3–10-GHz Ultra-Wideband Low-Noise Amplifier Utilizing Miller Effect and Inductive Shunt–Shunt Feedback Technique

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Abstract—In this paper, we demonstrate an SiGe HBT ultra-wideband (UWB) low-noise amplifier (LNA), achieved by a newly proposed methodology, which takes advantage of the Miller effect for UWB input impedance matching and the inductive shunt–shunt feedback technique for bandwidth extension by pole-zero cancellation. The SiGe UWB LNA dissipates 25.8-mW power and achieves S_{11} below -10 dB for frequencies from 3 to 14 GHz (except for a small range from 10 to 11 GHz, which is below -9 dB), flat S_{21} of 24.6 ± 1.5 dB for frequencies from 3 to 11.6 GHz, noise figure of 2.5 and 5.8 dB at 3 and 10 GHz, respectively, and good phase linearity property (group-delay variation is only ± 28 ps across the entire band). The measured 1-dB compression point ($P_{1\text{ dB}}$) and input third-order intermodulation point are -25.5 and -17 dBm, respectively, at 5.4 GHz.

Index Terms—HBT, inductive feedback, low-noise amplifier (LNA), Miller effect, SiGe, ultra-wideband (UWB).

I. INTRODUCTION

SINCE THE U.S. Federal Communications Commission (FCC) approved the application of ultra-wideband (UWB) technology for commercial use in early 2002, ranging from 3.1 to 10.6 GHz (bandwidth: 7.5 GHz), the implementation of a UWB system and its sub-blocks has been a challenge for integrated circuit (IC) designers. In UWB receiver design, the UWB low-noise amplifier (LNA) is a critical block of the receiver front-end. Different from that operating in a narrowband communication system, a UWB LNA has to receive small signals from 3.1 to 10.6 GHz and amplify them with a good signal-to-noise ratio property over the entire UWB band. In addition, flat and high power gain (S_{21}), input impedance matching (i.e., low input-return loss S_{11}), and good noise figure (NF) performances across the entire UWB band are also required. Recently, several UWB LNAs have been reported [1]–[11]. For some UWB LNAs [1], [2], which are designed for an orthogonal frequency division multiplexing (OFDM)

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Fig. 1. RLC series network.

system, linearity is a tight requirement for suppressing adjacent channel interferences. On the other hand, for UWB pulse radio systems [3], [4], the amplitude linearity requirements are relaxed; instead, in order to keep the shape of the pulse while receiving radio signals from an antenna, good phase linearity is required.

In this paper, a new methodology to implement a UWB LNA for UWB pulse radio system applications is demonstrated by 0.35- μm SiGe HBT technology. The proposed methodology takes advantage of the Miller effect for UWB input impedance matching and the inductive shunt–shunt feedback technique for bandwidth extension by pole and zero cancellation. Although inductive shunt–shunt feedback topology has been seen [12]–[14] before, their major intention is for transistor neutralization or peaking rather than circuit pole-zero cancellation, which will be detailed later in this paper. This paper is organized as follows. Section II describes the circuit design principles of UWB input impedance matching and UWB S_{21} . Section III introduces the proposed UWB LNA with four cascaded stages. Transfer function of S_{21} of this circuit is derived for the purpose of illustrating the proposed methodology. In addition, theoretical analysis of NF and linearity is also included. Finally, the experimental results of the SiGe UWB LNA are discussed in Section IV, which is followed by conclusions in Section V.

II. PRINCIPLES OF CIRCUIT DESIGN

A. UWB Input Impedance Matching

To discuss UWB input impedance matching, consider the frequency response of S_{11} of an RLC series network shown in Fig. 1 first. The input impedance Z_{IN} of it is given by

$$Z_{\text{IN}} = sL_{\text{IN}} + R_{\text{IN}} + \frac{1}{sC_{\text{IN}}}. \quad (1)$$

According to (1), S_{11} can be expressed as follows:

$$\begin{aligned}
 S_{11} &= \frac{Z_{IN} - R_S}{Z_{IN} + R_S} \\
 &\approx \frac{s^2 + s \frac{R_{IN} - R_s}{L_{IN}} + \frac{1}{C_{IN} L_{IN}}}{s^2 + s \frac{R_{IN} + R_s}{L_{IN}} + \frac{1}{C_{IN} L_{IN}}} \\
 &\equiv \frac{s^2 + \omega_O^2}{s^2 + s \frac{\omega_O}{Q_{IN}} + \omega_O^2} \left(\text{or } \frac{s^2 + \omega_O^2}{s^2 + sB + \omega_O^2} \right) \quad (2)
 \end{aligned}$$

in which ω_o is called the pole frequency (or center frequency) and $Q_{IN} = \omega_O L_{IN} / (R_{IN} + R_s)$ is called the pole Q factor [15]. Now it is clear that $|S_{11}|$ is a standard notch function with -3 -dB bandwidth $\Delta\omega_{3\text{dB}}$ equal to $B = \omega_o / Q_{IN}$ [16]. However, in RF design, $|S_{11}|$ should be smaller than -10 dB over the band of interest, i.e.,

$$20 \log |S_{11}| \leq -10 \text{ dB}. \quad (3)$$

By substituting the expression of S_{11} in (2) into (3) and after some calculations, it was found that the input impedance matching bandwidth $\Delta f_{10\text{dB}}$ is equal to

$$\Delta f_{10\text{dB}} = \frac{\Delta\omega_{10\text{dB}}}{2\pi} = \frac{B}{6\pi} = \frac{\omega_O}{6\pi Q_{IN}} \approx \frac{100}{6\pi L_{IN}}. \quad (4)$$

That is, the input matching bandwidth $\Delta f_{10\text{dB}}$ is inversely proportional to the value of L_{IN} (or inversely proportional to the value of Q_{IN} if the value of the center frequency ω_o is fixed). From (4), for a 3.1–10.6-GHz UWB LNA, the inductance L_{IN} should not be larger than 0.707 nH (or $Q_{IN} < 0.25$). In the traditional LNA design using series-series inductive feedback, as shown in Fig. 2(a), the input resistance R_{IN} (usually 50Ω) is generated by the emitter (or a source) degenerative inductor L_S [17], and the base inductor L_{GD} and the emitter inductor L_S ($L_{GD} + L_S = L_{IN}$) are used to resonate in the center frequency (f_o), as shown in Fig. 2(a). Recently, several circuit topologies have been developed for UWB input impedance matching based on the LC bandpass filter theory [1] [2], in which the necessary 50Ω is still mainly generated by the emitter (or source) inductor.

