# On the Design of Broadband Power-to-Current Low Noise Amplifiers

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Abstract-Aiming for the simultaneous realization of constant gain, accurate input impedance match, and minimum noise figure over a wide frequency range, the circuit topologies and detailed design of two broadband dual-loop negative feedback power-to-current low noise amplifiers (LNAs) are presented in this paper: 1) a resistive indirect-feedback power-to-current LNA, which requires an active part with two output terminal pairs, and 2) a transformer feedback power-to-current LNA, which requires a transformer in its current feedback path having a high turn ratio with high magnetic coupling. For both LNAs the feedback networks and active part implementations are discussed in detail. It is shown that for this purpose a novel stacked transformer can be realized using only two metal layers. The two LNAs are designed to be implemented in a 0.2  $\mu$ m GaAs p-HEMT technology process to verify the theory presented. Counter measures are applied to deal with the effects of bond wires and the effects of transformer parasitics on the circuit performance are analyzed. Simulation results show that the resistive indirect-feedback power-to-current LNA exhibits a 0.6-0.8 dB noise figure, an input return loss well below -15 dB, a 200 mS voltage-to-current gain (which corresponds to 23 dB power gain for a 50  $\Omega$  load) from 0.3 GHz to 4 GHz, a -1 dBm third-order input intercept point (IIP3) and a 23 dBm second-order input intercept point (IIP2) at 2 GHz. It consumes 73 mA current from a 4 V power supply. The transformer-feedback power-to-current LNA achieves a 0.5-0.8 dB noise figure, an input return loss of less than -15 dB, a 22 dB power gain from 0.8 GHz to 4 GHz, and a 0 dBm IIP3 and 22 dBm IIP2 at 2 GHz while drawing 53 mA current from the 4 V power supply.

*Index Terms*—Broadband, double-loop negative feedback amplifier, GaAs, indirect feedback, input power matching, low noise amplifier (LNA), noise matching, power-to-current, transformer feedback.

#### I. INTRODUCTION

**T** HE NEW generation receivers, such as needed for the Square Kilometer Array (SKA) (0.6–1.6 GHz), a new radio telescope developed by the Netherlands Foundation for Research in Astronomy (ASTRON), and global navigation satellite systems (0.9–1.7 GHz), which offer compatibility and interoperability among GPS (USA), Galileo (Europe), GLONASS (Russia), and COMPASS (China), demand high

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Fig. 1. Dual-loop power-to-current amplifier.

performance low noise amplifiers operating across a wide radio frequency band.

Conventional solutions for broadband LNAs employ LC-ladders [1], distributed amplifiers [2], common-gate (CG) amplifiers [3], [17], and resistive shunt feedback (RSF) amplifiers [4], [5], [18], [20]. However, these amplifiers have significant trade-offs among noise, power matching, linearity, chip area, and power consumption. LNAs with multiple LC sections and distributed LNAs offer good input power matching over a reasonably wide bandwidth but at the expense of low linearity, large power consumption, and large chip area. In these amplifiers, noise is deteriorated by the series resistance of the low Q-factor LC sections, and the linearity is compromised due to the nonlinear transistor characteristics. Both CG amplifiers and RSF amplifiers exhibit good wideband input power matching, but significantly degraded noise performance due to either the unit current gain or the extra noise contributed by the resistive feedback network.

The two amplifiers described in this work are based on the dual-loop power-to-current (referred to as PI) circuit configuration shown in Fig. 1, which has its current-to-current feedback implemented by a transformer with a turn-ratio n, and a current-to-voltage feedback realized by a trans-impedance Z. The active part is a nullor, which is an ideal element with infinite transfer parameters [6]. As a result, the input impedance of the PI is given by

$$Z_{\rm in} = nZ,\tag{1}$$

which is dependent only on the two feedback elements. Therefore, good input power matching over a wide bandwidth can be obtained by properly designing the feedback networks. Further, a matching network at the output is not necessary since the output signal quantity is current. Noise in a PI can be optimized by choosing appropriate biasing current levels, and linearity can be fulfilled by ensuring a sufficient amount of loop gain.



Fig. 2. Possible circuit topologies of the power-to-current amplifier. (a)  $\alpha$  + Z. (b)  $\alpha$  + Active\_Z. (c)  $\alpha$  + TF\_Z. (d)  $\alpha$  + Z\_TF. (e) TF + Z. (f) TF + Active\_Z. (g) TF + TF\_Z (h) TF + Z\_TF.

