A single-to-differential low-noise amplifier with low differential output imbalance*

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Abstract: This paper presents a single-ended input differential output low-noise amplifier intended for GPS applications. We propose a method to reduce the gain/amplitude and phase imbalance of a differential output exploiting the inductive coupling of a transformer or center-tapped differential inductor. A detailed analysis of the theory of imbalance reduction, as well as a discussion on the principle of choosing the dimensions of a transformer, are given. An LNA has been implemented using TSMC 0.18 μm technology with ESD-protected. Measurement on board shows a voltage gain of 24.6 dB at 1.575 GHz and a noise figure of 3.2 dB. The gain imbalance is below 0.2 dB and phase imbalance is less than 2 degrees. The LNA consumes 5.2 mA from a 1.8 V supply.

Key words: GPS receiver; single-ended input differential output; balun; low-noise amplifier; transformer
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1. Introduction

In commonly used GPS receivers, differential signaling is preferred in order to reject power supply and substrate noise and reduce second order distortion. As a result, a single-ended RF signal from an antenna needs to be converted into a differential one. Usually, we can exploit an external off-chip balun to act as the converter. However, the external balun would cause extra gain loss and an NF increase of about 1 dB[1]. Alternatively, an on-chip single-ended input differential output low noise amplifier instead of a differential LNA as the second stage LNA can be adopted. However, a single-to-differential LNA cannot provide a precisely balanced output due to its asymmetric topology. The output imbalance would introduce an unwanted common mode signal, which will degrade the linearity of the following mixer. What's more, the phase imbalance will deteriorate the phase-modulated signal, making it difficult to demodulate. In this work, we focus on the design of a low output imbalance single-to-differential LNA. We first review several reported single-to-differential LNA topologies and output balancing methods, and propose our method. Then we give the analysis and design of the proposed LNA.

2. Design of single-to-differential LNA

The main single-to-differential LNA approach is the common source common gate (CSCG) combination[2] or its variant[1,3], as depicted in Figs. 1(a) and 1(b). In this approach, the differential output is difficult to balance when the process varies, because the CS and CG transistors operate in different conditions.

Another option to realize a single-to-differential LNA is a pseudo differential topology, as depicted in Fig. 1(c). The output imbalance can be improved by connecting the drain of M1 and the gate of M2 with a capacitor[4]. But the large parasite at the common node will cause intense imbalance at high frequency. In Ref. [5], an on-chip balun is integrated to complete the conversion, and a capacitive-cross-coupling common gate amplifier, which is often applied in differential LNAs, is used to ensure output balancing.

In this paper, we propose a simple but effective method to improve the differential output imbalance. The proposed

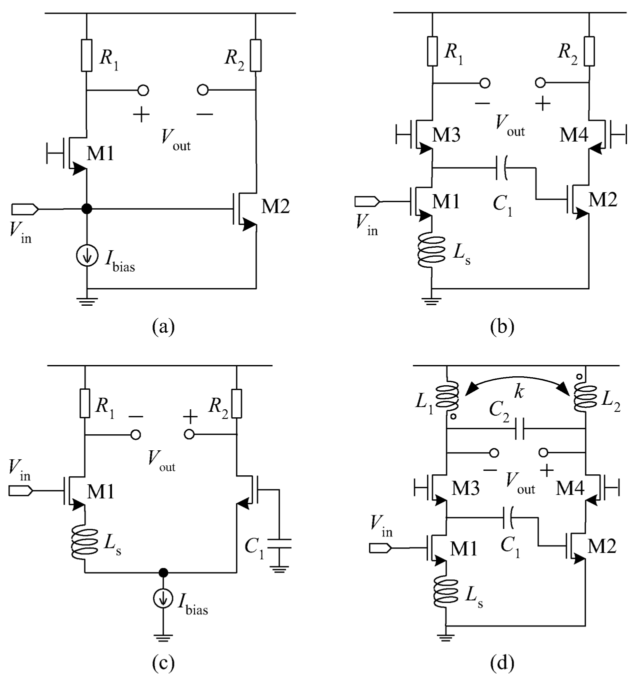

Fig. 1. (a) CSCG topology. (b) Variant of CSCG. (c) Pseudo differential topology. (d) Proposed low imbalance topology.

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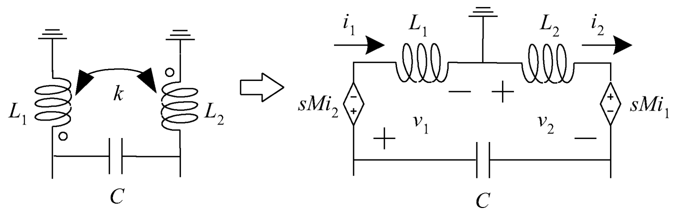


Fig. 2. Small signal equivalent model.