The capacitive feedback technique utilizing Miller effect has been used to achieve UWB input impedance matching [18], [19] instead of generating the necessary 50Ω (R_{IN}) by an emitter (or a source) degenerative inductor. For the common-emitter amplifier with shunt-shunt capacitive feedback, shown in Fig. 2(b), suppose $(1 + G_{M1} R_L) \gg |s R_L C_T|$ over the 3.1–10.6-GHz band of interest, which is usually the case. The input impedance Z_{IN} of it is then given by

$$Z_{IN} = s L_{GM} + R_{IN} + \frac{1}{s C_{IN}} \quad (5a)$$

Fig. 2. Input impedance matching achieved by the: (a) traditional series-series inductive feedback and (b) shunt-shunt capacitive feedback.

in which

$$C_{IN} = C_{BE1} + C_M = C_{BE1} + C_{bcl}(1 + G_{M1} R_L) \quad (5b)$$

$$R_{IN} = R_M \left(\frac{C_M}{C_{BE1} + C_M} \right)^2 \approx 50 \Omega \quad (5c)$$

$$R_M = \frac{R_L}{1 + G_{M1} R_L} \cdot \left(1 + \frac{C_T}{C_{bcl}} \right) \quad (5d)$$

$$R_L = R_1 \parallel R_T. \quad (5e)$$

In the above equations, G_{M1} is the transconductance and C_{BE1} is the base-emitter capacitance of transistor Q_1 . C_{bcl} is the equivalent base-collector capacitance, which can include the base-collector capacitance (C_{BC1}) and the intentionally added capacitance (C_F), if necessary, between the base and collector of transistor Q_1 . The parallel $R_T - C_T$ load represents the input impedance of the following stages of the circuit, which provides an appropriate load for generating the necessary 50Ω . Compared with the traditional series-series inductive input matching technique (two inductors), the shunt-shunt capacitive input matching method (one inductor) is more area efficient because no area consuming inductor (L_S) is needed to generate 50Ω .

B. UWB Gain Flattening Technique

As shown in Fig. 3, resistive shunt-shunt feedback is traditionally adopted for extending the bandwidth of an amplifier. However, the corresponding gain performance is usually not

Fig. 3. Traditional resistive shunt-shunt feedback amplifier and its equivalent circuit. The loading effect of the previous stage (R_1) and the output terminal (R_{L1}) is also shown.

Fig. 4. Multiband LNA consists of a shunt-shunt capacitive feedback stage followed by a gain stage.

high enough for UWB application due to the tradeoff between gain and bandwidth. The transimpedance gain Z_{RFB} and the pole P_1 arising from $C_{\pi 2}$ of the resistive shunt-shunt feedback amplifier can be expressed as follows:

$$Z_{RFB} = -\frac{R_{IN2} \parallel R_1}{1 + \frac{1}{P_1}} \times \left(G_{M2} - \frac{1}{R_F} \right) (R_F \parallel R_L) \quad (6)$$

$$P_1 = \frac{1}{C_{\pi 2} [R_1 \parallel R_{IN2}]} = \frac{1}{C_{\pi 2} \times \left[R_1 \parallel \left(\frac{R_F + R_L}{1 + G_{M2} R_L} \right) \right]} \quad (7)$$

where G_{M2} is the transconductance of transistor Q_2 , $\parallel$ denotes parallel connection, $R_L = R_3 \parallel R_{L1}$, and $R_{IN2} = (R_F + R_L)/(1 + G_{M2} R_L)$ is the input resistance of the amplifier without considering $C_{\pi 2}$. Note that the loading effect of the previous stage (i.e., R_1 in Fig. 3) and the output terminal (i.e., R_{L1} in Fig. 3) is included so that the result can be applied directly to the analysis of S_{21} of the amplifier shown in Fig. 4. From (6) and (7), we can see that decreasing R_F widens the bandwidth of the amplifier, but at the same time, the gain is degraded.

Fig. 5. Calculated characteristics of $|G_{M11}|$, $|Z_{RFB}|$, and $|S_{21}|$ versus frequency of a SiGe multiband LNA.

Although this circuit topology in conjunction with the above-mentioned capacitive feedback technique is appropriate as a multiband LNA (see Fig. 4) for a multiband or multistandard application [18], [19], it is difficult to achieve high and flat UWB gain at the same time, which is explained as follows. For the multiband LNA shown in Fig. 4, the transfer function of S_{21} , which is equal to two times of the voltage gain ($A_V = v_{out}/v_s$), can be derived as follows:

$$\begin{aligned} S_{21} &= 2 \frac{v_{out}}{v_s} \\ &\approx 2 \frac{1}{sC_{IN}} \frac{1}{R_S + sL_{GM} + \frac{1}{sC_{IN}} + R_{IN}} \\ &\quad \cdot (G_{M1} - sC_{bc1}) \cdot \left(R_1 \parallel R_{IN2} \parallel \frac{1}{s(C_{\pi 2} + C_{bc1})} \right) \\ &\quad \cdot \left(G_{M2} - \frac{1}{R_F} \right) \cdot (R_F \parallel R_L) \\ &\equiv \frac{2 \cdot G_{M1}}{s^2 L_{GM} C_{IN} + 1 + sC_{IN}(R_{IN} + R_S)} \cdot Z_{RFB} \\ &\equiv G_{M11} \cdot Z_{RFB} \end{aligned} \quad (8)$$

in which G_{M11} is proportional to $1/(s^2 L_{GM} C_{IN} + 1 + sC_{IN}(R_{IN} + R_S))$, which is a transfer function of a second-order low- Q low-pass RLC circuit. For example, $Q_{IN} = \sqrt{L_{GM}/C_{IN}}/(R_{IN} + R_S) \approx 0.22$, i.e., over-damped response, if $L_{GM} = 0.6$ nH and the center frequency $f_o = 5.7$ GHz over the band of interest are specified. Besides, the zero ($= G_{M1}/(2\pi C_{bc1}) \approx 122.5$ GHz) contributed by the capacitive-feedback first-stage is much higher than the 3.1–10.6-GHz band of interest and, hence, is negligible.

