A Frequency-Selective Nested Dual-Loop Broadband Low-Noise Amplifier in 90 nm CMOS

Sumit Bagga∗†, Andre L. Mansano∗, Wouter A. Serdijn∗, John R. Long∗, Kathleen Philips‡ and John J. Pekarik§

∗ERL/DIMES, Delft University of Technology, 2628 CD Delft, The Netherlands
‡IMEC-NL, Holst Centre, 5605 KN Eindhoven, The Netherlands
§IBM Microelectronics, Essex Junction, Vermont 05452, USA

Abstract—A broadband, frequency-selective low-noise amplifier (LNA) with at least 25 dB of rejection at frequencies below the L-band (includes GPS and GSM carriers) is fabricated in a 90 nm standard CMOS process. The proposed LNA can be used for broadband impulse-radio ultra-wideband (IR-UWB) and frequency modulated FM-UWB. The frequency-selective (3.5-10.5 GHz) LNA is power-to-current (P-I) configured and comprises nested reactive feedback loops: a positive current-to-current (I-I) feedback loop to boost the power gain and a negative I-I feedback loop for impedance and noise matching. The measured gain of the P-I LNA is 15 ± 3 dB. It has a noise figure (NF) of 2.4 ± 0.8 dB and a 1-dB gain compression point (P_{1dB}) of -17.5 ± 2.5 dBm. The die area of the LNA is 0.7x0.8 mm² and it consumes 9.6 mW from a 0.8 V power supply. The proposed P-I LNA is most suitable for a sub-1 V single-cell radios.

I. INTRODUCTION

Broadband wireless systems transmitting at low power spectral densities tend to overlap and share bandwidth with existing narrowband systems. Narrowband interference (NBI) mitigation remains unresolved, as existing solutions are too complex or are ineffective in rejecting narrowband interferers [1]. Recent studies show that non-coherent type receivers are especially vulnerable to NBI. For a non-coherent auto-correlation receiver (ACR), the narrowband interference term, \( i(t) \) may be defined as

\[
i(t) = \mu(i) + \mu(n) + \mu(s),
\]

where \( \mu(i) \), \( \mu(n) \) and \( \mu(s) \) are the interference-interference, interference-noise and interference-signal correlation terms, respectively [1]. Through digital signal processing, \( \mu(i) \) can be reduced by several orders of magnitude. However, terms \( \mu(n) \) and \( \mu(s) \) may not be completely removed. These unwanted interference terms must be suppressed in the analog RF frontend to limit bit-error-rate (BER) degradation. The average bit-error probability (BEP) of an ACR with and without a notch (of bandwidth \( B_{NF} \)) is shown in Fig. 1 for different values of the signal-to-interference ratio (SIR), \( C/I = E_s/T_i P_i \), where \( T_i \) is the bit duration, \( E_s \) is the energy per bit of the signal, and \( P_i \) is the power of the narrowband interferer. NBI is modeled by a single tone sinusoidal interferer.

For practical reasons, a passive filter (as designed in [2]) is often placed at the RF input of the receiver’s front-end in order to reject out-of-band interferers. The drawback with this approach is that the insertion loss of passive filters adds to the overall noise figure of the receiver. An alternative solution is to distribute the required interference rejection in the RF frontend by means of a notch antenna that offers attenuation in the passband [3] in conjunction with a frequency-selective broadband amplifier designed to reject out-of-band signals. The frequency-selective LNA is the topic of this work. The P-I LNA is based on [4].

The next section begins with a brief discussion on single and dual-loop feedback systems, followed by a detailed description of the proposed LNA (Section II). Measurement results are presented and compared to recently published results in Section III.

