Design of a 1-V 90-nm CMOS Folded Cascode LNA for Multi-Standard Applications

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Abstract—This paper analyses the use of folded cascode Low Noise Amplifiers (LNAs) for the implementation of multi-standard wireless transceivers. The proposed LNA consists of a two-stage topology made up of a folded cascode and simple-stage amplifiers that use NMOS-varactor based tuning networks to make the operating frequency continuously programmable. The circuit has been designed and implemented in a 90-nm CMOS technology in order to cope with the requirements of GSM, WCDMA, Bluetooth and WLAN (IEEE 802.11b/g) standards. Practical design issues are analysed, considering the effect of circuit parasitics associated to both the chip package and integrated inductors, capacitors and varactors; as well as technology parameter deviations. The circuit design is optimized using genetic algorithms to achieve the required specifications with adaptive power consumption. Layout-extracted simulation results demonstrate a correct operation of the proposed circuit, showing a continuous tuning of Noise Figure (NF) and S -parameters within the 1.85-2.48GHz band, featuring $NF < 3.8\text{dB}$, $S_{21} > 12\text{dB}$ and $IIP3 > -12\text{dBm}$, with an adaptive power dissipation between 13.3mW and 23.1mW from a 1-V supply voltage.¹

I. INTRODUCTION

The increasing number of personal wireless applications demands for Radio Frequency (RF) front-ends capable to handle different standard specifications, signal conditions and battery status [1]. The trend is towards a maximum hardware reuse, by making as many transceiver building blocks as possible, digitally programmable, reconfigurable and compatible with mainstream nanometer CMOS technologies [2].

Among other RF circuits, the design of the Low Noise Amplifier (LNA) is specially critical, due to its early position at the very beginning of the receiver chain, what makes this block a determinant factor in the overall system performance. The LNA has to simultaneously match the antenna and to amplify weak input signals with reduced noise contribution, high linearity and isolation from the rest of the receiver chain. This problem is aggravated in the case of multi-standard applications, in which LNAs must operate over different frequency ranges, whereas keeping reduced number of passive circuit elements to increase integration [1].

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The majority of multi-standard LNAs are based on the use of switchable passive networks to select the resonance frequency, thus preserving immunity to out-of-band interferers, although only one signal band is received at one time [3], [4]. Besides, the use of switches forces a discrete frequency selection and introduces parasitic switch-on resistances and capacitances, with the subsequent trade-off between linearity, noise and speed. This limitation can be partially solved by using concurrent multi-band LNAs [5], [6]; which allows a simultaneous reception of multiple signal bands without using switches. However, the spurs in one band may corrupt signals in other band due to the LNA non-linear operation.

A common issue in most reported multi-standard LNAs is the increase of the number of integrated passive elements (basically capacitors and inductors) as compared to their mono-standard counterparts, not offering a clear advantage with respect to using separate mono-standard LNAs [1], [2]. Moreover, most circuit topologies are based on stacked amplifiers (usually simple cascode) in either common-source or common-gate configuration, which are not very suited for low-voltage operation. Recently, the use of folded cascode topology has been applied to the design of LNAs, although very little has been done for multi-standard applications [7].

This paper contributes to this topic and presents the design, layout implementation and electrical characterization of a reconfigurable two-stage folded cascode LNA in a 90-nm CMOS technology. The circuit is based on NMOS-varactor tuning networks to make the resonance frequency continuously programmable without needing any switches. An optimized design methodology has been used, that takes into account realistic equivalent circuit models for passive elements as well as the impact of package and technology parasitics. Post-layout simulations are shown to validate the presented approach.

