

(10) **Patent No.:** US 7,345,547 B2
(45) **Date of Patent:** Mar. 18, 2008

- | | | | | |
|-----------|------|---------|---------------------|---------|
| 6,300,837 | B1 * | 10/2001 | Sowlati et al. | 330/296 |
| 6,333,677 | B1 * | 12/2001 | Dening | 330/285 |
| 6,486,739 | B1 | 11/2002 | Luo | 330/296 |
| 6,529,080 | B1 | 3/2003 | Seymour et al. | |
| 7,154,336 | B2 * | 12/2006 | Maeda | 330/285 |

- ## OTHER PUBLICATIONS

- Chen, Shiao-Shien, et al. "Investigation of ESD Devices in 0.18- μ m SiGE BiCMOS Process," IEEE 03CH37400 41st Annual International Reliability Physics Symposium, Dallas, Texas (2003) pp. 357-360.

- Dunn, J., et al., "Trends in silicon Germanium BiCMOS Integration and Reliability," IEEE 00CH37059, 38th Annual International Reliability Physics Symposium, San Jose, CA (2000) pp. 237-242.

- Iwai, Taisuke, et al., "42% High-Efficiency Two-Stage HBT Power-Amplifier MMIC for W-CDMA Cellular Phone Systems," IEEE Transactions on Microwave Theory and Techniques, vol. 48, No. 12, (Dec. 2000) pp. 2567-2572.

- Kawamura, H., et al., "A Miniature 44% Efficiency GaAs HBT Power Amplifier MMIC for the W-CDMA Application", IEEE GaAs Digest (2000) pp. 25-28.

- (Continued)

- Primary Examiner*—Khanh Van Nguyen

- (74) *Attorney, Agent, or Firm*—Morgan, Lewis & Bockius LLP

- (57) **ABSTRACT**

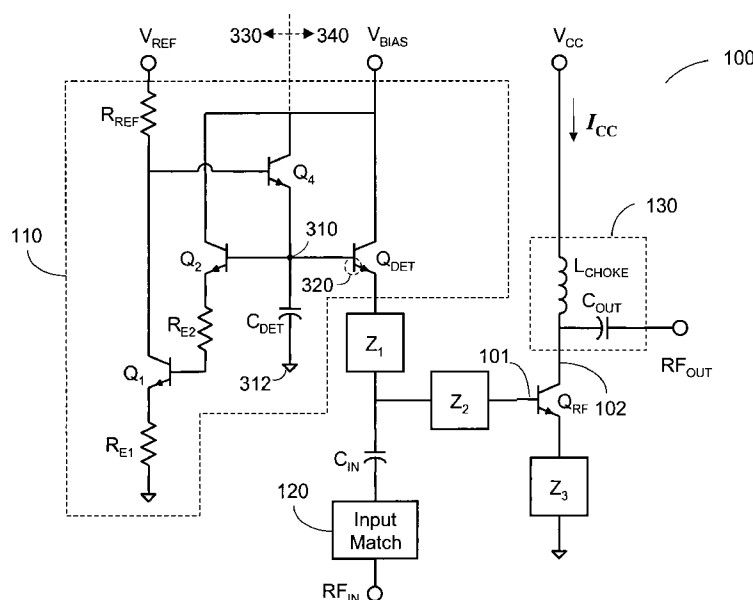
See application file for complete search history.

- The embodiments of the present invention include a bias circuit for a power-amplifying device, which receives and amplifies an input RF signal having a series of RF cycles within a modulation envelop. The bias circuit compensates odd-order distortion processes by detecting the power in the input signal and providing a dynamic adjustment to a bias stimulus for the power-amplifying device within a time scale of the modulation envelope.

U.S. PATENT DOCUMENTS

- | | | |
|--------------|---------|------------------|
| 1,850,187 A | 3/1932 | Thewes |
| 2,114,971 A | 4/1938 | Segeler |
| 2,107,910 A | 8/1938 | Schumacher |
| 5,590,411 A | 12/1996 | Sroka et al. |
| 5,710,523 A | 1/1998 | Kobayashi |
| 5,850,187 A | 12/1998 | Carrender et al. |
| 6,107,910 A | 8/2000 | Nysen |
| 6,114,971 A | 9/2000 | Nysen |
| 6,275,687 B1 | 8/2001 | Lloyd |

9 Claims, 10 Drawing Sheets



OTHER PUBLICATIONS

Jeon, Yong-Joon, et al., "Improved HBT Linearity With a 'Post Distortion'-Type Collector Linearizer," IEEE Microwave and Wireless Components Letters, vol. 13, No. 3, (Mar. 2003) pp. 102-104.

Kang, Sanghoon, et al., "Linearity Analysis of CMOS for RF Application," IEEE Radio Frequency Integrated Circuits Symposium, (2002), pp. 363-366.

Kim, Hye-Ryoung, et al., "A 5-GHz LNA for Wireless LAN Application Based on 0.5 μ m SiGe BiCMOS," 3rd International Conference on Microwave and Millimeter Wave Technology Proceedings, IEEE (2005) pp. 50-53.

Kim, Joon H., et al., "High Linear HBT MMIC Power amplifier with Partial RF Coupling to Bias Circuit for W-CDMA Portable Application," 3rd International Conference on Microwave and Millimeter Wave Technology Proceedings, IEEE (2005) pp. 809-812.

Ko, Jin-Su, et al., "Effect of Bias Scheme on Intermodulation Distortion and its Use for the Design of PCS Tx Driver," IEEE Radio Frequency Integrated Circuits Symposium, (2000) pp. 105-105.

Noh, Youn Sub, et al., "PCS/W-CDMA Dual-Band MMIC Power Amplifier With a Newly Proposed Linearizing Bias Circuit," IEEE Journal of Solid-State Circuits, vol. 37, No. 9, (Sep. 2002) pp. 1096-1099.

Taniguchi, Eiji, et al., "Dual Bias Feed SiGe HBT Low Noise Linear Amplifier," IEEE Radio Frequency Integrated Circuits Symposium, (2001) pp. 227-230.

Trask, Chris, "High Efficiency Broadband Linear Push-Pull Power Amplifiers Using Linearity Augmentation," IEEE International Symposium on Circuits and Systems, (2002) pp. II432-II435.

Voldman, Steven H., et al., "The Influence of Process and Design of Subcollectors on the ESD Robustness of ESD Structures and Silicon Germanium Heterojunction Bipolar Transistors in a BiCMOS SiGe Technology," IEEE 03CH37400, 41st Annual International Reliability Physics Symposium, Dallas, Texas (2003) pp. 347-356.

