Low mismatch UHF LNA for cellular infrastructure

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The non-coincidence of noise $\Gamma$opt and conjugate $S_{11}^*$ matching points is the biggest hurdle in low noise amplifier (LNA) design. Because of this, the LNA design process inescapably entails trading off input match for noise figure and vice-versa. Unfortunately, this trade-off may be unavailable in cellular infrastructure application because of an aerial matching requirement; i.e. to achieve a 14 dB aerial mismatch when the LNA is preceded by a bandpass filter of -18 dB return loss (assumed lossless), the input return loss (IRL) has to be better than -23 dB. An isolator can cure the high reflectivity in the noise-matched amplifier but it is costly and heavy.

Cheaper and lighter than isolators is the balanced amplifier topology developed at Bell Labs in the 60’s [1-2]. It channels the energies reflected from a pair of amplifiers to the quadrature 3 dB couplers (also known as the 90° or hybrid coupler) where self-cancelation can take place. Since the port match will be excellent regardless of the constituent amplifiers’ actual reflectivity, the designer is free to tune the input networks for minimum noise. Additionally, the balanced configuration has better reliability, linearity and bandwidth than its single-ended counterpart and, most amazingly, it is inherently self-stabilizing; i.e. perfect stability, both in and out-band, can ensue even when it comprises two potentially unstable amplifiers [3].

On the negative side, the balanced LNA needs twice the current and the components of its single-ended counterpart. The quadrature couplers for signal splitting and combining also add cost and PCB area, especially for distributed implementations, and their insertion losses degrade RF performances. If commercial drop-in couplers are used, their RF performances are generally proportional to their size and cost. Furthermore, the confined space atop cellular towers, i.e. tower mounted amplifiers (TMA), disfavors balanced LNAs because they are roughly twice the weight and size of their single-ended counterpart. A reading of the literature reveals two broad approaches to solving the balanced LNA’s size issue – shrinking the couplers [4-5] and higher circuit integration [6-7, 13-14]. At the same time, the TMA’s need for cutting edge sensitivity and linearity hampers miniaturization.

To improve on the size and component count of previous 900 MHz balanced LNA designs, we designed around miniature multilayer couplers and a MMIC that integrates dual amplifiers, biasing and shutdown functions. To our knowledge, this is the industry’s first dual-amplifier MMIC with an integrated shutdown function. To enhance the design’s usefulness, we target substantial size reduction while either equaling or exceeding the best noise and linearity performances recorded by the prior arts (see table 3). This article describes how the design accomplishes the conflicting size and performance goals and then summarizes the most important results.

**Materials and methods**

This section first describes the on-chip functions, then, the off-chip circuit, followed by the circuit modelling of the prototype balanced amplifier and finally, validates the designed source ($\Gamma'$) and load ($\Gamma''$) impedances against source and load-pull data.

As the first step to size reduction, a new MMIC that integrates dual amplifiers, electrostatic discharge (ESD) protection, active bias and shutdown functions (Figure 2, inside yellow box) was designed. Aside from less external components, combining the bias circuits with the amplifier on the same chip also beneficially stabilizes the operating current against gate threshold voltage and temperature variations. The chip is fabricated using a 0.25 μm enhancement-mode pseudomorphic high electron mobility transistor (ePHEMT) on 6” wafer process because it...
has a suitable balance of cost and performance. Firstly, the process has previously enabled a single-ended LNA to attain 0.3 dB NF at 900 MHz [8], so a 0.5 dB NF in balanced mode might be reachable after factoring in coupler’s loss. Secondly, the process’ relatively high fT and peak transconductance — >30 GHz and -615 mS/mm, respectively — reduces the number of gain stages required to reach 17.6 dB gain to one. Lastly, this process is suitable for low voltage operation because its linearity does not drop appreciably until the Vds is reduced below 2 V [9]. The MMIC, which integrates 6 transistors, 26 diodes, 12 resistors and 2 capacitors, is epoxy encapsulated in a 16-pin 4 x 4 x 0.85 mm flat no-lead (QFN) package.