II. PROPOSED RECONFIGURABLE FOLDED CASCODE LNA

Fig. 1 shows the schematic of the proposed reconfigurable LNA including the matching networks and biasing circuits. It consists of a two-stage topology with separate tuning networks. The input stage is a folded cascode amplifier formed by transistors M_{n1} and M_{p2} and inductor L_d , which resonates with interstage parasitic capacitance, thus providing the necessary high-impedance at the operation frequency. Inductor L_s

Fig. 1. Schematic of the proposed reconfigurable folded cascode LNA.

is used to implement source degeneration whereas L_o is the load impedance. Folded cascode amplifier is very suited for low-voltage applications – since bias voltage can be reduced to one transistor overdrive – with reasonable good linearity, stability and Noise Figure (NF) [7]. Thus, in order to get the required forward gain, S_{21} , an additional amplifier stage – formed by PMOS transistor M_{p3} and load inductor L_{o2} – is used. The circuit is biased by dc voltages V_{b1} and V_{b2} , which are respectively connected at the gates of M_{n1} and M_{p2} through the impedances Z_{b1} and Z_{b2} – implemented by diode-connected MOS transistors.

As the LNA is fully integrated as a stand-alone circuit, a termination of 50Ω is needed at both the input and the output terminals connected at the bonding PADs (see Fig. 1). For this purpose, C_i and L_g are used to provide the required input impedance, Z_{in} , whereas decoupling capacitors C_o and C_{op} are used at the output node. The tuning mechanism of the LNA is achieved by varying the resonance frequencies of the different passive networks, through accumulation NMOS varactors, C_{t1-3} .

A. Effect of Circuit Parasitics

The schematic in Fig. 1 assumes ideal circuit elements, particularly integrated capacitors, varactors and inductors. In practice, circuit parasitics cause these elements to behave as RLC networks that may severely degrade the performance of the LNA [8]. This is specially important in the case of reconfigurable/adaptive RF circuits like the one proposed in this paper, because of the many different specifications to be covered by the same circuit.

In order to evaluate the impact of circuit parasitics, the proposed LNA was analyzed by replacing every inductor and varactor in Fig. 1 with the equivalent circuit shown in Fig. 2 [8], [9]. In the case of the inductor (Fig. 2(a)), the nominal inductance is given by $L_{s1} + L_{s2}$, whereas the remaining elements are circuit parasitics: C_{o1-3} and R_{b1-3} are respectively substrate capacitances and resistances; R_{s1-2} are series resistances and C_{s1-2} are embedded coupling capacitances. In the case of varactors (Fig. 2(b)), C_p is the nominal capacitance and C_c is the variable voltage controlled capacitance, whereas the rest of circuit elements are parasitics, with L_{1-2} and R_{p1-2} being series inductances and resistances, respectively, and R_w is the well parasitic resistance.

B. Circuit analysis

Replacing the inductors and varactors in Fig. 1 with the RLC equivalent circuits shown in Fig. 2, it can be shown that,

Fig. 2. Equivalent RLC circuit for (a) inductors and (b) varactors.

after some approximations, the input impedance (Z_i) and the noise factor (F) of the proposed LNA are respectively given by:

$$Z_i(s) \simeq \frac{Z_{i2}(g_{mn1} + sC_{gsn1}) + Z_{i4}}{sC_i + sC_i Z_{i2}(g_{mn1} + sC_{gsn1}) + sZ_{i1}Z_{i4}} \quad (1)$$

$$F \simeq 1 + \left(\frac{1}{g_{mn1}R_{RF}} \right) \left(\frac{R_{RF} - Z_i}{sC_i Z_{b1} Z_i} \right)^2 \cdot \left(\frac{\gamma}{\alpha} - \frac{s^2 C_{gsn1}^2 Z_{i1}^2 \alpha \delta}{5} + (2|c|sC_{gsn1}Z_{i1}) \sqrt{\frac{\gamma\delta}{5}} \right) \quad (2)$$