II. LNA WITH NESTED REACTIVE FEEDBACK

A. Single and Dual Feedback Loops

Negative feedback promotes insensitivity to process and supply variations, stabilization of gain, lower distortion, larger bandwidth (at the expense of gain), and orthogonal noise and impedance matching [5] in broadband amplifiers. In principle, a single current-to-current (I-I) negative feedback loop together with a transistor’s transconductance, \( g_m \), can define its input impedance. Due to the limited loop-gain at

\[ C/I \]
higher frequencies, and to facilitate the trade-off between impedance matching and gain, a second positive feedback loop is introduced (as in [4]). A power-to-current (P-I) configuration is the preferred choice, as the proposed LNA is to be interfaced with a mixer or an IF filter.

B. P-I LNA with Negative I-I and Positive I-I Feedback

Fig. 2 shows the topology of the frequency-selective P-I LNA. The frequency-selective P-I LNA comprises a single common-source stage (M1), and two reactive networks made up of current-to-current transformers (T1 in concentric configuration with weak mutual coupling and T2 in stacked configuration with strong mutual coupling) followed by a common-gate stage (M2). Transistor M2 with the reactive feedback networks forms a high impedance output node. To keep the noise figure to a minimum while maintaining sufficiently high gain, the LNA is biased (using bias-T networks) between optimum noise and fT points.

With the intermediate I-I positive feedback loop (as shown in Fig. 2), the input impedance (Zi) is made less susceptible to the turns ratio and coupling coefficient of the gain boosting transformer. To maintain stability, the carefully controlled positive feedback loop must be stabilized by the negative (balancing) feedback loop. This loop works as follows: the output current (io) at the primary winding Lp2 of T2 is sensed and added to the drain current (i2) of M1, thereby boosting the transconductance of the first stage (Gm ≜ im/o) without increasing the bias current or the aspect ratio. For the negative feedback loop (as in [2]), the output current (and not the intermediate current, i2) is sensed by the primary winding Lp1 of T1 and added to the gate of the common-source stage M1. This facilitates the trade-off in gain and impedance matching. Inductance L3 resonates with the parasitic capacitances of M2 (provides gain peaking) to compensate for the high frequency gain roll-off. The bondwires and bondpads are modeled using Lb andCb, respectively.

Transformer non-idealities are neglected to simplify the analysis. The role of the positive feedback transformer, T2 is to provide additional current gain, which in turn boosts the transconductance of the first stage, gm to Gm (Fig. 3), thereby increasing the power gain and sets Zi to 50 Ω (4).

\[
G_m = \frac{g_m}{1 - (k_2/n_2)}
\]

where \(k_2\) is the coupling coefficient and \(n_2\) is the turns ratio of transformer T2.

![Fig. 2. Schematic of the frequency-selective P-I LNA with nested (negative and positive I-I) reactive feedback loops.](image)

![Fig. 3. Transconductance boosting with a positive I-I feedback loop.](image)
The second feedback loop (i.e., variable $k_2/n_2$) allows for more control over the input impedance.

To realize a power gain $\geq 14$ dB, an input impedance of $50 \, \Omega$ and a noise figure $\leq 3$ dB, the transformer parameters presented in Table I are used.

### TABLE I

<table>
<thead>
<tr>
<th>Trans.</th>
<th>$k$</th>
<th>$L_s$ (nH)</th>
<th>$L_p$ (nH)</th>
<th>$N^\dagger$</th>
<th>$Q_\ell^\ddagger$</th>
<th>OD (mm$^2$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$T_1$</td>
<td>0.3</td>
<td>2.1</td>
<td>0.33</td>
<td>8.4</td>
<td>23/19</td>
<td>250x325</td>
</tr>
<tr>
<td>$T_2$</td>
<td>0.65</td>
<td>1.2</td>
<td>1.5</td>
<td>1.4</td>
<td>21/18</td>
<td>225x225</td>
</tr>
</tbody>
</table>

$^\dagger$ Physical turns ratio, $N = (n/k) = \sqrt{L_s/L_p}$.
$^\ddagger$ Q-factor of $T_1$ and $T_2$ simulated at 6.5 GHz.

