A NOVEL ADAPTIVE CURRENT BIASED LINEAR RADIO-FREQUENCY POWER AMPLIFIER ON SIGE HBT PROCESS

TUO WU¹ and HONGYI CHEN
Institute of Microelectronics,
Tsinghua University, Beijing, 100084, China
¹wutuo@tsinghua.org.cn

DAHONG QIAN
Accel Semiconductor Corp. (Shanghai),
Shanghai, 201203, China

Received 13 January 2009
Accepted 27 January 2010

A novel adaptive current biased CLASS-A/shallow AB RF power amplifier is demonstrated in this paper. By theoretical deduction, a prototype is described to improve the linearity of a linear PA. With the realization on Jazz 0.35 µm SiGe HBT process and test verification, the novel adaptive current biased RF power amplifier shows 3 ~ 7 dB improvement of the ACPR at the output power of 19 dBm to meet the demand of CDMA IS95 spectrum mask without debasing the efficiency.

Keywords: Bipolar; power amplifier; adaptive bias; SiGe HBT; linearity.

1. Introduction

In recent years, SiGe HBT is a quite promising competitor with GaAs process in RFPA (Radio Frequency Power Amplifier) design.¹-³ The SiGe HBT shows comparable linearity performance and PAE (Power-Added Efficiency) in many PA modules, and the good consistency with standard Si technology could not only reduce the cost but also increase the flexibility to be integrated with other circuits, such as transceiver and power control circuits.

The deterioration of linearity induced by DC offset in PA has been observed for years, some researchers proposed several adaptive-bias circuits to alleviate the offset to enhance the linearity of PA.⁴ ⁵ But to our knowledge, there is no theoretic analysis

*This paper was recommended by Regional Editor Krishna Shenai.
in literatures to guide the design of the offset-compensate circuits, so only simulation could be carried in RFPA design, which is hardly optimized in depth for lacking of theoretic analysis.

In this paper, the effect of DC offset on linearity is analyzed theoretically. Moreover, a novel adaptive current bias circuitry is proposed to fully compensate the DC offset and improve the linearity of linear RFPAs. First, the DC offset of common biased PA is analyzed and a guideline of adaptive bias theme is put forward to improve the linearity in Sec. 2. According to the theoretical derivation, a novel adaptive current biased linear RF PA is concretely described in Sec. 3. The fabrication and test results on CDMA IS95 modulation are demonstrated in Sec. 4. Finally, a conclusion is drawn in Sec. 5.

2. Linearity of Adaptive Bias Bipolar Power Amplifier

2.1. DC offset of voltage biased bipolar power amplifier

The bias of a bipolar power amplifier is usually a voltage source with a resistance, as shown in Fig. 1(a). While its simplified small-signal equivalent circuit based on Gummel–Poon BJT model is shown in Fig. 1(b). The collector and base voltage-controlled current sources are respectively as follow:

\[ i_C = I_S \left( e^{\frac{v_{be}}{V_T}} - 1 \right) = I_S \left( e^{\frac{v_{be}}{V_T}} - 1 \right) \approx I_0 e^{\frac{v_{be}}{V_T}} \]

\[ \approx I_0 \left[ 1 + \frac{v_{be}}{V_T} + \left( \frac{v_{be}}{V_T} \right)^2 + \left( \frac{v_{be}}{V_T} \right)^3 \right] = I_0 + i_c \]

\[ i_B \approx \frac{i_C}{\beta} \approx \frac{I_0}{\beta} e^{\frac{v_{be}}{V_T}} \approx \frac{I_0}{\beta} \left[ 1 + \frac{v_{be}}{V_T} + \left( \frac{v_{be}}{V_T} \right)^2 + \left( \frac{v_{be}}{V_T} \right)^3 \right] = \frac{I_0}{\beta} + i_b \]

![Fig. 1](image-url)
where
\[ I_0 = I_S e^{\frac{v_{be}}{V_T}} \cdot i_c = I_0 \left[ \frac{v_{be}}{V_T} + \left( \frac{v_{be}}{V_T} \right)^2 + \left( \frac{v_{be}}{V_T} \right)^3 \right] \cdot i_b. \]

Here \( \beta \) is the current gain of bipolar Q1, \( i_c \) and \( i_b \) are the collector and base AC current respectively, while \( I_0 \) represents the DC collector current, but not exactly the quiescent current, since DC offset could be induced by even-order nonlinearity. A 3rd polynomial expansion of exponent function is usually sufficient for linear PAs (Class A/Shallow AB). So the DC equation of Fig. 1(b) could be written as (ignoring the emitter resistance \( R_e \))

\[ v_{be|dc} + \frac{R_{Bias}}{\beta} (I_0 + i_c|dc) - V_{Bias} = 0. \] (3)

According to Volterra’s theory,\(^6,7\) the nonlinearity can be written as following AC equations.