First, Section II introduces eight possible circuit topologies of the PI. The noise performance of each configuration is evaluated by its total noise power spectral density (NPSD), resulting in two optimal candidates for broadband applications: one is a resistive indirect-feedback PI and the other is a transformer-feedback PI. Practical implementations for the

|     |    |   |  | I-V   |   |
|-----|----|---|--|---|---|
|     |    | Ζ   | Active_Z   | TF_Z  | Z_TF  |
|     |    | $(S_{nv} + 4KTR_f)(1 + \frac{R_s}{R_1 + R_2})^2$              | $(S_{nv} + 4KTR_f)(1 + \frac{R_s}{R_1 + R_2})^2$   | $(S_{nv} + 4KTR_f)(1 + \frac{R_s}{R_1 + R_2})^2$              | $(S_{nv} + \frac{4KTR_f}{m^2})(1 + \frac{R_s}{R_1 + R_2})^2$                    |
| II  | α  | $+S_{ni}(R_{s}+R_{f})^{2} + \frac{4KTR_{s}^{2}}{R_{1}+R_{2}}$ | $+S_{ni}R_{s}^{2} + \frac{4KTR_{s}^{2}}{R_{1} + R_{2}}$ $+S_{nv}(1 + \frac{R_{s}}{R_{1} + R_{2}})^{2}$ | $+S_{ni}(R_{s}+R_{f})^{2} + \frac{4KTR_{s}^{2}}{R_{1}+R_{2}}$ | $+S_{ni}\left(\frac{R_{f}}{m^{2}}+R_{s}\right)^{2}$ $+\frac{4KTR_{s}^{2}}{R+R}$ |
| 1-1 | TF | $S_{nv} + 4KTR_f + S_{ni}(R_s + R_f)^2$                       | $+S_{ni}R_{f}^{2}$ $S_{nv} + 4KTR_{f}$ $+S_{ni}R_{s}^{2}$ $+S_{nv'} + S_{ni'}R_{f}^{2}$                | $S_{nv} + 4KTR_f + S_{ni}(R_s + R_f)^2$                       | $S_{nv} + \frac{4KTR_f}{m^2}$ $+ S_{ni}(R_s + \frac{R_f}{m^2})^2$               |

 TABLE I

 The NPSD of the Power-to-Current Amplifiers

employed nullor are also proposed. Section III illustrates the transformer design dedicated to the transformer-feedback PI, targeting both a high magnetic coupling and a high turn ratio in a two-metal layer technology. In Section IV, the specific circuit designs of the two PIs are presented. Section V deals with the influence of bond wires. Frequency compensation is applied to mitigate their effects on the signal transfer function and input impedance. Section VI analyzes the performance of the designed resistive indirect-feedback PI and transformer-feedback PI. Finally, Section VII shows the simulation results.

#### II. CIRCUIT TOPOLOGIES OF THE POWER-TO-CURRENT LNA

#### A. Feedback Networks

The dual-loop negative feedback PI has two feedback loops: a current-to-current (I-I) feedback loop and a current-to-voltage (I-V) feedback loop [6], [13]. The I-I network can be implemented simply by means of a passive current divider (denoted as  $\alpha$ ) or a transformer (TF). The alternatives of the I-V network can either be a (single) passive element (denoted as Z), an active network with a (or some) passive element(s) (denoted as Active\_Z), or a transformer together with a (or some) passive element(s). Furthermore, in the last case, the output current can either be fed back by a transformer first and then transferred to a voltage by the passive element (denoted as TF\_Z), or *vice versa* (denoted as Z\_TF). Therefore, there are at least eight possible PI configurations, which are grouped into two categories:

 $\alpha$  group: 1)  $\alpha$  + Z, 2)  $\alpha$  + Active\_Z, 3)  $\alpha$  + TF\_Z, 4)  $\alpha$  + Z-TF

TF group: 1) TF + Z, 2) TF + Active\_Z, 3) TF + TF\_Z, 4) TF + Z\_TF

In the  $\alpha$  group, it is hard for a typical two-port network to deliver its output signal directly to the load since both output terminals of the nullor will be used as feedback networks. A promising solution to overcome this problem is to introduce an additional output terminal (or an output terminal pair), which copies the output signal and then either feeds it back or delivers it directly to the load, thereby resulting in a so-called indirect-feedback network. Consequently, the nullor presented in the  $\alpha$  group should have four output terminals (or two output terminal pairs). On the other hand, a transformer can be directly connected in series with one of the output terminals and therefore the active parts of the amplifiers in TF group do not need additional output terminals.

