Chapter 2
Design and Implementation of Ultra-Low-Power ZigBee/WPAN Receiver

In recent years, the proliferation of short-range wireless applications for Internet of Things and personal healthcare calls for ultra-low power and cost CMOS radios [1]. Ultra-low voltage (ULV) designs have been one of the key directions to approach a better power efficiency [2–5]. Regrettably, an ULV supply will limit the voltage swing, and device’s $f_T$ and overdrives, deteriorating the spurious-free dynamic range (SFDR) while necessitating area-hungry inductors (or transformers) to assist the bias and tune out the parasitic capacitances. This chapter describes the design and implementation of a compact, low-power and high-SFDR receiver suitable for ZigBee or wireless personal area network (WPAN) applications. The research background can be outlined as follows.

Four potential ultra-low-power receiver architectures are shown in Fig. 2.1. The first (Fig. 2.1a) employs a single low-noise transconductance amplifier (single-LNTA) followed by two passive I/Q mixers and transimpedance amplifiers (TIAs). If a 50 %-duty-cycle local oscillator (50 % LO) is applied, this topology can suffer from image current circulation between the I and Q paths, inducing I/Q crosstalk, unequal high-side and low-side gains, IIP2 and IIP3 [6]. Lowering the LO duty cycle to 25 % (Fig. 2.1b) can alleviate such issues [7], at the expense of extra sine-to-square LO buffers and logic operation. Another alternative is to add two signal buffers before the mixers (Fig. 2.1c), but they must be linear enough (i.e., more power) to withstand the voltage gain of the low-noise amplifier (LNA) [8, 9].

The basis of our proposed solution (Fig. 2.1d) is to split the LNTA into two, such that a single-ended RF input is maintained, while allowing isolated passive mixing that facilitates the use of a 50 % LO for power savings.

This chapter is organized as follows: Sect. 2.1 will give an overview of the operating principle of the proposed “split-LNTA + 50 % LO” receiver. An analytical comparison of it with the existing “single-LNTA + 25 % LO” architecture will be presented in Sect. 2.2. In Sect. 2.3, a number of circuit techniques will be proposed, including: (1) a low-power voltage-mode transimpedance amplifier (TIA) to enhance the out-channel linearity both at RF and baseband (BB); (2) a mixed-supply ($V_{DD}$) design approach [10] to alleviate the design trade-offs in RF LNTA (power, gain and noise) and BB TIA (power, linearity and signal swing); (3) a low-power LO generation scheme that consists of a LC voltage-controlled oscillator (VCO) and an input-impedance-boosted Type-II RC-CR network. They
optimize the VCO’s output swing with the LC tank’s quality factor, while offering adequate I/Q accuracy at low power. The measured experimental results will be reported in Sect. 2.4.

2.1 Proposed “Split-LNTA + 50 % LO” Receiver

The split-LNTA (Fig. 2.2) is based on two self-biased inverter-based amplifiers (M₁, M₂ and R_F), which have no inner parasitic pole. They also can take the speed advantage of fine linewidth CMOS to lower the device overdrive voltages, featuring a high g_m-to-I_d efficiency at low V_DD (V_DD06 = 0.6 V). Its single-ended RF input avoids the RF balun and its associated insertion loss. In front of the split- LNTA, a proper co-design between the RF input capacitance (C_in) and bond wire (L_bw) facilitates the input impedance matching, while offering a passive pre-gain (Av) decisively important to the NF and power efficiency. The two LNTAs convert the RF signal (v_in) into two equal currents i_out,I and i_out,Q for the I and Q channels, respectively. To avoid the parasitic and area impact from AC coupling, i_out,I and i_out,Q, are directly DC-coupled to the passive mixers (M₃ and M₄). As long as the DC current passing through M₃ and M₄ is kept small, the 1/f noise induced by the mixers can be minimized [11]. This aim can be achieved by matching the output common-mode level of the LNTA to that of the BB TIA.

Fig. 2.1 Four potential receiver architectures: a Single-LNTA + 50% LO. b Single-LNTA + 25% LO. c Single-LNA + 50% LO + signal buffers. d Split-LNTA + 50% LO (proposed)
The 50% 4-phase LO (LO\textsubscript{Ip,n} and LO\textsubscript{Qp,n}) is generated by a 2.4-GHz LC VCO followed by a new type-II RC-CR network, which features a capacitor divider at the input to boost the input impedance. When driving the LO to the mixers (M\textsubscript{3} and M\textsubscript{4}), a proper DC level (V\textsubscript{LO,b}) can optimize the switching time. The down-converted low-IF (2 MHz) signal is further amplified by a common-gate TIA (M\textsubscript{5}-8) and RL, which uses a 1.2 V (V\textsubscript{DD12}) supply to accommodate more signal swing and enhance linearity. Here, we assume a complex low-IF filter will follow the BB TIA, rendering the 1/f noise and IIP2 not significant and will not be further addressed.

