine resolution in [8]. Similarly to [7], the PVT variations become a headache too. The D-LDO presented in [9] using a SAR-type ADC and a separate DAC driving a single power transistor. But, this design is not power-efficient since its complex design consumes additional quiescent current. Another work shown in [10] adopts a TDC-based 4-b ADC and a PID controller for stability compensation. In that design, the performance of fast transient response is achieved through dynamic clock scaling from normally 250 MHz to 1 GHz. However, it consumes large quiescent current of 2.5 mA. To trade-off between transition response and steady performance, the coarse-fine tuning technique is widely used such as in [5, 11]. Briefly, coarse-fine tuning technique mainly includes two aspects, namely the coarse-tuning and fine-tuning. The coarse-tuning with more power MOS units and high sampling frequency enhances transient performance while the fine-tuning with small power MOS units and slow sampling frequency achieves high steady and low quiescent current. Unfortunately, the coarse/fine-tuning algorithm leads to complexity of controlled block.

In this paper, a novel all digital LDO (D-LDO) with voltage peak detecting and push-pull feedback loops is presented. The proposed D-LDO can achieve small voltage spikes, low quiescent current, and up to 200-mA load current capacity. This brief is organized as follows. The architecture and core circuit of our proposed D-LDO is illustrated in Section 2. Steady and small signal model analysis is presented in Section 3. Simulation results are shown in Section 4. Finally, conclusions are made in Section 5.

2 Implementation and design considerations

2.1 Architecture of proposed D-LDO

The architecture of the proposed digital LDO is presented in Fig. 1. Push-pull topology as in [12] is used for multi-loop operation. Generic digital LDO is comprised of load RC network, clock comparator CMP₂, shift register S/R and P-channel MOSFET (PMOS) array (PG₀ to PG₁₅₉) acting as the power transistors. In the proposed architecture, a push-pull feedback loop trigger (PP-FLT) scheme is embedded in the generic digital LDO to control the push-pull output stage that consists of PMOS array (PG₀ to PG₁₅₉) and N-channel MOSFET (NMOS) transistor N_G. As shown in Fig. 1, there are three feedback loops in the proposed

architecture, namely L_{SR} , L_P and L_N . The shift register (S/R) feedback loop L_{SR} that includes load RC network, clock comparator CMP_2 , shift register S/R and the PMOS array is widely used in generic D-LDO to regulate the output voltage to a desired value under difference load conditions, while the proposed push-pull feedback loops L_P and L_N consisting of load RC network, PP-FLT scheme and the push-pull output stage are employed to reduce the variations of V_{OUT} .

Fig. 1. Architecture of proposed D-LDO with voltage peak detecting and push-pull feedback loop control.

The undershoot and overshoot detection boundaries are $(2V_{REF} - \Delta V_L)$ and $(2V_{REF} + \Delta V_H)$, where $2V_{REF}$ denotes the desired value of V_{OUT} in steady state, ΔV_L is the predetermined undershoot detection threshold, and ΔV_H is the overshoot detection threshold. When V_{OUT} is within the range between $(2V_{REF} - \Delta V_L)$ and $(2V_{REF} + \Delta V_H)$, the D-LDO works only at L_{SR} -turning mode, at which only shift register (S/R) feedback loop L_{SR} is available to ensure that V_{OUT} incessantly tracks the desired voltage. On the other hand, once V_{OUT} exceeds undershoot or overshoot detection boundaries $(2V_{REF} - \Delta V_L)$ or $(2V_{REF} + \Delta V_H)$, the L_P -turning or L_N -turning mode is quickly triggered while L_{SR} -turning mode is always available. In this scenario, the push-pull feedback loop L_P/L_N is enable to prevent a large undershoot/overshoot on V_{OUT} while the L_{SR} regulates the V_{OUT} to $2V_{REF}$ with a desired load current.

To achieve wide bandwidths of push-pull feedback loops (L_P, L_N), a high frequency (1 GHz) clock (CK_{HF}) is employed in the PP-FLT scheme, in addition, the push-pull output stage is directly controlled by PP-FLT scheme. Though a high frequency clock is needed, it is only used to trigger two clock comparators in PP-FLT scheme. Thus the increment of power consumption caused by the high frequency clock is negligible. For lower quiescent current and higher accuracy in

steady state, the D-LDO operates only on the L_{SR} -turning mode, with a small power MOS strength and low frequency (50-kHz) clock (CK_{LF}). Therefore, with the proposed PP-FLT scheme, the D-LDO could effectively improve transient response and achieve low quiescent current simultaneously.