Voldman, Steven H., et al., "Silicon Germanium Heterojunction Bipolar Transistor Electrostatic Discharge Power Clamps and the Johnson Limit in RF BiCMOS SiGe Technology," Elsevier Science, Journal of Electrostatics, vol. 56, (2002) pp. 341-362.

Voldman, Steven H., "The State of the Art of Electrostatic Discharge Protection" Physics, Technology, Circuits, Design, Simulation, and Scaling, IEEE Journal of Solid-State Circuits, vol. 34, No. 9 (1999) pp. 1272-1282.

Wang, Albert, "Recent Developments in ESD Protection for RFIC's," IEEE (2003) pp. 171-178.

Yoshimasu, Tosjijiko, et al., "An HBT MMIC Power Amplifier with an Integrated Diode Linearizer for Low-Voltage Portable Phone Applications," IEEE Journal of solid-State Circuits, vol. 33, No. 9, (Sep. 1998) pp. 1290-1296.

* cited by examiner

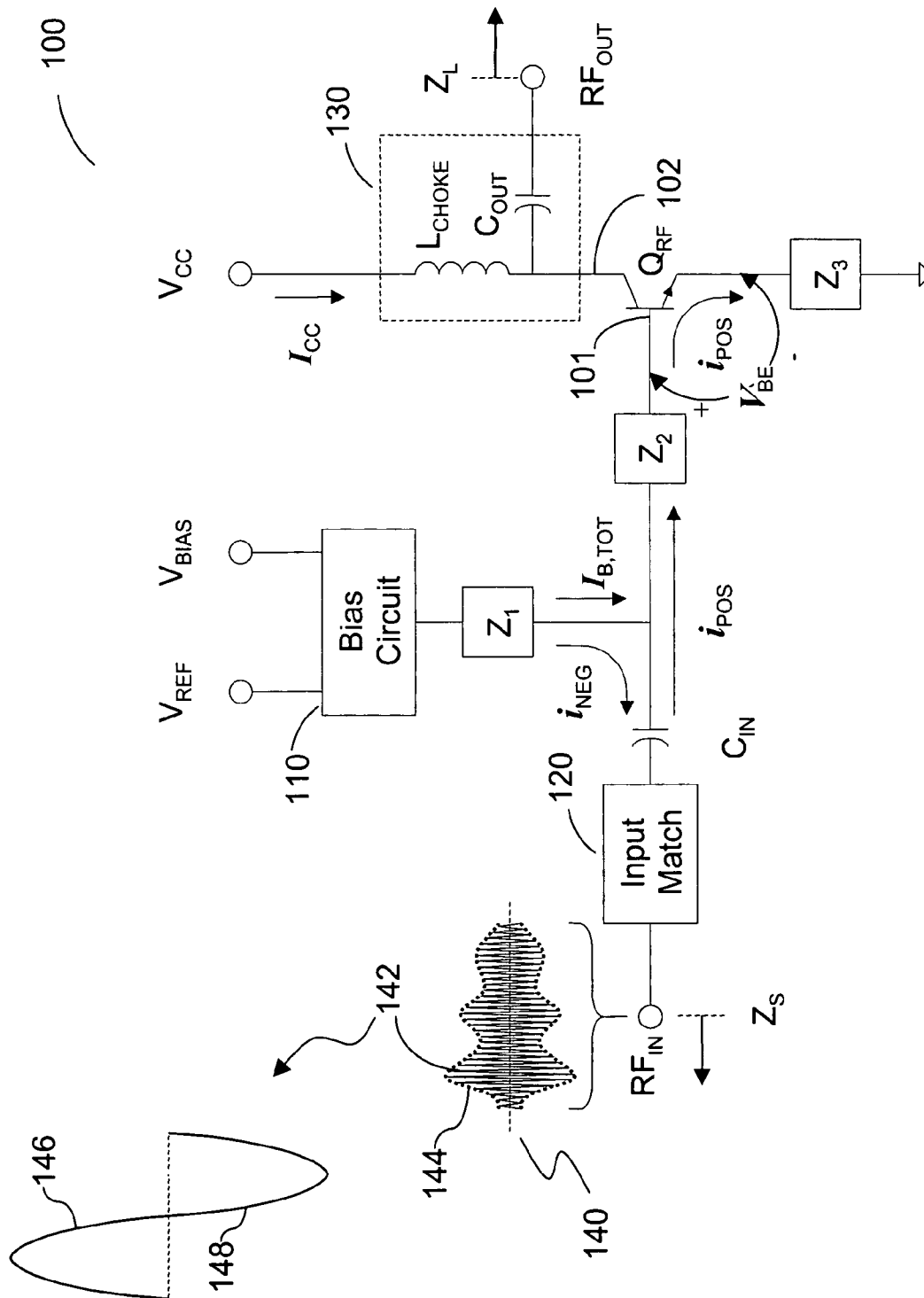


FIG. 1

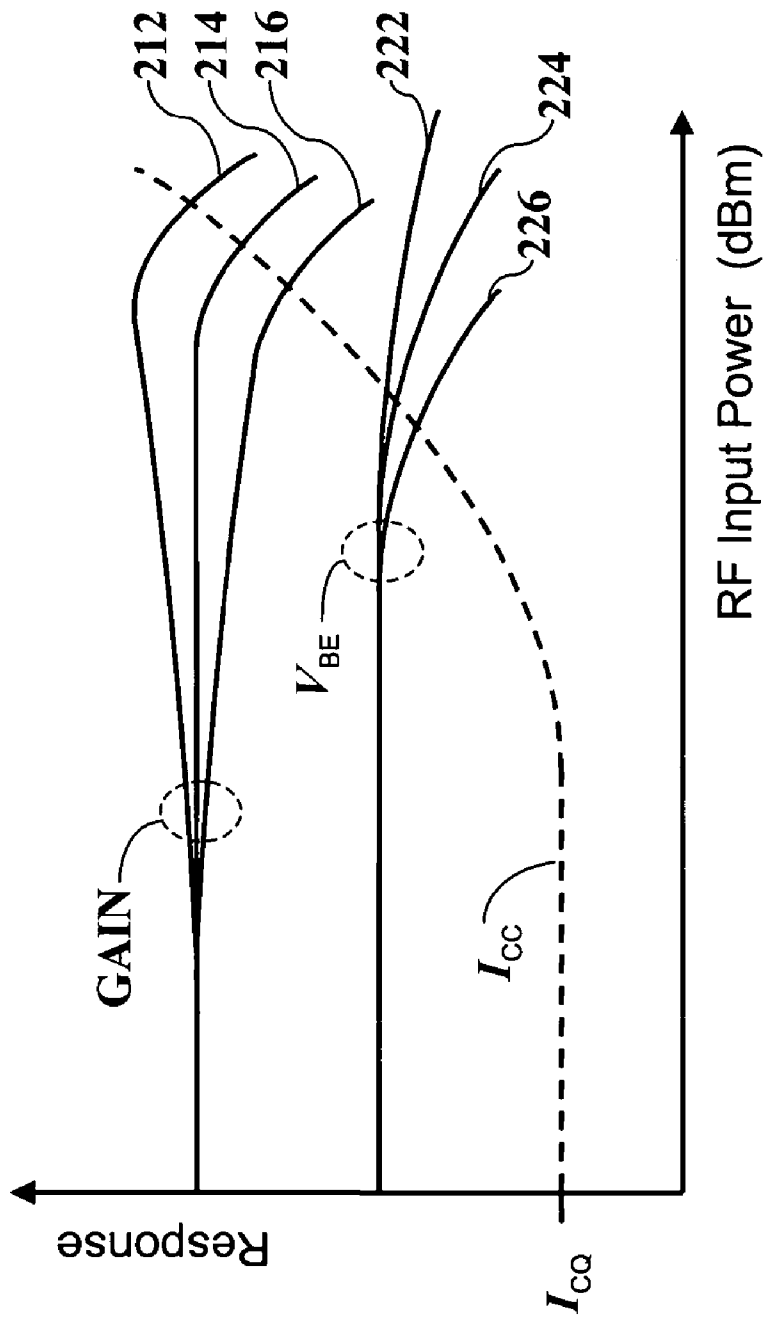


FIG. 2

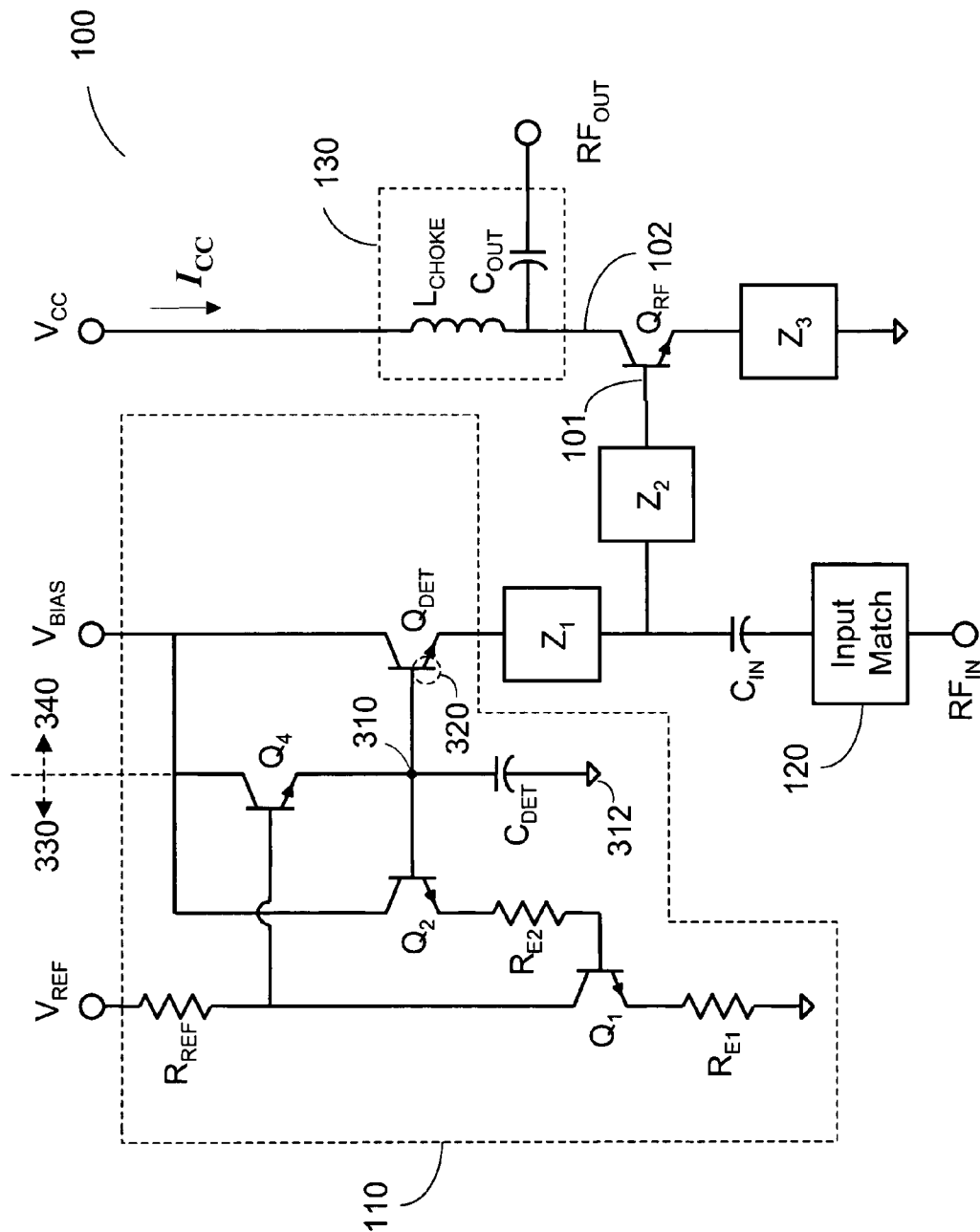


FIG. 3

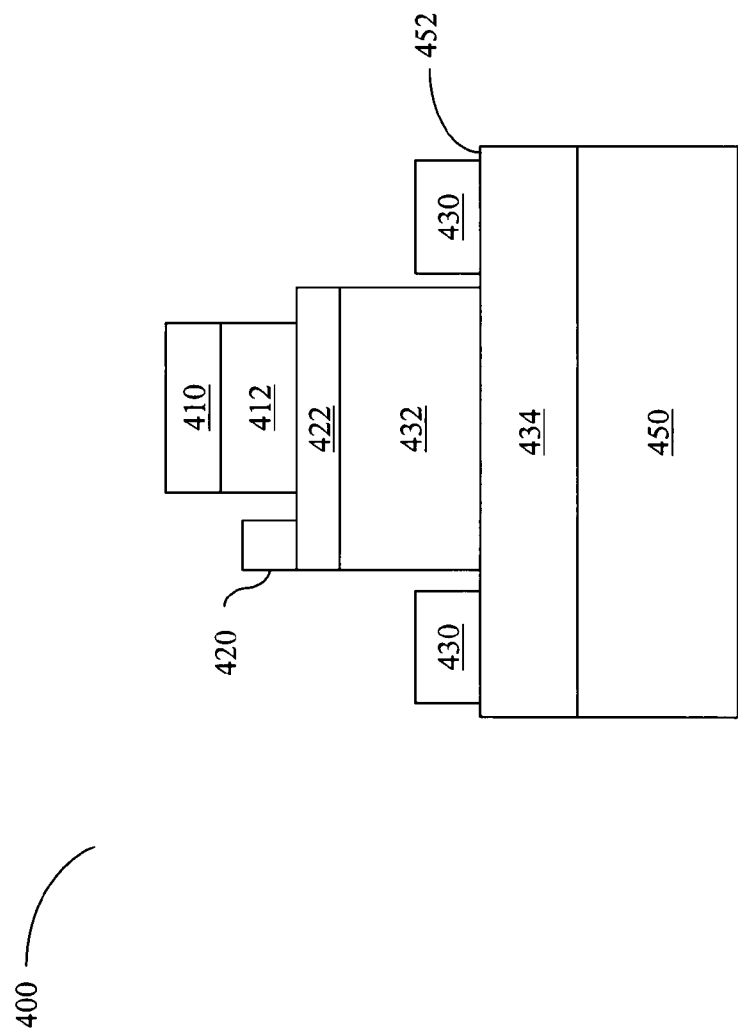


FIG. 4

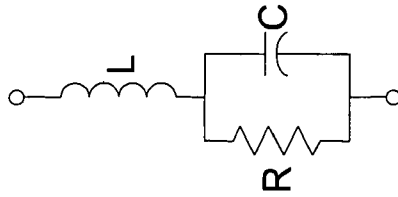


FIG. 5(f)

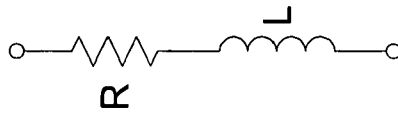


FIG. 5(e)



FIG. 5(d)

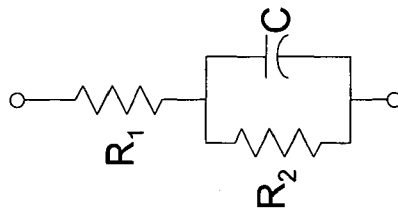


FIG. 5(c)

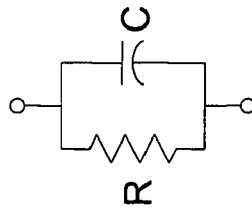


FIG. 5(b)

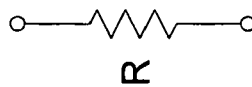


FIG. 5(a)