Fig. 2 presents the eight circuit topologies of the PIs mentioned above, where the first four configurations apply the indirect-feedback network. The active part in the figure is modeled by a nullor together with a voltage noise source  $v_n$  (with NPSD  $S_{nv}$ ) and an uncorrelated current noise source  $i_n$  (with NPSD  $S_{ni}$ ). Table I shows the calculated NPSD of each PI when assuming the transformer to be ideal, where  $R_s = \operatorname{Re}(Z_s), R_f =$  $\operatorname{Re}(Z_f), R_1 = \operatorname{Re}(Z_1), \text{ and } R_2 = \operatorname{Re}(Z_2).$  It can be seen in the table that the  $\alpha + Z$  and TF + Z topologies have the lowest NPSD, and the  $\alpha$  + Active\_Z and TF + Active\_Z have the highest NPSD in their groups. The  $\alpha$  + TF<sub>-</sub>Z (or TF + TF<sub>-</sub>Z) and the  $\alpha$  + Z\_TF (or TF + Z\_TF) have almost comparable NPSD. Therefore, the  $\alpha + Z$  PI (from now on referred to as  $\alpha$ -PI) and the TF + Z PI (from now on referred to as TF-PI) are favored in this work. For the  $\alpha$ -PI, noise mainly comes from the active part and two passive feedback loops. For the TF-PI, noise is mainly contributed by the active part and the passive element Z. The NPSD of the TF-PI is expected to be slightly lower than that of the  $\alpha$ -PI if the transformer is ideal.

# B. Nullor Implementation

As mentioned in the previous section, the nullor present in the  $\alpha$ -PI should have two output terminal pairs. Possible implementations of such a nullor are shown in Fig. 3. The nullor in Fig. 3(a) is formed by means of a single stage (denoted as M1), such as a common-source (CS) stage. Alternatively, single stages in the form of a common-emitter (CE) stage, cascode stage, or differential pair are also possible. An additional stage M2, which is identical to M1, is introduced to duplicate the output signal of M1. As a result, M1 and M2 provide two output terminal pairs. The drawback of this solution is that the loop gain is small and the overall noise of this stage will be increased by a factor of  $\sqrt{2}$ . Therefore, for sufficiently large (absolute value of the) loop gain, a two-stage configuration is usually preferred.

Fig. 3(b) shows a two-stage realization of the active part of the LNA where both the input stage and the output stage are single-ended (i.e., a CS or CE stage, or a cascode stage—here a



Fig. 3. Nullor implementation of the  $\alpha$ -PI. (a) Single stage. (b) Two single-ended stages. (c) Single-ended input and differential output. (d) Differential input and single-ended output. (e) Two differential stages.

CS stage is shown). Output transistors M2 and M3 are identical. It must be noted that this active part has a positive transfer and thus cannot produce a negative loop gain.

Fig. 3(c) shows a two-stage realization of the active part of the LNA where the input stage is single-ended (i.e., a CS or CE stage, or a cascode stage—here a CS stage is shown), while the output stage is a differential pair. Transistor pairs M2, M3 and M4, M5 are identical.

Fig. 3(d) shows a two-stage realization of the active part of the LNA where the input stage is a differential pair, while the output stage is single-ended (i.e., a CS or CE stage, or a cascode stage—here a CS stage is shown). Output transistors M3 and M4 are identical.

Finally, Fig. 3(e) shows a two-stage realization of the active part of the LNA where both the input stage and the output stage are differential pairs. Transistor pairs M3, M4 and M5, M6 are identical.

The current feedback transformer offers some design flexibility when implementing the nullor, since it offers both inverting and noninverting feedback possibilities. Therefore, the only requirement on nullor alternatives in the TF-PI is that it should provide sufficient contribution to the loop gain. The possible configurations are analogue to that of the  $\alpha$ -PI but without the replica stage. Specifically, transistor M2 in Fig. 3(a), M3 in Fig. 3(b), and M4 in Fig. 3(d), and transistor pairs M4, M5 in Fig. 3(c) and M5, M6 in Fig. 3(e), respectively, are not necessary in the TF-PI.

#### **III. TRANSFORMER DESIGN**

Practically, the transformer is realized by two magnetically coupled inductors with a turn ratio n and a magnetic coupling factor k [7]–[9]. There are some constraints on the physical transformer utilized as current feedback. First, a high turn ratio n is desired for high gain and a high k-factor for wide bandwidth. Secondly, a high self-inductance of the primary winding is demanded for good input power matching, and a high Q-factor

for low noise. Last but not least, the size of the transformer should be as small as possible to minimize the parasitic resistance of the two coupled inductors and parasitic capacitance of both port-to-substrate and port-to-port. Based on a technology with two metal layers, two possible transformer layout schemes are proposed to be used in the TF-PI. One is the hybrid interleaved and tapped transformer (HIT) we have reported in [10], and the other is a modified stacked transformer (MST) shown in Fig. 4. The primary winding of the MST is a spiral inductor formed with the top metal layer. In order to obtain both high k and high n while minimizing parasitic resistance, the secondary winding (with doubled line width) is constructed with two single-turn inductors, which are connected in parallel and formed by the bottom metal layer. As shown in Fig. 4(b), the two terminals of both the primary ports and secondary ports are located on the same metal layer (on the bottom metal in this case). Moreover, the terminals of the primary ports are placed in the inner side of the terminals of the secondary ports so that they do not crossover each other.

