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A 40nm CMOS Ultra-Wideband Low Noise Amplifier Design

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Abstract—An ultra-wideband low noise amplifier(LNA) based on the cascode configuration with resistive feedback is presented in this paper. A shunt-shunt feedback resistor and a pre-π matching network is employed to achieve wideband input impedance matching, and to enhance gain response and reduce noise for using a post-cascode inductor LP. In this paper, the SMIC 40nm CMOS process is used. The circuit simulation results show that S11 is lower than -10dB, and the gain S21 is around 10±2dB,and the NF is lower than 4.5dB in the 3.5-31-GHz frequency range.

Keywords—CMOS, feedback; flatness; Low Noise Amplifier(LNA); Noise Figure(NF); bandwidth

I. Introduction

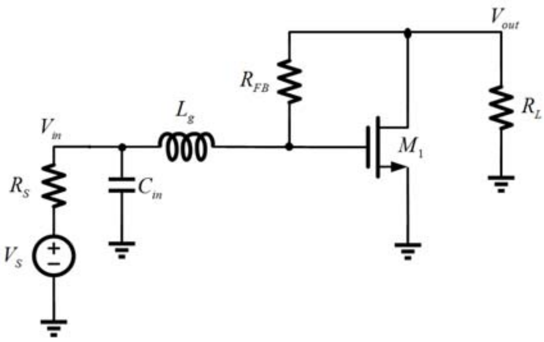
Low noise amplifiers (LNAs) with multi-decade gigahertz bandwidth have been one of the most crucial components in applications of mobile communications, radar and multi-standard software-defined radios [1],[2]. And LNA is the first active circuit of the receiving link in T/R module. It amplifies the received signal with as little noise as possible to reduce the influence of the noise on the subsequent stage. At the same time, the LNA also exhibits high linearity for providing high gain to reduce the influence of nonlinear characteristics on signal quality. A dual-feedback topology is a typical choice to extend impedance matching bandwidth. And using a shunt peaking inductor to extend 3-dB bandwidth [3]. However, this structure suffers from high noise figure (NF). Using asymmetrical T-coils and gate-inductive peaking, to enhance the wideband gain and decrease noise figure [4].

In this paper, the design principles and analytical equations of an ultra-wideband LNA is presented. The rest of this paper is arranged as follows: Details of the input impedance matching, gain bandwidth, and frequency response of the NF are provided in section II; Section III describes the given wideband LNA circuit and its simulation results, and the simulation results are compared with the previous work results; Section IV gives the conclusion.

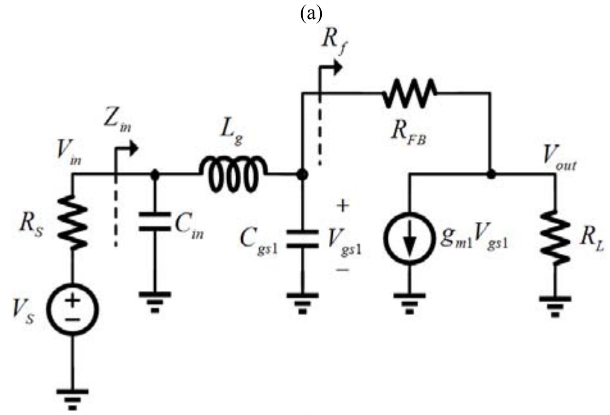
II. Circuits Design and Analysis

A. Wideband Input Matching Network

The resistive shunt-shunt feedback is a popular technique for enlarge the bandwidth of an amplifier. Figure 1.(a) and (b) shown the classical resistive shunt-shunt resistive feedback amplifier and its small-signal equivalent circuit. To simplify the analysis, the intrinsic gate-drain capacitance C_{gd1} is ignored.