The broadband notch in the stopband is a result of the series LC resonant network formed with the secondary winding, $L_{s1}$ of $T_1$ and AC coupling capacitor $C_1$.

### III. MEASUREMENT RESULTS

The microphotograph of the fabricated P-I LNA is shown in Fig. 4. The chip area is $0.56 \, \text{mm}^2$ $(0.7 \times 0.8 \, \text{mm}^2)$ including bondpads. The active area is approximately $0.3 \, \text{mm}^2$. All inductors, transformers and metal-insulator-metal (MIM) capacitors are implemented in the top two thick metallization layers. The windings of the non-inverting transformers, $T_1$ and $T_2$ are concentric and stacked, respectively.

![Fig. 4. Microphotograph of the frequency-selective P-I LNA in 90 nm CMOS.](image)

The measured and simulated results for the forward transmission ($S_{21}$) and reflection ($S_{11}$) coefficients are shown in Fig. 5. In the passband (3.5-10.5 GHz), the power gain is $15 \pm 3$ dB. The LNA presents 30 dB of rejection at frequencies below the L-band (includes GPS and GSM carriers). The transconductance of the cascode and the transformer parameters, such as the self-inductances of the windings, physical turns ratio and coupling coefficient, set the input impedance of the LNA (Fig. 5). An acceptable $S_{11}$ over a broad frequency range is $\leq -10$ dB. The measured $S_{11}$ varies from -7 to -16 dB.

A linear phase (or uniform group delay) response is paramount in broadband amplifier design. The measured phase response (of the $S_{21}$) of the P-I LNA is shown in Fig. 6. At the resonance frequency, the LNA demonstrates a phase jump of approximately 125 degrees. The group delay (including the test fixture) is approximately $330 \pm 40$ ps across the passband.

![Fig. 5. Forward transmission and reflection coefficients of the P-I LNA (after de-embedding). Gain peaking is from 7 to 9 GHz instead of 8 to 10 GHz as a result of larger self-inductance of $L_3$.](image)

The measured reverse transmission coefficient (isolation), $S_{12} \leq -20$ dB ($\Delta 10$ dB from simulated) is shown in Fig. 7. Discrepancy in $S_{12}$ is a result of unwanted parasitic feedback.

![Fig. 6. Phase response of the P-I LNA (including the test fixture).](image)

![Fig. 7. Reverse transmission coefficient of the P-I LNA.](image)

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The noise figure (2.4±0.8 dB) is plotted in Fig. 8. As the transformer coupling degrades at lower frequencies, the noise figure is higher. At higher frequencies, the noise figure is greater because of more substrate and parasitic losses. In broadband amplifier design, reactive feedback increases linearity without increasing thermal noise. It is often the case that linearity of an amplifier deteriorates as frequency increases, however, in transformer-based feedback systems, the effects are not as profound. The 1-dB compression point, P₁dB (-17±2.5 dBm) is an appropriate measure of the linearity for broadband circuits (see Fig. 8). The input-inferred third-order intercept point (IIP3) can be extrapolated from the P₁dB.

**Fig. 8. Noise figure and 1-dB compression point of the P-I LNA.**

Table II compares the frequency selective P-I LNA to recently published wideband LNAs in standard CMOS and SiGe HBT technologies. This prototype demonstrates superior design characteristics, such as a smaller silicon footprint, capacity to operate from a lower voltage supply, is least technology dependent (as a result of feedback) and provides excellent out-of-band rejection.

### IV. Conclusion

A γcommended frequency-selective LNA with nested reactive feedback loops in 90 nm standard CMOS is presented. Reactive feedback loops are constructed using on-chip concentric and stacked current-to-current transformers. A broadband notch is placed in the stopband to suppress narrowband interferers below the L-band (includes GSM carriers). The measured power gain of the LNA is 15±3 dB. The noise figure (2.4 dB) and the 1-dB compression point (-17.5 dBm) exhibit a 0.8 dB and 2.5 dBm variation across the passband. Total power dissipation is 9.6 mW from a 0.8 V supply; this LNA is intended for sub-1 V single-cell radios.