The designed MST is simulated with Momentum. The options of the thick conductor expansion of metal layers and the horizontal side current and edge mesh are enabled to obtain accurate results during simulation. Table II shows some key parameters of the designed MST together with the HIT based on a 0.2  $\mu$ m p-HEMT technology process [11]. The parameters of both the HIT and MST are derived using the methods described in [7], [10], [12]. From the table, it can be found that the MST occupies less area but achieves higher k (0.6) and higher effective turn-ratio n' (18) than the HIT (where k = 0.4 and n' = 14).

## **IV. CIRCUIT DESIGN**

To evaluate the performance of the power-to-current amplifiers proposed, both a  $\alpha$ -PI and a TF-PI are designed to be implemented in the process given in the previous section. The I-I feedback network of the  $\alpha$ -PI is implemented simply by means of a resistive current divider ( $R_1$  and  $R_2$ ), while the I-I feedback



Fig. 4. Layout of the transformer. (a) HIT. (b) MST and its cross section.

TABLE II Key Parameters of the Designed Transformer

| Parameters | OD $\mu$ m | W1 $\mu$ m | W2<br>µm | S<br>µm | $L_p$ nH | $R_p \ \Omega$ | L <sub>s</sub><br>nH | C <sub>f</sub><br>fF | M<br>nH | k   | n' |
|------------|------------|------------|----------|---------|----------|----------------|----------------------|----------------------|---------|-----|----|
| HIT        | 440        | 10         | 20       | 3       | 23.5     | 27.8           | 0.6                  | 114                  | 1.2     | 0.4 | 14 |
| MST        | 396        | 10         | 20       | 3       | 22       | 21             | 0.45                 | 147                  | 1.8     | 0.6 | 18 |

network of the TF-PI is implemented by means of the MST designed in the previous section. The I-V feedback of both amplifiers is formed by means of a resistor  $R_f$ , which is created by two 1  $\Omega$  resistors in parallel in order to handle the large bias current.

The nullor is constructed by the configuration of Fig. 3(c) with high electron mobility transistors (HEMTs). The input stage of the nullor consists of three HEMTs, M1, M2, and M3, each of them having eight gate fingers with a length of 50  $\mu$ m. M1 and M2 are in parallel in order to increase the gate width and reduce the optimal noise impedance, and M3 is cascaded with M1 and M2 to improve the reverse isolation and the output impedance of the input stage, which is initially low due to the large bias current. In addition, the gate-source voltage ( $V_{\rm gs}$ ) should be chosen in such a way that the HEMTs attain a high cutoff frequency ( $f_T$ ) at relatively low NFmin. From Fig. 5 it turned out that a  $V_{\rm gs}$  of 0.5 V results in a 60 GHz  $f_T$  and an NFmin below 0.3 dB at 4 GHz if the drain-source voltage is 1 V.

The output signal of the input stage is directly coupled to the output stage, which is a differential pair configuration to achieve sufficient negative loop gain. The output stage of the  $\alpha$ -PI contains HEMTs M4, M5, M6, and M7, each of them having six gate fingers with a length of 40  $\mu$ m. As a replica of M4 and M5, the differential pair M6 and M7 is used for indirect feedback. In contrast, the output stage of the TF-PI is formed only by the differential pair M4 and M5. Fig. 6(a) and (b) are the circuit diagrams of the  $\alpha$ -PI and TF-PI, respectively.



Fig. 5.  $f_T$  and NFmin as function of  $V_{gs}$  for n-HEMT.

## V. BOND WIRE EFFECTS AND FREQUENCY COMPENSATION

The circuit topology of the power-to-current amplifier makes it possible to constrain the influence of bond wires (with inductance  $L_{\rm BW}$ ) on the circuit performance in advance by applying proper frequency compensation techniques. For instance, the bond wire in series with  $R_f$  in a power-to-current amplifier has a significant effect on the power matching, gain, and stability. It will introduce a low frequency zero at  $R_f/L_{\rm BW}$  in the input impedance, a pole in the signal transfer function and a (phantom) zero in the loop gain at the same frequency. There are three measures to mitigate the effects of this bond wire. First, it is possible to shift this low frequency zero or pole to a higher frequency by putting multiple bond wires in parallel. Therefore, multiple pads should be placed between the termination of  $R_f$ and the ground in the layout.



Fig. 6. Circuit diagram. (a)  $\alpha$ -PI. (b) TF-PI.

Secondly, the influence of the bond wire on the input impedance in the  $\alpha$ -PI can be reduced by introducing a counteracting inductor  $L_1$  in series with  $R_1$ . The required inductance is determined approximately by

$$L_1 \approx \frac{R_1}{R_f} L_{\rm BW}.$$
 (2)

However, a large inductance may arise due to a large  $R_1$ . This constrains the effectiveness of  $L_1$ . Here we choose  $L_1 = 2.5$  nH.