\[ v_{be} = C_1(s) \cdot v_s + C_2(s_1, s_2) \cdot v_s^2 + C_3(s_1, s_2, s_3) \cdot v_s^3, \] (4)

\[ i_c = I_0 \left( \frac{v_{be}}{V_T} + \frac{1}{2} \frac{v_{be}^2}{V_T^2} + \frac{1}{6} \frac{v_{be}^3}{V_T^3} \right) \]

\[ = D_1(s) \cdot v_s + D_2(s_1, s_2) \cdot v_s^2 + D_3(s_1, s_2, s_3) \cdot v_s^3, \] (5)

where \( V_{s^n} \) is the \( n \)th power of the voltage source signal and \( C_n(\cdot)/D_n(\cdot) \) is the Volterra series coefficient which is a linear function of \( n \) number of frequencies. The operator “\( \circ \)” indicates multiplying each frequency component in \( V_{s^n} \) by the magnitude of \( C_n(\cdot)/D_n(\cdot) \) and shifting each frequency component in \( V_{s^n} \) by the phase of \( C_n(\cdot)/D_n(\cdot) \).\(^7\)

Owing to the even-order nonlinearity, the AC-induced DC collector current could be written as\(^8\)

\[ i_{c|dc} = \frac{v_{sm}^2 I_0}{4V_T^2} C_1(s)C_1(-s). \] (6)

Equation (3) could be rewritten as:

\[ V_T \ln \frac{I_0}{I_s} + \frac{R_{Bias}}{\beta} + I_0 \frac{v_{sm}^2}{4V_T^2} |C_1(s)|^2 \frac{R_{Bias}}{\beta} - V_{Bias} = 0. \] (7)

When \( v_{sm} = 0 \), \( I_0(0) = I_{0q1} \), which represents quiescent \( I_0 \) with non-AC-input. Equation (7) turns into as follow,

\[ V_T \ln \frac{I_{0q1}}{I_s} + \frac{R_{Bias}}{\beta} - V_{Bias} = 0. \] (8)
Equation (7) minus Eq. (8), and take the approximation of

\[
\ln \frac{I_0}{I_{0q}} \approx \frac{I_0 - I_{0q}}{I_{0q}},
\]

\[
I_0 = I_{0q} \left(1 + \frac{\beta V_T}{I_{0q} R_{Bias}}\right) \left(1 + \frac{\beta V_T}{I_{0q} R_{Bias}} + kv_{sm}^2\right),
\]

where \(k = \frac{|C_s(s)|^2}{4V_T^2}\).

So we can see \(I_0\) would reduce with the increase of \(v_{sm}\) and \(R_{Bias}\), which means that the linearity of current Bias PA is worse than voltage bias PA.

Especially, if the DC source is an ideal current source, \(R_{Bias} = \infty\), then

\[
I_0 \approx \frac{I_{0q}}{1 + kv_{sm}^2} \approx I_{0q}(1 - kv_{sm}^2).
\]

The output current

\[
i_{out} = F_1(s) \circ v_s + F_2(s_1, s_2) \circ v_s^2 + F_3(s_1, s_2, s_3) \circ v_s^3.
\]

If input is a 2-tone signal, the 3rd inter-modulation product (IMR3) could be written as\(^7,8\)

\[
|IMR3| = \frac{3}{4} \frac{F_3(s_1, s_1, -s_2)}{F_1(s_1)} \left|\frac{v_{sm}}{\sqrt{2}}\right|^2 = \frac{3|v_{sm}|^2}{8} f_1(I_0),
\]

where \(f_1(I_0) = \frac{F_3(s_1, s_1, -s_2)}{F_1(s_1)}\) is a function of \(I_0\).

By MATLAB simulation, Fig. 2 illustrates the variety of \(I_0\) and IMR3 with input power. It shows that, if the DC current \(I_0\) drops to nearly 70% of quiescent value, the offset-induced IMR3 would be 6dB (double) worse than the small-signal extrapolated IMR3 (\(f_1(I_0) = f_1(I_{0q})\)).

![Fig. 2. Coefficient \(f(I_0)\) of IMR versus \(I_0\).](attachment:image.png)
2.2. DC offset compensation of adaptive biased PA

From Eqs. (10) and (11) we can see that if an adaptive bias could be adopted to compensate the drop of DC current $I_0$, the linearity could be improved. As Fig. 3 shows, a constant DC current source $I_{B0}$ as well as an input-correlative DC current source $I_{ad}$ could be combined together as the current bias of RF PA.

Apparently, for optimum the input-correlative DC current source $I_{ad}$ should be as follow:

$$I_{ad} = K v_s^2,$$

where \( K = \frac{|C_1(s)|^2}{4V_T^2} \).

If Eq. (11) is satisfied, $I_0$ would not decrease with the increases of input signal, so that the linearity could be improved.

3. Design of the Proposed Adaptive Current Biased Linear RF PA

Based on a PVT-insensitive current bias linear PA, a novel adaptive bias PA is illustrated in Fig. 4. A fixed DC bias with an adaptive DC bias would be added to inject into the base current.

To realize the AC-to-DC relationship of Eq. (11), a power detector and a DC trans-conductance amplifier are added into the adaptive bias branch.