### V. Acknowledgment

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**REFERENCES**


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**Table II**

SUMMARY OF THE P-I LNA AND COMPARISON WITH PREVIOUSLY DESIGNED DESIGNS

<table>
<thead>
<tr>
<th>Specifications</th>
<th>This work</th>
<th>[6]</th>
<th>[7]</th>
<th>[8]</th>
<th>[9]</th>
<th>[10]</th>
<th>[11]</th>
<th>[12]</th>
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</thead>
<tbody>
<tr>
<td>BW (GHz)</td>
<td>3.3-9.25</td>
<td>3.25-10.25</td>
<td>3.1-10.6</td>
<td>3.1-10.6</td>
<td>2.4-9.5</td>
<td>3-10</td>
<td>3.4-11.4</td>
<td>10-5.2</td>
</tr>
<tr>
<td>S₁₁ (dB)</td>
<td>15±3</td>
<td>14.5±2.5</td>
<td>15.3±2.2</td>
<td>19±2</td>
<td>17.8±1.5</td>
<td>18.5±1.7</td>
<td>14.75±1.25</td>
<td>15±6</td>
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<tr>
<td>Notch (dB)</td>
<td>&gt; 30⁺</td>
<td>&gt; 20⁻</td>
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<td>n.a.</td>
<td>n.a.</td>
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<td>n.a.</td>
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<tr>
<td>S₁₂ (dB)</td>
<td>(-16) to (-8)</td>
<td>(-16) to (-10.5)</td>
<td>(-25) to (-8.6)</td>
<td>(-14) to (-9)</td>
<td>(-38) to (-15)</td>
<td>&lt; -7.2</td>
<td>(-40) to (-10)</td>
<td>&lt; -10</td>
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<tr>
<td>NF (dB)</td>
<td>300±40</td>
<td>225±125</td>
<td>102.5±27.5</td>
<td>n.a.</td>
<td>187.5±62.5</td>
<td>n.a.</td>
<td>n.a.</td>
<td>n.a.</td>
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<tr>
<td>NF (dB)</td>
<td>2.4±0.8</td>
<td>2.9±0.8</td>
<td>2.5±0.47</td>
<td>3.4±0.85</td>
<td>6.6±2.6</td>
<td>2.45±0.65</td>
<td>4.55±1.45</td>
<td>&lt; 3.5</td>
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<tr>
<td>IIP3 (dBm)</td>
<td>(-10) to (-5)</td>
<td>(-9) to (-1.8)</td>
<td>(-7.2) to (-4.3)</td>
<td>(-5.5) to 3</td>
<td>(-8.2) to (-5.6)</td>
<td>2.1 (6 GHz)</td>
<td>-7 (6 GHz)</td>
<td>&gt; 0</td>
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<td>V&lt;sub&gt;DD&lt;/sub&gt;/P (V/mW)</td>
<td>0.89/6</td>
<td>1.2/15</td>
<td>1.2/9</td>
<td>3.3/30</td>
<td>1.8</td>
<td>3.3/26</td>
<td>1.8/11</td>
<td>1.2/21</td>
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<tr>
<td>Area (mm²)</td>
<td>0.56</td>
<td>1.68</td>
<td>0.87</td>
<td>1.8</td>
<td>1.1</td>
<td>0.72</td>
<td>1.2</td>
<td>0.009</td>
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<tr>
<td>Tech. (nm)</td>
<td>C-90</td>
<td>C-130</td>
<td>C-130</td>
<td>SiGe-180</td>
<td>C-180</td>
<td>SiGe-180</td>
<td>C-180</td>
<td>C-65</td>
</tr>
</tbody>
</table>

⁺ Voltage gain; ⁻ Measured @ 1.5 GHz; ¹ Measured @ 5.25 GHz; ² Extrapolated IIP3 (i.e., measured P₁dB + 9.6 dB).