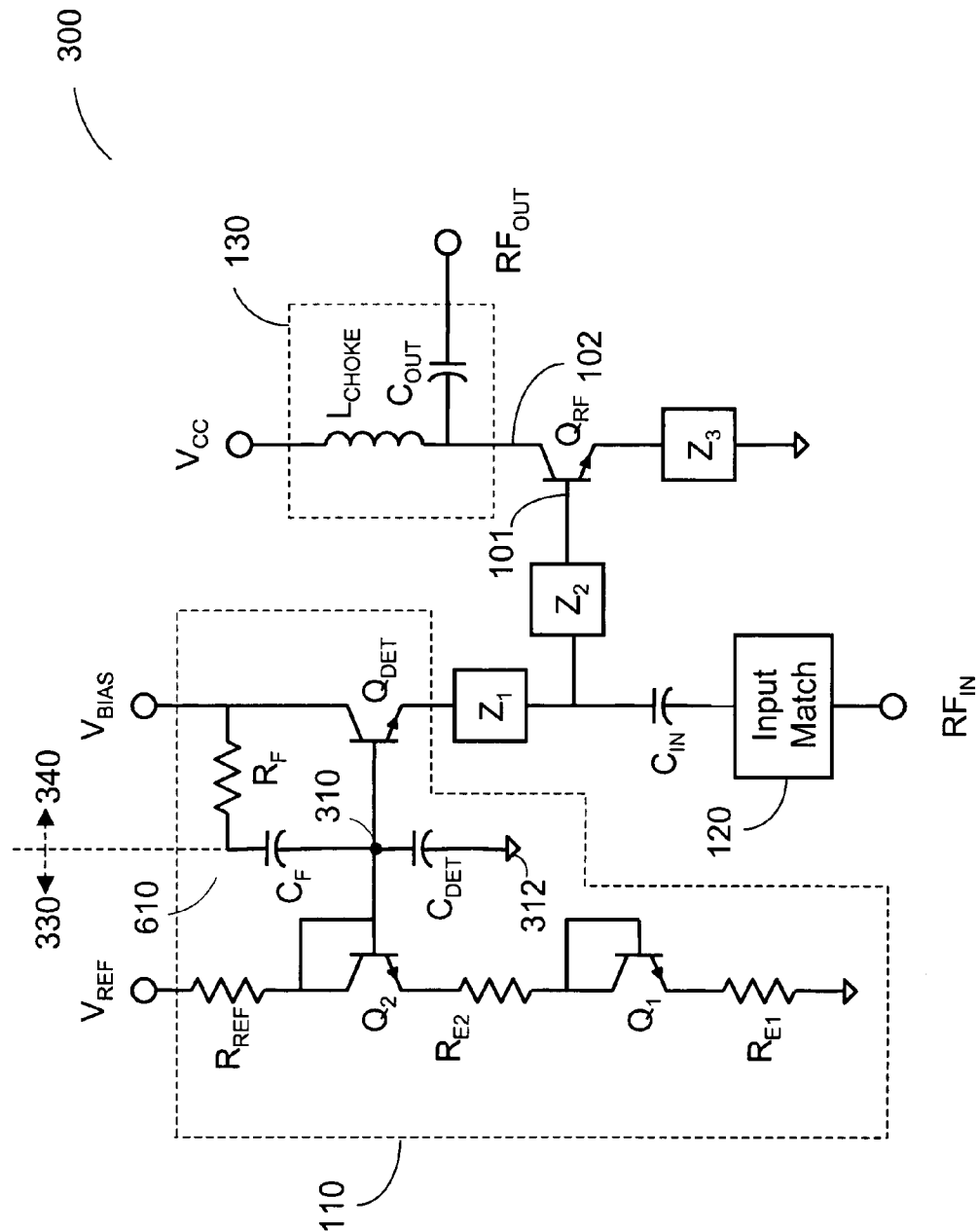


FIG. 6