Fig. 5 shows the characteristic of calculated $|G_{M11}|$, $|Z_{RFB}|$ and $|S_{21}|$ versus frequency of a typical SiGe multiband LNA, as published in [19]. As can be seen, over the 3.1–10.6-GHz band of interest, the characteristic of $|Z_{RFB}|$ versus frequency exhibits a negative slope close to 20 dB/decade because its pole is approximately 1.2 GHz. In addition, the characteristic of $|G_{M11}|$ versus frequency also exhibits a negative slope close to 20 dB/decade, which can be explained as follows: since

Fig. 6. (a) Relative position of the zeros (Z_1 and Z_2) and the poles (P_1 , P_3 , P_4 , and P_{IN}). (b) Corresponding frequency response of the UWB LNA proposed in this paper.

the input RLC circuit is a low- Q circuit and f_o is specified as 5.7 GHz, $s^2 L_{GM} C_{IN} + 1 \approx 0$ holds over the band of interest, i.e., $s^2 L_{GM} C_{IN} + 1 + s C_{IN}(R_{IN} + R_s) \approx s C_{IN}(R_{IN} + R_s)$ can be assumed, which, in turn, results in

$$G_{M11} \equiv \frac{2 \cdot G_{M1}}{s^2 L_{GM} C_{IN} + 1 + s C_{IN}(R_{IN} + R_s)} \approx \frac{2 \cdot G_{M1}}{s C_{IN}(R_{IN} + R_s)}. \quad (9)$$

This explains why the characteristic of $|G_{M11}|$ and $|S_{21}|$ versus frequency in Fig. 5 exhibits a negative slope close to 20 and 40 dB/decade, respectively, over the 3.1–10.6-GHz band of interest.

To flatten the S_{21} performance of the LNA in Fig. 4 for UWB applications, the poles contributed by G_{M11} [see $P_{IN} \sim 0$ in Fig. 6(a) and (b)] and Z_{RFB} [see P_1 in Fig. 6(a) and (b)], respectively, should be cancelled by introducing an additional two zeros [see Z_2 and Z_1 in Fig. 6(a) and (b)], which are located exactly at the same frequencies of the two poles.

This can be achieved by taking the following two actions. First, adding an inductive shunt–shunt feedback stage (see Fig. 7, in which L_1 is the feedback inductance and R_4 is the parasitic resistance of L_1 or the intentionally added small resistance) after the Q_2 gain stage of the LNA. A newly generated zero (Z_1) due to the inductive shunt–shunt feedback stage is specified to be equal to the pole (P_1) of Z_{RFB} to achieve pole-zero cancellation.

Second, adding a stage with an inductive load [see L_2 in Fig. 8(a)] to the end of the LNA to generate a zero (Z_2) at dc frequency so that the pole contributed by G_{M11} ($P_{IN} \sim 0$) can

Fig. 7. Proposed inductive shunt–shunt feedback amplifier and its equivalent circuit. The loading effect of the previous stage (R_3) and the next stage ($C_{\pi 4}$) is also considered.

Fig. 8. (a) Schematic of the proposed UWB LNA. (b) Corresponding equivalent circuit for calculating the input impedances seen at the input of the second, third, and last stages.

be cancelled. Consequently, because frequencies of the complex conjugate poles (P_3 and P_4 , which will be discussed in detail later) arising from the inductive shunt–shunt feedback stage are higher than that of P_1 originated from the Q_2 gain stage, the bandwidth of the LNA is extended.

In the following, we analyze the proposed inductive shunt–shunt feedback amplifier shown in Fig. 7. The transimpedance gain Z_{LFB} and input impedance Z_{IN3} can be derived as (10a), shown at the bottom of this page, and (10b),

$$Z_{LFB} = -(R_3 \parallel Z_{IN3}) \left(G_{M3} - \frac{1}{sL_1 + R_4} \right) \left((R_4 + sL_1) \parallel R_5 \parallel \frac{1}{sC_{\pi 4}} \right) \approx \frac{R_3 G_{M3} (R_4 + sL_1) R_5}{R_3 [(1 + sC_{\pi 4} R_5) + G_{M3} R_5] + [(1 + sC_{\pi 4} R_5)(R_4 + sL_1) + R_5] + sC_{\pi 3} R_3 [(1 + sC_{\pi 4} R_5)(R_4 + sL_1) + R_5]} \quad (10a)$$

respectively, where G_{M3} is the transconductance of transistor Q_3 ,

$$Z_{IN3} = \frac{(sL_1 + R_4) + \left(R_5 \parallel \frac{1}{sC_{\pi4}}\right)}{1 + G_{M3} \cdot \left(R_5 \parallel \frac{1}{sC_{\pi4}}\right)} \parallel \frac{1}{sC_{\pi3}}. \quad (10b)$$

Note that in (10a), $G_{M3}R_4 \gg 1$ is assumed. In addition, the loading effect of the previous stage (i.e., R_3 in Fig. 7) and the next stage (i.e., $C_{\pi4}$ in Fig. 7) is included so that the result can be applied directly to the analysis of the S_{21} of the proposed UWB LNA in Section III. Clearly, (10a) has one zero (Z_1) given by

$$Z_1 = \frac{R_4}{L_1}. \quad (11)$$

Note that all the poles and zeros in this paper are located at the left half-plane (LHP). However, we omit the minus sign of them for the purpose of simplification. As mentioned above, Z_1 is set to be equal to P_1 in (7) to achieve pole-zero cancellation. Thus, the $|S_{21}|$ performance of the cascaded amplifier, which consists of a gain stage followed by an inductive shunt–shunt feedback stage, is flat for frequencies up to some pole frequency of the inductive shunt–shunt feedback stage, if a maximally flat response is chosen.