(a)



(b)

FIGURE I. (A) SCHEMATIC OF SHUNT-SHUNT RESISTIVE FEEDBACK AMPLIFIER (B) SMALL-SIGNAL EQUIVALENT CIRCUIT

From Figure 1.(b), it can be obtained:

$$\frac{V_{out}}{R_L} + g_{m1}V_{gs1} = \frac{V_{gs1} - V_{out}}{R_{FB}} \tag{1}$$

So the current flows on R_{FB} can be written as:

$$I_x = \frac{V_{gs1} - V_{out}}{R_{FB}} \rightarrow R_f = \frac{V_{gs1}}{I_x} = \frac{R_{FB} + R_L}{1 + g_{m1}R_L} \tag{2}$$

So the input impedance of the amplifier is:

$$Z_{in} = \left[sL_g + \left(R_f \parallel \frac{1}{sC_{gs1}} \right) \right] \parallel \frac{1}{sC_{in}} \tag{3}$$

Finally the input impedance of the amplifier can be written as:

$$Z_{in} = \frac{s^2 R_f C_{gs1} L_g + sL_g + R_f}{s^3 R_f C_{in} C_{gs1} L_g + s^2 L_g C_{in} + sR_f \left(C_{in} + C_{gs1} \right) + 1} \tag{4}$$

Corresponding to the input impedance matching $Z_{in} = Z_O$, $R_f = Z_O$ is usually selected to achieve DC bandwidth matching. The $|S_{11}|$ of the amplifier can be expressed as follows:

$$|S_{11}| = \left| \frac{-s^3 L_g Z_O^2 C_{in} C_{gs1} + s^2 L_g Z_O \left(C_{gs1} - C_{in} \right) + s\left[L_g - Z_O^2 \left(C_{in} + C_{gs1} \right) \right]}{s^3 L_g Z_O^2 C_{in} C_{gs1} + s^2 L_g Z_O \left(C_{in} + C_{gs1} \right) + s\left[L_g + Z_O^2 \left(C_{in} + C_{gs1} \right) \right] + 2Z_O} \right| \tag{5}$$

If set $C_{in} = C_{gs1}$, then from equation (5) it can be seen clearly that the response of $|S_{11}|$ includes two zero points ω_{01} and ω_{02}.

$$\omega_{01} = 0 \tag{6}$$

$$\omega_{02} = \sqrt{\frac{2}{L_g C_{gs1}} - \frac{1}{Z_O^2 C_{gs1}^2}} \tag{7}$$

When L_g and C_{in} are not added, S11 can be expressed as follows:

$$|S_{11}| = \left| \frac{\frac{1}{R_S} - Y_{in}}{\frac{1}{R_S} + Y_{in}} \right| = \left| \frac{\frac{R_f}{R_S} - 1 - j\omega R_f C_{gs1}}{\frac{R_f}{R_S} + 1 + j\omega R_f C_{gs1}} \right| \tag{8}$$

As expected. That is, two of the predicted frequency responses are falling. Assuming $C_{gs1} = 137.075\,fF$, $L_g = 0.3nH$,the relationship between the calculated value of S11 and the frequency corresponding to the equations (5) and (8) shown in Figure 2 and the frequency can be obtained. It can be seen from equations (5), (8) and Fig.2 that when there is no series inductance and parallel capacitance at the input end, it will be greater than -10dB, and after adding π matching input network at the input end, $\omega_{02} = 26.3GHz$ is obtained, the original 15.5GHz input matching bandwidth is extended to 26.3GHz.

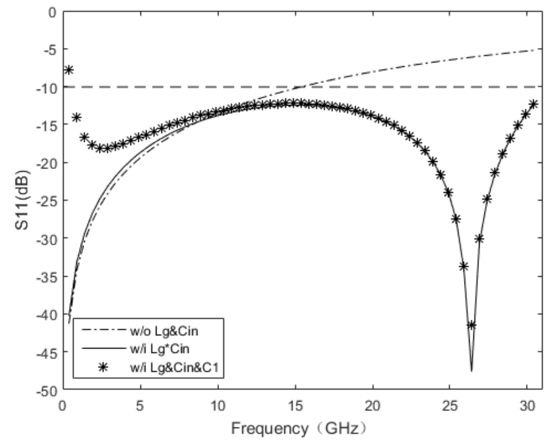
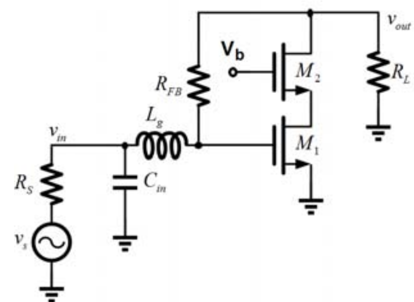