Thirdly, a Miller capacitor between the input and the output of the second stage is capable of bringing one of the loop poles to  $R_f/L_{\rm BW}$ , thereby cancelling the low frequency (phantom) zero introduced by the bond wire.

The loop gain of the  $\alpha$ -PI or TF-PI is given by (3) at the bottom of the next page, where  $R_d$  is the bias resistor of the first stage,  $C_{gsi}(i = 1, 2)$  and  $g_{mi}(i = 1, 2)$  are the gate-source capacitances and trans-conductance factors of the transistors in the *i*th stage. The two loop poles are located at  $P_1 = -(1 + g_{m1}R_f)/4\pi R_s C_{gs1} = -2.2$  GHz and



Fig. 7. Root locus of the power-to-current amplifiers. (a) Without frequency compensation. (b) With frequency compensation.

 $P_2 = -1/2\pi R_d C_{gs2} = -13.6$  GHz. The phantom zero is around  $P_z = -R_f/2\pi L_{\rm BW} \approx -300$  MHz when there are four bond wires in parallel and each has an inductance of 1 nH. Therefore, the Miller capacitor required can be approximately determined by

$$C_m \approx \frac{L_{\rm BW} - [C_{gs1} + (1 + g_{m1}R_f)C_{gs2}]R_dR_f}{(1 + g_{m1}R_f)R_dR_f}.$$
 (4)

The root locus and loop gain before and after frequency compensation are shown in Fig. 7. As a result, the Miller capacitor effectively increases the phase margin and thereby effectively improves the stability of the amplifier.

#### VI. PERFORMANCE ANALYSIS

#### A. Noise

The dominant noise in a HEMT is the channel thermal noise  $i_{nd}$ . It can be transferred to the input of the active part as a noise voltage

$$v_n = -\frac{1}{g_{m1}} i_{nd} \tag{5}$$

and a noise current

$$i_n = -\frac{j\omega}{\omega_T} i_{nd}.$$
 (6)

Substituting the above equations into Table I yields the NPSD of the  $\alpha$ -PI

$$S_{n\_\alpha\_PI}(\omega) = 4KT \left\{ \left(\frac{c}{g_{m1}} + R_f\right) \left(1 + \frac{R_S}{R_1 + R_2}\right)^2 + \frac{R_S^2}{R_1 + R_2} + cg_{m1} \left(\frac{\omega}{\omega_T}\right)^2 (R_S + R_f)^2 \right\},$$
(7)

where  $\omega_T = 2\pi f_T$ , K is Boltzmann's constant, T is the absolute temperature, and c is a process dependent constant. Equation (7) indicates that both a large drain current and a large cutoff frequency of the HEMT, and a small  $R_f$  lead to an optimal noise performance. In the TF-PI, the primary winding is connected in parallel with the input of the amplifier, and the secondary winding is connected in series with the output of the amplifier. The series resistance of the primary winding  $R_p$  provides a large noise current [13], which is

$$i_{n\_TF} = \frac{v_{n\_TF}}{R_p + j\omega L_p} \tag{8}$$

where  $v_{n_{-}TF}$  is the thermal noise voltage of  $R_p$ . Located at the output of the amplifier, the series resistance of the secondary winding is small and therefore is of no importance [14]. Taking the noise sources of the HEMTs, transformer, and  $R_f$  into account, the NPSD of the TF-PI becomes

$$S_{n\_TF\_PI}(\omega) = 4KT \left\{ \left( \frac{c}{g_{m1}} + R_f \right) \left( 1 + \frac{R_S^2}{\omega^2 L_p^2 + R_p^2} \right) + \frac{R_p R_S^2}{\omega^2 L_p^2 + R_p^2} + cg_{m1} \left( \frac{\omega}{\omega_T} \right)^2 (R_S + R_f)^2 \right\}.$$
 (9)

Fig. 8 shows remains that noise beyond frequency low in the TF-PI the  $\sqrt{[(c+g_{m1}R_f)(R_p^2+R_S^2)+g_{m1}R_pR_S^2]/(c+g_{m1}R_f)/2\pi L_p}.$ Moreover, a noise contribution analysis shows that the double-loop feedback networks of the  $\alpha$ -PI and TF-PI contribute 1.5% and 19% of the total noise, respectively. In both amplifiers, more than two-third of the noise is contributed by the first stage. The second stage of the  $\alpha$ -PI contributes much more noise (about 8%) than that of the TF-PI since it contains an indirect feedback network (M6 and M7). The

$$L(s) = -\frac{2g_{m1}g_{m2}R_sR_d}{(\alpha \text{or } n')(1+g_{m1}R_f)\left(1+sC_{gs1}\frac{2R_s}{1+g_{m1}R_f}\right)(1+sC_{gs2}R_d)},$$
(3)



Fig. 8. NPSD of the power-to-current amplifiers.

equivalent noise voltage (ENV) and noise contribution (NC) at 2 GHz of some elements are listed in Table III.