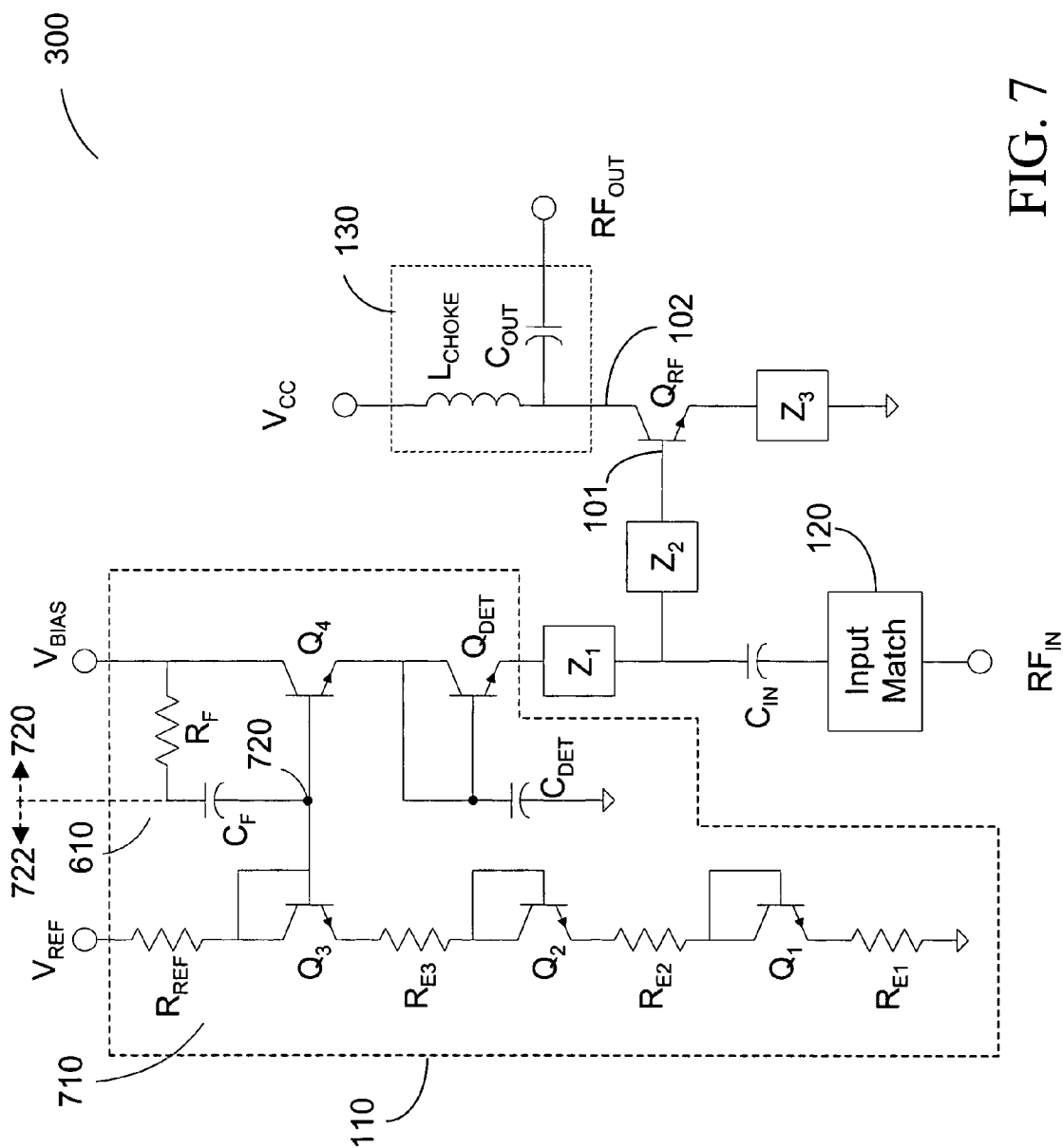
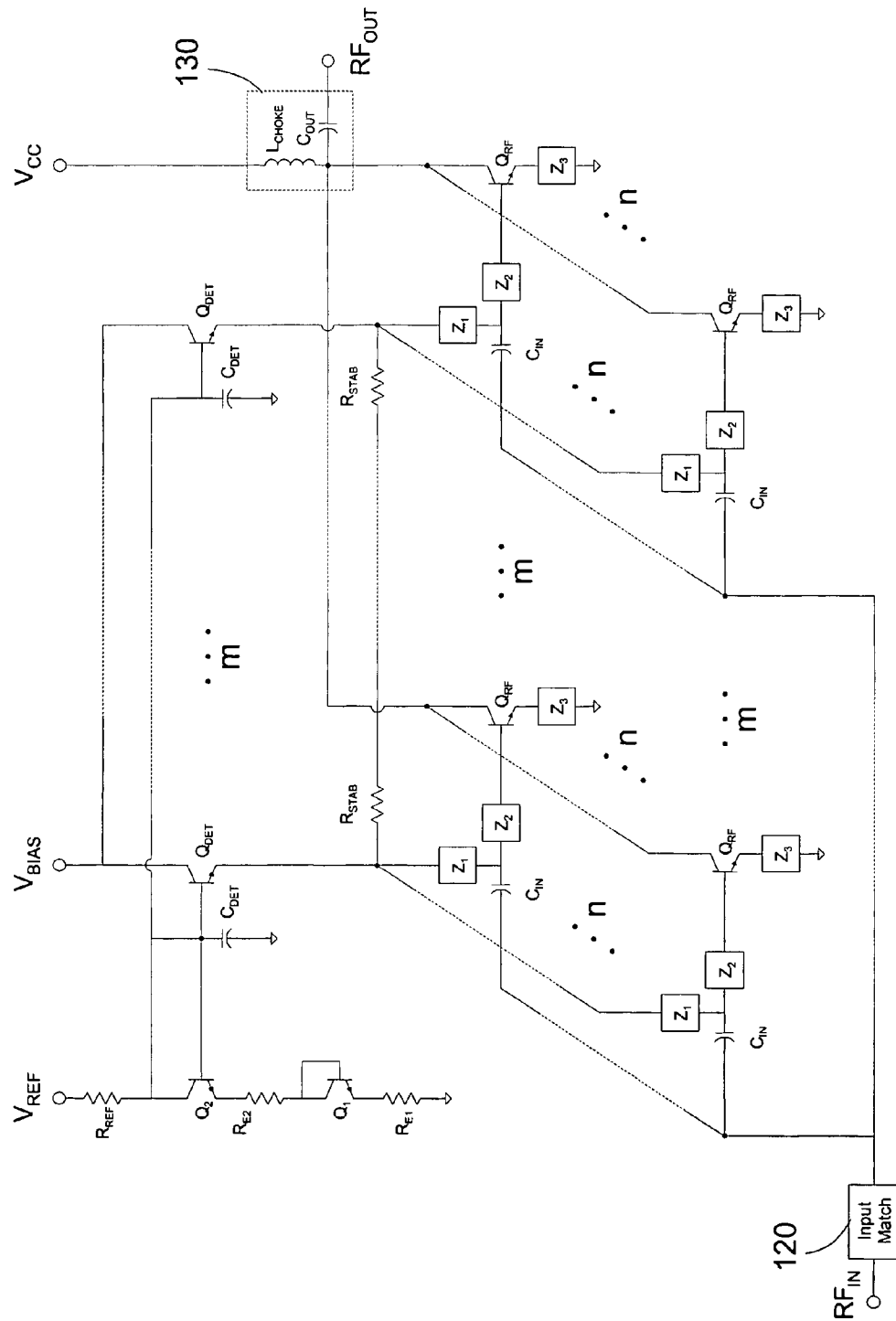


