A 900MHZ ACTIVE CMOS LNA WITH A BANDPASS FILTER

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ABSTRACT

A new active CMOS LNA with bandpass filtering is proposed. A NCG Q-enhancement technique is utilized to extend Q tuning range. By employing standard 0.5μm CMOS technology, center frequencies can be tuned between 559MHz to 970MHz with Q larger than 400. A LNA is designed and simulated for NADC and AMPS wireless standards to achieve a center frequency at 881MHz with Q=34, 15.7dB voltage gain, -12.4dBm IIP3, 6dB NF, and 52.5mW power dissipation.

1. INTRODUCTION

The increasing command for wireless communication necessitates low lost, high integration level, and high performance solutions to be implemented in radio frequency (RF) front-end systems. While GaAs and BiCMOS technologies are employed to achieve higher operating frequencies and low noise performance, considerable research has been conducted in developing CMOS monolithic RF systems as it would allow integration of other digital signal processing circuits, leading to a single-chip solution [1]. In the heterodyne receiver architecture, the development of a low noise amplifier (LNA) with a monolithic bandpass filter has received intense attention, and several proposed approaches have been reported in literatures [2], [3]. The most popular approach is to employ on-chip passive inductors in shunt with capacitors as bandpass filtering elements. The shortcomings of passive monolithic inductors are well known as larger required area and low inductor quality factor (Q) due to resistive loss and capacitive coupling to the substrate. A negative resistor generated by a positive feedback configuration is usually used to compensate for the inductor low-Q characteristic [4], [5].

This paper describes an alternate configuration based on attributes of active devices to implement LNA and bandpass functions with features of low noise, high gain, RF band center frequencies, and tunable Q. Noise performance is further derived as a useful vehicle to minimize the noise generated by the circuit itself and following stages. To demonstrate the feasibility of this proposed circuit, simulation results are presented and show that it meets the requirements of current communication systems .

2. CIRCUIT TOPOLOGY

The bandpass filter illustrated in [6] shows that it has a simple second-order transfer function with a pair of complex poles. The shortcoming of this simple filter configuration is that the filter Q is usually smaller than 2, but it can be circumvented by introducing another compensation circuit, called negative conductance generator (NCG).

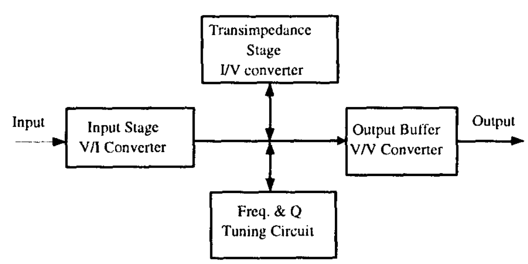

Figure 1: The block diagram of the circuit.

The form of the proposed circuit conceptually consists of four different signal processing stages, as illustrated in Figure 1. The first stage is a voltage-to-current converter. It features low noise and input matching property to antenna. The second stage is a transimpedance converter that transforms a current input signal to a voltage output signal at the same node. This is substantially different from the so-called the transimpedance amplifier, which has different input and output ports [7]. It features low driving input and output impedances at low frequencies, but exhibits very high impedance at the resonant frequency. The third stage features a center-frequency-and-quality-factor tuning structure, and a negative conductor is applied to electrically enhance and tune the filter characteristics. The final stage is a voltage buffer having an attenuated voltage gain, and delivering the voltage signal to the following stage. The resultant configuration emulates a biquad filter in the sense that it provides a second-order transfer function on the input/output characteristics.

Figure 2 shows the proposed LNA with a Q-enhancement circuit. In this figure, transistors M1 and M2 comprise the input amplifier stage. This common-gate configuration provides reasonable noise figure (NF) with a simple 50Ω input impedance matching and higher linearity in contrast to a common-source configuration without source degeneration. This common-gate approach also helps to increase the effective reverse isolation in heterodyne architectures due to the signal leakage of the local oscillator from the mixer to the antenna. This leakage arises from the capacitive path and substrate coupling.

