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Folded-cascode amplifier with efficient feedforward gain-boosting

J. Raman, P. Rombouts and L. Weyten

Abstract In this letter, an efficient implementation of a two-stage feedforward amplifier is proposed. The new amplifier topology boosts the DC gain of a folded-cascode amplifier by adding a gain path which makes use of the "free" transconductance of the folding current sources. Only a single differential pair is used to convert the input voltage in a current. A current splitter extracts the parts needed for feedforward and for the first gain stage. This turns out to be favorable in terms of noise when compared to more traditional feedforward amplifiers.

Introduction Many applications require a higher open-loop gain than a basic telescopic- or folded-cascode amplifier (FCA) can provide. In such situations, regulated (active) cascodes are often applied. Alternatively, one can turn to cascading, resulting in a second-order system which needs compensation. Miller and Ahuja compensation are two commonly applied techniques. Feedforward compensation is another approach, which tries to overcome some of the limitations of Miller compensation schemes [1, 2]. In literature, some examples can be found where both cascoding and cascading with feedforward compensation can be found (e.g., [1]). We present here an alternative feedforward-based gain-boosting technique which integrates efficiently with a folded-cascode amplifier.

Amplifier topology The proposed amplifier topology is shown in fig. 1. One easily recognizes in the output branch the basic components of the FCA: the current source T_1 cascoded by T_3, and the folding current source T_2 cascoded by T_4. Also, the input differential pair formed by $T_{in\pm}$ is readily identified, together with its tail current source T_0. The transconductance of this differential pair will be denoted as $g_{m,in}$. The biasing voltages are denoted V_{b0} up to V_{b6}. The cascode pair T_5/T_6 acts as a current splitter. Its purpose is to split the current of the input differential pair in two parts: $I_A = \alpha g_{m,in} V_{in}$ and $I_{ff} = (1-\alpha) g_{m,in} V_{in}$. Later on we will use the gate voltage of T_5 for common-mode feedback, but for the moment we can assume that $V_C = V_{b6}$. The fraction α of the current splitter depends on the relative sizing of T_5 and T_6, and can therefore be considered constant and well controlled. The current I_{ff} flows to the folding node (the extra current buffer T_8 will be discussed later). This signal path will play the role of the feedforward path. The current I_A is used in the intermediate gain stage. The output voltage V_A of this intermediate stage is connected to the folding current source T_2. In order to assure that the feedforward path and the gain path combine in the correct way, an extra sign inversion is needed to undo the inverting characteristic of the intermediate gain stage. In the proposed fully differential amplifier, this is solved by cross-coupling the connections.

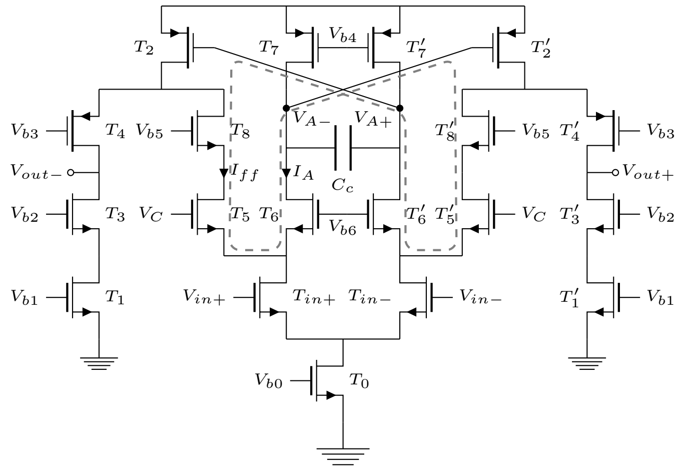

Figure 1: Proposed folded-cascode amplifier with feedforward gain-boosting.