The active bias circuits are connected to the voltage supplies Vdd1-2 via external resistors R3 and R8. Through these resistors, the gate bias voltages can be user controlled. Although each eHETM’s nominal drain current ID is 60 mA at 4.8 V Vdd, it can be varied 48-72 mA over the allowable range of values for these resistors. Inductors L1 & L16 and resistors, R1 & 16, serve as the gates’ (pins 1 & 4) bias networks. Although on-chip spiral inductors can perform this function, this design opted for external inductors because they have lower losses, the smaller chip is more economical, and the chip can be used at other frequencies. Since NF is critical in this application, we specifically used wire wound chip inductors instead of multilayer ones. Beside bias insertion, the aforementioned inductors and series capacitors, C3 and C16, also form high-pass networks to roll off unneeded low frequency gain. The long bondwires, Lw, which connect the drain to the package leads also serve as pre-matching.

The shutdown function block consists of transistor switch in series with the active bias. Shutdown is initiated by applying a high logic (≥2 V) at Vsd1/2 to open the switch. Conversely, a low logic, i.e. Vsd1/2 ≤500 mV, enables the amplifier. Transitioning from normal to shutdown takes less than 50 ns if the large (≥0.1 μF) decoupling capacitors C6/8/20/22/23/24 are omitted. However, these capacitors are generally recommended because they aid low frequency stability and dampen supply transients.

The balanced amplifier’s signal splitting and combining utilize commercially available multilayer couplers, X1-2, to save design effort and PCB area. A larger coupler measuring 6.4 x 5.1 mm is used at the input because of its lower loss, while a smaller (2.0 x 1.3 mm) coupler is used at the output to save space and cost. To ensure that the IRL ≤-21 dB requirement can be satisfied without fail in volume production, the critical parameters are identified through a Monte Carlo analysis (figure 3). Subsequently, two controls were instituted: correlating the amplifiers’ input match to [S11a-S11b] ≤0.025 and the input coupler’s isolation must be >23 dB. The first control is satisfied by using adjacent chips while the second requires specifying a high isolation input coupler [10]. The output coupler is non-critical.

The PCB utilizes the mid cost Rogers RO4350 material (figure 4). The microstrip width is dimensioned for 50 Ω characteristic impedance wherever practical. However the traces next to the input coupler’s mounting pads are necked down following the manufacturer’s recommendations to compensate for the pads’ parasitic capacitances. To the 10mil thick PCB, an FR4 layer is added to increase the stack height to 1.6 mm. The input and output microstrips are transitioned to coaxial using edge-launched SMA receptacles. All results are referenced to the coaxial ends.

To simulate in Agilent ADS2009, the circuit model is split into a two-level hierarchy (figure 5). The upper level consists of blocks representing the MMIC, the signal dividing/combining and the impedance matching functions. Each of the dual amplifier, Q1-2, is represented by an identical 2-port s-parameters (s2p). The s-parameters were previously extracted on a test fixture of similar material (10mil RO4350) and then applying the Thru Reflect Line (TRL) calibration to shift the reference planes to the package edges. Using the same test fixture, the device’s noise and
linearity parameters were extracted using automated source and load pull tuners (Focus). The ~0.2dB NF_{MIN} is particularly challenging to extract because it can be easily obscured by the combined losses of the mechanical tuner and the required connector adapter (APC7 to 3.5 mm). The inductors and hybrid couplers are modelled with the manufacturers’ s2p data. Other passive components are modelled using their equivalent circuit including their lower order parasitic.

To validate the designed input and output matching networks, the modelled source ($\Gamma_S$) and load ($\Gamma_L$) impedances are compared with the previously measured source and load pull contours. Both $\Gamma_S$ and $\Gamma_L$ are obtained by simulating with ADS’s “s-parameter” probe component. With this MMIC, it is easy to achieve good noise performance because the 0.3 dB constant NF circle is large and even encompasses the chart centre (figure 6). Because of this, the input networks comprising L1-C3 and L2-C16 are designed to function as bias-tees rather than as impedance/noise match. The absence of impedance transformation in the input LC network means that the bandwidth does not have to be constrained by the network’s loaded Q and the insertion loss is less sensitive to component Q. The $\Gamma_S$ offset from the chart centre is unintentional and is due to the lumped component parasitics; the NF associated with $\Gamma_S$ is 0.28 dB. Optimizing to the noise match $\Gamma_{OIP3}$ will improve NF by an insignificant 0.03 dB and so, is not necessary.