Due to the bidirectional transparency of passive mixers [7, 8], the BB capacitors (C\textsubscript{1} and C\textsubscript{M}) can enhance the selectivity at both RF (the output of the LNTA) and BB, improving the out-band linearity. The grounded C\textsubscript{M} also helps to suppress the common-mode RF feed through, which is limited by the bond wire inductance that appears in series with C\textsubscript{M} under common-mode operation.

### 2.2 Comparison of “Split-LNTA + 50% LO” and “Single-LNTA + 25% LO” Architectures

This Section presents an analytical comparison of the two architectures: “split-LNTA + 50% LO” and “single-LNTA + 25% LO”. For brevity, “50% LO” and “25% LO” are exploited to represent them, respectively. Figure 2.3a, b show their simplified equivalent circuits. For a fair comparison, the two LNTAs in Fig. 2.3a are modeled as \( g_m \) (transconductance) and \( 2R_{out} \) (output resistance),
where the single LNTA in Fig. 2.3b is modeled as $2g_{m}$ and $R_{out}$. These models are developed under the same approach described in [12–14], where the harmonic up-conversion in passive mixers is modeled as $R_{sh}$. The impedances looking into the 50 %-LO and 25 %-LO mixers are denoted as $Z_{MIX1}$ and $Z_{MIX2}$, respectively. Each mixer features an on-resistance of $R_{sw}$. $R_{TIA}$ is the input resistance of the TIA. The single-ended differential mode capacitance is denoted as $C_{d}$ ($=C_{M} + 2C_{1}$).

### 2.2.1 Gain

For Fig. 2.3a, we summarize in (2.1)–(2.5) the derived expressions of both $Z_{MIX1}$ and the voltage gain ($A_{Vx1}$) at $V_{x1}$ at the LO + IF frequency ($\omega_{LO} + \omega_{IF}$); the baseband output current ($I_{BB1}$) with respect to $v_{in}$; the voltage gain ($A_{Vy1}$) at $V_{y1p,n}$, and finally the voltage gain ($A_{Vout1}$) at $V_{out1p,n}$.

$$Z_{MIX1}@ (\omega_{LO} + \omega_{IF}) \approx R_{sw} + \left( \frac{2Z_{BB}}{\pi^2} / R_{sh} \right)$$  \hspace{1cm} (2.1)

where $Z_{BB} = \frac{1}{s(2C_{1}+C_{M})} / R_{TIA}$; $R_{sh} \approx \frac{2}{3}(2R_{out} + R_{sw})$

$$A_{Vx1}@ (\omega_{LO} + \omega_{IF}) \approx g_{m}(2R_{out}) / Z_{MIX1}$$  \hspace{1cm} (2.2)

$$\frac{I_{BB1}}{v_{in}@ DC} = \frac{I_{BB1p} - I_{BB1n}}{v_{in}} \approx g_{m} \frac{2R_{out}}{R_{TIA} + 2(2R_{out} + R_{sw})} \frac{4}{\pi} = G_{m1}$$  \hspace{1cm} (2.3)

$$A_{Vy1}@ DC = A_{Vy1p} - A_{Vy1n} \approx G_{m1}R_{TIA}$$  \hspace{1cm} (2.4)
Similarly, for Fig. 2.3b, we have (2.6)–(2.10) the derived expressions of both $Z_{\text{MIX2}}$ and the voltage gain ($A_{Vx2}$) at $V_{x2}$ at the LO + IF frequency ($\omega_{\text{LO}} + \omega_{\text{IF}}$); the baseband output current ($I_{\text{BB2}}$) with respect to $v_{\text{in}}$; the voltage gain ($A_{Vy2}$) at $V_{y2p,n}$, and finally the voltage gain ($A_{Vout2}$) at $V_{out2p,n}$.

$$Z_{\text{MIX2}}|_{\omega_{\text{LO}} + \omega_{\text{IF}}} \approx R_{\text{sw}} + \left(\frac{2Z_{\text{BB}}}{\pi^2} / R_{\text{sh}}\right)$$

(2.6)

where $Z_{\text{BB}} = \frac{1}{s(2C_1 + C_m)} / R_{\text{TIA}}; R_{\text{sh}} \approx 4(R_{\text{out}} + R_{\text{sw}})$

$$A_{Vx2}|_{\omega_{\text{LO}} + \omega_{\text{IF}}} \approx 2g_m\left(R_{\text{out}} / Z_{\text{MIX2}}\right)$$

(2.7)

$$\frac{I_{\text{BB2}}}{v_{\text{in}}}|_{\text{DC}} = \frac{I_{\text{BB2p}} - I_{\text{BB2n}}}{v_{\text{in}}} \approx 2g_m\frac{R_{\text{out}}}{R_{\text{TIA}} + 4(R_{\text{out}} + R_{\text{sw}})} \left(\frac{4\sqrt{2}}{\pi}\right) = G_{m2}$$