2.2 Push-pull feedback loop trigger (PP-FLT) scheme

As shown in Fig. 2, the PP-FLT scheme includes a voltage peak detector (VPD), a 2-input OR gate, a 2-input INAND gate and a 2-input AND gate, whereas VPD is composed of 2 clock comparators (CMP_0 , CMP_1) and a 2-input XNOR gate. The VPD is used to detect the undershoot/overshoot on V_{OUT} , which outputs 1 if V_{OUT} is not within the range between $(2V_{REF} - \Delta V_L)$ and $(2V_{REF} + \Delta V_H)$. With the output of VPD ($PPFL_{EN}$) becoming high, the signal $DIRE_{HF}$ generated by the OR gate indicates whether undershoot on V_{OUT} was detected or overshoot. Then signals $PPFL_{EN}$ and $DIRE_{HF}$ are fed to the INAND gate as well as AND gate to generate the control signals $CTRL_P$ and $CTRL_N$. The signals $CTRL_P$ and $CTRL_N$ determine which feedback loop of L_P and L_N will be triggered after undershoot/overshoot on V_{OUT} has been detected.

Fig. 2. Proposed push-pull feedback loop trigger (PP-FLT) scheme with voltage peak detector (VPD).

The operation processes of PP-FLT scheme are given in Fig. 3. In the first case [see Fig. 3(a)], under certain disturbances, the output voltage decreases by ΔV_L , resulting in a variation ΔV_{FBH} of feedback voltage V_{FBH} . When the V_{FBH} drops below reference voltage V_{REF} , the output of CMP_0 (denoted as q_0) changes from 1 to 0 while the output of CMP_1 (marked by q_1) remains constant 0 since the V_{FBL} is always below V_{REF} . Once q_0 becomes low, the output of VPD (illustrated as $PPFL_{EN}$) flips from 0 to 1 while signal $DIRE_{HF}$ changes to 0, indicating that undershoot on V_{OUT} has been detected. Then, the signal $CTRL_P$ becomes 0, which will trigger the push pull feedback loop L_P to suppress the undershoot voltage. The process of deriving ΔV_L is shown below.

Both of the V_{FBL} and V_{FBH} come from the load bias resistor ladder consisted of R_1 and R_2 , hence V_{FBH} is a function of V_{OUT} , as given by

$$V_{FBH} = \frac{2R_1 + R_2}{2(R_1 + R_2)} V_{OUT}. \quad (1)$$

Fig. 3. Operation processes of the proposed PP-FLT for (a) undershoot and (b) overshoot voltages detection.

At time t_0 , Eq. (1) is expressed as

$$V_{FBH}|_{t_0} = (V_{FBH@SS} - \Delta V_{FBH}) = \frac{2R_1 + R_2}{2(R_1 + R_2)} (V_{OUT@SS} - \Delta V_L), \quad (2)$$

where $V_{OUT@SS}$ and $V_{FBH@SS}$ denote the values of V_{OUT} and V_{FBH} at beginning steady state, respectively. The $V_{OUT@SS}$ is given by

$$V_{OUT@SS} = 2V_{REF}, \quad (3)$$

while the $V_{FBH@SS}$ is given by

$$V_{FBH@SS} = \left(\frac{2R_1 + R_2}{2(R_1 + R_2)} \right) V_{OUT@SS} = \left(\frac{2R_1 + R_2}{2(R_1 + R_2)} \right) 2V_{REF}, \quad (4)$$

hence $V_{FBH@SS}$ becomes

$$V_{FBH@SS} = k_1(2V_{REF}), \quad (5)$$

where

$$k_1 = \frac{2R_1 + R_2}{2(R_1 + R_2)}. \quad (6)$$

In order to trigger the push-pull feedback loop L_P , the feedback voltage V_{FBH} needs to satisfy the following condition

$$V_{FBH} < V_{REF}. \quad (7)$$

At time t_0 , Eq. (7) becomes

$$V_{FBH}|_{t_0} = \frac{2R_1 + R_2}{2(R_1 + R_2)} (V_{OUT@SS} - \Delta V_L) < V_{REF}. \quad (8)$$