FIG. 7



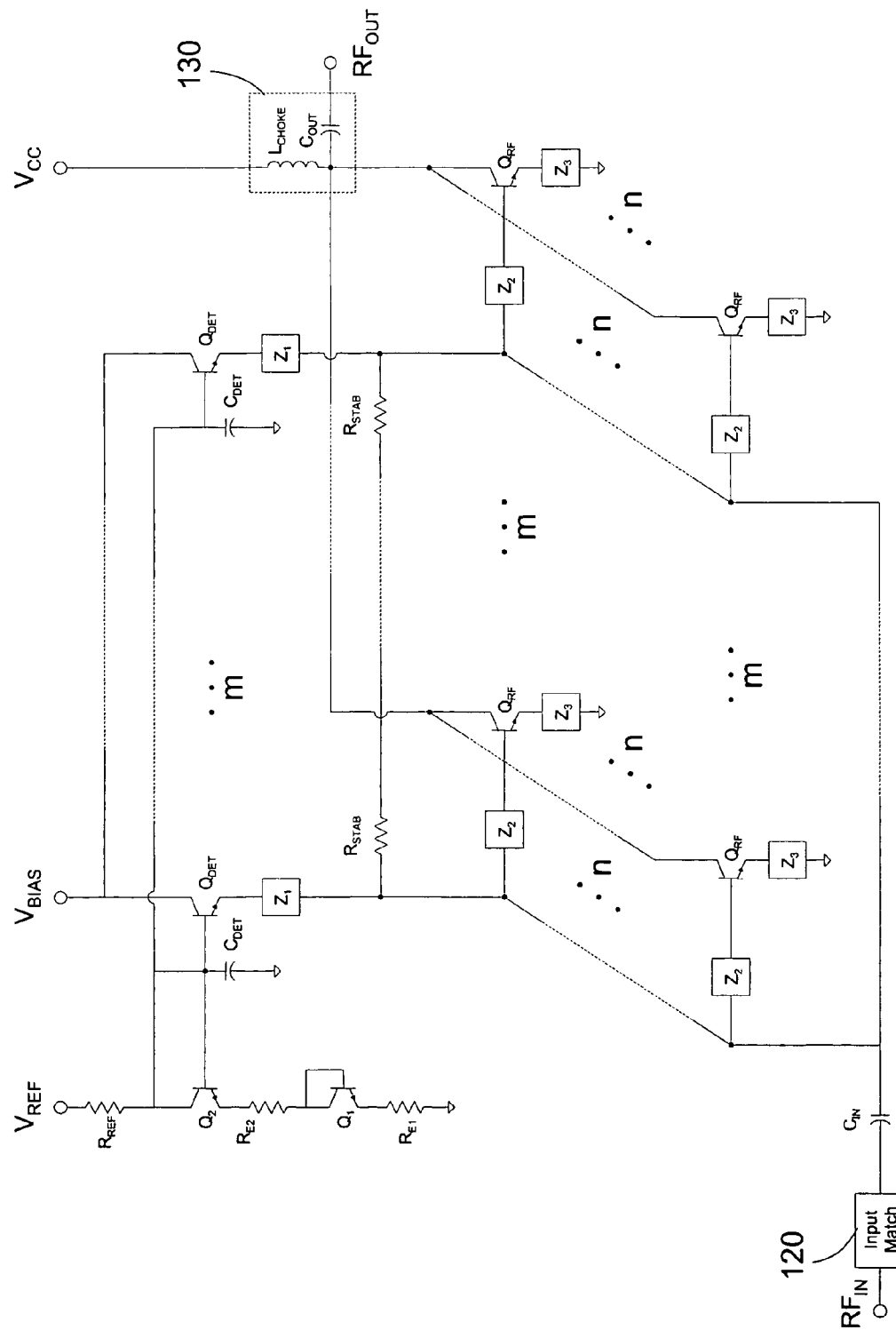


FIG. 9

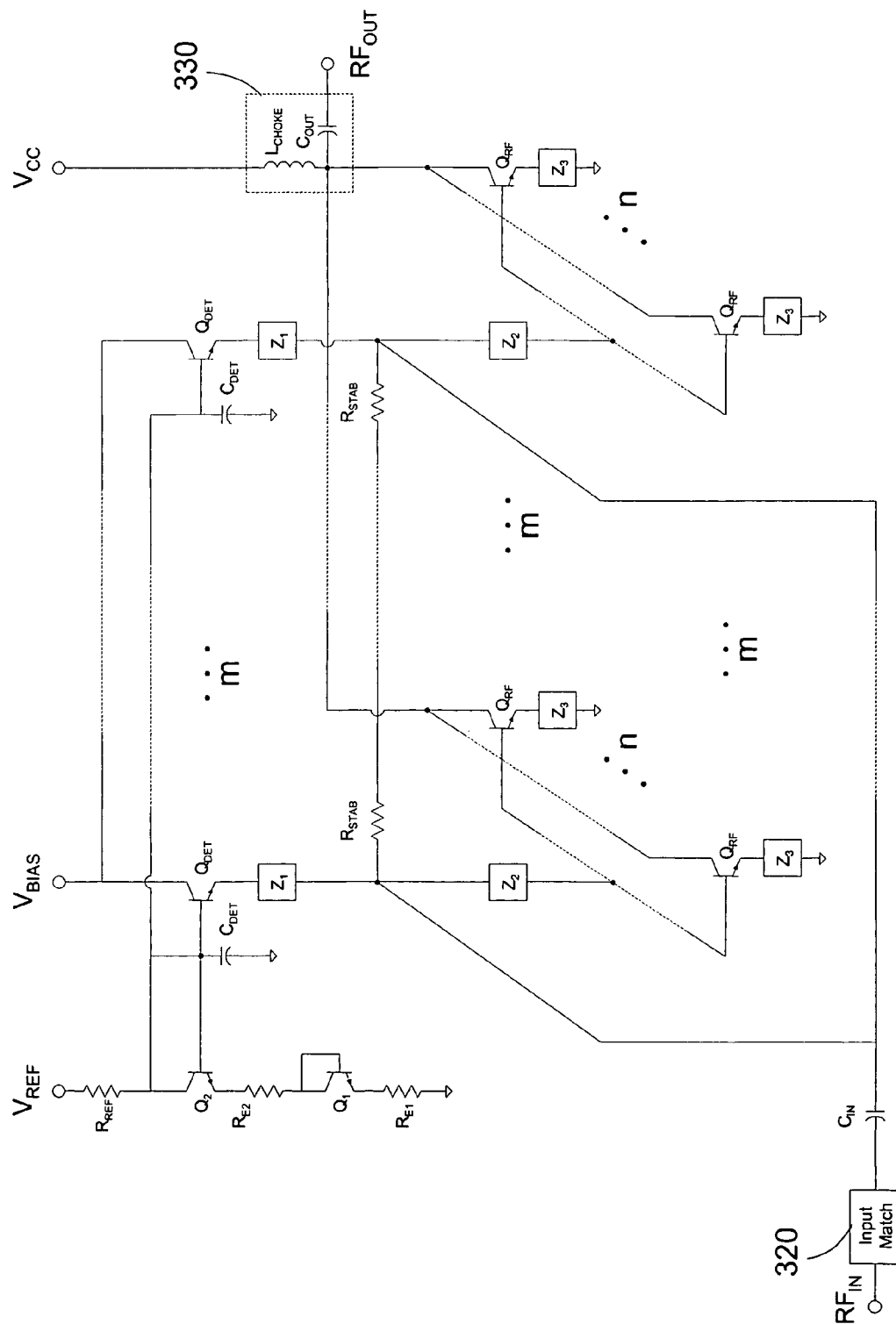


FIG. 10

1

BIAS CIRCUIT FOR BJT AMPLIFIER**FIELD OF THE INVENTION**

The present invention relates to radio-frequency (RF) power amplifier circuits and, in particular, to bias circuits that serve to minimize distortion in the amplifier circuits.

BACKGROUND OF THE INVENTION

Electronic equipment such as computers, wireless devices, broadband devices, radios, televisions and other similar devices communicate with each another by transmitting signals through air, space and guided media such as wire, cable, microstrip, waveguide, and optical fiber. These transmission signals undergo a variety of processes throughout their communication paths, one of which involves amplifying the signals using power amplifiers.

A radio frequency (RF) power amplifier is a circuit that is capable of receiving an RF input signal and amplifying it to produce an RF output signal that is a magnified version of the input signal. RF power amplifiers are frequently used in communications systems such as wireless telephony, satellite links, optical transceivers, and cable television distribution systems. An RF amplifier typically includes at least one power-amplifying device, such as a power-amplifying transistor, and a bias circuit that sets a quiescent operating point of the transistor. The transistor may be a field effect transistor (FET) device, such as a laterally-diffused metal-oxide-semiconductor (LDMOS) transistor, or a bipolar junctions transistor (BJT) device, such as a heterojunction bipolar transistor (HBT).

In a transistor-based RF amplifier, there are tradeoffs between maximizing efficiency and preserving the fidelity of the RF signal. The efficiency of the amplifier is defined as the output RF signal power divided by the power supplied to the amplifier from a power supply. The fidelity of an amplified signal is often described in terms of a deviation from an ideal linear noiseless process and is characterized using a variety of metrics including harmonic distortion, intermodulation distortion, adjacent channel power ratio (ACPR), cross-modulation, error vector magnitude, and bit error rate, etc., most of which are interrelated. For communication systems involving a modulated carrier signal whose modulation frequency is a small fraction of the carrier frequency, odd-order intermodulation distortion (IMD) processes are the primary sources of deleterious