In the following, we discuss the poles of (10a). The characteristic equation of (10a) can be written as follows:

$$\begin{aligned} & s^3 + s^2 \left(\omega_{P2} + \omega_{P3} + \frac{R_4}{L_1} \right) \\ & + s \left(\frac{(R_4 + R_5)\omega_{P2}}{L_1} + \omega_{P2}\omega_{P3} + \frac{(R_3 + R_4)\omega_{P3}}{L_1} \right) \\ & + \left[\frac{\omega_{P2}\omega_{P3}}{L_1} (R_3 + R_4 + R_5(1 + G_{M3}R_3)) \right] \\ & \equiv (s + P_2) \cdot (s + P_3) \cdot (s + P_4) \\ & = 0 \end{aligned} \quad (12)$$

in which $\omega_{P2} \equiv 1/(C_{\pi4}R_5)$ and $\omega_{P3} \equiv 1/(C_{\pi3}R_3)$ are the higher and lower frequency poles, respectively, of the open-loop amplifier, i.e., $L_1 \rightarrow \infty$. This is because all the terms on the left side of (12), which include L_1 , are equal to zero if $L_1 \rightarrow \infty$. P_2 and P_3 are the corresponding poles of ω_{P2} and ω_{P3} , respectively, if the effect of L_1 is considered. P_4 is a newly generated pole if the effect of L_1 is considered. Note that P_4 will be very close to the origin if L_1 is very large. From the coefficients of the second-order and constant terms of (12), we can obtain the following equations:

$$\omega_{P2} + \omega_{P3} + Z_1 = P_2 + P_3 + P_4 \quad (13)$$

$$\frac{\omega_{P2}\omega_{P3}}{L_1} (R_3 + R_4 + R_5(1 + G_{M3}R_3)) = P_2 \cdot P_3 \cdot P_4. \quad (14)$$

Fig. 9. Calculated root-locus diagram of the inductive shunt–shunt feedback amplifier in Fig. 8 by (10a) using the circuit parameters from the SiGe HBT UWB LNA.

Equation (13) implies that if a very large L_1 is added to the open loop amplifier, both Z_1 and P_4 will be created simultaneously around the origin since $P_2 \approx \omega_{P2}$, $P_3 \approx \omega_{P3}$, and $P_4 \approx 0$. This is consistent with the calculated root-locus diagram shown in Fig. 9. Furthermore, since P_2 and P_3 are very close to ω_{P2} and ω_{P3} , respectively, for a very large L_1 , P_4 can be approximated by the following equation based on (14):

$$P_4 = \frac{R_3 + R_4 + R_5(1 + G_{M3}R_3)}{L_1}. \quad (15)$$

Equation (15) predicts that the movement of P_4 to the left will be quicker than that of Z_1 to the left as the value of L_1 is reduced, which is indeed the case, as shown in Fig. 9. As indicated, decreasing L_1 from infinity moves the highest frequency pole (P_2) outward, while the other poles (P_3 and P_4) become coincident and then become complex and conjugate. The inductive shunt–shunt feedback amplifier used in this paper belongs to the latter situation (i.e., P_3 and P_4 are complex and conjugate) because a small value of L_1 (0.6 nH) is adopted.

It is interesting to note that if the value of R_4 is negligible and the value of L_1 is 0, corresponding to the case of shortening transistor base and collector terminals, then (12) can be simplified as follows:

$$s = - \frac{1}{(C_{\pi3} + C_{\pi4}) \left(R_3 \parallel R_5 \parallel \frac{1}{G_{M3}} \right)} \quad (16)$$

which is consistent with the result obtained by direct circuit analysis and also explains why P_2 approaches to a limiting frequency, as is evident in Fig. 9.

III. PROPOSED UWB LNA

A. Circuit Topology

Fig. 8(a) shows the proposed UWB LNA with four cascaded stages, including a common-emitter amplifier with Miller capacitance feedback for wideband input impedance matching,

a common-emitter gain stage, an inductive shunt–shunt feedback amplifier for bandwidth extension by pole-zero cancellation, and a buffer amplifier with an inductive load L_2 . Large resistances R_F and R_{F1} were used for self-biasing and C_1 , C_2 , and C_3 were used as dc blocking capacitors.

The equivalent loading impedance ($Z_{IN2} \parallel R_1$) of the first stage provides an appropriate load for generating the necessary 50Ω by the beneficial Miller capacitance $C_{bc1} \cdot (1 + G_{M1}R_L)$. Apart from the purpose of generating 50Ω , the Miller capacitance is also used to enlarge the input capacitance C_{IN} of the input low- Q RLC circuit [see (5b)], which, in turn, results in a larger input matching bandwidth. Fig. 8(b) shows the equivalent circuit of the second to the fourth stage of the proposed UWB LNA for calculating the Z_{IN4} , Z_{IN3} , and Z_{IN2} , which can be expressed, respectively, as follows:

$$Z_{IN4} = \frac{1}{sC_{\pi4}} \quad (17a)$$

$$Z_{IN3} = \frac{sL_1 + R_4 + (R_5 \parallel Z_{IN4})}{1 + G_{M3} \cdot (R_5 \parallel Z_{IN4})} \parallel \frac{1}{sC_{\pi3}} \\ = \frac{sL_1 + R_4 + \left(R_5 \parallel \frac{1}{sC_{\pi4}} \right)}{1 + G_{M3} \cdot \left(R_5 \parallel \frac{1}{sC_{\pi4}} \right)} \parallel \frac{1}{sC_{\pi3}} \quad (17b)$$

$$Z_{IN2} = \frac{1}{sC_{\pi2}}. \quad (17c)$$

In deriving the above equations, the two large self-biased resistors R_F and R_{F1} have been neglected. S_{21} of the four-stage cascaded UWB LNA, which is equal to two times of the voltage gain ($A_V = v_{out}/v_s$), can be expressed as (18a) and (18b), shown at the bottom of this page, where $R_{IN}(= R_S = R_L = 50 \Omega)$ and C_{IN} are the equivalent input circuit elements seen at the base of Q_1 with the capacitive feedback input matching method and

$$Z_{L1} = Z_{IN2} \parallel R_1 \parallel (1/(sC_{bc1})) \approx (1/sC_{\pi2}) \parallel R_1 \quad (19a)$$

$$Z_{L2} = Z_{IN3} \parallel R_3 \quad (19b)$$

$$Z_{L3} = Z_{IN4} \parallel R_5 \parallel (sL_1 + R_4) \quad (19c)$$

are the loading impedances, with the input impedance of the next stage being considered, of the first, second, and third stages, respectively.

