Multi-Band Combined LNA and Mixer

Miguel A. Martins, Jorge R. Fernandes, and Manuel M. Silva
Instituto Superior Técnico/INESC-ID
Rua Alves Redol, 9; 1000-029 Lisboa, Portugal
Emails: {miguel.martins; jorge.fernandes; manuel.silva}@inesc-id.pt

***Abstract*—In this paper we investigate the performance of the multi-band low noise amplifier (LNA) based on the concept of using multiple cascode transistors for band selection; thus, avoiding switches in the signal path. We also show that this circuit can perform mixing of two signal bands with two different local oscillator frequencies, which provides great flexibility in the design of dual-band RF front-ends. The operation of both as LNA and as combined LNA/mixer is demonstrated by simulation.**

I. Introduction

Conventional receiver architectures use a single narrowband low noise amplifier (LNA) which amplifies the input signal with minimum addition of noise, and has input impedance matching for maximum power transfer. The demand for increased functionality, such as reception with different wireless standards, has led to the research of multi-band LNAs [1]–[3].

In [3], a CMOS multi-band LNA was obtained by using a cascode stage with multiple output branches tuned to the different frequency bands; each branch has a cascode transistor that is used to select the branch into operation. In this paper we investigate the performance of a multi-band LNA based on this principle. We also show that the same circuit can perform mixing if the cascode transistors are driven by different local oscillators.

In section II we present the multi-band LNA/Mixer circuit. In section III we consider the performance of the circuit as a dual-band LNA and in section IV we demonstrate the operation as a dual-band mixer. In both sections III and IV simulations are presented to demonstrate the feasibility of the circuits. Finally, in section V some conclusions are drawn.

II. The Combined LNA/Mixer Circuit

Most narrowband CMOS LNAs are realized by the cascode stage with inductive degeneration shown in Fig. 1 [4], [5].

The inductive degeneration makes possible a real input impedance and signal gain in a required frequency band, without significant noise addition. Cascode transistor M_2 is used in this circuit to reduce the Miller effect due to the gate-drain capacitance of M_1 (C_{gd1}). When the effect of C_{gd1} is neglected, the input impedance can be written as:

$$Z_{in} = \frac{g_{m1} L_S}{C_t} + \frac{1}{sC_t} + sL_S \quad (1)$$

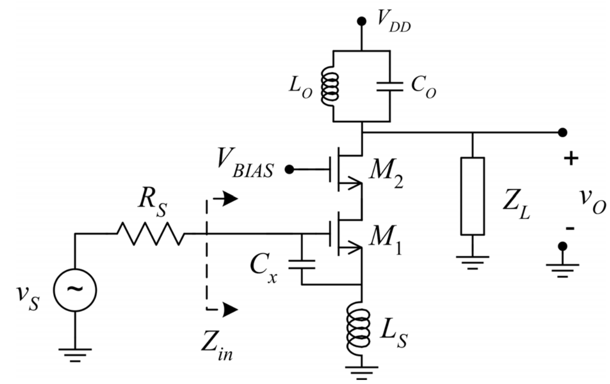

Fig. 1. Cascode LNA with inductive degeneration (biasing of M_1 not represented).

where $C_t = C_{gs1} + C_x$. When C_t and L_S resonate, the input impedance is real. Other advantages of using M_2 are improved reverse isolation between input and output, high output resistance, and better noise performance. This circuit has been extensively studied [6]–[8].

The circuit to be studied in this paper is represented in Fig. 2, with two output branches tuned to different frequencies and two cascode transistors. It should be noted that it is possible to have circuits with more than two cascode transistors and output branches.

