CMOS low dropout regulator with dynamic zero compensation

C.-L. Chen, W.-J. Huang and S.-I. Liu

The output capacitor and its equivalent series resistance (ESR) often limit the stability of a conventional low dropout regulator (LDR). A CMOS LDR with dynamic zero compensation is presented to tolerate the wide range of the output capacitor and the ESR. The stability constraints for the output capacitor with the ESR are derived. The measured LDR is stable for the output capacitor 2nF–47μF with ESR of 0.1–50 Ω. The maximum quiescent current of this LDR with a bandgap reference is 43.2μA and its maximum output current is 150 mA for the output voltage of 1.8V.

Introduction: Low dropout regulators (LDRs) [1–4] are widely used in power management integrated circuits and systems. A conventional LDR needs a large output capacitor to reduce output ripples. However, the output capacitor and its equivalent series resistance (ESR) often affect the stability of the LDR. In [4], it is indicated that the allowable range of ESR is limited, such as 0.2–9 Ω, for a typical LDR. To extend the allowable range of the output capacitor and its ESR, an LDR with dynamic zero compensation is presented.

Circuit description: The proposed LDR with dynamic zero compensation is shown in Fig. 1 (the bandgap voltage reference is not shown). It is composed of three gain stages, a current follower [3, 5, 6] with a Miller compensation capacitor C_a , and a feedback resistor network. The dynamic zero compensation is realised by a PMOS transistor, M_z , and a capacitor, C_z . It introduces a dynamic low-frequency zero, which is dependent on the output current, I_{out} . As the output current is zero, both the power PMOS transistor, M_p , and M_z are turned off, and this zero is not active. As the output current is increased, M_p and M_z are turned on. Then, the low-frequency zero is realised to enable the pole-zero elimination. It extends the useful range of the output capacitor and the ESR for the proposed LDR.

Fig. 1 Proposed LDR

To understand the design issues, the following analysis is derived. g_{m1} , g_{m2} and g_{mp} are the transconductances of the first, second and the output stage, respectively. R_1 , R_2 and R_p are output resistances of the first, second and the output stage, respectively. C_1 and C_2 are output parasitic capacitance of the first and second stage, respectively. g_{ma} and g_{mf} are the transconductances of the PMOS transistor, M_a , and the NMOS one, M_f , respectively. R_z is the equivalent source-to-drain resistance of the transistor M_z . C_z and C_a are the compensation capacitors. C_{out} is the output capacitor with the associated ESR, R_{esr} . R_{out} is the total equivalent output resistance, which is equal to $R_L // R_p // (R_{F1} + R_{F2})$, where R_L is the load resistor, and R_{F1} and R_{F2} are the feedback resistors.

Assume $C_{out} \gg (C_z, C_a, C_2) \gg C_1$, $g_{mp} \gg (g_{m1}, g_{m2}, g_{ma}, g_{mf})$, and $(g_{m1}R_1, g_{m2}R_2, g_{mp}R_{out}) \gg 1$. Based on the above assumptions, the loop transfer function of the proposed LDR in the light load (i.e. I_{out} is small) is given as

$$L_l(s) \simeq \frac{-A_{dc}(1 + sC_{out}R_{esr})(1 + s(g_{mf}C_1/g_{m1}g_{m2}))}{(1 + (s/P_{1_l}))(1 + s((C_aR_1g_{m2}R_2g_{mp}R_{esr} + C_2R_2 + C_1R_1)/(C_a/C_{out})(R_1g_{m2}R_2g_{mp} + 1)) + s^2((C_1R_1C_2R_2)/(C_a/C_{out})R_1g_{m2}R_2g_{mp} + 1))} \quad (1)$$

where $A_{dc} = (R_{F2}/(R_{F1} + R_{F2})) g_{m1}R_1g_{m2}R_2g_{mp}R_{out}$ is the DC loop gain and $p_{1_l} = 1/(C_aR_1g_{m2}R_2g_{mp}R_{out} + C_{out}R_{out})$ is the dominant pole of the LDR in the light load.

The unity-gain frequency is approximated as

$$UGF_1 = \frac{R_{F2}}{R_{F1} + R_{F2}} \cdot \frac{g_{m1}}{C_a} \cdot \frac{1}{1 + (C_{out}/C_aR_1g_{m2}R_2g_{mp})} \quad (2)$$

For the light load, two zeros, $1/R_{esr}C_{out}$ and $g_{m1}g_{m2}/g_{mf}C_1$, in (1) are much higher than the unity-gain frequency of (2) and their effects are neglected. From the second-order polynomial in the denominator of (1), the natural frequency and the damping factor are calculated as

$$\omega_n = \sqrt{\frac{C_ag_{m2}g_{mp} + (C_{out}/R_1R_2)}{C_{out}C_1C_2}} \quad \text{and} \quad \zeta = \frac{1}{2} \frac{C_ag_{m2}g_{mp}R_{esr} + (C_2/R_1) + (C_1/R_2)}{\sqrt{C_1C_2((C_a/C_{out})g_{m2}g_{mp} + (1/R_1R_2))}} \quad (3)$$

To guarantee the stability in the light load, the non-dominant poles should be higher than the unity-gain frequency, and the damping factor should be designed appropriately to ensure that the ESR and output capacitor do not affect the phase margin. From (2) and (3), the following constraint should be satisfied.

$$C_{out} > \frac{1}{2} C_ag_{m2}g_{mp}R_1R_2 \left[-1 + \sqrt{1 + 4 \left(\frac{g_{m1}}{C_a} \right)^2 \left(\frac{R_{F2}}{R_{F1} + R_{F2}} \right)^2 R_1R_2C_1C_2} \right] \quad (4)$$

Equation (4) indicates that there is a lower bound of the output capacitor in this proposed LDR. In this case, the output capacitor is designed to be around several nF.