Equation (18b) can be further discussed as follows. First, in order to achieve UWB input impedance matching, the series $L_{GM} - C_{IN} - R_{IN}$ input equivalent circuit of the UWB LNA is designed as a low- Q circuit, which is different from that (high- Q $L-C-R$ input equivalent circuit) used in the narrowband LNAs [17]. Since the input RLC circuit is a low- Q circuit and its center frequency f_o can be specified as the geometrical mean (5.7 GHz) of the upper (10.6 GHz) and lower (3.1 GHz) corner frequency, $s^2 L_{GM} C_{IN} + 1 + s C_{IN}(R_{IN} + R_s) \approx s C_{IN}(R_{IN} + R_s)$ holds due to $s^2 L_{GM} C_{IN} + 1 \approx 0$. This means that the pole ($P_{IN} \sim 0$) of $s^2 L_{GM} C_{IN} + 1 + s C_{IN}(R_{IN} + R_s) \approx s C_{IN}(R_{IN} + R_s)$ can be cancelled by the zero (Z_2) contributed by the impedance (sL_2) of L_2 . Moreover, the zero (Z_1) of Z_{LFB} is set to be equal to the pole (P_1) of Z_{RFB} , i.e., $L_1 \approx R_4 R_1 C_{\pi2}$, to achieve pole-zero cancellation. If the pole originates from the addition of L_2 [i.e., $1/(R_L + sL_2)$ in (18b)] is negligible (i.e., larger than 10.6 GHz), then the maximally flat response condition for P_3 and P_4 should be chosen. If the pole originating from the addition of L_2 cannot be neglected (i.e., smaller than 10.6 GHz), then a small peaking can be chosen to compensate the frequency dependency so that a flat response of S_{21} can still be achieved. It is interesting to note that the circuit architecture of the third stage of the UWB LNA, which is inductive feedback, is similar to that of a Colpitts oscillator [20] if the small feedback resistance R_4 is neglected. That means, in general, an inductive-feedback peaking extends the bandwidth of an amplifier at the expense of stability. Fig. 6(a) shows the relative position of the zeros (Z_1 and Z_2) and poles (P_1 , P_3 , P_4 , and P_{IN}) according to the proposed methodology. The corresponding frequency response of the UWB LNA is illustrated in Fig. 6(b).

An SiGe HBT UWB LNA was designed with a standard $0.35\text{-}\mu\text{m}$ SiGe HBT process based on the above-mentioned circuit design principle. The component parameters are as follows: $L_{GM} = 0.6 \text{ nH}$, $L_1 = 0.6 \text{ nH}$, $L_2 = 2.46 \text{ nH}$, $R_1 = 300 \Omega$, $R_F = 2010 \Omega$, $R_3 = 125 \Omega$, $R_4 = 3 \Omega$, $R_5 = 101 \Omega$, $R_{F1} = 1000 \Omega$, $C_1 = 3 \text{ pF}$, $C_2 = 3 \text{ pF}$, $C_3 = 3 \text{ pF}$, and $C_F = 18 \text{ fF}$. All transistors (Q_1 , Q_2 , Q_3 , and Q_4) have the

$$S_{21} = 2A_V \\ = 2 \frac{v_{out}}{v_s} \\ \approx 2 \frac{\frac{1}{sC_{IN}}}{R_S + sL_{GM} + \frac{1}{sC_{IN}} + R_{IN}} \cdot G_{M1} \cdot Z_{L1} \cdot G_{M2} \cdot Z_{L2} \cdot \left(G_{M3} - \frac{1}{sL_1 + R_4} \right) \cdot Z_{L3} \cdot G_{M4} \cdot \frac{R_L sL_2}{R_L + sL_2} \quad (18a) \\ \approx 2 \frac{1}{s^2 L_{GM} C_{IN} + 1 + s(R_S + R_{IN})C_{IN}} \cdot G_{M1} \cdot \frac{R_1}{1 + sR_1 C_{\pi2}} \cdot G_{M2} \cdot G_{M4} \cdot \frac{R_L \cdot sL_2}{R_L + sL_2} \\ \cdot \frac{R_3 G_{M3} (R_4 + sL_1) R_5}{R_3 [(1 + sC_{\pi4} R_5) + G_{M3} R_5] + [(1 + sC_{\pi4} R_5)(R_4 + sL_1) + R_5] + sC_{\pi3} R_3 [(1 + sC_{\pi4} R_5)(R_4 + sL_1) + R_5]} \quad (18b)$$

Fig. 10. Die photograph of the SiGe HBT UWB LNA.

Fig. 11. Complete equivalent circuit of Fig. 2(b) for noise calculation.

same emitter size of $0.3 \mu\text{m} \times 20.3 \mu\text{m} \times 1$ emitter finger. The die photograph of the finished circuit is shown in Fig. 10. The chip area is only $570 \mu\text{m} \times 390 \mu\text{m}$ excluding the test pads.

B. Noise Analysis

According to the Friis equation [21], the total NF is dominated by the NF of the first stage if the gain of the first stage is high enough. Fig. 11 shows the equivalent circuit of the first stage of the proposed UWB LNA for first-order approximation of noise calculation. Based on our previous study [17], the NF of the first stage can be given by (20), shown at the bottom of this page, where R_S is the signal source resistance (usually 50Ω), R_{E1}/R_{B1} is the parasitic emitter/base resistance of transistor Q_1 , R_{LGM} is the series resistance of the inductor L_{GM} , k is the Boltzmann constant, T is absolute temperature, f is frequency, i_{nB1} is the base noise current, and i_{nC1} is the collector noise current.

Note that the capacitance term $(C_{BE1} + C_{bc1})$ of (20) is different from that $(C_{BE1} + C_{bc1} \cdot (1 + G_{M1}R_L))$ of S_{11} [see (2) and (5b)] unless C_{bc1} is small, assuming $C_{bc1} = 0$ and $R_S + R_{LGM} + R_{B1} + R_{E1} \approx R_S$. To gain more insight into

(20), let us temporarily neglect the noise contribution from i_{nB1} and, hence, (20) can be rewritten as

$$\text{NF} \approx 1 + \frac{R_{LGM} + R_{B1} + R_{E1}}{R_S} + \frac{1}{2G_{M1}R_S} \cdot \left| \left(\frac{s}{\omega_{onf}} \right)^2 + s \frac{(R_S + R_{PAR})}{Q_{IN} \cdot \omega_{onf} \cdot (R_S + R_{IN} + R_{PAR})} + 1 \right|^2 \quad (21)$$

where $\omega_{onf} = 1/\sqrt{L_{GM}C_{BE1}}$. We can see that the numerator of the third term of (21) is similar to the square of the first denominator of the S_{21} expression in (18b) [or S_{11} expression in (2)]. From this, we conclude that the frequency response of the NF, in general, follows the inverse of the gain response of the input RLC circuit. That is, a low- Q circuit (over-damped) wideband input RLC circuit will yield a wideband NF response such as those in [1] and [18], while a high- Q (under-damped) narrowband input RLC circuit will yield a narrowband NF response such as the LNA in [22]. This general trend is true even if C_{bc1} is taken into account.