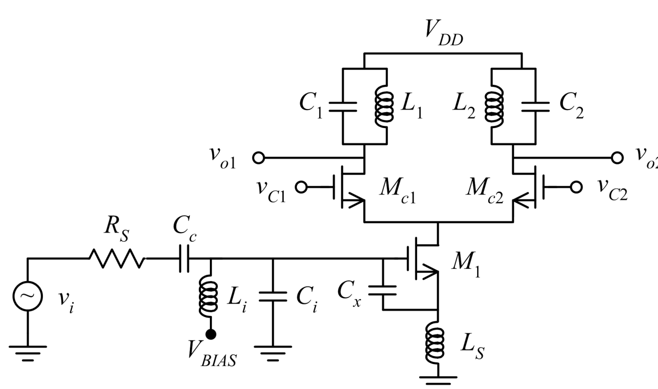

Fig. 2. Dual-band LNA/mixer.

The circuit has wideband input matching [9], covering all the wanted input frequency bands. The input matching network

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includes inductor L_i and capacitor L_i. The lower and upper 3 dB cutoff frequencies, ω_{lower} and ω_{upper} are:

$$\begin{cases} \omega_{lower} = \dfrac{R_{in}}{L_1} \\ \\ \omega_{upper} = \dfrac{1}{R_{in}C_1} \end{cases} \tag{2}$$

where $R_{in} = \frac{g_{m1}L_S}{C_t}$.

The circuit of Fig. 2, having only one input, can be connected to a multi-band antenna.

III. Operation as Dual-Band LNA

When the circuit of Fig. 2 is used as a dual-band LNA, each cascode transistor can be used to switch off the corresponding output branch by lowering the gate voltage (which is usually constant and typically equal to the supply voltage in the conventional circuit of Fig. 1). It is also possible to activate both outputs simultaneously.

When several input signal bands are present, each output network is only sensitive to the frequency band for which it is tuned. When only one cascode transistor is on, the LNA works as a normal single-band LNA and, apart from parasitic coupling effects, the inactive outputs will not affect the active one. This LNA does not have switches in the signal path, because each output is activated by the corresponding cascode transistor.

When both outputs are activated, the gain of one output can be controlled simply by varying the cascode transistor gate voltage (and consequently varying its quiescent current). This is an interesting feature that appears without the need of additional circuit elements.

To demonstrate the performance of the circuit of Fig. 2 as a dual-band LNA, a circuit was designed using AMS 0.35 μm CMOS technology with 3 V supply. The frequencies of 900 MHz and 1.8 GHz are chosen because they are two widely used frequencies. The cascode transistors have the same dimensions, so their bias currents are the same when both outputs are activated by equal gate voltages. All transistors are designed with the minimum length to improve the transition frequency. The output inductor series resistance affects the voltage gain characteristic, so the output network inductors L_1 and L_2 and their series resistances correspond to integrated spirals provided by the AMS technology. Inductors L_S and L_i of the input network are assumed to be ideal. The values used in the simulations are listed in table I. The circuit consumes 9 mA.

Fig. 3 shows the voltage gain at the two outputs, when only one output is active, i.e., when M_{c1} works in saturation and M_{c2} is cutoff, or vice-versa. The voltage gain has a peak at approximately the desired frequency, determined by the resonance of the corresponding tank circuit. There is no significant influence from the inactive output. If a cascode transistor is cutoff, the voltage gain at the corresponding output is below -70 dB for all the frequency range, which is low enough to consider that output inactive.

TABLE I
Parameter values used in simulations and to obtain the theoretical curves

Parameter	Value	Parameter	Value	
W_1	800 μm	L_S	1.2 nH	
W_{c1}	36 μm	L_1	6.4 nH	(r_1=7 Ω)
W_{c2}	36 μm	C_1	1.1 pF	
L_i	9.0 nH	L_2	13.2 nH	(r_2=14 Ω)
C_i	700 fF	C_2	2.2 pF	
C_x	1.2 pF	g_{m1}	101 mS	
c_{gs1}	743 fF	R_{g1}	4.5 Ω	

