A Wideband IM3 Cancellation Technique using Negative Impedance for LNAs with Cascode Topology

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Abstract — A negative impedance is used to enable distortion cancellation between the transconductor and the cascode transistor for LNAs with a cascode topology. As a proof of concept, a resistive feedback LNA using this IM3 cancellation technique in a standard 0.16 μm CMOS process shows that for 0.1GHz to 1GHz, improvements of 6.3dB to 10dB for IIP3 and 0.2dB to 1dB for gain are achieved without noise degradation. The power consumption for the LNA is increased by 2%, and the die area by about 700 μm^2 .

Index Terms — Low-noise amplifier (LNA), intermodulation distortion, output conductance nonlinearity, cascode transistor, IM3 cancellation, IIP3, CMOS.

I. INTRODUCTION

Much effort has been put in improving the linearity of LNAs. As a commonly used technique, an auxiliary path replicating the distortion of the main path is combined with the main LNA in a subtracting node. The auxiliary path can be transistors biased in weak inversion [1-2], transistors biased in saturation region as a nonlinear resistor [3] or current sources that injects IM2 to suppress IM3 [4]. The limitations of most reported linearization techniques mainly are due to:

- 1) only focusing on transconductance nonlinearity, neglecting nonlinearity related to output conductance [5].
- 2) neglecting the distortion generated by the cascode transistor.

Neglecting output conductance related distortion and neglecting the distortion contribution of cascode transistors is valid for older technologies with long-channel devices and high supply voltage. However, in deep submicron CMOS technologies, typically the output resistance of the transconductor stage is relatively low: the IM3 distortion contribution from the cascode transistor then may become dominant. On top of that, the low supply voltage together with high gain operation tends to push the cascode transistor out of the deep saturation region. This results in a very significant increase of the third-order output conductance nonlinearity term and the cross-modulation nonlinearities, which causes the increase of distortion generated by the cascode transistor [7].

In this paper we present a wideband IM3 cancellation technique that takes into account the distortion of the cascode transistor and all the third-order nonlinearity contributions related to the transconductance and the output conductance. A negative impedance is used to enable cancellation between the distortion current of the transconductor and the cascode transistor. Without loss of generality, we apply this IM3 cancellation technique to a resistive feedback wideband LNA. Section II presents an analysis to explain the proposed IM3 cancellation and presents a short discussion on the effect on gain and noise. Section III shows both simulation and measurement results to verify the theory.

II. THEORY OF IM3 CANCELLATION USING NEGATIVE IMPEDANCE

In the resistive feedback LNA shown in Fig. 1, M_{1a}/M_{1b} is the transconductor stage while the cascode transistor M_{2a}/M_{2b} increases the output impedance and improves the isolation between input and output. The shunt feedback resistor R_f is used to match to the source resistance R_s .

To improve the linearity we apply a negative impedance ($Y_{neg} = G_{neg} + j\omega C_{neg}$) between the drain of M_{1a}/M_{1b} as shown in Fig. 2. The IM3 distortion current of transistor M_x is due to the voltage swings at the transistor terminals, where we neglect the effect of a source-bulk signal for simplicity reasons only. This results in a 3rd order current $i_d^{Mx} = g_{m3}v_{gs}^3 + g_{x21}v_{gs}^2v_{ds} + g_{x12}v_{gs}v_{ds}^2 + g_{d3}v_{ds}^3$, for

Fig. 1. Schematic of a conventional differential resistive feedback LNA.

$$v_{out}^{IM3} = -\frac{2(R_f + R_s)R_L}{D_v} [g_m^{M_2} i_d^{M_1} + (Y_{ds}^{M_1} + 2Y_{neg}) i_d^{M_2}] \quad (1)$$

$$\text{where } D_v = g_m^{M_2}(R_f + R_L + R_s + g_m^{M_1}R_LR_s) + (R_f + R_L + R_s)(Y_{ds}^{M_1} + 2Y_{neg})$$

$$i_d^{M_1} = f_d^{M_1}(v_{gs}^{M_{1a}}, v_{ds}^{M_{1a}}) = f_d^{M_1} \left[\frac{(R_f + R_L)(g_m^{M_2} + Y_{ds}^{M_1} + 2Y_{neg})v_{in}}{2D_v}, \frac{-g_m^{M_1}(R_f + R_L)v_{in}}{2D_v} \right] \quad (2)$$

$$i_d^{M_2} = f_d^{M_2}(v_{gs}^{M_{2a}}, v_{ds}^{M_{2a}}) = f_d^{M_2} \left[\frac{g_m^{M_1}(R_f + R_L)v_{in}}{2D_v}, \frac{g_m^{M_1}(R_f + R_L - g_m^{M_2}R_fR_L) + R_L(g_m^{M_2} + Y_{ds}^{M_1} + 2Y_{neg})}{2D_v} \right] \quad (3)$$

$$G_{neg} \in \left[\frac{-g_{ds}^{M_1} - g_m^{M_2}}{2}, \frac{-g_{ds}^{M_1}}{2} \right], C_{neg} \approx -0.5C_{ds}^{M_1} \quad (4)$$