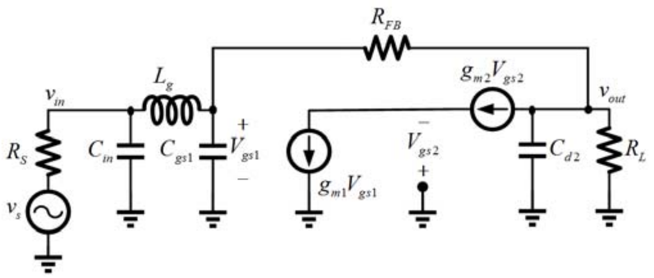
FIGURE II. CALCULATED S11 PARAMETERS AT DIFFERENT FREQUENCIES FOR RESISTIVE SHUNT–SHUNT FEEDBACK AMPLIFIER BOTH WITH AND WITHOUT SERIES INPUT INDUCTOR L_g AND PARALLEL INPUT CAPACITOR C_{in}.

B. *Frequency Response of S21*

The cascode structure is one of the most commonly used LNA topologies due to its low power consumption, high gain, and high reverse isolation. The schematic of cascade amplifier and its small-signal equivalent circuit are shown in Figure 3.(a) and (b) respectively.



(a)



(b)

FIGURE III. (A) SCHEMATIC OF CASCODE AMPLIFIER WITH RESISTIVE SHUNT–SHUNT FEEDBACK (B) SMALL-SIGNAL EQUIVALENT CIRCUIT.

For simplicity, assume $\left(s^2C_{gs1}L_g+1\right)sC_{in}\approx 0$, in this case the S_{21} of the amplifier can be expressed as:

$$S_{21}=2\cdot A_{vs}=2\cdot A_{vs,in}\cdot A_{vs,core}\approx 2\cdot\frac{1\Big/\left[L_g\left(C_{gs1}+\dfrac{C_{in}R_S}{R_f}\right)\right]}{s^2+s\left(\dfrac{\omega_{0,in}}{Q_{in}}\right)+\omega_{0,in}^2}\cdot\frac{\left(R_{FB}\parallel R_L\right)\left(\dfrac{1}{R_{FB}}-g_{m1}\right)}{1+\dfrac{s}{\omega_{core}}} \tag{9}$$

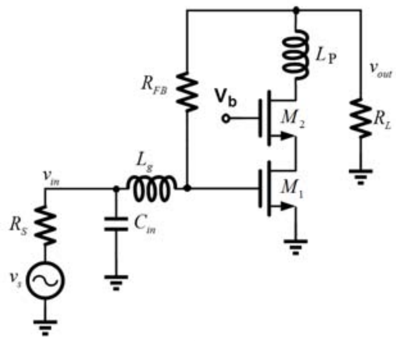
Where $A_{vs,in}$ and $A_{vs,core}$ represent the voltage gain V_{gs1}/V_S and V_{out}/V_{gs1} respectively, and

$$\omega_{0,in}=\sqrt{\left(1+\frac{R_S}{R_f}\right)\Big/\left[L_g\left(C_{gs1}+\frac{C_{in}R_S}{R_f}\right)\right]} \tag{10}$$

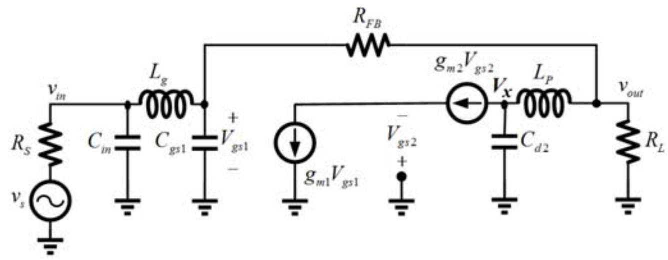
$$Q_{in}=\sqrt{L_g\left(C_{gs1}+\frac{C_{in}R_S}{R_f}\right)\left(1+\frac{R_S}{R_f}\right)}\Big/\left(\frac{L_g}{R_f}+R_SC_{gs1}\right) \tag{11}$$

$$\omega_{core}=1\big/\left[C_{d2}\left(R_{FB}\parallel R_L\right)\right] \tag{12}$$

Where $\omega_{0,in}$ and Q_{in} are the pole frequency and pole Q factor of the input network of the LNA, and ω_{core} stands for the pole frequency of the core circuit of the LNA, respectively. Obviously, the possible method of increasing the bandwidth f_{3dB} of the amplifier is to add a shunt peak inductor L_P as shown in Figure 4(a), (b) is its small-signal equivalent circuit.