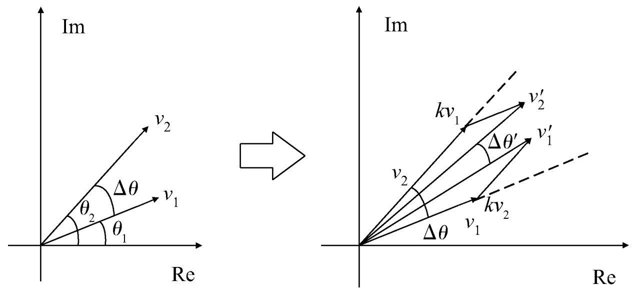


Fig. 3. v_1 and v_2 with and without inductive coupling.

single-to-differential LNA, which is based on the CSCG topology, is shown in Fig. 1(d). Here we use a symmetrical transformer[6] as the load instead of resistors or other separate elements as inductors. Although it is often used in some mixers, VCOs or even differential LNAs, none of them has been aware of the transformer improving the imbalance, to say nothing of the analysis on it.

3. Circuit analysis

3.1. Imbalance reduction

A transformer as a load can be modeled as shown in Fig. 2.

The transformer provided positive feedback at the input. In this notation, M is mutual inductance and equals $M = k\sqrt{L_1 L_2}$, where k is the coupling coefficient, and $0 < k < 1$.

In a symmetrical transformer, $L_1 = L_2 = L$, and mutual inductance will be $M = kL$. To analyze the amplitude and phase imbalance, in general, we can suppose $i_1 = I_1 \mathrm{e}^{-j(\omega t+\theta_1)}$ and $i_2 = I_2 \mathrm{e}^{-j(\omega t+\theta_2)}$. Without inductive coupling, the voltage on two inductors are $v_1 = sL_1 i_1$ and $v_2 = sL_2 i_2$. Thus, the amplitude imbalance is

$$A_{\text{imbalance}} = 20\left|\lg\frac{V_1}{V_2}\right| = 20\left|\lg\frac{LI_1}{LI_2}\right| = 20\left|\lg\frac{I_1}{I_2}\right|, \quad (1)$$

and the phase imbalance is

$$P_{\text{imbalance}} = \Delta\theta = \left|\theta_1 - \theta_2\right|. \quad (2)$$

But with inductive coupling, the voltages are $v_1' = sL_1 i_1 + sMi_2 = v_1 + kv_2$ and $v_2' = sL_2 i_2 + sMi_1 = v_2 + kv_1$. So, by intuitive analysis, we can perceive that the coupling/mutual inductance, represented by two CCVSs, will compensate for the gap between the amplitudes and phases of the two inductances, which is shown in Fig. 3. In the detailed analysis, quantitatively, the amplitude imbalance is

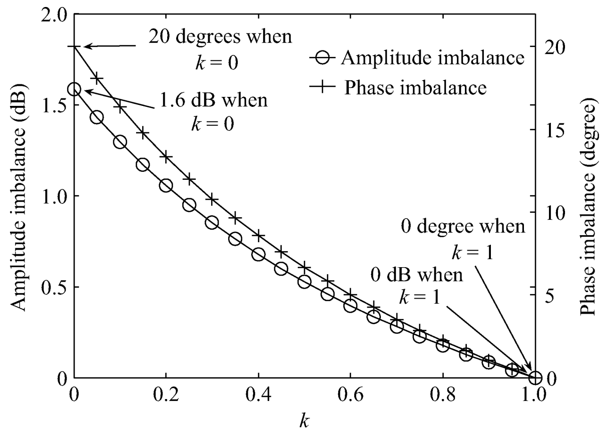


Fig. 4. Phase and amplitude imbalance versus k.

$$A_{\text{imbalance}} = 20\left|\lg\sqrt{\frac{I_1^2 + (kI_2)^2 + 2kI_1I_2\cos\Delta\theta}{I_2^2 + (kI_1)^2 + 2kI_1I_2\cos\Delta\theta}}\right|, \quad (3)$$

and the phase imbalance is

$$P_{\text{imbalance}} = \left|\arctan\frac{I_1\sin\theta_1 + kI_2\sin\theta_2}{I_1\cos\theta_1 + kI_2\cos\theta_2} - \arctan\frac{kI_1\sin\theta_1 + I_2\sin\theta_2}{kI_1\cos\theta_1 + I_2\cos\theta_2}\right|. \quad (4)$$

In Fig. 4, the initial amplitude imbalance is set to 1.6 dB, that's 1.2/1, and the initial phase imbalance is set to 20 degrees. As it shows, the imbalance of both amplitude and phase decrease as k increases from 0 to 1. And when k equals 1, the imbalance will be compensated completely. Besides the advantage of reducing the imbalance, inductive coupling can also improve the Q of the inductor, as the coupling can raise the effective inductor; however, the resistor in the inductor is still the same. Also we can see from the analysis that, the imbalance reduction effect is independent of the frequency.