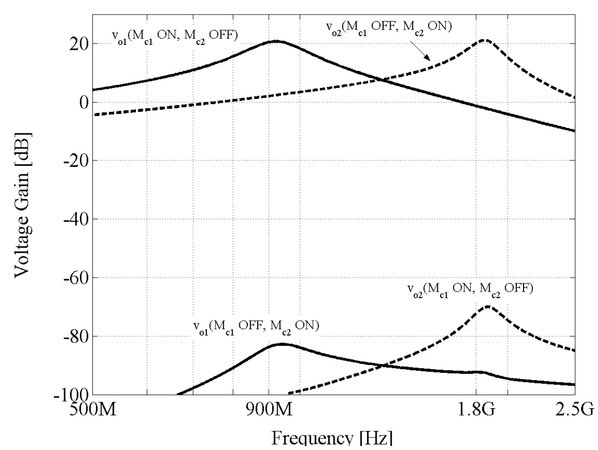

Fig. 3. Comparison of voltage gain at active and inactive outputs.

When both outputs are active, the gain at each output drops approximately 6 dB with respect to the case where only one output is active (the gain is reduced to a half). Since both cascode transistors are equal, the current from M_1 is divided equally by the two output branches.

Fig. 4 represents the voltage gain at output 1, when v_{c1} is varied. This shows that it is possible to vary significantly the gain, by varying the cascode gate voltages.

The input matching, measured by parameter S_{11}, is represented in Fig. 5 for three different operating modes (both outputs active, or only one). In all three cases S_{11} is below -10 dB at the two frequencies of interest, without significant differences between the three curves. This means that the input impedance is not significantly affected by the different operating modes.

In Fig. 6, the LNA noise figure (NF) obtained by simulation is compared with the theoretical value. The parameter values

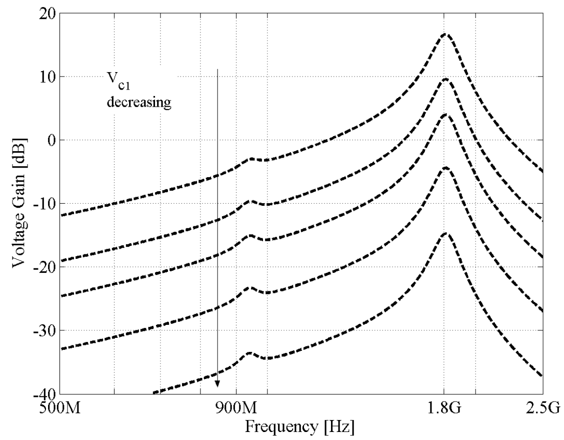

Fig. 4. Voltage gain at output 1 when V_{C1} varies and $V_{C2} = V_{DD}$.

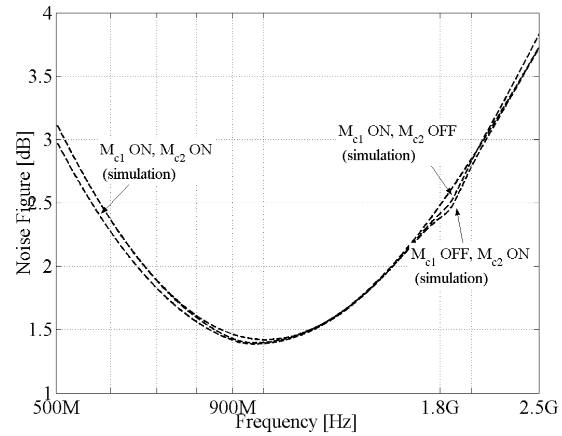

Fig. 6. Noise figure in three different operating modes.

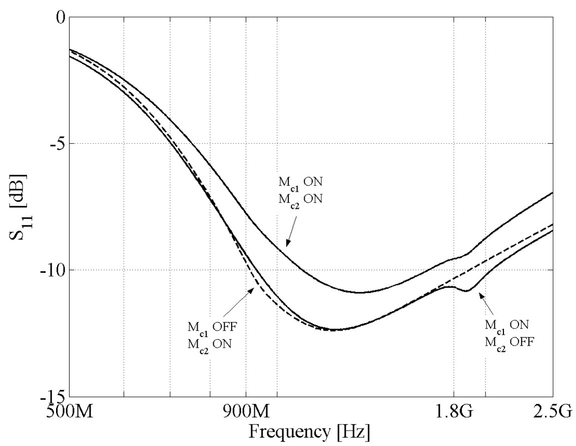

Fig. 5. S_{11} plots in three different operating modes.