When the LDR is in the heavy load (i.e. I_{out} is large), the loop transfer function is

$$L_h(s) \simeq \frac{-A_{dc}(1 + sC_zR_z)(1 + sC_{out}R_{esr})}{(1 + s(C_a/g_{ma}))(1 + s(g_{mf}C_1/g_{m1}g_{m2}))} \cdot \frac{1}{(1 + (s/P_{1_h}))(1 + (s/(g_{ma}g_{m2}g_{mp}R_{esr}(R_1//R_z)/C_2) + (g_{ma}/C_a))(1 + sC_1 + R_z)[1 + s(C_zR_z + C_{out}R_{esr}) + s^2(C_zR_zC_{out}R_{esr} + ((C_zR_z + C_2R_1)C_2C_{out}/C_aR_1g_{m2}g_{mp}))])} \quad (5)$$

Five poles and four zeros are given in (5). The dominant pole is $p_{1_h} = 1/C_aR_1g_{m2}R_2g_{mp}R_{out}$ and the unity-gain frequency in the heavy load is $UGF_h = R_{F2}g_{m1}/(R_{F1} + R_{F2})C_a$. For the heavy load, two zeros, g_{ma}/C_a and $g_{m1}g_{m2}/g_{mf}C_1$, and two non-dominant poles, $((g_{ma}g_{m2}g_{mp}R_{esr}(R_1//R_z)/C_2) + (g_{ma}/C_a))$ and $1/R_zC_1$, in (5) are much higher than the unity-gain frequency of UGF_h , and their effects are neglected.

Assume R_{esr} is large and the following equation is satisfied:

$$R_{esr} \gg \frac{C_2}{C_a(R_1//R_z)g_{m2}g_{mp}} \quad (6)$$

According to (6), the second-order polynomial in the denominator of (5) can be decomposed as $(1 + sC_zR_z)(1 + sC_{out}R_{esr})$. Then, two pole-zero pairs can be eliminated. Therefore, the stability of this LDR in the heavy load is ensured when R_{esr} is large.

When R_{esr} is small, one of the poles in the second-order polynomial in the denominator of (5) is shifted to a lower frequency. This pole is expressed as

$$P_{nondom_h_lowESR} = \frac{1}{C_{out}R_{esr} + (C_2C_{out}/C_ag_{m2}g_{mp}(R_1//R_z))} \quad (7)$$

To ensure stability, this non-dominant pole must be higher than the unity-gain frequency. It results in the following constraint as

$$C_{out} < \frac{R_{F1} + R_{F2}}{R_{F2}} \cdot \frac{C_a}{g_{m1}} \cdot \frac{1}{R_{esr} + (C_2/C_a g_{m2} g_{mp} (R_1//R_z))} \quad (8)$$

Equation (8) shows the upper bound of the output capacitor. Since this LDR only has a dominant pole within the unity-gain frequency, stability is guaranteed in the heavy load when R_{esr} is small. Based on (4) and (8), the allowable range of the output capacitor and the ESR for the LDR will be extended.

To verify the stability of the proposed LDR, Fig. 2 shows its root-locus plot with varying output currents. When the constraints in (4) and (8) are satisfied, the solid line shows that all the poles of the proposed LDR lie in the left-hand plane with the output current of $I_{out} = 0$ mA– $I_{out} = 150$ mA. When the constraints in (4) and (8) are not satisfied, the dotted line shows the complex poles move into the right-hand plane with the output current of $I_{out} = 0.3$ mA– $I_{out} = 30$ mA. It results in an unstable system as the output current varied.

Fig. 2 Root-locus of proposed LDR

Arrows indicate the direction, which the output current increases
Solid line: (4) and (8) satisfied; dotted line: (4) and (8) not satisfied

Experimental results: To verify the above analysis, the proposed LDR with a bandgap reference has been fabricated in a 0.35 μ m CMOS process.

For the proposed LDR, the measured line regulations at the output current $I_{out} = 0$ and $I_{out} = 150$ mA are 0.134 and 0.100%/V, respectively. The measured maximum load regulation is 92.54 ppm/mA at $V_{in} = 3.3$ V. Fig. 3 shows the measured transient response of the LDR at the supply of 3.3 V under the output current switching from 0 to 20, 50, and 150 mA. Fig. 3a shows the result for $C_{out} = 47$ μ F with $R_{esr} = 50$ Ω and Fig. 3b shows the result for $C_{out} = 2$ nF with $R_{esr} = 0.1$ Ω . It shows that this LDO can recover the output voltage within 20 μ s within the error of 0.5% for switching the output current of 150 mA.

Fig. 3 Transient response of LDR for $C_{out} = 47$ μ F and $R_{esr} = 50$ Ω

(Fig. 3a) $C_{out} = 2$ nF and $R_{esr} = 0.1$ Ω (Fig. 3b) V_{out} : 500 mV/div., 10 μ s/div.

Conclusions: A CMOS LDR with a dynamic zero compensation is presented. The experimental results confirm that this LDR with dynamic zero compensation can be stable for a wide range of the output capacitor and the ESR.

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