3.2. Dimension of inductor

In this work, two terminals of the transformer are connected to the source supply, so the transformer can be replaced with a center-tapped differential inductor. The layout and lumped circuit model of the inductor[7] is shown in Fig. 5. Figure 6 shows the relationship between k and the value of the inductor and the width of the metal wire. In Fig. 6, W is the width of the metal wire, and n_{r} is the number of round. Due to the limitations of technology, when n_{r} is 3, the maximum inductor value is only about 1.5 nH. We can still perceive from the trend of curve that, k increases with increasing n_{r} or a wider metal wire. In Ref. [5], k can be as large as 0.9. While in the technology we use, k reaches its peak at only 0.66 when $W = 15\ \mu$m and $n_{\text{r}} = 5$.

3.3. Dimensions of transistors

The left half of the proposed single-to-differential LNA is a conventional single-ended LNA. With a constrained amount

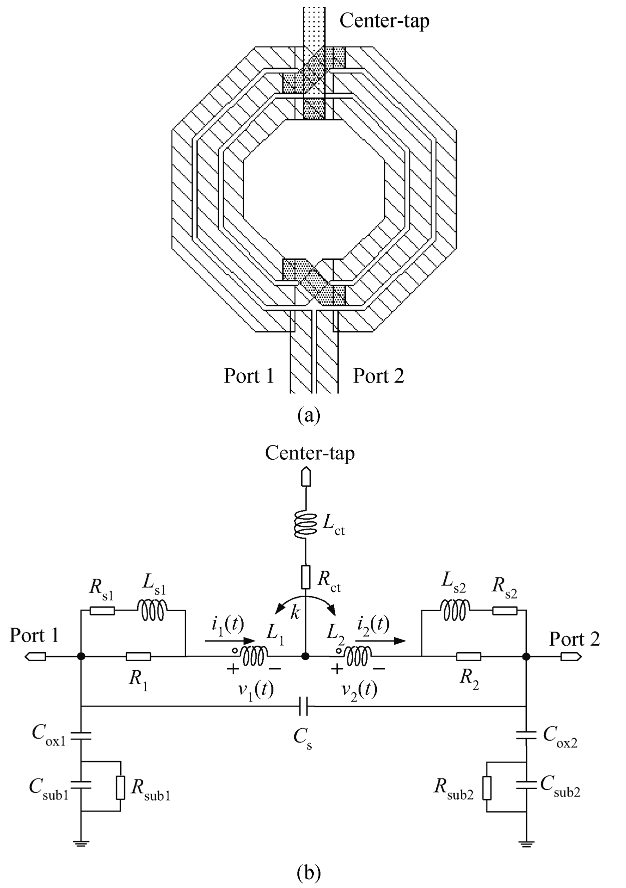


Fig. 5. (a) Layout of a center-tapped differential inductor. (b) Lumped circuit model.

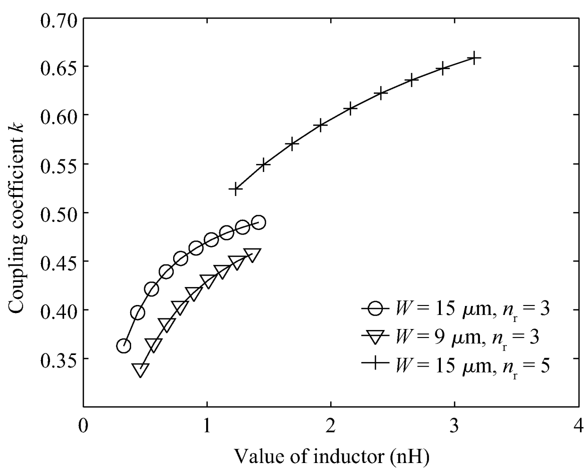


Fig. 6. Mutual inductor coefficient versus the value of the inductor.

of power dissipation, there exists a transistor size where the NF is at a minimum[8].