For a low- Q input RLC circuit, $s^2 L_{GM} C_{IN} \approx 0$ holds over the band of interest and, hence, (20) can be approximated without considering i_{nB1} by

$$\begin{aligned} \text{NF} &\approx 1 + \frac{R_{LGM} + R_{B1} + R_{E1}}{R_S} \\ &\quad + \frac{1}{2G_{M1}R_S} \cdot |1 + j\omega(C_{BE1} + C_{bc1}) \cdot R_S \\ &\quad - \omega^2(C_{BE1} + C_{bc1}) \cdot L_{GM}|^2 \\ &\approx 1 + \frac{R_{LGM} + R_{B1} + R_{E1}}{R_S} \\ &\quad + \frac{R_S}{2G_{M1}} \cdot (C_{BE1} + C_{bc1})^2 \cdot \omega^2. \end{aligned} \quad (22)$$

From (22), we can see that the NF increases with increasing $C_{BE1} + C_{bc1}$; i.e., in order to reduce the NF, it is better to reduce $C_{BE1} + C_{bc1}$. However, this is in conflict with the purpose of widening the input matching band, which increases with increasing C_{BE1} and/or C_{bc1} . Hence, there is a tradeoff between the input impedance-matching bandwidth and the NF. We can call this phenomenon matching-bandwidth-NF limitation. Fortunately, because of the Miller effect, the needed C_F is normally much smaller than $C_{BE1} + C_{bc1}$. This means the degradation in the NF due to the intentionally added C_F is normally not significant. Since the NF in (22) is inversely proportional to G_{M1} (or I_{C1}), power consumption can be increased, if necessary, to remedy the degradation of the NF. Note that, in fact, the total

$$\begin{aligned} \text{NF} &\approx 1 + \frac{R_{LGM} + R_{B1} + R_{E1}}{R_S} + |R_S + R_{LGM} + R_{B1} + R_{E1} + j\omega \cdot L_{GM}|^2 \cdot \frac{|i_{nB1}^2|}{4kTR_S\Delta f} \\ &\quad + \left| \frac{1 + j\omega(C_{BE1} + C_{bc1}) \cdot (R_S + R_{LGM} + R_{B1} + R_{E1} + j\omega \cdot L_{GM})}{G_{M1}} \right|^2 \cdot \frac{|i_{nC1}^2|}{4kTR_S\Delta f} \end{aligned} \quad (20)$$

NF also includes the noise contribution from i_{nB1} , which decreases with increasing $C_{BE1} + C_{bc1}$. Therefore, in real cases, the degradation of the NF due to the increase of $C_{BE1} + C_{bc1}$ is offset by the decrease of i_{nB1} .

The noise contributions from both i_{nB1} and i_{nC1} are now taken into account. Following the power-constrained noise-optimization method proposed by Shaeffer and Lee [23] and assuming $R_S + R_{LGM} + R_{B1} + R_{E1} \approx R_S$, NF (20) near resonance (i.e., $\omega_o^2 L_{GM} C_{IN} \approx 1$) can be rewritten as

$$\begin{aligned} NF \approx 1 + & \frac{R_{LGM} + R_{B1} + R_{E1}}{R_S} \\ & + \left(1 + \frac{Q_S^2}{(1 + \chi(1 + |A_{V1}|))^2} \right) \frac{G_{M1} R_S}{2\beta_{dc}} \\ & + \left\{ \left(\frac{\chi |A_{V1}|}{1 + \chi(1 + |A_{V1}|)} \right)^2 + \frac{(1 + \chi)^2}{Q_S^2} \right\} \frac{1}{2G_{M1} R_S} \end{aligned} \quad (23)$$

where A_{v1} is the voltage gain of the first stage, and Q_S and χ are defined as follows, respectively:

$$Q_S = \frac{1}{\omega_o C_{BE1}(R_S + R_{PAR})} = \frac{C_{IN}(R_S + R_{IN} + R_{PAR})}{C_{BE1}(R_S + R_{PAR})} Q_{IN} \quad (24)$$

$$\chi = \frac{C_{bc1}}{C_{BE1}}. \quad (25)$$

The typical value of χ is around 0.1–0.2. It follows that (23) includes terms that are proportional to Q_S^2 , as well as to $1/Q_S^2$. Therefore, a minimum NF exists for a particular Q_S (or device size) given by

$$Q_{S,opt,I_{C1}} = \sqrt[4]{\beta_{dc} \frac{\sqrt{(1 + \chi)(1 + \chi(1 + A_{V1}))}}{\sqrt{G_{M1} R_S}}}. \quad (26)$$

Similarly, a minimum NF also exists for a particular G_{M1} (or I_{C1}) representing power consumption because the NF includes terms that are proportional to G_{M1} (or I_{C1}), as well as to $1/G_{M1}$ (or $1/I_{C1}$). The optimum G_{M1} can be derived as

$$G_{M1,opt} = \frac{\sqrt{\beta_{dc}}}{R_S} \frac{\sqrt{\left(\frac{\chi A_{V1}}{1 + \chi(1 + A_{V1})} \right)^2 + \frac{(1 + \chi)^2}{Q_S^2}}}{\sqrt{1 + \frac{Q_S^2}{(1 + \chi(1 + A_{V1}))^2}}}. \quad (27)$$

The characteristics of the NF versus Q_{IN} using the power consumption (I_{C1} or G_{M1}) as a parameter with a typical dc gain of 20 are plotted in Fig. 12. Also remember that the requirement of wideband impedance matching sets an upper bound of $Q_{IN} = 0.25$ (see (4), $L_{GM} < 0.707$ nH for 3.1–10.6-GHz input matching). Note that a G_{M1} greater than $G_{M1,opt} \approx 0.22$ degrades the NF at $Q_{IN} = 0.25$ because the noise contribution from I_{nB1} becomes dominant beyond $G_{M1,opt}$.

Fig. 12. Characteristics of NF versus Q_{IN} using the G_{M1} as a parameter.