By solving the Eq. (3) and Eq. (8), ΔV_L is calculated as

$$\Delta V_L > \frac{1}{2 + R_2/R_1} (2V_{REF}). \quad (9)$$

The undershoot detection threshold is define as

$$\Delta V_L = \frac{1}{2 + R_2/R_1} (2V_{REF}). \quad (10)$$

On the other hand [see Fig. 3(b)], the output voltage V_{OUT} increases by ΔV_H , leading to a variation ΔV_{FBL} of feedback voltage V_{FBL} . Once the V_{FBL} is higher than V_{REF} , the $CTRL_N$ thus changes from 0 to 1, which enables L_N -turning mode to reduce the overshoot on V_{OUT} . Similarly to k_1 , parameter k_2 shown in Fig. 3(b) as well as Fig. 3(a) is expressed as

$$k_2 = \frac{R_2}{2(R_1 + R_2)}. \quad (11)$$

Analogously to ΔV_L , the overshoot detection threshold ΔV_H is define as

$$\Delta V_H = \frac{R_1}{R_2} (2V_{REF}). \quad (12)$$

Therefore the voltage shoot detection thresholds (marked by ΔV_L , ΔV_H) are proportional to the R_1/R_2 . In this Letter, $V_{REF} = 0.5$ V, $R_1 = 0.1$ K Ω , $R_2 = 4.9$ K Ω and $R_1/R_2 = 1/49$. Thus $\Delta V_L \approx 19.6$ mV, $\Delta V_H \approx 20$ mV.

2.3 Clocked comparator and S/R

As given in Fig. 4(a), the clocked comparator scheme introduced in [5] is used in the proposed D-LDO. A latch following the first stage restores the output signal. The comparator operates with two phases, namely pre-charge phase and regenerative phase. When CLK signal is low (pre-charge phase), the internal nodes are pulled to high while the latch outputs keep unchanged; when CLK rises to high (regenerative phase), the comparator accumulates the difference between V_{INP} and V_{INN} and rapidly amplifies the small difference to full swing, then the output latch will restore the output signal. Table I is the true table of the clocked comparator.

Table I. True table of clocked comparator

	CMP _{OUT}
$V_{INP} < V_{INN}$	“1”
$V_{INP} > V_{INN}$	“0”

Fig. 4(b) shows the scheme of the 160-bit shift register (S/R), which consists of D-type flip-flops and multiplexers. The Q[159:0] are the outputs of S/R, which control the PMOS array PG₁₅₉~PG₀. When DIRE is low, all the bits Q are shifted to the right by one count at the rising edge of the sampling clock CLK and vice versa, indicating one more PMOS transistor to turn on/off.

2.4 Operation principles

The operation principles of the proposed D-LDO are depicted in Fig. 5. In steady state, the D-LDO works only at L_{SR} -turning mode, with a small PMOS transistor step and low frequency clock CK_{LF}. As shown in Fig. 5(a), when the load current (I_L) changes abruptly from 20 to 200 mA, V_{OUT} starts to drop. The S/R feedback

Fig. 4. Schemes of (a) clocked comparator and (b) shift register.

loop (L_{SR}) senses the V_{OUT} drop and generates more logic low control signal among $Q[159:0]$, which will turn on more extra PMOS transistors to suppress V_{OUT} drop. However, since the transition speed of L_{SR} is quite slow, V_{OUT} keep going downward. At time t_0 , when V_{OUT} exceeds the undershoot detection boundary ($2V_{REF} - \Delta V_L$), the output of PP-FLT scheme (i.e. $CTRL_P$) immediately set to logic 0. So, L_P -turning mode is triggered and all of the PMOS transistors (PG_0 to PG_{159}) are turned on to charge the C_L with theatrically strength current. Then V_{OUT} recovers to the level above ($2V_{REF} - \Delta V_L$) quickly. As thus, signal $CTRL_P$ reset to logic 1 disabling L_P -turning mode at time t_1 , which leads V_{OUT} to go downward and exceed ($2V_{REF} - \Delta V_L$) once more at time t_2 . Then, L_P -turning mode is enabled and regulates V_{OUT} to the level above ($2V_{REF} - \Delta V_L$) again. Next, L_P -turning mode is disabled and enabled alternately to limit V_{OUT} drop until enough PMOS transistors are turned on by S/R. Hence the V_{OUT} successfully recovers to common level with greatly small undershoots. Finally, S/R keeps regulating V_{OUT} closer to the desired value and makes the D-LDO enter steady state.