Simultaneous gain and noise matching can be achieved by proper selection of L_s. M2 can reduce the Miller effect of C_{gd} of M1, improve the output impedance and enhance the isolation[8]. The other half is a cascode amplifier. In order to make the imbalance low enough, the dimensions of M3 and M4 should be the same to ensure that the parasite at the load is the same. The drain currents through them should also be the same,

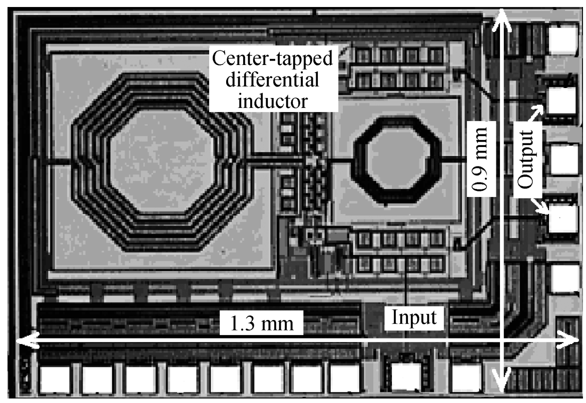


Fig. 7. Die photo of the single-to-differential LNA.

to ensure the M3 and M4 are operating in the same condition. In a simple analysis, the g_m of M2 and M3 should be the same to bring about the same small current through the load[1]. But from another point of view, we should reduce the g_m of M2 to make sure M2 produces a unity gain. In fact, the loads after the drains of M2 and M3 are different, so even if the same g_m in M2 and M3 produce the same small current, the loads split a different current from M2 and M3. So according to

$$g_m = \sqrt{2I_D\mu_n C_{ox}\frac{W}{L}}, \tag{5}$$

the W/L ratio of M2 should be properly reduced.

4. Experimental results

Based on the analysis above, the width of the inductor is set to 15 μm and the number of round is set to 5, to maximize the coupling coefficient k. Much effort is taken to make sure that the layouts of the differential output branches are as symmetrical as possible. For measurement purposes, source followers as buffers are added to the outputs to achieve 50 Ω output impedance, which is not needed in practice. Fabricated and packaged LNAs are tested on board. Figure 7 shows the die photo of the LNA.

The noise figure is measured with the help of an Agilent N9020A spectrum analyzer and an Agilent N4000A noise source. Figure 8 shows that the measured noise figure (NF), which has been raised by the noise introduced by buffers, is 3.2 dB at 1575.42 MHz. S-parameters are measured by means of an Agilent E5062A network analyzer. As shown in Fig. 9, S_{11} is better than −12 dB and S_{12} is better than −40 dB. Single-ended input to differential output S-parameter gains S_{ds21}, are measured at the two outputs respectively. Figure 10 shows they are both around 15.5 dB. In practice, the LNA will be followed by an on-chip mixer with a voltage-type input, and matching to 50 Ω at the outputs is not needed. The most meaningful gain parameter then is the voltage gain. To convert S_{ds21} into a voltage gain, 9 dB should be added[2], so the voltage gain is 24.5 dB. Figure 10(a) shows that the amplitude imbalance is less than 0.2 dB. Figure 10(b) shows the phase imbalance is less than 2 degrees. The LNA consumes 8.4 mA totally, including 3.2 mA consumed by buffers.

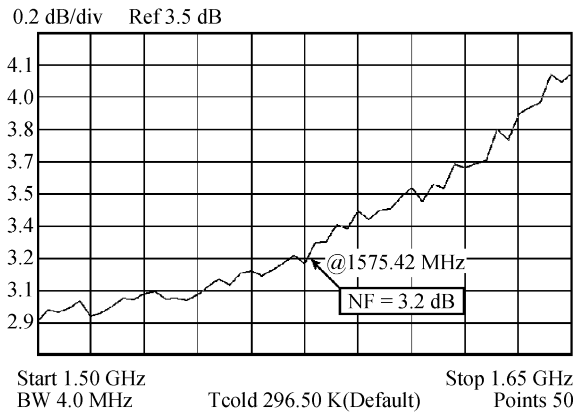


Fig. 8. Measured noise figure.

