A Compact Wideband CMOS Low-Noise Amplifier Using Shunt Resistive-Feedback and Series Inductive-Peaking Techniques

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Abstract—A wideband low-noise amplifier (LNA) with shunt resistive-feedback and series inductive-peaking is proposed for wideband input matching, broadband power gain and flat noise figure (NF) response. The proposed wideband LNA is implemented in 0.18- μm CMOS technology. Measured results show that power gain is greater than 10 dB and input return loss is below -10 dB from 2 to 11.5 GHz. The IIP3 is about +3 dBm, and the NF ranges from 3.1 to 4.1 dB over the band of interest. An excellent agreement between the simulated and measured results is found and attributed to less number of passive components needed in this circuit compared with previous designs. Besides, the ratio of figure-of-merit to chip size is as high as $190 \text{ (mW}^{-1}/\text{mm}^2)$ which is the best results among all previous reported CMOS-based wideband LNA.

Index Terms—CMOS, feedback, low cost, low-noise amplifier (LNA), series peaking, small chip size, wideband.

I. INTRODUCTION

WIDEBAND systems have recently gained much attention due to their capability of high data rate transmission. The so called ultrawideband (UWB) technology provides attractive high speed wireless signaling schemes, which use the frequency bands in 3.1 ~ 10.6 GHz. To interface with the antenna and pre-select filter, the low-noise amplifier (LNA) input impedance should be close to 50Ω across the band from 3.1 to 10.6 GHz.

Up to now, many wideband techniques have been proposed for the UWB LNA. The distributed amplifier (DA) is widely used for wideband application due to its intrinsic broadband frequency response going all the way down to dc along with good input and output impedance matching. Yet, so far, high power consumption and large die area have hampered its widespread applications [1], [2]. Recently, a cascode CMOS LNA with a bandpass response at the input for wideband impedance matching has been reported in [3]. The bandpass-filter (BPF) based topology incorporates the input impedance of the cascode amplifier as a part of the filter. However, the adoption of the filter at the input requires a number of reactive elements, which inevitably results in a larger chip area and noise figure

(NF) degradation for on-chip implementation. An efficient way to achieve a broadband matching with small power dissipation and chip area is the common-gate input topology. However, it has been reported that a common-gate LNA exhibits a lower gain and higher NF than a common-source amplifier [4].

In this letter, a wideband LNA with wide bandwidth, low NF, low power consumption, broadband impedance matching and small chip-area is reported. Both the shunt resistive-feedback and series inductive-peaking techniques are adopted to design the 2 ~ 11.5 GHz UWB LNA. These techniques have the merits of achieving wideband frequency response, sufficient power gain and small NF without using many passive components that require large die area.

II. WIDEBAND LNA DESIGN

The conventional cascode-type LNA with inductive source-degeneration is one of the most popular LNA topologies due to its merits of low power consumption, high gain, and high reverse isolation. For wideband application, this topology becomes unsuitable because the bandwidth of the input series resonance circuit is limited by its quality factor. Hence, a 3–5 GHz cascode-type UWB LNA with resistive feedback [5] has been proposed to reduce the quality factor of the input series resonance circuit. But the bandwidth of that design is still limited by the quality factor of the input series resonance circuit and capacitive effect, namely, C_{gs} , C_{gd} , C_{db} , and C_{sb} of M1 and M2, and thus the input return loss and gain began to roll-off at 4.2 GHz. Fig. 1 shows the proposed UWB LNA, which contains the conventional cascode-type LNA enclosed in the dash lines of the figure. A smaller feedback resistor R2 (300Ω) compared to that ($1 \text{ k}\Omega$) of the conventional design [5] is used to reduce the quality factor of the input series resonance circuit further so that a wider input matching can be achieved. However, from the simulation, the feedback resistor R2 degrades the NF of LNA by approximately 1.2 dB. Moreover, the power gain of this amplifier at higher frequencies is also degraded due to the capacitive effect of large size M1 and M2 chosen. To counterbalance this gain degradation, a series-peaking inductor L2 as shown in Fig. 1 is added to make the gain response flatter. The series-peaking inductor L2 can also resonate with the parasitic shunt capacitor C_p at the drain of M2. The value of L2 is chosen to be 1.13 nH for peaking at the frequency of 10 GHz, and C2 is used for dc decoupling. The degeneration inductor TL1 was chosen to be small (0.12 nH) for matching at higher band and implemented by a microstrip (MS) line with a line width of 6 μm and a line-length of 162 μm . Compared to the spiral inductor,

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Fig. 1. Wideband LNA schematic.