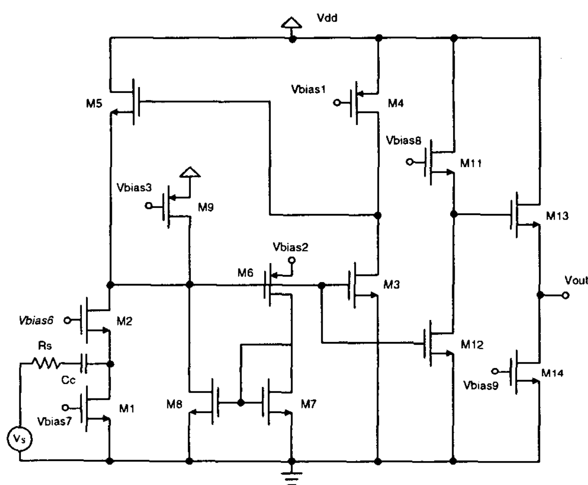

Figure 2: The proposed LNA with NCG circuit.

The transimpedance stage consists of transistors M3, M4, and M5. This negative feedback configuration constrains the gate of the transistor M3 to a very low impedance node, which is much lower than the input impedance of the common-gate amplifier. The effect of the feedback starts reducing while operating frequencies increase, and accordingly the impedance level at this node rises, thereby acting inductive. At higher frequencies, the impedance level drops due to the complex poles generated by the feedback circuit, and the impedance is capacitive at this range of frequencies. Therefore, this node exhibits resonant bandpass behavior across the full frequency range. It is worth noting that the drain of the transistor M3 is also a low impedance node due to the negative feedback path.

The NCG circuit is constructed by transistors M6 ~ M9,

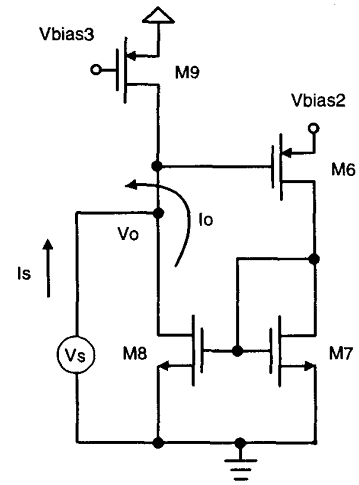

Figure 3: A negative conductance generator.

as illustrated in Figure 3. The principal idea is to generate an out-of-phase output current I_o respected to the input current I_s and feed it back to the applied input voltage V_s at the same node, called the positive feedback configuration. Transistor M6 is a transistor that provides transconductance gain g_{m6}. Transistors M7 and M8 form a simple current mirror, providing the out-of-phase function and gain if different (W/L) ratios are employed. Transistor M9 provides a required DC current through the NCG, transistor M5, and input stage.

Assuming $g_m \gg g_{ds}$ for all transistors and ignoring all non-dominant high-order terms, the conductance of the NCG can be expressed as

$$g_{neg} = -\frac{g_{m6}\, g_{m8}}{g_{m7}} \tag{1}$$

where g_{mi} is the transconductance of the transistor M_i.

This LNA is sensitive to the parasitic capacitances, and any following stage loading will modify the overall circuit response. To minimize the loading effect, an output buffer amplifier is attached to the output terminal of the transimpedance converter. Another merit of this output stage is that it also provides a range of gain or attenuation to suit many different application requirements. In wireless communication, the output power of the LNA needs to be large enough to compensate the loss of the image-rejection filter and reduce the noise contributed by the following stages, but it can not be so large as to overdrive the mixer and degrade the third-order input intercept point (IIP_3) of the system. In addition, the output port may need to exhibit 50Ω output impedance to drive the image-rejection filter. The transistor M12 is a common-source amplifier with the drain connected to the source of the transistor M11. It offers an attenuated gain and delivers the signal to a common-drain output buffer stage.