where

$$\begin{aligned} Z_{i1}(s) &\simeq \frac{C_{ct1} + C_{pt1}}{1 + 2sR_{p1t1}(C_{ct1} + C_{pt1})} \\ Z_{i2}(s) &\simeq \frac{4(R_{s1s} + sL_{s1s})(2 + 5sC_{o3s}R_{b3s})}{4 + sC_{o3s}(10R_{b3s} + 3(R_{s1s} + sL_{s1s}))} \\ Z_{i3}(s) &\simeq \frac{4(R_{s1g} + sL_{s1g})(2 + 5sC_{o3g}R_{b3g})}{4 + sC_{o3g}(10R_{b3g} + 3(R_{s1g} + sL_{s1g}))} + Z_{i2} \\ Z_{i4}(s) &\simeq sC_i Z_{b1}(1 + g_{mn1}Z_{i2} + sC_{gsn1}Z_{i3}) \\ \omega_0 &\simeq \frac{1}{\sqrt{(L_g + L_s)C_{gsn1}}}; \quad Z_1 = Z_{b1} + Z_{i2} + sC_{it1}Z_{i3}Z_{b1} \end{aligned} \quad (3)$$

with c being the correlation factor; δ and γ are technology parameters; R_{RF} is the RF source resistance; $\alpha = g_{mn1}/g_{dsn1}$; and g_{dsn1} , g_{mn1} and C_{gsn1} are respectively the small-signal drain-source conductance, transconductance and gate-source capacitance of M_{n1} . Parasitic resistors, capacitors and inductors (R_p , R_s , C_p , etc.) are labelled accordingly to the name of their associated varactor/inductor in Fig. 1.

The real part of Z_i is chosen to be equal to R_{RF} – usually 50Ω – in the band of interest, 1.85-2.48GHz in this case. In order to achieve proper impedance matching within this band, inductor L_g and varactor C_{t1} are properly sized to get the required frequency tuning. This is illustrated in Fig. 3 by showing the trajectories of the values of these elements towards impedance matching within the band of interest, with Z_{si} being the input impedance seen at the right of L_g .

The same tuning mechanism – based on the use of varactors $C_{t2,3}$ – is used to vary the output impedance and the voltage gain, approximately given by:

Fig. 3. Input impedance matching trajectories in the Smith chart.

$$A_v \simeq \left(\frac{g_{mn1} Z_{v2} Z_{v3} (g_{mp3} + s C_{gsp2} g_{dsp2} Z_{b2})}{g_{mp2} Z_{v2} + n_2 + n_1 (1 + s C_{gsp2} (Z_{b2} + Z_{v2}))} \right) \cdot \left(\frac{g_{mp3} R_L Z_{o2}}{R_L + (1 + s R_L Z_{o1}) Z_{o2}} \right) \quad (4)$$

where

$$\begin{aligned} Z_{o1}(s) &\simeq \frac{C_{ct3} + C_{pt3}}{1 + 2sR_{p1t3}(C_{ct3} + C_{pt3})} \\ Z_{o2}(s) &\simeq \frac{4(R_{s1o2} + sL_{s1o2})(2 + 5sC_{o3o2}R_{b3o2})}{4 + sC_{o3o2}(10R_{b3o2} + 3(R_{s1o2} + sL_{s1o2}))} \\ Z_{v1}(s) &\simeq \frac{C_{ct2} + C_{pt2}}{1 + 2sR_{p1t2}(C_{ct2} + C_{pt2})} \\ Z_{v2}(s) &\simeq \frac{4(R_{s1d} + sL_{s1d})(2 + 5sC_{o3d}R_{b3d})}{4 + sC_{o3d}(10R_{b3d} + 3(R_{s1d} + sL_{s1d}))} \\ Z_{v3}(s) &\simeq \frac{4(R_{s1o} + sL_{s1o})(2 + 5sC_{o3o}R_{b3o})}{4 + sC_{o3o}(10R_{b3o} + 3(R_{s1o} + sL_{s1o}))} \\ n_1 &= 1 + sZ_{v1}Z_{v3}; \quad n_2 = g_{dsp2}Z_{v3}(1 + sC_{gsp2}Z_{b2}) \end{aligned} \quad (5)$$

III. DESIGN METHODOLOGY

The proposed LNA has been designed to fulfill the requirements of four wireless standards: GSM, Bluetooth, WCDMA and WLAN (IEEE 802.11b/g), including continuously-tuning operation within the band of interest. In order to cope with the different sets of specifications, the design equations described in previous section were used to find out an initial set of design parameters, as well as proper biasing to guarantee all transistors operating in the saturation region.