The device has a third order output intercept point (OIP3) in excess of 42 dBm at optimum output match. The area encompassed by the 42 dBm constant linearity circle will require some impedance transformation to arrive at because it located away from the chart centre (figure 6 right). The next circle at 40 dBm readily includes the chart centre. Since a 40 dBm OIP3 is sufficient for this application, we designed the output network without impedance transformation; i.e. $\Gamma_L=50\Omega$. However $\Gamma_L$ is shown slightly offset from the exact centre because component parasitics.

**Results and discussion**

**A: Size, component count and function integration**

This design is probably the smallest and the most highly-integrated among balanced LNA designs for MFA application. Compared to previous designs, the PCB is 110% to 40% smaller (table 1). The part count is higher than the uMUT chip on board (MCOB) design with integrated...
Wireless Infrastructure

\[ \Gamma_S & NF \text{ contours at 950MHz} \]

\[ \Gamma_L & OIP3 \text{ contours at 950MHz} \]

Figure 6: Good noise and linearity performances can be obtained without impedance matching because the Smith chart centre is encircled by the 0.3 dB constant noise figure circle (left) and at the output side, by the 40 dBm constant IP3 circle (right).

<table>
<thead>
<tr>
<th>Ref.</th>
<th>year</th>
<th>PCB area, mm²</th>
<th>Part count</th>
<th>device packaging</th>
<th>supplies, V</th>
<th>Dual amp.</th>
<th>bias</th>
<th>matching</th>
<th>shutdown</th>
</tr>
</thead>
<tbody>
<tr>
<td>Piper*</td>
<td>2002</td>
<td>945 (+110%)</td>
<td>34</td>
<td>SOT-334</td>
<td>5.0</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Lee [3]</td>
<td>2008</td>
<td>760 (+69%)</td>
<td>10</td>
<td>MCOB/SV6</td>
<td>2.8, 5.0</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Avago [4]</td>
<td>2009</td>
<td>630 (+40%)</td>
<td>34</td>
<td>QFN4x4d</td>
<td>0.6, 5.0</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Omnii [5]</td>
<td>n. a.</td>
<td>n. a.</td>
<td>27</td>
<td>QFN4x4d</td>
<td>-0.5, 5.0</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Omnii [6]</td>
<td>n. a.</td>
<td>n. a.</td>
<td>27</td>
<td>QFN4x4d</td>
<td>-0.9, 2.5</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>This work</td>
<td>2012</td>
<td>450 (0%)</td>
<td>32</td>
<td>QFN4x4d</td>
<td>4.8</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 1: This work has the smallest PCB among 900 MHz balanced LNAs. Moreover, its single 4.8 V saves on components that are otherwise required for the second supply’s regulation and filtering. Its MMIC has one of the most integrated functions, including the never before integrated shutdown. Although this design reports 1.9 GHz performances, it is included in this survey because it is readily adaptable to 900 MHz by swapping the drop-in multilayer couplers with another model.

<table>
<thead>
<tr>
<th>Section</th>
<th>input</th>
<th>MMIC</th>
<th>output</th>
<th>total</th>
</tr>
</thead>
<tbody>
<tr>
<td>NF (dB)</td>
<td>0.2</td>
<td>0.3</td>
<td>0.2</td>
<td>0.5</td>
</tr>
<tr>
<td>G (dB)</td>
<td>-0.2</td>
<td>18.4</td>
<td>-0.2</td>
<td>18</td>
</tr>
</tbody>
</table>

Table 2: Breakdown of gain & noise figure contributions from various LNA sections at 900MHz.

matching [13], but its external matching enables lower NF and its monolithic fabrication is cheaper. Additionally, its single voltage requirement can potentially save many more components that will be required for the second voltage supply’s regulation and filtering. The MMIC also integrates as many functions as the best of the prior arts. To our knowledge, this is the first dual-amplifier MMIC that integrates the shutdown function.