By ignoring all non-dominant high-order terms, the transfer function of the LNA with NCG is approximately equal to

$$A_v(s) = \frac{g_{m2}\, g_{m12}\, (c_2 + c_3)}{2\, g_{m11}\, c}\; \frac{s + D}{s^2 + s\,A + B} \quad (2)$$

where

$$\begin{aligned}
A &= \frac{c_1\, g_2 + c_2\,(g_1 + g_2 + g_{m3}) + c_3\,(g_1 + g_{m5})}{c} \\
B &= \frac{g_1\, g_2 + g_{m5}\,(g_2 + g_{m3})}{c} \\
D &= \frac{g_2}{c_2 + c_3} \\
c &= c_1\, c_2 + c_2\, c_3 + c_1\, c_3 \\
c_1 &= c_{gd2} + c_{db2} + c_{gs3} + c_{gb3} + c_{sb5} \\
&\quad + c_{gs6} + c_{gb6} + c_{db8} \\
&\quad + c_{gd9} + c_{db9} + c_{gs12} + c_{gb12} \\
c_2 &= c_{gd3} + c_{gs5} \\
c_3 &= c_{db3} + c_{gd4} + c_{db4} + c_{gd5} + c_{gb5} \\
g_1 &= \frac{g_{ds2}}{2} + g_{ds5} + g_{ds8} \\
&\quad + g_{ds9} - g_{neg} \\
g_2 &= g_{ds3} + g_{ds4}
\end{aligned}$$

c_1, c_2, and c_3 represent accumulated parasitic capacitances in the circuit. g_1 and g_2 represent the equivalent output conductances. A negative conductance is explicitly introduced in g_1 to circumvent the capacitive and resistive loading at the gate of the transistor M3.

The zero in Eq. 2 can be minimized by cascading another PMOS transistor to reduce g_2, or increasing channel length of the transistor M4. The center frequency of the bandpass filter ω_o is therefore given by

$$\omega_o = \sqrt{\frac{g_{m3}\, g_{m5}}{c}} \quad (3)$$

Eq. 3 reveals that the center frequency is determined by transconductances of the transistors M3, M5, and parasitic capacitances exhibited in the circuit. The transconductance of the transistor M5 is chosen as a prime candidate to be tuned by varying Vbias3 while Vbias1 remains constant and the transistor M2 is biased to provide 50Ω input matching. The Q tuning is accomplished by changing Vbias2 to tune g_{m6}, as shown in Eq. 1. An observation can be made from Eqs. 2 and 3 is that for the same ω_o the Q can be independently tuned by varying the g_{neg} without altering the center frequency. It is worth noting that an iterative tuning approach for Vbias2 and Vbias3 needs to be simultaneously applied to maintain the same current through the transistor M5 to have the same ω_o.

3. NOISE ANALYSIS

Noise performance is another essential consideration that governs LNA design in communication applications. Noise determines the smallest input signal that the circuit can effectively process. In RF design, Noise Figure (NF) is used as a figure of merit to evaluate the noise performance. If only thermal noise is considered, the NF can be approximately expressed as

$$\begin{aligned}
NF_{LNA} \approx\; & 1 + \gamma + \gamma g_{m1} R_s + \gamma \frac{4}{g_{m2}} \Big(\frac{g_{m3}\, g_{m5}^2}{g_2^2} \\
& + \frac{g_{m4}\, g_{m5}^2}{g_2^2} + g_{m5} + g_{m8} + \frac{g_{neg}^2\, g_{m8}}{g_{m7}} \\
& + \frac{g_{neg}^2}{g_{m6}} + \frac{g_{neg}^2}{g_{m12}} + g_{m9} + \frac{g_{ds9}^2}{g_{m10}} \Big) \quad (4)
\end{aligned}$$