The initial design parameters are used as start point of an optimized design process, which combines electrical simulations using Cadence SpectreRF with genetic-based algorithms [10] in order to find out the best design parameters that fulfill the required specifications with minimum power consumption. For that purpose, the circuit in Fig. 4 was used. This circuit includes the package and the external components to be included in the PCB. A 4mmx4mm 12-pin QFN plastic package has been used. This package has been modeled using Cadence PKG tool in order to take into account their associated parasitics during the design process.

Fig. 5(a) shows the flow diagram of the optimization process, where after a random variation of one of the design

Fig. 4. Equivalent circuit including chip-package parasitics.

Fig. 5. Optimization procedure: (a) Flow diagram. (b) Adaptation function.

parameters – referred to as *mutation* – a set of simulations are performed to evaluate the performance of the LNA. The data extracted from simulations are used to evaluate a cost function – called *adaptation* – defined as:

$$\begin{aligned} adp &\simeq |S_{11i} - S_{11r}| + |S_{21i} - S_{21r}| + |S_{22i} - S_{22r}| \\ &\quad + |NF_i - NF_r| + |f_{cS11i} - f_{cS11r}| \\ &\quad + |f_{cS21i} - f_{cS21r}| + |f_{cS22i} - f_{cS22r}| \\ &\quad + |f_{cNF_i} - f_{cNF_r}| + (S_{11max} - S_{11min}) \\ &\quad + (S_{21max} - S_{21min}) + (S_{22max} - S_{22min}) \\ &\quad + (NF_{max} - NF_{min}) \end{aligned} \quad (6)$$

where the subscripts *i* and *r* denote the ideal and real values of a given parameter; and f_c stands for the central frequency of the frequency response of that parameter. For instance, f_{cNF_i} is the central frequency of the ideal NF function. Once *adp* is computed for a given set of design parameters, a new mutation is generated and the process is repeated again until a minimum value of *adaptation* is found as illustrated in Fig. 5(b). In order to find out a global minimum, local minima are stored in a best-designs table, which *remembers* these minima in order to avoid them in ulterior mutations.

The outcome of the design procedure described above is the sizing and biasing of the LNA – summarized in Table I. As stated above, the performance of the circuit is adapted to the different standards specifications by varying the values of varactor capacitances – also shown in Table I.

TABLE I
LNA SIZING

Capacitors (pF)	Inductors (nH)	Transistors	W/L ($\mu\text{m}/\mu\text{m}$)
$C_i = 1.9; C_o = 2.5$	$L_d = 3.13$	M_{n1}	113/0.3
$C_{op} = 0.48$	$L_g = 5.60$	M_{p2}	455/0.3
$C_{t1} = (0.08 - 0.33)$	$L_s = 1.23$	M_{p3}	32/0.3
$C_{t2} = (0.08 - 0.31)$	$L_o = 4.60$		
$C_{t3} = (0.10 - 0.35)$	$L_{o2} = 1.95$		

IV. LAYOUT AND SIMULATION RESULTS

The LNA has been implemented using a 90-nm CMOS technology with a single 1-V supply voltage. Fig. 6 shows the layout of the chip highlighting their main parts, namely integrated inductors, capacitors, NMOS varactors and CMOS active area. Integrated inductors have octagonal shape with a patterned ground shield. Input/output capacitors are implemented by M-O-M structures, which are based on the combination of stacked and finger metal-metal capacitors. All pads are ElectroStatic Discharged (ESD) protected. The die area, including pads, is 0.64mm².