signal distortion in power amplifiers. In a given communication system, the contributing effects of such distortion can be related to other performance metrics such as bit error rate and error vector magnitude. Thus, in the following description, odd-order IMD and its related ACPR characteristic are mainly used as the metrics for the fidelity of amplified signals. See Cripps, "Power Amplifiers for Wireless Communications," Artech House: Norwood, Mass., 1999, and Pedro and Carvalho, "Intermodulation Distortion in Microwave and Wireless Circuits," Artech House: Norwood, Mass., 2003.

Given the inherent tradeoff between efficiency and linearity (fidelity), practitioners in the art of power amplifier design have developed a broad portfolio of circuit concepts and methods for exercising this tradeoff. One primary technique for increasing the efficiency of a power amplifier is to control the conduction angle of the transistor device. Various classes of amplifiers have been devised to manage the conduction angle, such as class-A, class-AB, class-B, class-C, class-D, class-F, and class-S amplifiers, listed roughly in the order of reduced conduction angle. See Clark and Hess,

2

"Communication Circuits: Analysis and Design," Wiley: New York, 1971, and H. L. Krauss et al., "Solid State Radio Engineering," Wiley: New York, 1980.

The reduced conduction angle allows the amplifier to approach the efficiency of a switching device at the expense of signal distortion and gain. Furthermore, BJT operation at high power (e.g., 2-10 W) and high voltage (e.g., >10V) requires substantial thermal ballasting in either or both of the base and emitter terminals of the power-amplifying transistor. See Anholt, "Electrical and Thermal Characterization of MESFETs, HEMTs, and HBTs," Artech House: Norwood, Mass., 1995. Such thermal ballasting helps to increase the current-handling capability and thermal stability of the power-amplifying device but at the same time degrade the linearity of the amplifier circuit. See Pedro and Carvalho, supra, and Vuolevei and Rahkonen, "Distortion in RF Power Amplifiers," Artech House: Norwood, Mass., 2003. The engineering challenge in modern RF power amplifier design is to devise circuits with an optimum configuration of transistor device, tuning, conduction angle and bias control to maximize efficiency while controlling odd-order distortion processes to meet the linearity specifications for a particular communication signal or system.