used to determine the NF curves are those in table I. The noise figure at 900 MHz and 1.8 GHz is, respectively, 1.7 dB and 2.5 dB, which is well below the 3 dB reference value, giving a margin for other noise sources to be considered. The difference between the theoretical and simulated curves is small, due to the high transconductance of M_1, that attenuates significantly the influence of other noise sources such as those of the cascode transistors. There is no significant difference between the different modes of operation. This is due to the much higher transconductance of M_1 with respect to that of the cascode transistors.

IV. Operation as a Dual-Band Mixer

The circuit of Fig. 2 can also be used to obtain dual-band mixing by making v_{C1} and v_{C2} the voltages of two local oscillators with different frequencies f_{L1} and f_{L2}. In this mode of operation the circuit of Fig. 2 can be viewed as a 2-quadrant variable transconductance multiplier, composed of a differential pair (M_{c1}, M_{c2}) and a controlled current source M_1; transistors M_{c1} and M_{c2} must remain in the active zone of operation.

If the input signal v_i has two frequency bands centered at f_{i1} and f_{i2}, the currents of M_{c1} and M_{c2} will contain 8 modulation products centered at frequencies $f_{L1} \pm f_{i1}$, $f_{L1} \pm f_{i2}$, $f_{L2} \pm f_{i1}$, $f_{L2} \pm f_{i2}$. Two of these can be selected to appear at the outputs (v_{o1} and v_{o2}) by tuning the output branches to the desired frequencies. This provides a great design flexibility, since the two frequencies are chosen from a set of eight.

To demonstrate this mode of operation, a dual-band mixer was designed. As in the previous section, the frequencies of 900 MHz and 1.8 GHz are choosen for the RF input signal. The LO frequencies are 800 MHz and 1.6 GHz (these can be obtained by a single oscillator followed by a "frequency divider by 2"). The output IF frequencies are 100 MHz and 200 MHz, obtained by using the following element values: $C_1 = 100$ pF, $L_1 = 25.3$ nH, $C_2 = 80$ pF, and $L_2 = 7.9$ nH. The remaining parameters have the values listed in table I. In the simulations, the RF signals have -20 dBm of power and the LOs have 40 mV of voltage amplitude.

In figs. 7 and 8 the spectra of the outputs are represented. It is visible that at each output the corresponding IF frequency is dominant, and it is much stronger (more than 45 dB) then the nearest unwanted mixing frequency. There are strong components at the input and LO frequencies, but they are far away and can easily be suppressed by low pass filtering.

V. Conclusion

We have evaluated the performance of a dual-band LNA based on a cascode stage with multiple outputs, each one with a cascode transistor and a tuned output branch [3]. This circuit has a common input that can be connected to a single antenna, and has the advantage that there are no switches

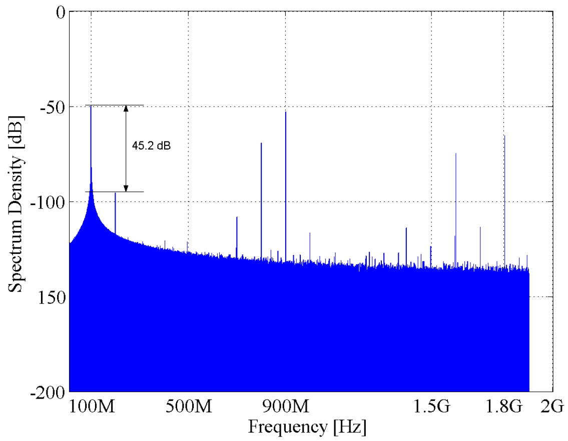

Fig. 7. Spectral density at output 1.