Fig. 7(a) represents NF vs. input frequency for all standards under study. The overall minimum value of NF is 3.1 dB, obtained at 2.44 GHz, which corresponds to the WLAN operation mode. Fig. 7(b)–(d) show respectively S_{21} , S_{11} and S_{22} . The minimum value of S_{21} within the band of interest is above 12 dB, corresponding to WCDMA, whereas S_{11} and S_{22} are below -8.5 dB for all standards.

The operation of the circuit can be continuously tuned within the band of interest. This feature – illustrated in Fig. 8 for S_{21} – is a direct consequence of using NMOS varactors in the resonance tuning networks, which provides approximately 700-MHz tuning range. Indeed, this is a peculiarity of the presented chip as compared to previously reported multi-standard LNAs – mostly based on using switchable LC tanks to select the operating frequency in a discrete way and using varactors to fine tuning the selected band. The difference in the presented design is that it achieves a coarse continuous frequency tuning capability, just using varactors. Finally, Table II sums up the simulated performance of the LNA by showing the values of the different performance figures for each standard.

CONCLUSIONS

The design and electrical implementation of a multi-standard 1-V 90-nm CMOS folded cascode LNA has been presented. The use of NMOS-varactor based tuning networks allows the amplifier to adapt its performance to the specifications of different wireless standards. An optimization-based design methodology has been used to find out the target requirements with adaptive power dissipation. Layout-extracted simulations including packaging circuit parasitics demonstrate a correct performance of the circuit.

Fig. 6. Layout of the LNA.

Fig. 7. Simulated NF and S-parameters.

Fig. 8. Illustrating the reconfiguration of S_{21} .

REFERENCES

- [1] M. Brandolini *et al.*, "Toward Multistandard Mobile Terminals - Fully Integrated Receivers Requirements and Architectures," *IEEE Trans. on Microwave Theory and Techniques*, vol. 53, pp. 1026–1038, 2005.
- [2] P.-I. Mak *et al.*, "Transceiver Architecture Selection: Review, State-of-the-Art Survey and Case Study," *IEEE Circ. and Syst. Mag.*, 2007.
- [3] V. Vidojkovic *et al.*, "Fully-integrated DECT/Bluetooth multi-band LNA in 0.18μm CMOS," *Proc. of IEEE ISCAS*, pp. 565–568, 2004.
- [4] Y. Koolivand *et al.*, "A new technique for design CMOS LNA for multi-standard receivers," *Proc. of IEEE ISCAS*, vol. 4, pp. 3231–3234, 2005.
- [5] H. Hashemi *et al.*, "Concurrent multiband low-noise amplifiers-theory, design, and applications," *IEEE Transactions on Microwave Theory and Techniques*, vol. 50, pp. 288–301, Jan 2002.
- [6] C.-W. Ang *et al.*, "A Multi-band CMOS Low Noise Amplifier for Multi-standard Wireless Receivers," *Proc. of IEEE ISCAS*, pp. 2802–2805, 2007.
- [7] H.-H. Hsieh *et al.*, "Gain-Enhancement Techniques for CMOS Folded Cascode LNAs at Low-Voltage Operations," *IEEE Transactions on Microwave Theory and Techniques*, vol. 56, no. 8, pp. 1807–1816, 2008.
- [8] T. H. Lee, *The Design of CMOS Radio-Frequency Integrated Circuits*. Cambridge University Press, 2 ed., 2004.
- [9] P. Viztmuller, *RF Design Guide*. Artech House, 1995.
- [10] M. D. Vose, *The Simple Genetic Algorithm: Foundations and Theory*. The MIT Press, 1999.

TABLE II
LNA PERFORMANCE

Standard	IIP3 (dBm)	NF (dB)	S_{21} (dB)	S_{11} (dB)	P_{DC} (mW)
GSM	-9.9	3.5	12.9	-8.5	23.1
WCDMA	-11.9	3.8	12.0	-9.3	16.0
Bluetooth	-10.5	3.3	13.2	-15.2	13.3
WLAN	-11.6	3.1	14.3	-15.4	16.0