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A 0.3–3.5 GHz active-feedback low-noise amplifier with linearization design for wideband receivers

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ABSTRACT

LNAs for wideband receivers usually require a high linearity to protect the desired signals from out-band interference. Active feedback LNAs always suffer from the nonlinear feedback of source follower, and present a poor linearity. In order to solve this problem, a complementary source follower (CSF) is proposed, which utilizes the different characteristic of NMOS and PMOS to linearize the source follower, leading to an improvement of LNA's IIP_3 and IIP_2 by about 10 dBm and 21 dBm respectively. In addition, a post-distortion technique is also used on the common source stage, which further enhances the IIP_3 by about 2 dBm and IIP_2 by 11 dBm. After using the two techniques, the noise figure (NF) does not deteriorate; instead it achieves a 0.3 dB improvement. A prototype is designed in TSMC 0.18 µm CMOS process, and a 14.8 mW power is dissipated from a 1.6 V supply. In typical process corner, across 0.3 to 3.5 GHz, this LNA achieves a 14.6 dB gain, a 2.9 dB minimum NF, and an IIP_2 larger than +22 dBm and IIP_3 larger than +1.2 dBm.

1. Introduction

Recent years, since a variety of wireless standards are emerging and meanwhile the spectrum source is less and less, the research on software defined radio (SDR) and cognitive radio has drawn much attentions. Mitola proposed the ideal SDR scheme [1], but the stringent requirements on A/D convertors makes it infeasible [2]. The receiver architecture composed of low-noise amplifier (LNA) and down-conversion mixer can mitigate that problem and is widely adopted nowadays [3]. In terms of LNA, a single wideband one is preferable due to its better performance of flexibility, area and power, comparing to a reconfigurable LNA [4] and a multi-LNAs solution [5]. But for a wideband receiver, in order to hold its configurability and software programmability, the external SAW filter is not desired due to its fixed center frequency, which leads to the concurrent reception of unfiltered multiple signals by LNA. Therefore, wideband LNAs for SDR should have a high linearity, to protect the desired signal from signals mutual intermodulation, mainly the 2nd and 3rd intermodulation. So, different from narrow-band LNAs, not only IIP_3 but also IIP_2 is critical for wideband LNAs.

As is well known, LNA design is a process of tradeoff among several parameters, such as input matching, noise figure (NF) and linearity. The most familiar wideband LNAs include common-gate LNA [6], resistive feedback LNA [7], active feedback LNA, etc. By contrast, the active feedback LNA can alleviate the tie among gain, input matching and NF, especially in nanoscale technology [8], and thus obtains more favor. Fig. 1(a) depicts the typical topology of active feedback LNA, in which the source follower (SF) composed of M_F and bias current source is used as feedback network. It has been proved in the paper [9], that the feedback is nonlinear and causes a large fraction of distortion, because of the M_F's nonlinearity and the largest signal swing at output, which usually leads to a poor linearity and makes the active feedback LNA not suitable for wideband receivers [10]. Therefore, for practicability, some circuit tricks to enhance linearity are needed. One of the effective methods is using a resistor R_F connected between M_F and the input, as depicted in Fig. 1(b), to lower M_F's gate-source voltage and nonlinear terms. But that method sacrifices the NF and power [11]. In this paper, we propose a complementary source follower (CSF), which can cancel the nonlinear terms of NMOS and PMOS transistors to linearize LNA, and meanwhile partly improve the NF. The detailed discuss will be given in the following sections.