(2.8)

$$A_{Vy2}|_{\text{DC}} = A_{Vy2p} - A_{Vy2n} \approx G_{m2}R_{\text{TIA}}$$

(2.9)

$$A_{Vout2}|_{\text{DC}} = A_{Vout2p} - A_{Vout2n} \approx G_{m2}R_L$$

(2.10)

Note that the output capacitance of the LNTA was neglected. In fact, the output capacitance of LNTA will induce $C_{\text{out}}$ and $2C_{\text{out}}$ for the $g_m$ and $2g_m$ LNTA stages, respectively. This will render the output impedance ratio at $V_{x1}$ and $V_{x2}$ slightly larger than 2. Besides, the parasitic capacitor will affect $R_{\text{sh}}$ too. The proposed separated $g_m$ stage imposes a smaller $C_{\text{out}}$ and thus lowers the degradation of gain and NF when compared with those predicted by Eqs. (2.11) and (2.12). With proper sizing, it would be possible to achieve $R_{\text{sw}} \ll R_{\text{out}}$ and $R_{\text{sw}} \ll R_{\text{TIA}}$ and $R_L$, such that the gain difference between 25 % LO and 50 % LO at different RF and BB nodes can be estimated as,

$$\Delta A_{Vx1,2}|_{\omega_{\text{LO}}} = 20\log A_{Vx2} - 20\log A_{Vx1} \approx 20\log\left(\frac{2(R_{\text{out}} / R_{\text{TIA}})}{2R_{\text{out}} / R_{\text{TIA}} / 4R_{\text{out}}}ight) = 6\text{ dB}$$

$$\Delta A_{Vy1,2}|_{\text{DC}} = 20\log A_{Vy2} - 20\log A_{Vy1} = 20\log\left(\sqrt{\frac{2}{2R_{\text{TIA}} + 4R_{\text{out}} + 2R_{\text{sw}}}}\right) \approx 3\text{ dB}$$

$$\Delta A_{Vout1,2}|_{\text{DC}} = 20\log A_{Vout2} - 20\log A_{Vout1} = 20\log\left(\sqrt{\frac{2}{R_L + 4R_{\text{out}} + 2R_{\text{sw}}}}\right) \approx 3\text{ dB}$$

(2.11)

From (2.11), the 25 % LO should have a higher gain at both RF and BB nodes than the 50 % LO. However, as analyzed in Sect. 2.3.3, a higher gain at RF will penalize the IIP3, while a higher BB gain can be achieved easily by using a larger $R_L$. Regarding the impact of these gain differences to the NF it will be analyzed next.
2.2.2 NF

The NF is analyzed according to the equivalent LTI noise model \([12–14]\). As shown in Fig. 2.4a, b, the four noise sources are the thermal noises from \(R_n V_{n,RS} = 4kT R_s\), LNTA \((I_{n,g_m}^2 = 4kT \gamma_1 g_m\) or \(I_{n,2g_m}^2 = 4kT \gamma_2 g_m\)), \(R_{sw} (V_{n,SW}^2 = 4kT R_{sw})\) and the noise from TIA is \(V_{n,TIA}^2 \approx 4kT \gamma_2 g_m TIA \approx 4kT \gamma_2 R_{TIA}\), given that the output impedance of the mixer is sufficiently large. Here, \(g_m TIA\) is the transconductance of the bias transistor for the TIA, while the noise from the CG device is degenerated. An accurate model of the TIA noise can be found in \([11]\). The noise of \(R_f\) is ignorable and the noise coupling between the \(I\) and \(Q\) paths under a 50 % LO is minor (confirmed by simulations), easing the NF calculation of each path separately. The noise factor \((F)\) can be found by dividing the total output noise by the portion related with \(R_s\) contribution,

\[
F = 1 + \frac{\gamma_1}{R_s A_v G_m} + \frac{R_{sw}}{R_s A_v^2 G_m^2 R^2} + \frac{(R + R_{sw})^2}{R_s A_v^2 G_m^2 R^2} + \frac{a \gamma_1}{R_s A_v^2 G_m} + a \]

where \(\beta = \frac{2}{\pi}\) is the down conversion scaling factor and \(a\) is the harmonic folding factor,

\[
a = \left(\frac{\pi}{4} - 1\right), G_m = g_m \text{ and } R = 2R_{out} \text{ for Fig. 2.4(a)}
\]

\[
a = \left(\frac{\pi}{8} - 1\right), G_m = 2g_m \text{ and } R = R_{out} \text{ for Fig. 2.4(b)}
\]

In (2.12), the 2nd term is from the LNTA, the 3rd term is from the mixer, and the 4th term is from the TIA. The rest of the terms are the noise folding from the odd harmonics of the LO for LNTA, \(R_s\) and \(R_{SW}\), respectively. The NF calculated from