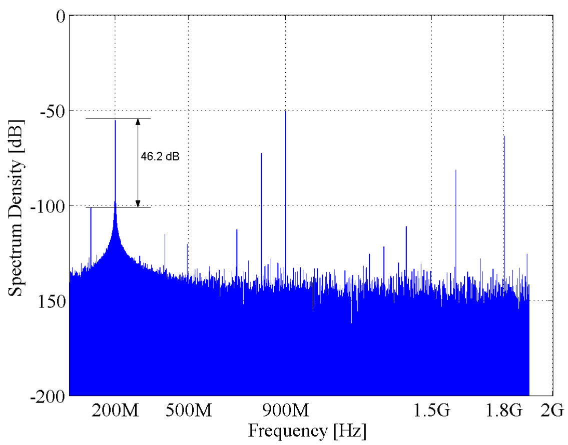

Fig. 8. Spectral density at output 2.

in the signal path, since the band selection is performed by transistors that are either switched off or operate as normal cascode transistors.

We have shown that, when one output is deactivated by switching off the corresponding cascode transistor, the effect of parasitic couplings on the active output is negligible. When both outputs are active, their voltage gain can be varied by changing the biasing of the cascode transistors. The input matching and the noise figure are very insensitive to whether one, the other, of both outputs are activated.

We also show that this LNA circuit can perform multi-band mixing by driving the cascode transistors with different local oscillators. There is great flexibility in the choice of intermediate and local oscillator frequencies. This flexibility can be used to ease the filtering out of the unwanted mixing frequencies. The resulting combined LNA/Mixer can be used to obtain a low area and low power RF front-end.

ACKNOWLEDGMENT

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REFERENCES

[1] S. Wu and B. Razavi, "A 900-MHz/1.8-GHz CMOS Receiver for Dual-Band Applications", *IEEE Journal of Solid-State Circuits*, vol.12, no. 12, pp. 2178-2185, December 1998.
[2] A. Hajimiri and H. Hashemi, "Concurrent Multiband Low-Noise Amplifiers - Theory, Design and Applications", *IEEE Transactions on Microwave Theory and Tecniques*, vol. 50, no. 1, pp. 288-301, January 02.
[3] C. Garuda and M. Ismail, "A Multi-Band CMOS RF Front-End for 4G WiMAX and WLAN Applications", In *Proc. IEEE International Symposium on Circuits and Systems (ISCAS 2006)*, vol. 1, pp. 3049-3052, May 2006.
[4] Thomas H. Lee, *The Design of CMOS Radio-Frequency Integrated Circuits*, Cambridge University Press, 2004.
[5] D. Shaeffer and T. Lee, "A 1.5-V, 1.5-GHz CMOS Low Noise Amplifier", *IEEE Journal of Solid-State Circuits*, vol. 32, no. 5, pp. 745-759, May 1997.
[6] J. Janssens P. Leroux and M. Steyaert, "A 0.8-dB NF ESD-Protected 9-mW CMOS LNA Operating at 1.23 GHz", *IEEE Journal of Solid-State Circuits*, vol. 37, no. 6, pp. 760-765, June 2002.
[7] D. J. Cassan and J. R. Long, "A 1-V Transformer-Feedback Low-Noise Amplifier for 5-GHz Wireless LNA in 0.18-μ m CMOS", *IEEE Journal of Solid-State Circuits*, vol. 38, no. 3, pp. 427-435, March 2003.
[8] L. Belostotski and J. W. Haslett, "Noise Figure Optimization of Inductively Degenerated CMOS LNAs with Integrated Gate Inductors", *IEEE Transactions on Circuits and Systems - I: Fundamental Theory and Applications*, vol. 53, no. 7, pp. 1409-1422, July 2006.
[9] A. Ismail and A.A. Abidi, "A 3-10 GHz Low-Noise Amplifier with Wideband LC-Ladder Matching Network", *IEEE Journal of Solid-State Circuits*, vol. 39, no. 12, pp. 2269-2277, December 2004.