In order to obtain the minimum NF, one should choose $G_{M1,opt} \approx 0.22$. However, as is evident in Fig. 12, the R_{IN} achieved by setting bias at $G_{M1,opt}$ is usually too small. R_{IN} generated by feedback capacitance [see (5c)] must be close to 50 Ω and can be approximated by

$$G_{M1,50} \approx \frac{1}{50} \left(1 + \frac{1}{\chi} \right). \quad (28)$$

Since χ is around 0.1–0.2, G_{M1} is near 0.12–0.22. In this design, $G_{M1} \approx 0.1$ was chosen for low power consumption. Once G_{M1} is given, the optimum Q_{IN} can be determined. Nevertheless, as mentioned previously, Q_{IN} cannot be greater than 0.25 for wideband input matching.

From Fig. 12, it is clear that the NF is insensitive to Q_{IN} from 0.15 to 0.7 and, hence, Q_{IN} of 0.22 was selected because of the available standard inductors provided by the foundry. We thus conclude that the optimum NF, Q_{IN} (or device size), gain, and power consumption can be determined from Fig. 12. Also note that since (20) was derived from the NF with inductive degeneration, good noise match is also possible with the proposed technique.

C. Linearity

In the following, the linearity characteristic and the gain partitioning in the cascaded stages of the LNA are explained. Fig. 13 shows the simulated gain (S_{21}) of each stage of the amplifier. As can be seen, in the 3–10-GHz band of interest, total gain larger than 25 dB is obtained. In addition, an average gain of 9.7, 8.8, –4.8, and 12 dB is achieved for the first, second, third, and fourth stages, respectively. The high gain of 9.7 dB in the first stage and 8.8 dB in the second stage is for achieving reasonable low NF. The low gain of the third stage is for better input third-order intermodulation point (IIP3). Since the bias resistance $R_{F1} (= 1000 \Omega)$ is large enough, the fourth stage of the LNA can be regarded as a typical common-emitter stage without

Fig. 13. Simulated S_{21} of the first, second, third, and fourth stages of the SiGe HBT UWB LNA. The overall S_{21} of the LNA is also shown for comparison.

a feedback resistance R_{F1} between its base and collector terminals. This means IIP3 of the proposed UWB LNA can be approximated as follows [20]:

$$\text{IIP3} \approx \frac{\text{IIP3}_{4}}{G_1 \cdot G_2 \cdot G_3} \approx \frac{4V_T^2/R_{S3}}{G_1 \cdot G_2 \cdot G_3},$$

or

$$\begin{aligned} \text{IIP3 (in dBm)} &\approx \frac{4V_T^2}{R_{S3}} \text{ (in dBm)} - G_1 \cdot G_2 \cdot G_3 \text{ (in dB)} \\ &= \frac{4V_T^2}{R_{S3}} \text{ (in dBm)} - G \text{ (in dB)} + G_4 \text{ (in dB)} \end{aligned} \quad (29)$$

in which G is the overall gain of the LNA. G_1, G_2, G_3 , and G_4 are the gain of the first, second, third, and fourth stages, respectively, of the LNA. IIP3_{4} is the IIP3 of the fourth stage of the LNA. R_{S3} is the real part of the input impedance if we look into the output terminal of the third stage, which is equal to 6.28, 4.88, and 1.06 Ω , respectively, at 3, 5.4, and 10 GHz. Since the LNA exhibits UWB characteristics and the frequency dependence of R_{S3} is not strong, IIP3 and 1-dB compression point ($P_{1\text{dB}}$) of the LNA will also exhibit UWB characteristics if G_4 exhibits UWB characteristics.

IV. MEASUREMENT RESULTS AND DISCUSSIONS

The noise and scattering parameters were measured on-wafer using an automated NP5 measurement system from ATN Microwave Inc., North Billerica, MA. The SiGe UWB LNA was biased at 2.74 mA and 1.5 V, 4.88 mA and 1.7 V, 4.3 mA and 1.7 V, and 6.1 mA and 1.0 V for the first, second, third and fourth stages, respectively; i.e., the power consumption is 25.8 mW. As shown in Fig. 14(a), the measured S_{11} is below -10 dB for frequency from 3 to 14 GHz (except for a small range from 10 to 11 GHz, which is below -9 dB), indicating a very broadband input matching characteristic. This is mainly attributed to the adoption of the capacitive feedback matching technique, which uses the Miller effect to enlarge the input capacitance C_{IN} . Simulation results show that the ripple around 10 GHz originates from the series resonance of inductance

$L_{GM} + L_1$ and the equivalent capacitance in the series path [see Fig. 14(a)]. Besides, the simulated S_{11} under both the small- and the large-signal condition (in the presence of -10-dBm interferer) is also shown in Fig. 14(a). As can be seen, S_{11} is below -10 dB from 3 to 14 GHz for both the small- and the large-signal condition. This means the impedance matching still holds even under the large-signal condition. Besides, from the fast-fast (FF) and slow-slow (SS) curves in Fig. 14(a), we can see that the impedance matching in the 3–14-GHz band still holds across standard 3- σ process variation.

The measured and calculated S_{21} based on (18b) are shown in Fig. 14(b). As can be seen, the calculated results agree well to the measured data. Peak S_{21} of 26.1 dB (at 9.6 GHz) and 3-dB bandwidth of 8.6 GHz (from 3 to 11.6 GHz) were achieved. In addition, from the FF and SS curves in Fig. 14(b), we can see that the proposed pole-zero cancellation approach is robust enough so under nominal 3- σ conditions, the bandwidth is still extended. These results demonstrate that the inductive shunt-shunt feedback technique can indeed achieve high and flat UWB S_{21} characteristics. The four-stage circuit configuration also guarantees a high reverse isolation. The measured S_{12} is below -42 dB from 2 to 15 GHz (not shown here), preventing LO leakage signals from transmitting to the antenna.

Fig. 14(c) shows the measured and calculated [by (20)] NF of the SiGe UWB LNA. This LNA achieves an NF of 2.5, 5.8, and 7.1 dB at 3, 10, and 12 GHz, respectively. These values are very close to the calculated results of 2.1, 5.5, and 6.8 dB, respectively, at 3, 10, and 12 GHz. The consistency between the measured and calculated ones verifies our noise analysis in Section III. Since the input matching bandwidth in this study (3–14 GHz) is wider than the specification (3.1–10.6 GHz), the NF frequency response can be flattened by reducing the capacitance term ($C_{\text{BE1}} + C_{\text{bc1}}$) [see (22)] at the expense of the input matching bandwidth.