#### B. Input Power Matching

The input impedance of the  $\alpha$ -PI or TF-PI, in case a nullor and an ideal transformer (with turn ratio n) are present in the circuit, equals

$$Z_{\rm in}(\omega) = (\alpha \text{ or } n)R_f \approx R_S \left(\alpha = \frac{R_1 + R_2}{R_1}\right)$$
(10)

However, when taking the bond wire  $L_{\rm BW}$  and its counter measure  $L_1$  into account, the input impedance of the  $\alpha$ -PI is changed into

$$Z_{\text{in}\_\alpha\_PI}(\omega) = \frac{(R_f + sL_{BW})(R_1 + R_2 + sL_1)}{R_1 + sL_1}$$
(11)

The nonidealities of a physical transformer, such as the selfinductance and the parasitic capacitance, will affect the input impedance of the TF-PI. At low frequencies, it is

$$Z_{\text{in_TF_PI_L}}(\omega) = \frac{n'(R_f + sL_{\text{BW}})(R_p + sL_p)}{n'R_f + R_p + s(L_p + L_{\text{BW}})}$$
(12)

while at high frequencies it is approximately

$$Z_{\text{in\_TF\_PL}H}(\omega) \approx \frac{n' R_f (1 + s^2 C_f L_s)}{1 + (n' + 1) s^2 C_f L_s}.$$
 (13)

Fig. 9 shows both the calculated and simulated input impedance of the  $\alpha$ -PI and TF-PI. It can be seen that although the input impedance of the  $\alpha$ -PI is degraded by the bond wire, it is able to achieve good power matching in the angular frequency

band  $[R_f/L_{\rm BW}, (R_1 + R_2)/L_1]$  or equally [0.3 GHz, 53 GHz] for the  $\alpha$ -PI, and  $[(R_S + R_p)/(L_p + L_{\rm BW}), 1/\sqrt{(n'+1)L_sC_f}]$ or equally [0.5 GHz, 6 GHz] for the TF-PI. Note that the deviation from the calculated value and the simulated result in the figure comes from the difference between the nullor (which provides an infinite loop gain) and its practical implementation with HEMTs.

#### C. Signal Transfer Function and Bandwidth

With a nullor and an ideal transformer, the voltage-to-current signal transfer function of the  $\alpha$ -PI or TF-PI is determined by

$$H(\omega) = \frac{i_o}{v_s} = \frac{1}{2R_f}.$$
(14)

However, with the bond wire and its counter measure, the signal transfer function of the  $\alpha$ -PI is changed to

$$H_{\alpha\_PI}(\omega) = \frac{(R_1 + R_2 + sL_1)}{R_S(R_1 + sL_1) + (R_f + sL_{BW})(R_1 + R_2 + sL_1)}.$$
(15)

With a physical transformer, the frequency responses of the signal transfer function of a TF-PI at both low and high frequencies are shown in (16) and (17) at the bottom of the page, respectively. Fig. 10 shows both the calculated and simulated signal transfer function of the  $\alpha$ -PI and TF-PI. The signal transfer function is almost flat in the angular frequency band [DC,  $(R_1 + R_2)/L_1$ ] or equally [DC, 53 GHz] for the  $\alpha$ -PI, and [ $(R_S + 2R_p)/(L_p + L_{\rm BW})$ ,  $2/\sqrt{(n'+2)L_sC_f}$ ] or equally [0.7 GHz, 8.8 GHz] for the TF-PI. Note that the deviation from the calculated value and the simulated result in the figure is also due to the nonideal active part.

The maximum attainable bandwidth of the  $\alpha$ -PI and TF-PI is approximately given by

$$\omega_m \approx \omega_T / \sqrt{(\alpha \text{ or } n')} \tag{18}$$

It is a fraction of the cutoff frequency of the HEMT. For  $\alpha$  or n' approximately equal to 18, the maximum attainable bandwidth  $f_m$  is about 14 GHz. However, the effective bandwidth of a PI should also take into account the (flat) bands of the input power matching, NPSD, and signal transfer function. Consequently, the effective bandwidth of the  $\alpha$ -PI and TF-PI are  $[R_f/L_{\rm BW}, 0.1\omega_T]$  or equally [0.3 GHz, 6 GHz], and

$$H_{TF\_PI\_L}(\omega) = \frac{n'(R_p + sL_p)}{R_S[R_S + R_p + s(L_p + L_{BW})] + n'(R_f + sL_{BW})(R_p + sL_p)}$$
(16)  
and  
$$H_{TE\_PI\_H}(\omega)$$

$$\approx \frac{n'(1+s^2C_fL_s)}{R_S[2+(n'+2)s^2C_fL_s]},$$
(17)