(a)



(b)

FIGURE IV. (A) CASCODE AMPLIFIER WITH A SERIES PEAKING INDUCTOR, (B) SMALL-SIGNAL EQUIVALENT CIRCUIT LET THE VOLTAGE AT THE INTERSECTION OF L_P AND C_{d2} BE V_x, THEN:

$$\frac{V_{in}-V_{gs1}}{sL_g}=V_{gs1}\cdot sC_{gs1}+\frac{V_{gs1}-V_{out}}{R_{FB}} \qquad (13)$$

$$\frac{V_{out}-V_x}{sL_P}=g_{m1}V_{gs1}+V_x\cdot sC_{d2} \qquad (14)$$

$$\omega_{core}=1/\left(C_{d2}\left(R_{FB}\parallel R_L\right)\right) \qquad (15)$$

Finally the gain of the amplifier is given by:

$$S_{21}=2\cdot A_{vs}=2\cdot A_{vs,in}\cdot A_{vs,core\_a}\approx 2\cdot\frac{1\Big/\left[L_g\left(C_{gs1}+\frac{C_{in}R_S}{R_f}\right)\right]}{s^2+s\left(\frac{\omega_{0,in}}{Q_{in}}\right)+\omega_{0,in}^2}\cdot\frac{\frac{1}{C_{d2}L_P}\left(R_{FB}\parallel R_L\right)\left(\frac{1}{R_{FB}}-g_{m1}\right)}{s^2+s\left(\frac{\omega_{core\_a}}{Q_{core\_a}}\right)+\omega_{core\_a}^2} \qquad (16)$$

ω_{core_a} and Q_{core_a} represent the pole frequency and pole Q factor, respectively, of the core circuit of the LNA with a post-cascode series-peaking inductor L_P, $L_P=0.306nH$, $C_{d2}=31.74\,fF$. Since both $A_{vs,in}$ and $A_{vs,core\_a}$ are normal second-order low-pass filtering functions, the gain of the amplifier is depended on $\omega_{0,in}$, Q_{in}, ω_{core_a} and Q_{core_a}. Figure 5 shows different S21 values at different conditions. L_P* is series in series after resistance feedback.

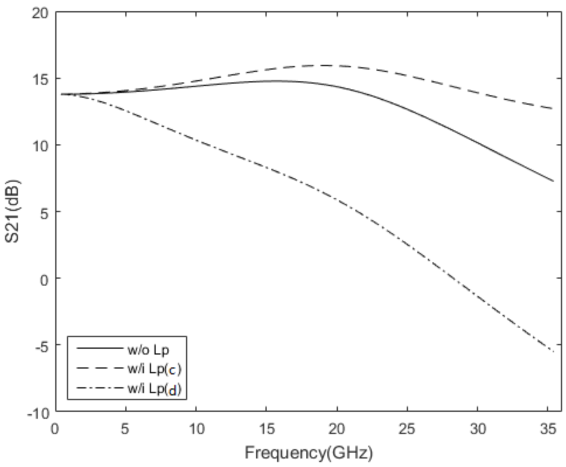


FIGURE V. SIMULATION S21 VERSUS RFB OF CASCODE AMPLIFIER WITH RESISTIVE SHUNT–SHUNT FEEDBACK WITH AND WITHOUT POST-CASCODE INDUCTOR L_P.

III. SIMULATION RESULT AND ANALYSIS

The presented ultra-wideband LNA schematic is shown in Fig.6, it employs a resistive shunt-shunt feedback with cascade structure. In Figure 6, a small inductor (50 pH) is increased to achieve good linearity and stability.

Figure 7-9 show the S11, S21 and NF simulation results of the ultra-wideband respectively.