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Fig. 2.4 Equivalent LTI noise model with pre-gain for a 50 % LO (Fig. 2.3a) and b 25 % LO (Fig. 2.3b)
(2.12) for 50 % LO is single sideband (SSB). For a double sideband (DSB) NF, it is 3 dB less. Since the harmonic’s power of 50 % LO is larger than that of 25 % LO, the folding terms of 50 % LO are also higher. From (2.12), the DSB NF of 50 % LO and 25 % LO are plotted in Fig. 2.5 as a function of $A_V$, where $\Delta N_F = N_F_{50\%} - N_F_{25\%}$, $R_{sw} = 50 \Omega$, $\gamma_1 = \gamma_2 = 1$, $g_m = 9 \text{ mS}$, $R_{out} = 200 \Omega$ and $R_{TIA} = 2.5 \text{ k\Omega}$. It can be seen that $\Delta N_F$ is reduced to 0.91 dB (0.51 dB) when $A_V$ is just 2 V/V (3 V/V), which is easily achievable in practice. In fact, a moderated $A_V$ can even eliminate the need of the LNTA (or LNA) [3]. However, when considering also the input matching and LO-to-RF isolation, both pre-gain and LNTA should be employed concurrently. The simulated LO-to-RF isolation is $<-100$ dBm. Due to the passive pre-gain, the IIP3 of the receiver is more demanding than the NF, promoting the use of a 50 % LO. Together with its power advantage (i.e. lower VCO frequency and no divider), our proposed topology (i.e., pre-gain + split-LNTA + 50 % LO) should ease the tradeoff between NF, IIP3, area and power.

2.2.3 IIP3

The 3rd-order intermodulation (IM3) distortion is analyzed to assess the linearity. The aim is to find the in-band IIP3 of the receiver under 50 % LO and 25 % LO in response to two-tone excitation. Assuming that the nonlinearity of the receiver is dominated by the LNTA, its nonlinearity contributions are considered as:

(a) 3rd-order LNTA nonlinearity due to input excitation $v_{in}$ [$\alpha_2 (I/V^3)$].
(b) 3rd-order LNTA nonlinearity due to output excitation $v_x$ [$\alpha_3 (I/V^3)$].

Thus, $i_{ds} = \alpha_1 v_{in} + \alpha_2 v_{in}^3 + \alpha_3 v_x^3$. If the coefficients $\alpha_1$, $\alpha_2$ and $\alpha_3$ are assumed to be proportional to the device $W/L$,

For 50 % LO, $\alpha_1 = g_m$, $\alpha_2 = g_{m3}$, $\alpha_3 = g_{o3}$;
For 25 % LO, $\alpha_1 = 2g_m$, $\alpha_2 = 2g_{m3}$, $\alpha_3 = 2g_{o3}$.

where $g_{m3}$ and $g_{o3}$ are the 3rd-order nonlinear transconductance and conductance, respectively. With a two-tone excitation of amplitude $A$ and the 1st-order voltage
gain and current gain given in (2.1)–(2.11), the IM3 output voltage for each of the nonlinear coefficients listed above can be written as,

\[ v_{o3x2} = \frac{3}{4} g_m A^3 I_{BB1} R_L; \quad v_{o3x3} = \frac{3}{4} g_o A^3 V_{X1} A^3 I_{BB1} R_L \]

for a 50 % LO. Thus,

\[ \text{IM3}_{50\%} = \frac{v_{o3x2} + v_{o3x3}}{v_{01x1}} = \frac{\frac{3}{4} g_m A^3 I_{BB1} R_L + \frac{3}{4} g_o A^3 V_{X1} A^3 I_{BB1} R_L}{A g_m I_{BB1} R_L} \]

Let \( \text{IM3}_{50\%} = 1 \rightarrow \text{IIP3}_{50\%} = \sqrt{\frac{4 g_m}{3(g_m + g_o A^3 V_{X1})}} \) (2.13)

Following the same procedure, the IIP3 for 25 % LO can be derived as,

\[ \text{IIP3}_{25\%} = \sqrt{\frac{4 g_m}{3(g_m + g_o A^3 V_{X2})}} \] (2.14)

Since \( A_{Vx2} > A_{Vx1} \), we can find that, from (2.13)–(2.14), the LNTA’s 3rd-order nonlinearity term is larger for a 25 % LO. Thus, the IIP3 of 50 % LO should be better than that of 25 % LO, benefiting the SFDR since both architectures will feature a similar NF after adding the pre-gain.

### 2.2.4 Current- and Voltage-Mode Operations

Both 25 % LO and 50 % LO architectures can be intensively designed for current-mode or voltage-mode operation. For a high-performance design like [7, 8, 12], \( R_{TIA} \ll R_{out} \) and \( R_{sw} \ll R_{out} \) are preferred to keep the signals in the deep current mode. As such, (2.3) and (2.8) can be simplified as \( G_{m1} = \frac{2 g_m}{\pi} \) and \( G_{m2} = \frac{2 \sqrt{2} g_m}{\pi} \), respectively. Both of them are higher when compared to themselves in the voltage-mode operation. In terms of IIP3 and NF, the current mode is also preferable since \( A_{Vx1} \approx g_m (R_{sw} + \frac{2}{\pi} R_{TIA}) \) and \( A_{Vx2} \approx 2 g_m (R_{sw} + \frac{2}{\pi} R_{TIA}) \) will be lower, and the noise due to the folding term and TIA will be also smaller as noted in (2.12).