Fig. 2. Die photograph of the wideband LNA.

the MS inductor is less susceptible to process variation when the inductance is very small.

III. EXPERIMENTAL RESULTS

The wideband LNA was implemented in a 1P6M 0.18 μm CMOS technology. The die photograph is shown in Fig. 2. The die area including testing pads is only $0.48 \times 0.69 \text{ mm}^2$, which is very area-efficient and consequently cost effective due to the techniques adopted. The circuit was tested on wafer for RF characterization. The LNA draws 11.2 mA from bias-T and the drain voltages of M2 is 1.2 V.

Fig. 3 shows the simulated and measured results of the input return loss and power gain of the wideband LNA. The power gain is above 10 dB with a 3-dB frequency band up to 12.5 GHz while the return losses are below -10 dB from 2 to 11.5 GHz. Simulation shows that the bandwidth of power matching shrinks from 9.5 GHz to only ~ 2.5 GHz if R2 is removed from the circuit. These results unequivocally prove the effectiveness of the broadband matching realized by using the small R2 shunt resistive-feedback technique in conjunction with the source degenerative inductor-TL1. Fig. 3 also shows the simulated frequency responses of the LNA with different values of L2. It is found that $L2 = 0.619 \text{ nH}$ corresponds to the maximally flat condition. In our work, L2 is chosen to be 1.13 nH for peaking at 10 GHz. An L2 smaller than 1.13 nH (but larger than 0.619 nH) can result in a flatter gain response (and even a wider bandwidth). However, in this work, a larger L2 value corresponding to under-damped rather than maximally flat condition is chosen

Fig. 3. Simulated and measured results for input return loss and power gain of wideband LNA versus frequency.

Fig. 4. Simulated and measured results for NF of wideband LNA versus frequency.

because the LNA is designed to be integrated with a mixer behind it and the intentional peaking at 10 GHz can pre-compensate the usual overdamped response of the next stage. Besides, it can also be seen that the power gain begins to roll-off from 3 GHz if L2 is removed from circuit, which evidences that the series inductive-peaking technique is useful for extending the bandwidth of the resistive feedback cascode-LNA.

Fig. 4 shows the simulated and measured results of the NF of this wideband LNA. Based on Fig. 4, the measured minimum and maximum NFs are 3.1 dB (at 9 GHz) and 4.1 dB (at 11.5 GHz), respectively, over the band of 2–11.5 GHz. From the simulation it is clear that the shunt feedback resistor R2 for the purpose of wideband matching degrades the NF of LNA by ~ 1.2 dB. Nevertheless, the achieved NFs are still better than the previously measured values of 3.1 ~ 10.6 GHz CMOS LNAs [1]–[9]. In addition, this circuit also has the flattest NF across the band of all UWB LNAs, which is explained as follows. Based on [10] and neglecting gate noise, we can show that the NF of an LNA using a source degenerative inductor can be approximated by

$$\text{NF} \approx 1 + \frac{R_{\text{par}}}{R_S} + \frac{\gamma}{G_{M1}R_S} \cdot \left| \left(\frac{s}{\omega_{\text{Onf}}} \right)^2 + \frac{s}{\omega_{\text{Onf}}} \frac{1}{Q_{\text{in}}} + 1 \right|^2 \quad (1)$$

where R_{par} and $Q_{\text{in}} = \omega_{\text{Onf}}(L1 + TL1)/(R_S + R_{\text{par}})$ are the series parasitic resistance, and quality factor of the input R - L - C circuit, respectively, $\omega_{\text{Onf}} = 1/\sqrt{(L1 + TL1)(C_{gs} + C_{gd})}$, $R_S = 50 \Omega$ and γ the channel noise parameter. (1) is verified by comparison with simulation results (see Fig. 4). According to (1), the frequency responses (under-, over-, and critical-damping) of NF are controlled by Q_{in} and in this work a Q_{in} as close to the maximally flat condition as possible is chosen, which explains the excellent NF flatness of our LNA.