Because the proposed topology is of second-order, we study the dominant poles. The differential small signal scheme of the amplifier is equivalent to fig. 2. In this, $g_{mA} = \alpha g_{m,in}$, $g_{m,ff} = (1-\alpha) g_{m,in}$ and $g_{mB} = g_{m2}$. Note that by splitting the folding current source T_2 in a constant part and a controlled part (not shown on the figure), g_{mB} can actually be designed lower than g_{m2}. This creates some additional design freedom. We have $r_{oA} \approx r_{o6} \| r_{o7}$. The output resistance of the amplifier amounts to $r_{oB} \approx (g_{m3} r_{o3}) r_{o1} \| (g_{m4} r_{o4}) r_{o2}$. This is typically higher than the standard FCA, because the double-cascode $T_5 + T_8$ neutralizes the finite output resistance of T_{in} (which can be substantial if T_{in} has minimal length). An important advantage of the proposed approach is that transconductance g_{mB} is realized by re-using (part of) the folding current, hence there is no extra current consumption. The capacitance C_A at node V_A is in many cases dominated by the gate-source capacitance C_{GS2} and explicit compensation capacitance added at this point (e.g., C_c in fig. 1). Capacitance C_B will in most cases be dominated by the load capacitance. It is elementary to include other parasitic capacitances.

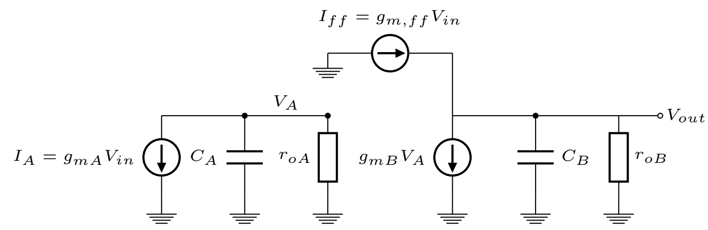

Figure 2: Small-signal model of a second-order feedforward amplifier.

Noise advantage The main reason for using a single differential pair and a current splitter instead of separate differential pairs (as is the case in a more traditional feedforward topology) is the improved noise performance. To illustrate this, we consider the white noise contribution of the input differential pair(s) at frequencies up to the open-loop bandwidth of the amplifier. In this bandwidth, the second-order gain path is dominant, and the feedforward path can be neglected. The current-noise generated in the input MOSFETs propagates to the current splitter, and introduces a current-noise component i_A in I_A of fig. 1. We have the following spectral density: $\overline{di^2_{A,split}}/df = \alpha^2 4kT\frac{2}{3}g_{m,in}$. In contrast, when using separate differential pairs, the contribution of the gain-dominant path becomes (assuming the same transconductances are used): $\overline{di^2_{A,sep}}/df = 4kT\frac{2}{3}g_{mA} = 4kT\frac{2}{3}\alpha g_{m,in}$. We deduce that current-splitting gives a noise improvement of $10\log_{10}(\alpha)$ dB at low frequencies. Likewise, we can compare the situation at higher frequencies, when the feedforward path becomes dominant. Here, we obtain a noise improvement of $10\log_{10}(1-\alpha)$ dB. For $\alpha = 1/2$, an overall improvement of a factor two (3dB) is established.

Signal range The signal range available at the output of the amplifier is completely determined by the FCA topology. However, compared to a standard FCA, there are more strict requirements on the input common-mode level ($V_{CM,in}$). For this reason, the amplifier is best used in an inverting feedback configuration, with predetermined $V_{CM,in}$. One derives the condition

$$V_{Tn}+V_{ds,in}+V_{ds0} \leq V_{CM,in} \leq V_{DD}+V_{Tn}-V_{ds2}-V_{ds8}-V_{ds5} \quad (1)$$

to guarantee that there is enough voltage headroom for T_5 and T_8. Also, a minimum supply voltage V_{DD} consisting of one threshold and four saturation voltages is required:

$$V_{DD} \geq V_{Tp} + V_{ds0} + V_{ds,in} + V_{ds5} + V_{ds2} \quad (2)$$

This expression can be derived by stating that the voltage at the V_A should be high enough to let T_6 operate properly. In most cases, we can safely assume $V_{Tp} > V_{ds8}$, and the latter condition on the supply voltage automatically implies that the range (1) is nonempty. Actually, this expresses in mathematical terms what can be observed at the circuit level: because the voltage at the V_A node should be able to get low enough to operate T_2, and because $V_{ds5} = V_{ds6}$, we can see that there is plenty of voltage headroom (a full threshold voltage V_{Tp}) to insert the extra cascode transistor T_8 in the feedforward path. Why these extra current buffers are needed is explained in the next section.