Nevertheless, the current-mode operation also brings up two sizing constraints being less attractive for low-power design: (1) a low \( R_{sw} \) entails a large device \( W/L \) and a higher overdrive voltage for the mixers; both calling for a larger power budget in the LO path, and (2) a low \( R_{TIA} \) implies that the TIA has to draw a large bias current. For example, if a low \( R_{TIA} \) of 50 \( \Omega \) is required from the 1.2-V TIA (a common-gate amplifier), its bias current is as high as \( I_{bias} = 2 \) mA for a typical overdrive voltage of 200 mV. Thus, for ultra-low-power applications like
ZigBee/WPAN that has relaxed NF and linearity requirements, higher $R_{sw}$ and $R_{TIA}$ are preferable to operate the receiver more on the voltage mode. A summary of performance differences in current- and voltage-mode operations is given in Table 2.1.

### 2.3 Circuit Techniques

#### 2.3.1 Impedance Up Conversion Matching

From Sect. 2.2, we expect a passive pre-gain $A_v$ of 2 to 3 V/V. As shown in Fig. 2.6a, $A_v$ can be derived under $R_{in} = R_s$,

$$\frac{V_{out}^2}{2R_{out}} = \frac{V_s^2}{8R_s}, V_{out} = V_{in}A_v, V_{in} = 0.5V_s \Rightarrow A_v = \sqrt{\frac{R_{out}}{R_{in}}}$$

Thus, an up-conversion matching network is entailed to ensure $A_v > 1$. A convenient way to achieve it is to use $L_{bw}$ to resonate with $C_{in}$. The schematic is shown in Fig. 2.6b. The parallel connection of $C_{in}$ and $R_{out}$ can be transformed into

![Input impedance matching](image)

**Fig. 2.6** Input impedance matching: a $A_v$ converts $R_{out}$ to $R_{in}$ to match with $R_s$, b $L_{bw}C_{in}$ as an impedance conversion network and its c narrowband equivalent circuit

### Table 2.1 Proposed Receiver under current- and voltage-mode operations

<table>
<thead>
<tr>
<th>Mode</th>
<th>Gain</th>
<th>NF</th>
<th>In-Band</th>
<th>Power</th>
<th>Suitable for</th>
</tr>
</thead>
<tbody>
<tr>
<td>Current mode (Small $R_{sw}$ &amp; $R_{TIA}$)</td>
<td>↗↘</td>
<td>↗</td>
<td>↗</td>
<td>↗</td>
<td>High performance</td>
</tr>
<tr>
<td>Voltage mode (Large $R_{sw}$ &amp; $R_{TIA}$)</td>
<td>↘↗</td>
<td>↘</td>
<td>↘</td>
<td>↘</td>
<td>Ultra low power</td>
</tr>
</tbody>
</table>
a series connection of $C_{\text{ser}}$ and $R_{\text{ser}}$, as shown in Fig. 2.6c. At $L_{bw}C_{\text{ser}}$ resonance, and with $R_{\text{ser}} = R_s$ and $i = \frac{V_s}{2R_{\text{ser}}}$, we have,

$$V_{\text{out}} = V_{R_{\text{ser}}} + V_{C_{\text{ser}}} = \frac{V_s}{2} \left( 1 - j \frac{Q_C}{2} \right)$$

where,

$$V_{R_{\text{ser}}} = -j \frac{Q_C V_s}{2} \frac{C_{\text{ser}} R_{\text{ser}}}{2} = \frac{V_s}{2}$$

$$V_{C_{\text{ser}}} = \frac{1}{j \omega_0 C_{\text{ser}}} \frac{V_s}{2R_{\text{ser}}} = -j \frac{Q_C}{2} V_s,$$

$$\omega_0 = \frac{1}{\sqrt{L_{bw} C_{\text{ser}}}} \text{ and } Q_C = \frac{\sqrt{L_{bw} C_{\text{ser}}}}{R_{\text{ser}}}$$

Interestingly, such a voltage boosting factor $\sqrt{1 + \frac{Q_C^2}{4}}$ is larger than the conventional inductively-degenerated LNA, which is only $\frac{Q_C}{2}$. In fact, when the capacitance of the PCB trace is accounted, the $Q$ of the matching network will be higher, easing the impedance matching.