Fig. 5(b) shows the case of load current decreasing abruptly from 200 to 20 mA. Once V_{OUT} exceeds the overshoot detection boundary ($2V_{REF} + \Delta V_H$), the PP-FLT scheme triggers the L_N -turning mode by setting $CTRL_N$ to 1, which turns on the NMOS transistor N_G to discharge the C_L with strength current, hence, V_{OUT} immediately drops to below level of ($2V_{REF} + \Delta V_H$). To limit the overshoots on V_{OUT} , the L_N -turning mode toggles between on and off alternatively depending on whether V_{OUT} exceeds ($2V_{REF} + \Delta V_H$) or not. Simultaneously, L_{SR} -turning mode is available and S/R generates more logic high control signal among $Q[159:0]$ to turn off extra PMOS transistors. After most of the extra PMOS transistors have been turned off, L_N -turning mode is not need to be triggered any more. Finally, D-LDO continues to work at L_{SR} -turning mode which adjusts V_{OUT} closer to $2V_{REF}$ and steady state is hence achieved.

Fig. 5. Transient responses of proposed D-LDO with (a) positive and (b) negative load current step.

These operations prevent large spikes on V_{OUT} and significantly improve transition response.

3 Steady analyses

Fig. 6 is the illustration of unsteady push-pull feedback loop control. The region from t_0 to t_1 depicts that all PMOS transistors are turned on at time t_0 to charge C_L with charged current I_{PGMAX} and pull V_{OUT} up. However, since either the C_L is too small or I_{PGMAX} is too strong, V_{OUT} quickly goes upward to $(2V_{REF} + \Delta V_H)$ within a period of CK_{HF} . At time t_1 , the D-LDO directly toggles from L_P -turning mode to L_N -turning mode without entering steady state. Hence, with L_P releasing control of PMOS array, NMOS transistor N_G is turned on immediately by feedback loop L_N to discharge C_L with discharged current I_{NGMAX} , leading to V_{OUT} goes downward. Unfortunately, the V_{OUT} unexpectedly drops to $(2V_{REF} - \Delta V_L)$ at next clock rise edge (namely at time t_2) leading to L_P -turning mode be enable again. As thus, C_L is charged and discharged alternatively within adjacent clock period of CK_{HF} . V_{OUT}

thus oscillates between $(2V_{REF} - \Delta V_L)$ and $(2V_{REF} + \Delta V_H)$ with a period of $2T_{HF}$, resulting in unsteady problems. To avoid that unwanted unsteady state, the variation of charged and discharged voltage (ΔV_{CL}) within a period (T_{HF}) of CK_{HF} should be smaller than $(\Delta V_L + \Delta V_H)$. Hence the steady conditions of push-pull feedback loop control are given as follows

$$\frac{I_{PGMAX}T_{HF}}{C_L} \leq (\Delta V_L + \Delta V_H) \quad (13)$$

$$\frac{I_{NGMAX}T_{HF}}{C_L} \leq (\Delta V_L + \Delta V_H), \quad (14)$$

where I_{PGMAX} and I_{NGMAX} are maximum current available to charge and discharge the C_L , respectively. The value of C_L can be given by solving Eq. (13) and Eq. (14).

$$C_L \geq \max \left\{ \left(\frac{I_{PGMAX}T_{HF}}{\Delta V_L + \Delta V_H} \right), \left(\frac{I_{NGMAX}T_{HF}}{\Delta V_L + \Delta V_H} \right) \right\}. \quad (15)$$

In the proposed D-LDO, the I_{PGMAX} and I_{NGMAX} are approximately 200 mA, T_{HF} is 1 ns, and $(\Delta V_L + \Delta V_H)$ is approximately 40 mV. Thus C_L becomes as follows

$$C_L \geq 5 \text{ nF}. \quad (16)$$

Fig. 6. Illustration of unsteady push-pull feedback loop control.