| NOISE CONTRIBUTION IN POWER-TO-CURRENT AMPLIFIERS |                |        |              |        |  |  |  |  |  |
|---|----------------|--------|--------------|--------|--|--|--|--|--|
|   | α-PI @         | 2GHz   | TF-PI @      | 2GHz   |  |  |  |  |  |
|   | ENV (pV)       | NC (%) | ENV (pV)     | NC (%) |  |  |  |  |  |
| The first stage                                   | 129 (M1–M3)    | 67     | 116 (M1– M3) | 69.6   |  |  |  |  |  |
| The second stage                                  | 49 (M4–M7)     | 9.8    | 20 (M4, M5)  | 2      |  |  |  |  |  |
| I-I feedback                                      | 10 (α)         | 0.5    | 59 (TF)      | 18     |  |  |  |  |  |
| I-V feedback                                      | 44 ( $R_{f}$ ) | 1      | 44 $(R_f)$   | 1      |  |  |  |  |  |
| Total   | 158            | 100    | 139          | 100    |  |  |  |  |  |

TABLE III



Fig. 9. Simulated input impedance. (a)  $\alpha$ -PI. (b) TF-PI.



Fig. 10. Simulated voltage-to-current frequency response (magnitude only). (a)  $\alpha$ -PI. (b) TF-PI.

$$\frac{[\sqrt{[(c+g_{m1}R_f)(R_p^2+R_S^2)+g_{m1}R_pR_S^2]/(c+g_{m1}R_f)}/L_p, 1/\sqrt{(n'+1)L_sC_f}]}{\sqrt{(n'+1)L_sC_f}}$$
 or equally [0.8 GHz, 6 GHz], respectively.

#### D. Linearity and Stability

The linearity of a power-to-current amplifier is determined by the linearity of the output stage and by the amplifier's loop gain. The larger the loop gain, the higher the linearity. The calculated DC loop gain from (3) is

$$L(0) = -\frac{2g_{m1}g_{m2}R_sR_d}{(\alpha \text{ or } n')(1+g_{m1}R_f)} = -25,$$
(19)

which is sufficiently large and in accordance with our objective.

The stability factor of both PIs is larger than one. Their phase margin is around 90 degrees. A sharp step transient response also shows that there is no oscillation in the circuit. Therefore, both the  $\alpha$ -PI and TF-PI are unconditionally stable.

## VII. SIMULATION RESULTS

Both the  $\alpha$ -PI and TF-PI were simulated with ADS using 0.2  $\mu$ m GaAs p-HEMT technology process parameters and models [11]. The transformer was designed and modeled by Momentum. As shown in Fig. 11, the input return loss is well below -15 dB in the frequency range from 0.3 GHz to 4 GHz for the  $\alpha$ -PI and in the frequency range from 0.8 GHz to 4 GHz



Fig. 11. Simulated S11 as a function of frequency of both the  $\alpha$ -PI and TF-PI.



Fig. 12. Simulated noise figure as a function of frequency of both the  $\alpha$ -PI and TF-PI.



Fig. 13. Simulated S21 as a function of frequency of both the  $\alpha$ -PI and TF-PI.

for the TF-PI, validating the suitability of both power-to-current circuit topologies for wideband radio applications.

In the band of interest, the  $\alpha$ -PI achieves a 0.6–0.8 dB noise figure (Fig. 12). Its voltage-to-current gain is 200 mS (Fig. 10), which corresponds to a 23 dB power gain for a 50  $\Omega$  load (Fig. 13). The gain fluctuation is  $\pm 1$  dB. A two-tone test predicts that the IIP3 is -1 dBm (Fig. 14) and the IIP2 is 23 dBm at 2 GHz.

The TF-PI exhibits a 0.5–0.8 dB noise figure, a 200 mS voltage-to-current gain or  $22 \pm 1$  dB power gain from 0.8 GHz to 4 GHz, and a 0 dBm IIP3 and a 22 dBm IIP2 at 2 GHz. The minimum noise figure is obtained at 2.4 GHz.

The  $\alpha$ -PI and TF-PI draw 73 mA and 53 mA current from a 4 V power supply, respectively. It can be seen that the two



Fig. 14. Simulated IIP3 of both the  $\alpha$ -PI and TF-PI at 2 GHz.

power-to-current amplifiers exhibit almost comparable performance. Due to the fact that the manufacturer provides realistic element models, together with the measures to constrain the effects of the bond wires, we expect the simulation results well predict the actual performance of both power-to-current amplifiers when fabricated. Table IV summarizes the performance of various wideband LNAs for radio applications, where the figure of merit (FOM) is defined as [15]

$$\mathbf{FOM} = \frac{G(\mathbf{mag}) \times BW(\mathbf{GHz}) \times IIP3(\mathbf{mW})}{VSWR \times (F-1) \times P_{DC}(\mathbf{mW})}.$$
 (20)

The FOM evaluates the input return loss (by the voltage standing wave ratio, VSWR), maximum power gain (G), -3 dB bandwidth (BW), excess noise factor (F), linearity (IIP3) and the power consumption  $(P_{\rm DC})$  of the wideband LNA. It can be seen from the table that the FOMs of both the  $\alpha$ -PI and TF-PI are superior to those reported in other works except [20]. Table IV clearly states the advantages of the power-to-current amplifier for the design of broadband LNAs that should simultaneously attain good input power matching, low noise, high gain, and high linearity over a wide bandwidth.