### 2.3.2 Mixer-TIA Interface Biased for Impedance Transfer Filtering

For the employed single-balanced passive mixers, the RF-to-IF feed through has to be addressed. Based on Fig. 2.7, we can calculate the currents $i_{M7}$ and $i_{M8}$ with respect to the RF current $i_{RF}$ as given by,

$$i_{M7} = \frac{i_{RF}}{2} \left[ 1 - \text{sign}(\cos \omega_{LO} t) \right]$$

$$i_{M8} = \frac{i_{RF}}{2} \left[ 1 + \text{sign}(\cos \omega_{LO} t) \right]$$

They imply that the currents can be decomposed into the differential mode (Fig. 2.7a) with amplitude of $2i_{RF}/\pi$ at BB, and into the common mode (Fig. 2.7b) with amplitude of $0.5i_{RF}$ at RF. To suppress the latter, $C_M$ was added to create a lowpass pole ($C_M//R_{\text{TIA}}$). For the differential IF signal, the pole is located at $(C_M + 2C_1)//R_{\text{TIA}}$, which suppresses the out-of-channel interferers before they enter the TIA. As such, the TIA can be biased under a very small bias current. The resultant high input impedance of the TIA, indeed, benefits both BB and RF
filtering because of the bidirectional impedance-translation property of the passive mixers [7, 8]. Figure 2.8 shows the simulated out-band IIP3, which is subject to the allowed total capacitance of $C_M + 2C_1$. For instance, when $C_M + 2C_1$ is increased from 16 to 42 pF, the out-band IIP3 raises from +2.5 to +4.7 dBm, at the expense of the die area. For the on-resistance of the mixer switches ($R_{sw}$), it involves a tradeoff of the LO path’s power to the out-band IIP3 and NF. As shown in Fig. 2.9, if $R_{sw}$ is increased from 50 to 150 $\Omega$ for power savings, the NF and out-band IIP3 will be penalized by $\sim 1$ dB.

**Fig. 2.7** Equivalent circuits of the mixer-TIA interface for a the differential low-IF signal and b the common-mode RF feed through

**Fig. 2.8** Out-band IIP3 can be improved by allowing more total capacitance of $C_M + 2C_1$

**Fig. 2.9** The on-resistance of the mixer switches represents a tradeoff among the LO-path’s power, out-band IIP3 and NF
2.3.3 RC-CR Network and VCO Co-Design

The LC VCO (Fig. 2.10a) employs a complementary NMOS-PMOS \((M_{1-4})\) negative transconductor. For power savings, \(M_1\) and \(M_2\) are based on AC-coupled gate bias \((V_{\text{vco,b}})\) to lower the supply to 0.6 V. Here, we implement a capacitive divider \((C_{M1} \text{ and } C_{M2})\) to boost the input impedance of its subsequent two-stage RC-CR network (Fig. 2.10b). The optimization details are presented next.

RC-CR network is excellent for low-power and narrowband I/Q generation. With a Type-II architecture, both phase balancing and insertion loss can be better optimized than its Type-I counterpart [15]. For instance, the simulated insertion loss of a two-stage Type-II RC-CR network is roughly 2 dB as shown in Fig. 2.11, which will be raised to 4 to 5 dB if a Type-I topology is applied (not shown). For low-power LO buffering, the amplitude balancing is critical because its imbalance will lead to inconsistent zero-crossing points, resulting in AM to duty-cycle

<table>
<thead>
<tr>
<th></th>
<th>(M_{1,2})</th>
<th>(M_{3,4})</th>
<th>(L_p)</th>
<th>(C_{M1})</th>
<th>(C_{M2})</th>
<th>(R_{N1})</th>
<th>(C_{N1})</th>
<th>(R_{N2})</th>
<th>(C_{N2})</th>
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<tbody>
<tr>
<td>(W)</td>
<td>12</td>
<td>0.12</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>(L)</td>
<td>24</td>
<td>0.12</td>
<td>4.6nH</td>
<td>530fF</td>
<td>280fF</td>
<td>450\Omega</td>
<td>35fF</td>
<td>900\Omega</td>
<td>120fF</td>
</tr>
</tbody>
</table>

Fig. 2.10 a LC VCO and b the proposed input-impedance-boosted two-stage Type-II RC-CR network for 4-phase 50 % LO generation

![Simulated time-domain signals at the output of the VCO (V\(_{\text{vco}}\)), capacitor divider (V\(_p\)) and the RC-CR network (V\(_{\text{RC1}}\))]
distortion. Figures 2.12 (V_{RC1-4}) and 2.13 (LO_{Ip,n} and LO_{Qp,n}) are the simulated transient waveforms, showing the consistent duty cycle and zero-crossing points achieved in the proposed design.