Usually, when SF is linearized to a certain extent, the common source (CS) stage will become a main limiting factor for LNA's linearity. Many techniques which can be used to optimize CS-stage's linearity have been published in recent years, such as multiple gated transistor technique (MGTR) [12] and post-distortion technique [13]. MGTR

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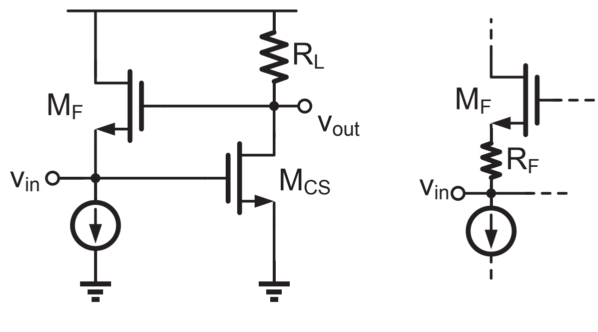

Fig. 1. (a) Traditional active feedback LNA, (b) using R_F to improve linearity.

technique may usually deteriorate input matching due to the introduced capacitance, and additionally suffer from the inaccuracy model in sub-threshold region. By contrast, post-distortion technique is more robust and has no effect on input matching, for its auxiliary transistor works at saturation region and locates at output end. In our previous work [14], a scheme using p-MOSFET as auxiliary transistor was proposed for balun-LNA. Comparing to n-MOSFET, the p-MOSFET can make the 2nd and 3rd nonlinear terms decrease simultaneously. Thus in this work, we adopts the similar scheme to linearize the CS-stage. The detailed analysis and design steps are also given.

This paper is organized as follows: Section 2 discusses the theory and design method of CSF and post distortion technique. In section 3, the full circuit and simulated results are presented. And finally, Section 4 concludes this paper.

2. Analysis of linearization technique

Firstly, let's make a brief review on the active feedback LNA in Fig. 1(a). Denoting the trans-conductance of M_{CS} and M_F as g_{mCS} and g_{mF} respectively, the voltage gain can be expressed as $A = -g_{mCS}R_L$, and the input impedance is $Z_{in} = g_{mF}^{-1}/(1 + |A|)$. Obviously, the gain is determined by the CS-stage, and with the gain defined, g_{mF} should be fixed to a certain value for matching. The noise factor can be written as follows:

$$F= 1 + \left(\frac{2-A}{1-A}\right)^2 \cdot \frac{\gamma_{CS}}{g_{mCS}R_S} + \frac{\gamma_F}{1-A} + \left(\frac{2-A}{1-A}\right)^2 \cdot \frac{1}{g_{mCS}R_S A} + \gamma_C g_{mC} R_S \quad (1)$$

where γ_{CS}, γ_F, and γ_C are respectively the thermal noise coefficients of M_{CS}, M_F and the transistor used to realize the bias current source.

Next is the discussion of linearity. Since it has been pointed out in the literature [9] that the SF in Fig. 1(a) is the main nonlinearity contributor, we first assume that the CS-stage is linear to simplify the analysis. Fig. 2 gives an equivalent circuit model in which R_S is the source impedance and i_F represents the nonlinear current of M_F. Using a three order power series, i_F can be expressed as:

$$i_F = g_{mF}(v_{out}-v_{in}) + g_{2F}(v_{out}-v_{in})^2 + g_{3F}(v_{out}-v_{in})^3 \quad (2)$$

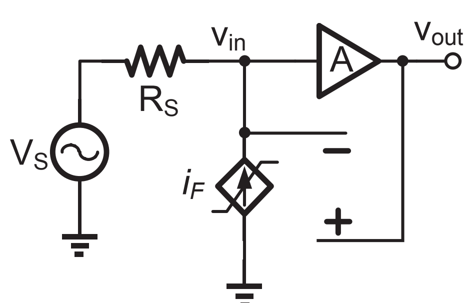

Fig. 2. The equivalent circuit to analyze SF's effect on linearity.

where g_{2F} and g_{3F} are the 2nd and 3rd nonlinear coefficient respectively. According to (2) and the topology in Fig. 2, it can be derived out that

$$v_{out} = \frac{A}{2}v_S + \frac{g_{2F}}{8g_{mF}}A(1-A)v_S^2 + \frac{A(1-A)^2}{16}\left(\frac{g_{2F}^2}{g_{mF}^2} - \frac{g_{3F}}{g_{mF}}\right)v_S^3. \quad (3)$$