**B: RF performances**

The gain and noise performances are not compromised by the substantial miniaturization. The experimental gain and noise figure are 0.5 dB and 18 dB, respectively at 900 MHz (figure 7). The experimental NF is comparable to competing designs which used more expensive processes or packaging (table 3). The NF result is also notable for being equal to our last design using 55% larger couplers [14]. The circuit model developed for this design has good predictive value - the maximum error for gain and NF are 0.2 dB and 0.1 dB, respectively. Assuming ideal input coupler and matching networks, the balanced LNA has similar NF to its constituent amplifiers [17]. Since the device-level NF is -0.3 dB, a figure of 0.2 dB can be inferred for the combined loss of the coaxial connector, hybrid coupler and input matching (table 3). Over 600-1050 MHz, the NF changes less than a tenth of a decibel from the minimum, hence, the usable bandwidth is sufficient to serve existing and planned 3G/4G bands worldwide. The NF increases abruptly outside this frequency range due to the input coupler’s characteristics.

In the event of one arm completely failing, the balanced LNA’s midband gain remains a useful -12 dB. To simulate the failure of one amplifier, the voltage supplies to each amplifier, Vdd1 and Vdd2, are disconnected alternately. With either one amplifier powered at a time, the midband gain is alternately 11.8 dB and 12.7 dB (figure 8). The gain in this semi-operational state is within tenths of a decibel from the 6 dB drop predicted by theory. Because there is no complete loss of gain when one amplifier completely fails, the balanced amplifier is more reliable than a single ended one; this is an important advantage for cellular applications because service outage is intolerable.

The experimental result surpasses the -21 dB input mismatch required by TMs. Moreover, the good matching is maintained over a wide bandwidth; i.e. the bandwidth equals 83% of centre frequency (fc) at IRLs = -20 dB (figure 9). The experimental IRL is best around the input coupler’s centre frequency. Likewise, the output coupler determines the ORL response. The model has semi-quantitative agreement with the experimental result over most of the passband. However, the experimental ORL’s unexpected dip at 500 MHz was not predicted by the model and this could be due to failure to model a component’s parasitics. As previously mentioned, the quadrature couplers enable the extremely wide matched bandwidth. Both input and output RL are limited by the couplers’ finite isolation and by the microstrip discontinuities.

The balanced LNA is unconditionally stable even when its constituent amplifiers are not. Both modelled and measured mu stability factor exceed unity over 50 MHz to 20 GHz (figure 10). Therefore, the balanced topology’s self-stabilizing promise is validated because the individual amplifiers are potentially unstable (mu <1) at several frequencies over the evaluated range. The calculated stability factor has the same general trend as the experimental result but the peaks do not converge exactly; the errors are probably caused by oversimplification of the passive component models.

The design’s gain compression and linearity performances are best-in-class; its IIP3 is 5.6 dB higher than the nearest competitor (table 3). The 1 dB gain compression, P1dB, measures -24 dBm at midband and is relatively constant over frequency, varying less than a dB over a 1 GHz span (figure 11). Using -20 dBm tones spaced 1 MHz apart as the test stimuli, the third-order input intercept point IIP3 measures -21.6 dBm at midband. But unlike the flat P1dB response, the IIP3 exhibits a pronounced peak (21.9 dBm) at the approximate centre of the couplers’
Figure 7: Modelled and tested noise figure (F) and gain (G) show good agreement over a wide swatch of frequencies. The experimental NF is among the best for balanced LNAs.

Figure 8: Gain versus frequency when both amplifiers are functioning (BOTH) and when only either one of the amplifier is functioning (amp1/amp2 only). Hence, useable amplification is assured even when one amplifier completely fails.

Figure 9: Modelled and measured input and output return loss versus frequency.

Figure 10: The balanced LNA’s modelled and tested mu stability factors point to unconditional stability. To demonstrate the balanced amplifier’s self-stabilizing property, the constituent amplifiers’ poorer stability factor is also shown for comparison.

Figure 11: The ~24 dBm P1dB and 21.6 dBm IIP3 are the highest among 900 MHz balanced LNAs.

Figure 12: The good match (RL≤-20 dB) during shutdown can obviate the need for external bypass.
Table 3: Among 900 MHz balanced LNA designs, this work has the best linearity (IIP3). Its input return loss and noise figure are among the best.

and better performing tower-mounted balanced LNAs.

Recently, balun-less connections of balanced aerials with balanced LNAs have been demonstrated [20-21]. Because eliminating the balun can beneficially cut cost and RF losses, one tantalizing question is how performance and cost benefits this work’s highly integrated MMIC can bring to such application?

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References