#### VIII. CONCLUSION

Design strategies for power-to-current LNAs are presented that offer flat gain, accurate input impedance match, and minimum noise figure for broadband radio applications. The circuit topologies, transformer layout schemes, and bond wire effects are discussed. Both the  $\alpha$ -PI and TF-PI are designed to be implemented in a 0.2  $\mu$ m GaAs p-HEMT process and their performances are analyzed to verify the theory presented.

Operating across a frequency range from 0.3 GHz to 4 GHz, the  $\alpha$ -PI achieves 0.6–0.8 dB NF,  $23 \pm 1$  dB power gain, less than -15 dB input reflection coefficient, -1 dBm IIP3 and 23 dBm IIP2 at 2 GHz, and dissipates 292 mW power.

From 0.8 GHz to 4 GHz, the TF-PI shows 0.5–0.8 dB NF,  $22 \pm 1$  dB power gain, less than -15 dB input return loss, 0 dBm IIP3 and 22 dBm IIP2 at 2 GHz, and consumes 212 mW power.

| Literatures               | Technology              | Freq<br>[GHz] | NF<br>[dB] | Gain<br>[dB] | S11 [dB]<br>(VSWR) | IIP3<br>[dBm]    | P <sub>DC</sub><br>[mW] | FOM | Topology |
|---------------------------|-------------------------|---------------|------------|--------------|--------------------|------------------|-------------------------|-----|----------|
| [4]                       | $0.13 \ \mu$ m CMOS     | 2-5.2         | 4.7-5.7    | 16           | <-9 (2.1)          | -14 <sup>a</sup> | 38                      | 1   | RSF      |
| Our previous<br>work [10] | $0.2 \ \mu$ m GaAs      | 1-2           | 0.8        | 15           | <-12 (1.7)         | +1.8             | 210                     | 1   | TF-PI    |
| [16]                      | 0.18 $\mu$ m AlGaN/GaN  | 0.3-4         | 1.2        | 17.7         | <-10 (1.9)         | +10 <sup>b</sup> | 1000                    | 4   | RSF      |
| [17]                      | $0.18 \mu$ m CMOS       | 0.05-0.88     | 3          | 14           | <-9 (2.1)          | +3               | 34.8                    | 1   | CG       |
| [18]                      | $0.18 \ \mu$ m CMOS     | 0-1.2         | 2.1        | 16.4         | <-9 (2.1)          | 0                | 14.4                    | 3   | RSF      |
| [19]                      | $0.25 \mu$ m GaAs       | 1.5-6         | 1.5        | 17.5         | <-6 (3)            | -4               | 25.2                    | 3   | RSF      |
| [20]                      | $0.2 \ \mu$ m GaAs      | 2-10          | 2.2        | 11.7         | <-9.6 (2.0)        | +13 <sup>a</sup> | 94.5                    | 19  | RSF      |
| [21]                      | $0.2 \ \mu$ m GaAs      | 1-4           | 1.2-1.4    | 16.6         | <-12 (1.7)         | -11.4            | 84                      | 1   | RSF      |
| This work 1               | <b>0.2</b> $\mu$ m GaAs | 0.3-4         | 0.6-0.8    | 23           | <-15 (1.4)         | -1               | 292                     | 7   | α-PI     |
| This work 2               | 0.2 $\mu$ m GaAs        | 0.8-4         | 0.5-0.8    | 22           | <-15 (1.4)         | 0                | 212                     | 9   | TF-PI    |

TABLE IV Comparison With Previously Reported Broadband LNAs

<sup>a</sup>Estimated according to -1dB compression point <sup>b</sup>Estimated according to power consumption

The noise performance of the TF-PI at frequencies below 0.8 GHz is deteriorated by the nonidealities of the physical transformer. Therefore, the  $\alpha$ -PI is suitable for radio applications operating in the P-band, L-band, and S-band, such as the Square Kilometer Array, and the TF-PI is optimal for radio applications operating in the L-band and S-band, such as new receivers for global navigation satellite systems. The obvious advantage of these power-to-current amplifiers is that they accomplish good power matching, low noise, high gain, and high linearity in wideband applications simultaneously.

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