A good phase linearity property is achieved, i.e., group-delay variation is only ± 28 ps across the entire band [see Fig. 14(d)]. By definition, group delay is the derivation of the phase of transfer function S_{21} and, hence, any resonance in the signal path (or pole in S_{21}) will contribute distortion in the group delay [24]. This means the good group-delay variation characteristic of the proposed UWB LNA architecture is attributed to the pushing of poles P_3 and P_4 [see Fig. 6(b)] outside of the 3–10-GHz-band of interest by the shunt-shunt feedback inductor. These results again demonstrate that this UWB LNA is very suitable for UWB pulse radio system applications. An expression often used to characterize the stability of an LNA is the Stern stability factor (K -factor), defined in [20]. In addition, it has been shown that $\mu > 1$ (or $\mu' > 1$ alone is necessary and sufficient for a circuit to be unconditionally stable [25]. Fig. 15 shows the measured K -factor, μ , and μ' of the SiGe HBT UWB LNA. Clearly, the LNA is unconditionally stable from 2 to 14 GHz.

Microwave power performances were measured by a load-pull ATN system with automatic tuners. The LNA achieves IIP3 of -18, -17, and -18 dBm at 3, 5.4, and 10 GHz, respectively, which conform well with those (-18.3, -16.1, and -15.7 dBm at 3, 5.4, and 10 GHz, respectively) calculated by (29), and close to those (-23.3, -21.6, and -17.8 dBm at

Fig. 14. (a) Measured S_{11} and simulated S_{11} under process variation and small- and large-signal condition. (b) Measured S_{21} and calculated S_{21} under process variation. (c) Measured and calculated NF [by (20)]. (d) Measured group delay of the SiGe UWB LNA.

Fig. 15. Measured K -factor, μ , and μ' versus frequency characteristics of the SiGe UWB LNA.

Fig. 16. Measured P_{1dB} and IIP3 of the SiGe UWB LNA.

3, 5.4, and 10 GHz, respectively) simulated by Agilent's Advanced Design System (ADS). In addition, the LNA achieves P_{1dB} of -27 , -25.5 , and -26 dBm at 3, 5.4, and 10 GHz, respectively, which conform well with the calculated results

(-27.9 , -25.7 , and -25.3 dBm at 3, 5.4, and 10 GHz, respectively) [20], and close to those (-32 , -30.5 , and -27 dBm at 3, 5.4, and 10 GHz, respectively) simulated by Agilent's ADS. The UWB IIP3 and P_{1dB} characteristics of the UWB

TABLE I
SUMMARY OF THE IMPLEMENTED SiGe UWB LNA, AND THE RECENTLY REPORTED STATE-OF-THE-ART SiGe AND CMOS UWB LNA

	Tech.	Maximum V_{DD} (V)	Input matching BW (GHz)	Peak S_{21} (dB)	3-dB BW of S_{21} (GHz)	Minimum NF (dB)	Max. Group delay variation (ps)	P_{1dB} (dBm)	IIP3 (dBm)	Power consumption (mW)
This work	0.35 μm SiGe	1.7	3 – 14	26.1	3-11.6	2.5	± 28	-26*	-17*	26
[1]	0.18 μm SiGe	3	2.2-8	21	3-10	2.5	NA	-14.7*	-1*	30
[2]	0.18 μm CMOS	1.8	2.6-11.7	9.3	2.3-9.2	4	± 40	-15**	-6.7**	9
[4]	0.25 μm SiGe	1	4-6	10	3.4-6.9	4.5	± 30	-10*	NA	3.5
[5]	0.18 μm SiGe	3.3	0.1-13.6 (< -7.2 dB)	20.3	0.1-13.6	1.8	NA	-12.5**	2.1**	26
[6]	0.13 μm CMOS	1.2	2.2-10.6	16.5	3.1-10.6	2.07	± 27.5	NA	-5.1 ⁺	9
[7]	0.25 μm SiGe	2.7	3.1-10.6	21	3.1-10.6	2.8	NA	-19.5 [#]	-8 [#]	29.7
[8]-[9]	0.18 μm CMOS	1.8	2.8-10.8	8.5	1.3-10.7	4.4	NA	NA	8.3 ⁺⁺	4.5
[10]	0.18 μm SiGe	2.5	3-10 (< -8.1 dB)	20.8	3-10	3.05	NA	NA	-11.7 [#]	42.5
[11]	0.18 μm CMOS	1.8	3.1-10.6 (< -9 dB)	17.5	3.1-10.6	3.1	± 48	NA	NA	33.2

#: 3 GHz *: 5 GHz **: 6 GHz +: 8 GHz ++: 10 GHz

LNA are attributed to the last stage of the LNA, which exhibits UWB characteristics, as explained in Section III. In addition, the low or mediocre IIP3 and P_{1dB} performances of the SiGe HBT UWB LNA are mainly due to its high gain of 25 dB [see (29)]; i.e., the IIP3 and P_{1dB} performances of the LNA can be improved at the expense of its gain. Fig. 16 shows the measured IIP3 and P_{1dB} characteristics of the SiGe HBT UWB LNA at 5.4 GHz.

Table I is a summary of this study along with other reported state-of-the-art UWB amplifiers. Based on Table I, it is found that UWB LNAs with gain higher than 20 dB ([1], [5], [7], [10], and this study) consume more power (30 mW in [1], 26 mW in [5], 29.7 mW in [7], 42.5 mW in [10], and 26 mW in this study) than those with gain near 10 dB. Clearly, if a more advanced SiGe HBT technology is adopted and a lower S_{21} is specified, such as those in [4], the power consumption of our proposed LNA is expected to be largely reduced. In addition, the overall performances of our proposed LNA are expected to be compatible with those in [5] and [6] if a more advanced 0.18- μm SiGe HBT (or 0.13- μm CMOS) technology is adopted.

V. CONCLUSION

A new methodology of UWB LNA design, which takes advantage of the Miller effect to achieve UWB input impedance matching and the inductive shunt-shunt feedback technique to achieve flat and high UWB S_{21} , has been proposed. Based on the proposed methodology, an SiGe UWB LNA was designed and implemented. Very good input matching, power gain, NF, and phase linearity performances were achieved for the

SiGe UWB LNA. These results demonstrate that the proposed methodology is very suitable for UWB LNA design.

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