3.1. Current bias of power amplifier

The current bias scheme is based on the circuitry of Ref. 9. This structure could separate the mirror current and actual bias current, so that the adaptive bias could be fully added into the bias branch. As shown in Fig. 5, current mirror of $M5/M6$ would pull $V_{fb}$ down and then turn on $M1/M2$, as well as $Q2$ and $M3/M4$ without any additional start-up sub-circuit. The negative feedback scheme forces
While $I_{ds}(M_3) = nI_{ds}(M_4) = nI_{Bias}$, while $I_d(M_1) = mI_d(M_2)$, $R_b(Q_2) = mR_b$ ($Q_1$) and $Q_1 = mQ_2$ could guarantee $I_C(Q_1) = mI_C(Q_2)$, so once the transistor match is perfect, $I_C(Q_1) = mnI_{Bias}$. Since the feedback is used to set up the DC point, the dominative pole could be put towards the origin sufficiently to ensure the stability. A capacitance is added to node $V_{fb}$, so that the dominative pole should be $s_1 = -\frac{1}{(r_{ds4}+r_{ds5})C_0}$.

### 3.2. Novel power detector

Power detectors have been studied in some papers.\textsuperscript{10} Figure 6(a) shows a conventional scheme of a power detector. Theoretical deduction concluded that in small signal region the input (AC) to output (DC) voltage transfer function is in square relation,\textsuperscript{10}

$$V_o \approx V_T \ln \left( 1 + \frac{v_i^2}{4V_T^2} \right) \approx \frac{v_i^2}{4V_T},$$

(13)
Moreover, AC fluctuation on DC bias would affect the linearity of PA, depending on its phase, so the AC signal leakage between input and output of the power detector should be minimized as possible. In conventional detector circuit, the forward biased BE junction induced a leakage from $V_i$ to $V_0$. A novel power detector circuit (Fig. 6(b)) is adopted to enhance the isolation between input and output without increasing the complexity of circuits. The DC output voltage is also proportional to the square of AC input voltage at small-small region.

$$V_o \approx \frac{2\beta R_C}{R_b} V_T \ln \left( 1 + \frac{v_i^2}{4V_T^2} \right) \approx \frac{\beta R_C}{2R_bV_T} v_i^2.$$

HB (Harmonic Balance) simulation result shows that the leakage could be depressed for $2 \sim 3$ dB due to the novel power sensor circuit, as shown in Fig. 7.
3.3. DC trans-conductance amplifier

The output of the power sensor is a DC voltage, so the current driver is actually a DC trans-conductance, as Fig. 8 shows. When $V_{in} = 0$, the PMOS and NMOS of the 2nd driven branch is matched by the current mirror so that $I_{out} = 0$. As the $V_{in}$ increases, the $V_{o1}$ decreases so that the PMOS current of the 2nd branch is larger than the NMOS transistor. The key issue of the DC trans-conductance amplifier is the headroom. The DC point should be carefully designed to avoid the transistors departing saturate region within the whole detecting range.

4. Simulation and Test Results

A two-stage power amplifier is fabricated on Jazz 0.35 SiGe BiCMOS process. Each stage adopts adaptive bias to improve the linearity. The area of the first and the second stage is $1500 \times 0.3 \mu m^2$ and $4100 \times 0.3 \mu m^2$, respectively. The inter-stage match is on chip for simplicity, and the total area (including PADS) is within $1.65 \times 0.8 mm^2$.

The fabricated PA is assembled and tested at 1.95 GHz, which is the application spectrum of CDMA IS95. Figure 10 is the test board and Fig. 11 illustrates comparative 1-tone and 2-tone test results of different bias PAs. Figure 11(a) indicates that adaptive bias could suspend the PA into compression to improve the linearity. The swept IMR test result is shown in Fig. 11(b), in which the IMR improvement could be clearly observed.

The modulated signal (CDMA IS95) is also added into the test board, and the test result is shown in Figs. 12 and 13. When output power is 19 dBm, the comparative spectrum mask shows that the ACPR could be improved for 3–7 dB.

The PAE of adaptive-bias PA is slightly lower than fixed-bias one in low power range. But as shown in Fig. 14, the PAE at $P_{1dB}$ (saturation power) are both slightly

![Fig. 8. Schematic of DC trans-conductance amplifier.](image-url)
A Novel Adaptive Current Biased Linear Radio-Frequency Power Amplifier

Fig. 9. (a) scheme and (b) layout of the novel two stage power amplifier.

Fig. 10. Test board of fabricated PA.
Fig. 11. (a) 1-tone and (b) 2-tone comparative test results of different biased PAs.

Fig. 12. ACPR of adjacent channel (885 kHz) @ (P_{out} = 19 \text{ dBm}).

Fig. 13. ACPR of alternative channel (1.98 MHz) @ (P_{out} = 19 \text{ dBm}).
enhanced from 24% (35%) to 28% (40%), because the $P_{\text{sat}}$ (saturation power) has been increased, as Fig. 11(a) illustrates.

5. Conclusions

In this paper a novel adaptive biased RF power amplifier is demonstrated. By analyzing the bias affection on linearity, an adaptive bias circuitry proportional to the input power is added to improve the linearity. Test result shows that the ACPR could be improved for $3 \sim 7$ dB with CDMA IS95 modulation without debasing the efficiency.

References