The small signal model of S/R feedback loop L_{SR} in the proposed D-LDO is shown in Fig. 7(a) which is made up of a sampled comparator CMP_2 , a shift register (S/R), a zero-order hold (ZOH), and a first-order plant representing the continuous time plant consisting of the PMOS transistor array and the RC load. A feedback coefficient $\beta = 1/2$ has been inserted into the feedback loop because the feedback voltage V_{FB} is generated by the load bias resistor ladder and $V_{FB} = 1/2V_{OUT}$. The open-loop transfer function $G(z)$ is written as Eq. (17) according to [13]

$$G(z) = K_{DC}(1 - e^{-\alpha T}) \frac{z}{(z - 1)(z - e^{-\alpha T})}, \quad (17)$$

where K_{DC} is the loop DC gain, α is the time constant of the output pole, and T is the sampling time interval of S/R. In the synchronous S/R feedback loop, sampling time interval T equals to the period of CK_{LF} which is 20 μ s. Since the feedback coefficient $\beta = 1/2$, the overall closed loop transfer function of the digital LDO in the z -domain can be written as Eq. (18)

$$H(z) = G(z) / \left(1 + \frac{1}{2} G(z) \right). \quad (18)$$

Eq. (18) provides insight into the stability of the proposed D-LDO. The root loci of this system can be plotted in Fig. 7(b). All of the poles in z -domain are within the unit circle to guarantee the steady of the proposed D-LDO system under the heavy and light load conditions.

Fig. 7. (a) Small signal model and (b) root loci of proposed D-LDO under the conditions of heavy load ($I_L = 200$ mA) and light load ($I_L = 20$ mA).

4 Simulated results

The proposed D-LDO has been designed and simulated in the SMIC 65 nm process with 1.2 V supply and 1.0 V output voltage. The HSPICE is chosen as the SPICE simulator. According to the transistor level simulation results shown in Fig. 8, the output overshoot and undershoot voltages are 27 and 26 mV, respectively, with load steps of 20 to 200 mA with a 10-ns edge time, while they are 183 and 1363 mV, respectively, when the push-pull feedback loop is not engaged. On the other hand, with the push-pull feedback loop control, the steady voltage ripples are not degraded, which are simulated to be 15 and 7 mV under 20 and 200 mA load current, respectively. And the setting time is approximately 1.25 ms. In addition, the quiescent current is only 400 μ A with maximum load current up to 200 mA. When $I_L = 200$ mA, the current peak efficiency is 99.8%. The output capacitance is 1 μ F.

With the help of Layout XL in Virtuoso, the estimated layout area is around 0.42 mm² except for output capacitor.

The proposed D-LDO is compared with prior arts, as shown in Table II. The output voltage variation (ΔV_{OUT}) enhancement over previous work is notable. In order to compare the key performances of the proposed D-LDO with other state-of-art designs, the figure-of-merit for response time (FOM_{RESP}) presented in [14] is employed. A larger FOM_{RESP} implies faster response performance.

Fig. 8. Simulated load transient response.

Table II. Comparison of performance

	[2]	[9]	This work
Technology (nm)	130	180	65
Active area (mm ²)	N/A	0.81	≈0.42
Input voltage (V)	1.0	0.9–1.8	1.2
Output voltage (V)	0.8	0.8–1.5	1.0
Maximum load current (mA)	100	200	200
Output Capacitor	0–100 pF	1 μF	1 μF
Quiescent current (μA)	1.8–67	750	400
Peak current eff. (%)	99.93 ^(a)	99.6	99.8
Edge time (ns)	300	500	10
Load step (mA)	0.1–100	1–100	20–200
ΔV _{OUT} (mV)	80	70	27
FOM _{RESP} ^(c)	24.9	1.62	32.7

(a) Calculated by using maximum load and quiescent current.

(c) $FOM_{RESP} = I_{STEP} / (I_{MIN} \times \Delta V_{OUT} \times T_{EDGE})$ is shown in [14].

5 Conclusions

This paper presents a novel D-LDO with voltage peak detecting and push-pull feedback loop operation. The proposed D-LDO operates at 1.2 V supply and 65 nm technology. The simulation results show that the proposed D-LDO's transient response performance has been significantly improved owing to the proposed voltage peak detecting technique and push-pull feedback loop control. In addition, the proposed D-LDO achieves maximum load capability up to 200 mA.

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