In the above equation, R_s represents the output impedance of the preceding stage and is usually fixed at 50Ω. γ is the coefficient of the channel thermal noise. For comparisons, noise performance is evaluated by maintaining the same negative conductance g_{neg} and center frequency ω_o, as indicated in Eqs. 1 and 3, respectively. As a result, NF can be minimized by properly adjusting the transconductances of the transistors: increasing $g_{m2}, g_{m3}, g_{m6}, g_{m7}, g_{m10}$, and g_{m12}; decreasing g_{m1}, g_{m4}, g_{m5}, and g_{m9}. The transistor M2 of the input stage plays a critical role in the noise figure representation, and a large transistor size W/L is required, but with a constraint of offering 50Ω input matching there is little option to increase g_{m2}. The most critical noise contribution in Eq. 4 is caused by transistors M1, M3, and M5. Reasonable low noise attribute can be obtained by minimizing the transconductances g_{m5} and g_{m1}, but enlarging g_{m3}. For the same bias current, a better strategy of minimizing the transconductance is to increase the gate-to-source voltage of a transistor without using large transistor size, and thus reduce the associated parasitic capacitances.

In Figure 4, simulations show that by tuning Vbias2 and Vbias3 a filter with a center frequency ranging from 559MHz to 970MHz and Q extending to over 400 can be obtained. A circuit has been designed to emulate the receiver path of the North American Digital Cellular (NADC) and AMPS, and shows that by properly sizing the transistors we achieve a center frequency at 881MHz, Q equal to 34, 15.7dB voltage gain at ω_o, -12.4dBm IIP3, 6dB noise figure, and 52.5mW power dissipation. The linearity of the this narrow band filter is tested by applying two-tone signals at 879MHz and 883MHz so that the third-order intermodulation products fall into the passband of the filter without much attenuation by the circuit frequency response. By tuning Vbias2 and Vbias3, the Q variations at 881MHz center frequency are between 1.2 to 400. Simulation results and

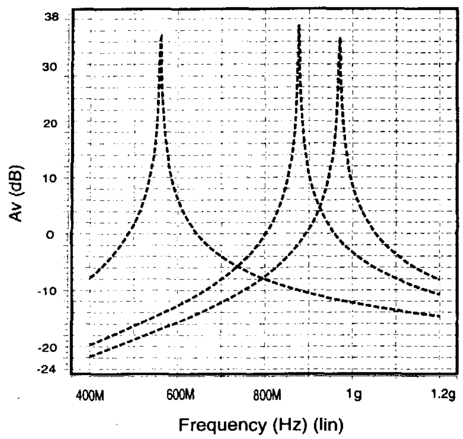

Figure 4: The frequency responses at 559MHz, 877MHz, and 970MHz with high-Q capabilities.

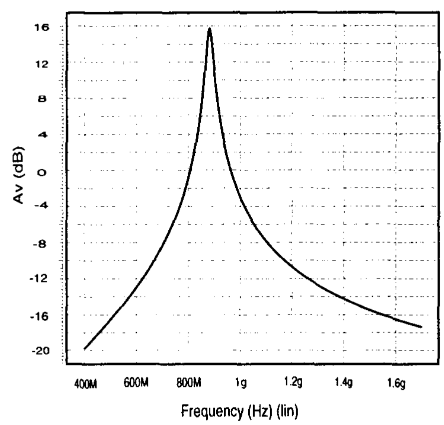

Figure 5: The LNA frequency response with ω_o=881MHz and Q=34

performance metrics are shown in figure 5 and Table 1, respectively.

4. CONCLUSION

A continuous-time CMOS tunable LNA with a Q-tuning circuit for bandpass filtering at 900MHz band communication channel has been analyzed, designed and simulated. Active elements in RF front-end systems are usually plagued by severe noise contributions compared with those of passive devices. The proposed circuit demonstrates reasonable NF can be obtained by employing active components. Automatic tuning circuitry is required to be added to this proposed filter due to manufacturing variations. In addition, the fluctuations of the temperature further aggravate the situation.

Table 1: The performance metrics of LNA with bandpass filtering.

	LNA and BP filter
Center frequency (ω_o)	881MHz
Quality factor (Q)	34
Q-tuning range	1.2 $\sim$ 400
Linearity (IIP_3)	-12.4dBm
Noise figure (NF)	6dB
Voltage gain (Av)	15.7dB
Negative conductance generator (NCG) power dissipation	15.8mW
Total power dissipation	52.5mW

Therefore, automatic tuning circuitry will be further investigated and developed in the future.

5. REFERENCES

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