Parasitic feedback loop As a side-effect of cross-coupling the connection from $V_{A\pm}$ to T_2/T_2', a parasitic local feedback loop is created (see fig. 1). To examine this loop, we set $V_{in+} = V_{in-}$ and break the loop at the gate of T_2. A change of voltage δ_V at this gate creates a current $g_{m2}\delta_V$. The input resistance of T_4 is approximately $r_{in4} \approx \frac{g_{m3}r_{o3}}{g_{m4}r_{o4}}r_{o1}$, a value much higher than $\frac{1}{g_{m4}}$ due to the high loading resistance (T_1 cascoded by T_3). When aiming for maximum gain in the FCA, we have $(g_{m3}r_{o3})r_{o1} = (g_{m4}r_{o4})r_{o2}$, from which we derive the simpler result $r_{in4} \approx r_{o2}$. The current $g_{m2}\delta_V$ splits into three parts: a current through r_{o2}, one through $r_{in4} \approx r_{o2}$, and one going through the output resistance of the $T_5 - T_8$ branch, which we will denote further by r_{out}. The latter current is almost completely collected by T_6, which acts as a cascode transistor with low input impedance. The DC gain from the gate of T_2 to V_{A-} is then $A_x = -g_{m2}\frac{r_{o2}}{r_{o2}+2r_{out}}(r_{o6}\|r_{o7})$. Since the transfer from V_{A+} to V_{A-} is exactly the same, we clearly have positive feedback, and the DC loop gain is equal to A_x^2. A necessary condition for stability is now $|A_x| < 1$, or equivalently

$$r_{out} > g_{m2}r_{o2}\frac{r_{o6}\|r_{o7}}{2} \quad (3)$$

Apparently, we need a high output resistance r_{out} for the feedforward current-path, which explains why the extra cascodes T_8/T_8' are needed. The condition (3) is in most cases also sufficient for stability. Indeed, for increasing frequencies, the input impedance of T_4 quickly drops from r_{o2} to a much lower value (of the order $1/g_{m4}$). In this case, virtually no current flows through r_{out}, and the loop gain of the parasitic positive feedback loop drops to zero.

The above analysis of the parasitic loop applies to the amplifier in open loop. This again turns out to be the worst case situation. In practice, negative feedback will be applied to the amplifier. In many cases, the global negative feedback path overpowers the parasitic positive feedback path, and good closed-loop stability is obtained (even without the extra cascodes). These conclusions are supported by SPICE simulations on an actual design, where both open-loop and closed-loop simulations (with capacitive feedback) have been evaluated.

Common-mode feedback A very convenient way to control the output common-mode of the amplifier is by changing the gate voltage of T_5/T_5' according to $V_C = V_{b6} + \frac{V_{out+} - V_{CMref}}{2} + \frac{V_{out-} - V_{CMref}}{2}$. This can for instance be done in the usual way by a standard switched-capacitor circuit. The voltage V_C controls the current-split factor α in such a ways as to make $\alpha\frac{I_{D0}}{2} = I_{D7}$. This fixes α as a ratio of well-defined currents. Also the common-mode voltage level of V_A/V_A' is well defined, because this corresponds to the require gate-source voltage of T_2/T_2' to make $I_{D2} = (1-\alpha)\frac{I_{D0}}{2} + I_{D1}$.

Conclusion We have described a new efficient amplifier topology which boosts the DC gain of an FCA by means of cascading and feedforward compensation. By using a single input differential pair in combination with a current-splitter, we have obtained a better noise performance compared to a more traditional amplifier with feedforward compensation. Also, the transconductance of the folding current sources has been used to realize the transconductance g_{mB} of fig. 2 without extra current consumption.

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