For a RC-CR network operated at 2.4 GHz, if we select R_{N1} = 1 k\Omega, C_{N1} is just 66 fF, which benefits the area, VCO tuning range and phase noise, but the I/Q accuracy over PVT variations should be considered [16]

\[
\frac{\sigma(\text{Image Out})}{\text{Desired Out}} = 0.25 \sqrt{\left(\frac{\sigma_R}{R}\right)^2 + \left(\frac{\sigma_C}{C}\right)^2}
\]

(2.17)

Since ZigBee/WPAN applications call for a low image-rejection ratio (IRR) of 20–30 dB [17], according to (2.17), the matching of the resistors (\(\sigma_R\)) and capacitors (\(\sigma_C\)) can be relaxed to 2.93 % for a 30-dB IRR (3\(\sigma\)). The sizes of C_{N1,2} and R_{N1,2} are summarized in Fig. 2.10. The poles from C_{N1,2} and R_{N1,2} are distributed around 2.4 GHz to cover the PVT variations. The impact of R_{N1} to the VCO can be analyzed as follows:

When the VCO’s inductor is 4 nH with a Q of 20 (\(R_P \approx 1.2 \text{ k}\Omega\)), we have \(R_{\text{tank}} \approx 0.5R_P/0.5R_{N1}\). Thus, directly connecting the RC-CR network to the VCO will limit the LC tank’s Q_{tank} degrading the phase noise [18, 19]. To alleviate this, we boost up the equivalent input resistance of the RC-CR network (R_{eq}) by adding a capacitive divider (C_{M1} and C_{M2}). For the total tank capacitance C_{tank}, it can be approximated as
\[ C_{\text{tank}} \approx 2C_{\text{var}} + \frac{(C_{M2} + 2C_{N1})C_{M1}}{C_{M1} + C_{M2} + 2C_{N1}} \]  

(2.18)

By defining an input-impedance boosting factor \( n \),

\[ n = \frac{C_{M1}}{C_{M1} + C_{M2} + 2C_{N1}} \]  

(2.19)

we have

\[ V_{P1} \approx nV_{\text{VCOp}} \]  

(2.20)

It means that the signal swing \( (V_{P1}) \) delivered to the \( RC-CR \) network are in trade-off with \( n \). Handily, in our VCO, sweeping \( V_{\text{vco,b}} \) can track the phase noise with the output swing (Fig. 2.14). Given a bias current and a phase noise target, \( R_{\text{tank}} \) can be set from \( V_{\text{VCOp}} \approx 2I_{\text{bias}}R_{\text{tank}} \), and \( n \) can be set from (2.21) with a specific \( R_p \) and \( R_{\text{eq}} \),

\[ R_{\text{tank}} \approx \frac{R_{\text{eq}}}{n} \parallel \frac{R_p}{2} \]  

(2.21)

In this work, \( n = 0.6 \) is selected to balance the output swing with \( C_{\text{tank}} \) and the total tank resistance \( (R_{\text{tank}}) \).

### 2.4 Experimental Results

The receiver (Fig. 2.15) fabricated in 65-nm CMOS occupies an active area of 0.14 mm\(^2\) and is encapsulated in a 44-pin CQFP package for PCB-based measurements. The estimated bond wire inductance is \( \sim 7 \) nH for the provided package (13.5 × 13.5 mm). Figure 2.16 shows that the measured \( S_{11} \) is \( -8 \) dB within 2.24–2.46 GHz (for a different package, external inductor or capacitor can be added to optimize \( S_{11} \)). The simulation results with and without considering the PCB trace capacitances are also given. The measured voltage gain is 32.8–28.2 dB and the DSB NF is between 8.6–9 dB for an IF spanning from 1 to 3 MHz, as shown in Fig. 2.17. We also measured the gain and NF from 2.2 to 2.6 GHz (Fig. 2.18).
2.4 Experimental Results

**Fig. 2.15** Chip micrograph of the fabricated receiver

![Chip micrograph of the fabricated receiver](image)

**Fig. 2.16** Measured $S_{11}$, and simulated $S_{11}$ with and without $C_{pcb}$

![Measured S11, and simulated S11 with and without Cpcb](image)

**Fig. 2.17** Measured receiver gain and NF versus BB frequency

![Measured receiver gain and NF versus BB frequency](image)

**Fig. 2.18** Measured receiver gain and NF versus input signal frequency

![Measured receiver gain and NF versus input signal frequency](image)
For a narrowband receiver, the linearity is mainly justified by the out-channel linearity tests. According to the case given in [17, 20], two tones are applied at \([f_{LO} + 10 \text{ MHz}, f_{LO} + 22 \text{ MHz}]\) with a power level sweeping from \(-24\) to \(-32\) dBm. Because of the RF and baseband filtering associated with the bidirectional property of passive mixers, the out-band IIP3 (Fig. 2.19) achieves \(-7\) dBm and the \(P_{1dB}\) is \(-26\) dBm.

For the VCO, it measures 21% tuning range from 2.623 to 2.113 GHz, as shown in Fig. 2.20. At 3.5-MHz offset, the phase noise (Fig. 2.21) is \(-112.46\) dBc/Hz.

![Fig. 2.19 Measured out-of-band IIP3](image1)

![Fig. 2.20 Measured VCO turning range](image2)

![Fig. 2.21 Measured VCO phase noise at 2.4 GHz](image3)
fulfilling the specification (−102 dBc/Hz [17, 20]) with an adequate margin. From frequency 100 kHz to 1 MHz, the result fits the $1/f^3$ slope, and from 1 to 10 MHz, it starts to be saturated, primarily limited by the small output amplitude (−28.31 dBm) of the test buffer.