TABLE I
RECENTLY REPORTED PERFORMANCES OF WIDEBAND LNAs

	Freq. [GHz]	S_{11} [dB]	Gain-max [dB]	NF-min [dB]	NF-flatness (-3.1-10.6 GHz) [dB]	IIP3 [dBm]	P_{1dB} [dBm]	Power [mW]	Die Size [mm ²]	FoM/Size [(mW ⁻¹)/(mm ²)]	Technology
[1], VLSI 2003	0.5-14	<-11	10.6	3.4	4.4 ± 1	+20	+10	52@1.3-V	1.6	12.37/1.6=7.7	CMOS 0.18-um
[2], JSSC	0.04-7	<-16	8.6	5.7	-	+3	-8.5	9@1.3-V	1.16	12.7/1.16=10.9	CMOS 0.18-um
[3], JSSC	2.3-9.2	<-9.9	10.4	4	6.75 ± 2.75	-6.7	-15	9@1.8-V	1.1	28/1.1=25.5	CMOS 0.18-um
*[4], CAS-I	3.1-10.6	<-9	17.5	3.1	4.4 ± 1.3	-	-	33.2@1.8-V	0.5	26.8/0.5=53.7	CMOS 0.18-um
[5], JSSC	2-4.6	<-9	9.8	2.3	-	-7	-	12.6@1.8-V	0.9	15.3/0.9=17	CMOS 0.18-um
[6], ISSCC 2006	3-5	<-10.5	16	1.8	-	-9	-23	7.6@1.2-V	0.63	54/0.63=85.6	CMOS 0.18-um
[7], CICC 2005	1.2-11.9	<-12	9.7	4.5	6 ± 1.5	-6.2	-	20@1.8-V	0.59	14.9/0.59=25.3	CMOS 0.18-um
[8], MWCL	2-4.6	<-10	9.5	3.5	-	-0.8	-	16.5@1.5-V	1.1	5.43/1.1=4.9	CMOS 0.13-um
[9], RFIC 2006	3.1-10.6	<-11.4	12.02	4.7	5.15 ± 0.45	-10	-20	10.5@1.5-V	0.665	24.7/0.665=37.1	CMOS 0.18-um
This work	2-11.5	<-10	14.8	3.1	3.58 ± 0.41	+3	-8.5	13.4@1.2-V	0.33	62.8/0.33=190	CMOS 0.18-um

* [4] is the simulated results

Fig. 5. Measured results for IIP3 under the two tones input of 5.2 and 5.8 GHz.

From Figs. 3 and 4, it is found that the simulation and measurement results agree very well, which is attributed to careful layout and less number of passive components needed in the proposed LNA compared with previous designs.

To characterize the nonlinear behavior, two-tone signals with equal power levels at 5.2 and 5.8 GHz, which are commonly used frequencies in wireless local area network applications, were applied to the proposed UWB LNA. The measured results in Fig. 5 indicate that this LNA has an IIP_3 of 3 dBm and a P_{1dB} of -8.5 dBm. These results demonstrate that the proposed wideband LNA achieves good linearity, too.

To evaluate the performance of UWB LNAs for their maximum power gain, 3 dB-bandwidth, excess NF, power dissipation, and cut-off frequency (representing technology cost), the figure of merit (FoM) is defined as [11]

$$\text{FoM}[\text{mW}^{-1}] = \frac{S21[1] \times BW[\text{GHz}]}{(NF - 1)[1] \times P_{DC}[\text{mW}] \times f_t[\text{GHz}]}$$

This FoM includes the most relevant parameters in order to evaluate a UWB LNA for low-cost and low-power applications. A summary of the state-of-the-art results is given in Table I. Our proposed LNA achieves the highest FoM of 62.8 (mW⁻¹) among all CMOS-based wideband LNAs. Besides, the chip size of our LNA including testing pads is only 0.33 mm², and thus the ratio of FoM to chip size can be even as high as 190 (mW⁻¹/mm²), which is at least a factor of 2 better than any of the previously published results.

IV. CONCLUSION

A compact wideband LNA for 2 ~ 11.5 GHz application has been proposed. The shunt resistive-feedback in conjunction with series inductive-peaking provides a method for achieving a high performance wideband LNA with low cost. Because less component are used with this method, the proposed wideband LNA has shown a good agreement between simulation and measurement results. Besides, this wideband LNA reveals the best FoM of 62.8 (mW⁻¹) and moreover the highest ratio (190 mW⁻¹/mm²) of FoM to chip-size among all CMOS-based wideband LNAs.

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