And then, LNA's IIP_2 and IIP_3 can be directly calculated out as:

$$IIP_2 = \left|\frac{4g_{mF}}{(1-A)g_{2F}}\right| \quad (4)$$

$$IIP_3 = 4\sqrt{\frac{2}{3}}\frac{1}{|1-A|}\sqrt{\left|\frac{1}{(g_{2F}/g_{mF})^2-(g_{3F}/g_{mF})}\right|} \quad (5)$$

The Eqs. (4) and (5) indicate that, to simultaneously increase IIP_2 and IIP_3, both of g_{2F} and g_{3F} should tend to be zero. However, that is always impossible, for M_F is always biased in saturation region.

2.1. Complementary source follower (CSF)

To linearize the SF, the different nonlinear characteristic of NMOS and PMOS transistors can be taken advantage of. As shown in Fig. 3, when working in saturation region, the n-MOSFET always has a 2nd nonlinear coefficient g_2 with positive polarity while p-MOSFET has a negative one. And along with the overdrive voltage increasing, both the 3rd coefficients g_3 of n-MOSFET and p-MOSFET exhibit a flip of polarity. According to the phenomenon observed above, a CSF is proposed to realize a more linear SF, as shown in Fig. 4(a). Fig. 4(b) presents its equivalent circuit, in which i_n and i_p are the nonlinear current of M_n and M_p, respectively. Replacing the traditional SF in Fig. 1(a) with the CSF, the equations of IIP_2 and IIP_3 should be updated as:

$$IIP_2 = \left|\frac{4(g_{mn} + g_{mp})}{(1-A)(g_{2n} + g_{2p})}\right| \quad (6)$$

$$IIP_3 = 4\sqrt{\frac{2}{3}}\frac{1}{|1-A|}\sqrt{\left|\frac{1}{(g_{2n} + g_{2p})^2/(g_{mn} + g_{mp})^2-(g_{3n} + g_{3p})/(g_{mn} + g_{mp})}\right|} \quad (7)$$

where g_{mn} and g_{mp} are the trans-conductance, g_{2n}, g_{2p} and g_{3n}, g_{3p} are the 2nd and 3rd nonlinear coefficients respectively, and the subscript "n" denotes M_n while "p" M_p. According to Eqs. (6) and (7), if $g_{2n} + g_{2p} = 0$ and $g_{3n} + g_{3p} = 0$, the IIP_2 and IIP_3 of LNA will be towards infinity theoretically. But of course it is impossible to entirely cancel the 2nd

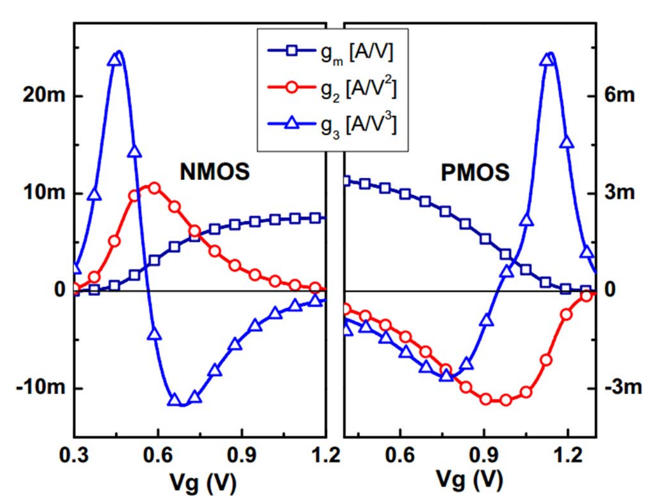

Fig. 3. Nonlinear coefficients of MOSFET versus gate-source voltage (TSMC 0.18 µm process, W/L = 15 µm/0.18 µm for NMOS, the source is connected to ground while for PMOS connected to a 1.6 V supply).