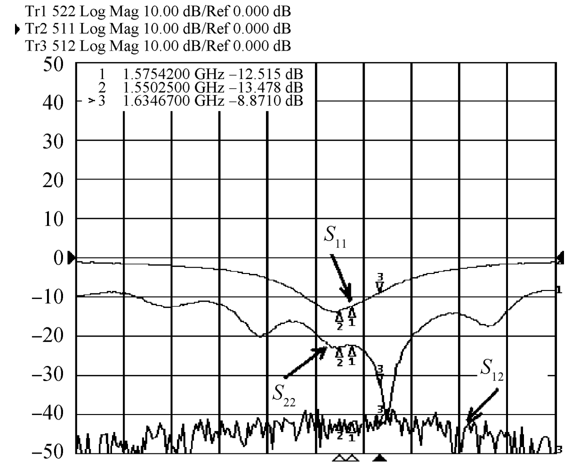
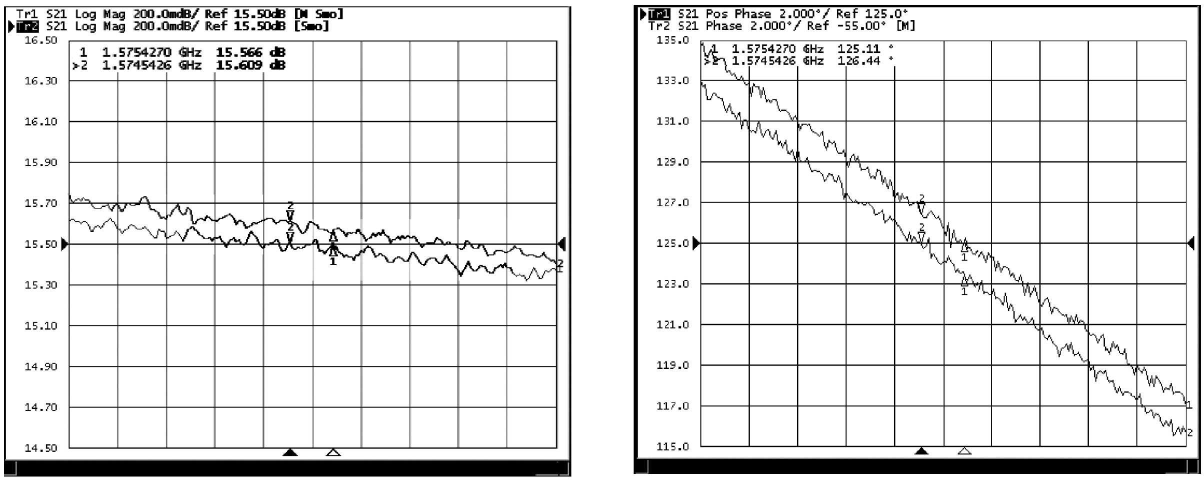


Fig. 9. Measured S_{11}, S_{22} and S_{12}.



(a) S_{21} and amplitude imbalance (b) Phase imbalance

Fig. 10. Measured S_{21}, amplitude and phase imbalance.

Table 1. Performance summary and comparison.

Specification	This work	Ref. [2]	Ref. [3]	Ref. [4]	Ref. [5][4)]	Ref. [9][5)]
Tech (μm)	0.18	0.065	0.13	0.18	0.13	0.18
Freq (GHz)	1.575	0.2–5.2	2–5.2	1.575	2.4	0.8–2.5
S_{11}	–12	–10	–9	–14	–12.5	—
Gain (dB)	24.5	15.6	16	20	30	−4
NF (dB)	3.2	3.5	4.7–5.7	2.3[1)]	2.6	4
Power (mW)	9	21	38	18[2)]	3.6	—
Amp. Imbal. (dB)	$<$ 0.2	$<$ 0.7	0.9	0.2[3)]	0.4	0.4
Phase Imbal. (degree)	$<$ 2	2	4–10	0.5[3)]	2	3.2

[1)]Estimated, not measured directly; [2)]Including mixer and PGA; [3)]Simulation results; [4)]Post simulation results; [5)]Passive balun.

Table 1 shows a comparison of the LNA with other single-to-differential LNAs or passive baluns reported. The proposed single-to-differential LNA shows a better amplitude and phase imbalance.

5. Conclusion

In this paper, we proposed a single-to-differential LNA with a low differential output imbalance exploiting inductive coupling. The proposed LNA exhibits an amplitude imbalance of less than 0.2 dB and a phase imbalance of less than 2 degrees, comparable to a passive external balun. Fabricated with a 0.18 μm TSMC process, the die occupies an area of 1.17 mm^2 with ESD pads. The LNA exhibits a noise figure of 3.2 dB, S_{11} less than –12 dB and S_{22} less than –20 dB. The voltage gain is 24.5 dB. Totally the circuit draws 8.4 mA from a 1.8 V supply, including 3.2 mA due to buffers only for test purposes.

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