$$A_v = \frac{R_L[g_m^{M_2}(1 - g_m^{M_1}R_f) + Y_{ds}^{M_1} + 2Y_{neg}]}{D_v} \approx \frac{R_L(-g_m^{M_1}g_m^{M_2}R_f + Y_{ds}^{M_1} + 2Y_{neg})}{D_v} \quad (5)$$

$$NF = 1 + \frac{[g_m^{M_2}R_L(R_f + R_s)]^2 \gamma g_m^{M_1}}{R_s D_n^2} + \frac{[R_L(g_m^{M_2} + g_m^{M_1}g_m^{M_2}R_s + Y_{ds}^{M_1} + 2Y_{neg})]^2 R_f}{R_s D_n^2} + \frac{[(R_f + R_s)(g_m^{M_2} + Y_{ds}^{M_1} + 2Y_{neg})]^2 R_L}{R_s D_n^2} \quad (6)$$

$$+ \frac{[R_L(R_f + R_s)(Y_{ds}^{M_1} + 2Y_{neg})]^2 \gamma g_m^{M_2}}{R_s D_n^2} + \frac{[g_m^{M_2}R_L(R_f + R_s)]^2 |G_{neg}|}{R_s D_n^2}$$

$$\text{where } D_n = g_m^{M_2}(1 - g_m^{M_1}R_f) + Y_{ds}^{M_1} + 2Y_{neg} \approx -g_m^{M_1}g_m^{M_2}R_f + Y_{ds}^{M_1} + 2Y_{neg}$$

Fig. 2. Circuit model for the distortion and gain analysis.

which a short hand notation $i_d^{M_x} = f_d^{M_x}(v_{gs}^{M_x}, v_{ds}^{M_x})$ will be used in the remainder of this paper. The exact function that describes the transistor nonlinearity is not relevant in this paper. Given the differential circuit topology we assume $i_d^{M_{1a}} = -i_d^{M_{1b}} = i_d^{M_1}$ and $i_d^{M_{2a}} = -i_d^{M_{2b}} = i_d^{M_2}$.

To determine the LNA's IM3, a symbolic harmonic balance like analysis with truncated terms was carried out, yielding the closed form solution for the IM3 shown in (1). In (1) the 3rd order distortion currents $i_d^{M_1}$ and $i_d^{M_2}$ are according to the previously introduced $i_d^{M_x} = f_d^{M_x}(v_{gs}^{M_x}, v_{ds}^{M_x})$, which in this circuit expands to (2-3). It now follows from (2-3) that for $Y_{neg} > -0.5(g_m^{M_2} + Y_{ds}^{M_1})$ the polarity of the voltage swings for M_{1a}/M_{1b} and M_{2a}/M_{2b} are unchanged, and hence so is the phase of the IM3 currents of the transistors. At the same time, (1) shows that for $Y_{neg} < -0.5Y_{ds}^{M_1}$ the distortion components of the transistors are subtracted. Combining these results yields a region defined by (4) in which IM3 contributions of M_1 and M_2 (partially or fully) cancel. Note that doing so, the distortion of the transconductor transistors is canceled using the distortion contribution of the cascode transistors.

The small signal voltage gain can be derived from the model in Fig. 2, resulting in (5). Since Y_{neg} increases the overall output impedance of the transconductor M_{1a}/M_{1b} , a higher gain is resulted. The noise figure (NF) is calculated by the model shown in Fig. 3, where we include the thermal noise current of M_{1a}/M_{1b} , M_{2a}/M_{2b} and Y_{neg} ($i_{th}^{Y_{neg}} = 4kT|G_{neg}|$) and the thermal noise voltage of R_f , R_s and R_L . The calculated NF is given by (6), where the five terms account for the thermal noise from respectively M_{1a}/M_{1b} , R_f , R_L , M_{2a}/M_{2b} and Y_{neg} . Compared to the situation without an Y_{neg} , when the IM3 cancellation condition (4) is met, the denominator D_n^2 is increased and the nominator of the second term and the third term are decreased. This results in decreases of NF contribution from M_{1a}/M_{1b} , R_f and R_L . Since Y_{neg} increases the overall output impedance of the transconductor M_{1a}/M_{1b} , less noise contribution can come from the cascode transistor M_{2a}/M_{2b} . Therefore, it can be concluded that although a Y_{neg} circuit introduces extra noise (last term in (6)), it also reduces the NF contribution from M_{1a}/M_{1b} , M_{2a}/M_{2b} , R_f and R_L . As a result, the effect on NF by a Y_{neg} circuit can be small.

Fig. 3. Circuit model for calculating noise.

III. LNA DESIGN AND EXPERIMENTAL RESULTS

To prove this IM3 cancellation concept, the LNA in Fig. 4 is implemented in a standard 0.16 μ m CMOS process. The negative impedance is implemented by the cross-coupled pair M_{4a}/M_{4b} with source degeneration provided by capacitor C_s (1.6pF) and current source M_{5a}/M_{5b} [8]. The capacitor C_f (1.57pF) and R_f (370 Ω) provide the shunt feedback. The Y_{neg} circuit provides an almost-constant negative resistance and a frequency dependent negative capacitance (decreasing with frequency).