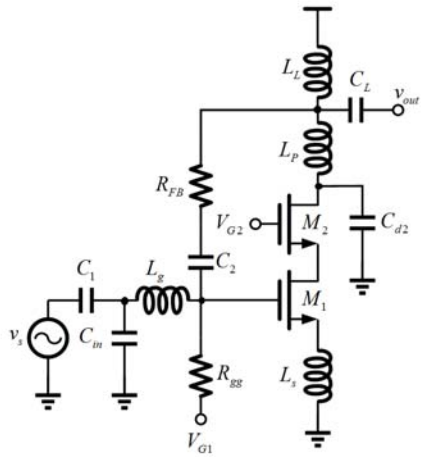


FIGURE VI. THE PRESENTED ULTRA-WIDEBAND LNA

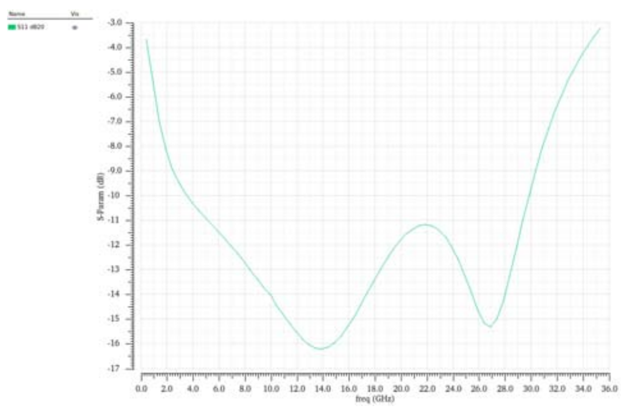


FIGURE VII. S11 SIMULATION OF THE ULTRA-WIDEBAND LNA

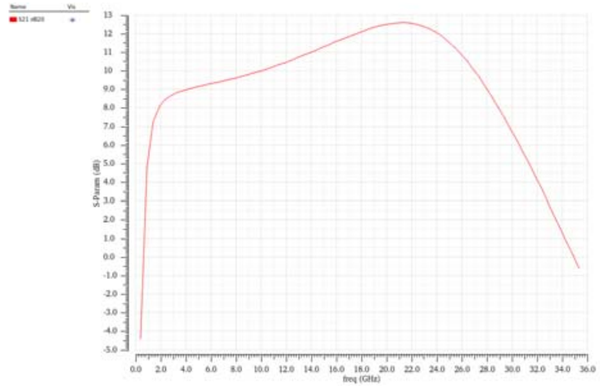


FIGURE VIII. S21 SIMULATION OF THE LNA

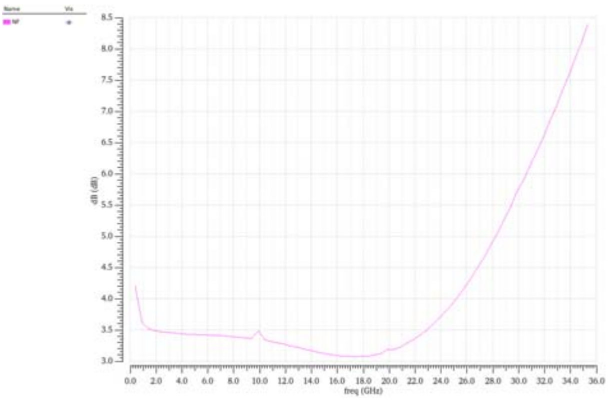


FIGURE IX. NF SIMULATION OF THE LNA

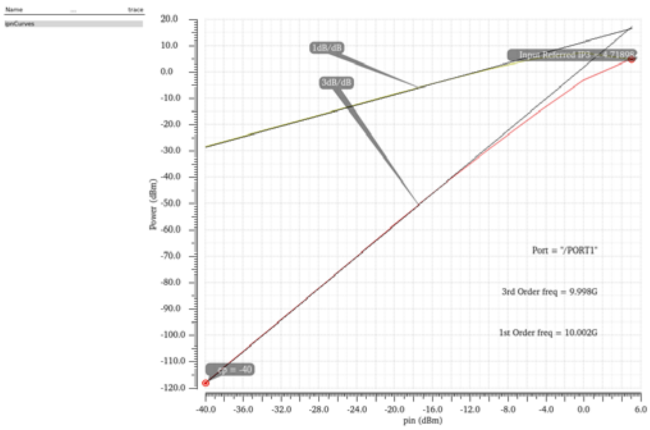


FIGURE X. IIP3 SIMULATION OF THE LNA

IV. CONCLUSION

This paper demonstrates the compact ultra-wideband LNA design for broadband applications in theory and experimental simulation. Adding the post-cascode shunt-peak inductor to get high and flat gain response. The simulation results are consistent with the analysis results, indicating that the proposed LNA topology is very suitable for low-cost and high-performance broadband LNA.

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