Based on transient measurements, the I/Q BB differential outputs (Fig. 2.22) has $\sim 0.08$ dB gain mismatch and 2° phase match, corresponding to an IRR of $\sim 25$ dB.

The performance summary and benchmark are given in Table 2.2 [5, 17, 21–27]. This work [28] succeeds in achieving the highest power and area efficiencies via proposing a mixed-$V_{DD}$ topology co-optimized with a number of circuit techniques. Only one on-chip inductor is entailed in the VCO. The achieved NF and out-band IIP3 correspond to a competitive SFDR of 59.4 dB according to [17, 19],

$$\text{SFDR} = \frac{2(P_{IIP3} + 174\text{dBm} - \text{NF} - 10\log B)}{3} - \text{SNR}_{\text{min}}$$ (2.22)

where $\text{SNR}_{\text{min}} = 4$ dB is the minimum signal-to-noise ratio required by the application, and $B = 2$ MHz is the channel bandwidth. As presented in Figs. 2.8 and 2.9, the SFDR can be further optimized by allowing more budgets in area (bigger $C_M + 2C_1$) and/or power (smaller on-resistance of the mixer switches), being a design-friendly architecture easily adaptable to different specifications.

Fig. 2.22 Measured I/Q BB transient outputs
Table 2.2 Performance summary and benchmark with the state-of-the-art

<table>
<thead>
<tr>
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<tbody>
<tr>
<td>Gain (dB)</td>
<td>35</td>
<td>67</td>
<td>24.5</td>
<td>51</td>
<td>22.5</td>
<td>30</td>
<td>83</td>
<td>55</td>
<td>32</td>
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<td>DSB NF (dB)</td>
<td>7.5</td>
<td>16</td>
<td>16.5 (SSB)</td>
<td>3.2</td>
<td>7 (SSB)</td>
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<td>6.1</td>
<td>9</td>
<td>8.8</td>
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<td>Out-band IIP3 (dBm)</td>
<td>−10</td>
<td>−10.5</td>
<td>−19 (in-band data)</td>
<td>−32 (in-band data)</td>
<td>−21.5 (in-band data)</td>
<td>−8</td>
<td>−21.5</td>
<td>−6</td>
<td>−7</td>
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<tr>
<td>SFDR (dB)</td>
<td>58.3</td>
<td>52.3</td>
<td>38.3</td>
<td>36.5</td>
<td>51</td>
<td>59.8</td>
<td>51.6</td>
<td>60</td>
<td>59.4</td>
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<td>VCO phase noise (dBc/√Hz)</td>
<td>N/A</td>
<td>−127 @ 3 MHz</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>−112 @ 1 MHz</td>
<td>−115 @ 3.5 MHz</td>
<td>−111.4 @ 3.5 MHz</td>
</tr>
<tr>
<td>Power (mW)</td>
<td>5.4 (w/o VCO)</td>
<td>32.5 (w/VCO)</td>
<td>2.52 (w/o VCO)</td>
<td>8.1 (w/o VCO)</td>
<td>1.06 (w/o VCO)</td>
<td>1.8 (w/o VCO)</td>
<td>1.6 (w/VCO)</td>
<td>2.7 (w/VCO)</td>
<td>1.4b (w/VCO)</td>
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<td>No. of inductor or transformer</td>
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<td>3</td>
<td>5</td>
<td>3</td>
<td>2</td>
<td>4</td>
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<td>1</td>
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<td>Die area (mm²)</td>
<td>0.23 (w/o VCO)</td>
<td>2.88 (w/VCO)</td>
<td>N/A</td>
<td>1.27 (w/o VCO)</td>
<td>1.1 (w/o VCO)</td>
<td>2.07a (w/o VCO)</td>
<td>2.5a (w/VCO)</td>
<td>0.26a (w/VCO)</td>
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<td>Supply (V)</td>
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<td>0.6</td>
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<td>1.8</td>
<td>1.2</td>
<td>1.8</td>
<td>0.3</td>
<td>0.6/1.2</td>
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<td>CMOS Tech.</td>
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<td>90 nm</td>
<td>0.18 μm</td>
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<td>0.18 μm</td>
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<td>65 nm</td>
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</table>

aInclude more BB gain stages and filters. bThe power breakdown is LNTA: 0.4 mW, TIA: 0.18 mW and VCO + Buffer: 0.82 mW
2.5 Conclusions

A mixed-$V_{DD}$ 2.4-GHz ZigBee/WPAN receiver measuring state-of-the-art performances has been described. It features passive pre-gain, a split-LNTA, a high-input-impedance BB TIA and a low-power 50% LO generation scheme. They together lead to improved power and area efficiencies, as well as a high SFDR while eliminating the need of a RF balun. These beneficial features render this work as a superior receiver candidate for cost and power reduction of ZigBee/WPAN radios in nanoscale CMOS.

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