We designed the Y_{neg} circuit for full IM3 cancellation at 1GHz. A buffer and a resistive attenuator are put in parallel on-chip after the LNA for noise/gain and IIP3 measurement. The IIP3 is extrapolated from -30dBm to -20dBm. The chip microphotograph is shown in Fig. 4b. The LNA occupies 0.00295 mm² active area, of which 25% is taken by the Y_{neg} circuit. Packaged chips were measured on PCB boards. Two off-chip baluns were used at the input and output of the chip for single-ended-to-differential conversion. By switching on/off of the Y_{neg} circuit, we measure the effect of Y_{neg} on LNA performance.

The measured and simulated S_{11} , NF and voltage gain is shown in Fig. 5. For 0.1GHz to 1GHz, the Y_{neg} circuit introduces no degradation on NF while it improves S_{11} by 1-3dB and improves gain by 0.2-1dB. Below 0.3GHz S_{11} becomes >-10dB because the impedance of C_f starts to block the shunt feedback. This can be improved by using a larger C_f . Note that this LNA is not optimized for very low NF as we only focus on demonstrating the IM3 cancellation technique.

To verify the robustness against process spread, the bias current of Y_{neg} (I_{Yneg}) is swept within $\pm 100\%$ variation of the optimal value. Fig. 6a shows the IIP3 improvement with respect to the circuit without Y_{neg} as a function of I_{Yneg} normalized to the optimum $I_{Yneg,opt}$; this parameter is denoted as NI_{Yneg} . In the optimum setting hence $NI_{Yneg}=100\%$. Fig. 6a shows that for wide bias variation ($NI_{Yneg}=-30\%$ to $NI_{Yneg}=+100\%$), +6dB IIP3 improvement is achieved at 1GHz. The power overhead of this technique is depicted in Fig. 6b.

Fig. 5. Measured and simulated (a) S_{11} , (b) NF and (c) Voltage gain.

Fig. 7a shows the frequency dependence of the IM3 cancellation technique, for $NI_{Yneg}=100\%$, on both IIP3 and on P_{1dB} . The measurements and simulation results in Fig. 7 show a weak frequency dependence in the IIP3 improvement and hence quite good robustness. Fig. 8 shows the measured HD1 and IM3 output at the optimal

Fig. 4. LNA using Y_{neg} for IM3 cancellation. (a) Schematic. (b) Microphotograph of the fabricated chip.

Fig. 6. Measured and simulated (a) IIP3 at 1GHz and (b) DC current consumption of LNA as a function of the nominated bias current of Y_{neg} .

Fig. 7. Measured (a) IIP3 and (b) input P_{1dB} as a function of RF frequency

Fig. 8. Measured HD1 and IM3 at 1GHz as a function of input power.

Fig. 9. Measured IIP3 at 1GHz as a function of two-tone spacing. Line for simulations and symbol for measurements.

Fig. 10. Effect of mismatches and process spread on IIP3 at 1GHz. (a) 200-time Monte Carlo simulation results of IIP3 with Y_{neg} ($NI_{yneg}=100\%$) and (b) measured IIP3 of ten dies for LNA with and without Y_{neg} .

Fig. 11. Simulated IIP3 as a function of temperature for LNA with and without Y_{neg} .

bias value of Y_{neg} ($NI_{yneg}=100\%$). The IM3 curve starts to show 5th order behavior for $P_{in} > -18$ dBm due to the transistors' higher-order nonlinearities that kick in at high input magnitudes. Fig. 9 presents IIP3 simulations results and IIP3 measurements as a function of the two-tone spacing, showing that the IM3 cancellation technique is not sensitive to two-tone spacing.

To estimate the overall effect of process spread and mismatch on this IM3 cancellation, a 200-time Monte-Carlo simulation is performed for an RF signal at 1GHz.

Fig. 10a shows that the mean IIP3 is 9.2dBm at the optimal bias value of Y_{neg} ($NI_{yneg}=100\%$) (the nominal value is 9.6dBm), which is 9dB higher than the LNA without Y_{neg} . This shows good robustness of this IM3 cancellation technique. The measurement results of ten LNA samples shows +6.2dB IIP3 improvement at 1GHz as shown in Fig.10b. The simulation results in Fig. 11 show that the LNA with the Y_{neg} circuit provides a constant IIP3 from -40°C to 40°C and starts to decrease as the temperature higher than 40°C .

IV. CONCLUSION

This paper presents a wideband IM3 cancellation technique using a negative impedance, applied to a wide band cascode LNA. Using a suitable negative impedance, the distortion current generated by the cascode transistor cancels the distortion from the transconductor. The negative impedance also increases gain while its effect on NF can be minimal. For a resistive feedback LNA fabricated in a standard $0.16\mu\text{m}$ CMOS process, for 0.1GHz to 1GHz this IM3 cancellation technique improves IIP3 by 6.3dB to 10dB, gain by 0.2 to 1dB and P_{1dB} by +3dB while NF is not degraded, at a low area and power penalty. Robustness of this cancellation technique is demonstrated both in simulation and in measurements.

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