SUMMARY

The embodiments of the present invention include a bias circuit for a power-amplifying device that receives and amplifies an input RF signal having a series of RF cycles within a modulation envelope. The bias circuit compensates odd-order distortion processes by detecting the power in the input signal and providing a dynamic adjustment to a bias stimulus for the power-amplifying device within a time scale of the modulation envelope. The present invention is applicable in general to all reduced conduction angle amplifiers. In the examples described herein, class-B and class-AB amplifiers employing BJT or HBT power-amplifying devices are mainly discussed. For ease of discussion, the terms 'BJT' and 'class-AB' are used with the understanding that such terms encompass both 'BJT' and 'HBT' and both 'class-AB' and 'class-B', respectively, and with the assertion that the techniques described herein can be applied to other reduced conduction angle power amplifiers.

In one embodiment of the present invention, an amplifier circuit for amplifying an input RF signal has an input terminal for receiving the RF input signal, an output terminal for outputting an amplified RF signal, and an RF power-amplifying device having a base coupled to the input terminal and a collector coupled to the output terminal. The amplifier circuit further comprises a bias circuit coupled to the base of the RF power-amplifying transistor through at least one impedance element. The bias circuit is also coupled to the input terminal and configured to couple a portion of the RF input signal into the bias circuitry. The bias circuit thus produces a bias circuit gain that depends on the RF power level in the RF input signal and that compensates for voltage drops along a current path through the power amplifying device by raising the DC voltage at the base of the RF power-amplifying device in response to an increase in the RF power level in the RF input signal.

In one embodiment, the bias circuit includes a PN junction diode having first and second terminals, the first terminal being coupled to the base of the RF amplifying transistor through at least one impedance element, and the second terminal being coupled to a bypass capacitor, which is coupled to an RF ground in the amplifier circuit. The bypass capacitor therefore holds the second terminal at a constant

3

voltage despite fluctuations in the RF signal. As the RF power level in the RF input signal is increased, the average voltage across the PN junction diode is decreased, causing the base voltage of the RF power-amplifying device to increase. The PN junction diode may be the base-emitter (or base-collector) junction of a detecting transistor and the bypass capacitor is coupled to the base of the detecting transistor.

The RF amplifier circuit further comprises a reference voltage terminal for connecting to a reference voltage supply, a bias voltage terminal for connecting to a bias voltage supply, a first bias transistor having a collector coupled to the reference voltage supply through a reference resistor, an emitter coupled to a circuit ground terminal through a first emitter degeneration resistor, and a base, and a second bias transistor coupled with the detecting transistor in a current mirror configuration, the second bias transistor having a base connected to the base of the detecting transistor at a common connected base node, an emitter coupled to the base of the first bias transistor through a second emitter degeneration resistor, and a collector coupled to the bias voltage terminal. The bypass capacitor is coupled between the common connected base node and the circuit ground.

In some embodiments, the power-amplifying device includes at least one BJT device, and the bias circuit employs BJT components so that the current-mirror serves two functions simultaneously. One function is to provide a current-mirror or level-shifting operation required for setting and controlling the quiescent conditions for the BJT in the power-amplifying device. The second function is to respond to the modulation envelope in the input RF signal by rectifying each negative half cycle in the input RF signal. The charge accumulated in each negative half cycle is deposited in the bypass capacitor and a properly amplified and delayed version of this charge is returned back through the bias circuit to affect, on the time scale of the modulation, the base-to-emitter voltage of the RF amplifying transistor. When this dynamic bias action is adjusted correctly through proper design of the impedance elements, the embodiments of the present invention achieve a substantial reduction in odd-order intermodulation distortion over a useful power range (approximately 6-10 dB) while simultaneously providing the necessary bias isolation, matching, power/phase distribution, and ballasting over the power-amplifying device. In this way, the higher efficiency of reduced conduction angle operation is realized while the linearity requirements of the communication signal are satisfied.

The amplifier circuit may include an array of transistor cells in the power-amplifying device, with some or all of the other components of the amplifier circuit effectively distributed to provide proper RF phasing and ballasting across the array of transistor cells.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is a functional block diagram of a bipolar junction transistor (BJT) amplifier circuit employing a reduced conduction angle amplifier, such as class-AB, class-B or class-C amplifier, according to one embodiment of the present invention.

FIG. 2 is a chart illustrating the power gain, base-emitter voltage (V_{BE}) and collector current (I_{CC}) characteristics as a function of input RF signal power commonly associated with a class-AB amplifier.

FIG. 3 is a circuit schematic diagram illustrating a bias circuit in the BJT amplifier circuit according to one embodiment of the present invention.

4

FIG. 4 is a cross-sectional view of a bipolar junction transistor (BJT) in the amplifier circuit shown in FIG. 3 according to one embodiment of the present invention.

FIGS. 5(a)-5(f) are circuit schematic diagrams illustrating several exemplary embodiments for the impedance elements in the BJT amplifier circuit.

FIG. 6 is a circuit schematic diagram illustrating a modified current mirror bias network according to an alternative embodiment of the present invention.

FIG. 7 is a circuit schematic diagram illustrating a modified current mirror bias network according to yet another alternative embodiment of the present invention.

FIG. 8 is a circuit schematic diagram illustrating a distributed current mirror bias network for a physical BJT array according to embodiments of the present invention.

FIG. 9 is a circuit schematic diagram illustrating a distributed current mirror bias network for a physical BJT array according to alternative embodiments of the present invention.

FIG. 10 is a circuit schematic diagram illustrating a distributed current mirror bias network for a physical BJT array according to yet another embodiment of the present invention.

DETAILED DESCRIPTION

FIG. 1 depicts a simplified block diagram of a single-stage transistor amplifier network (amplifier) 100, according to one embodiment of the present invention. As shown in FIG. 1, amplifier 100 includes an RF power-amplifying device Q_{RF} having an input 101 and an output 102, a bias circuit 110 coupled to the input 101 of the power-amplifying device Q_{RF} , an input impedance matching section 120 also coupled to the input 101 of the power-amplifying device Q_{RF} through a blocking capacitor C_{IN} , an output bias and impedance matching section 130 coupled to the output 102 of the power-amplifying device Q_{RF} , and terminals RF_{IN}, RF_{OUT}, V_{REF} , V_{BIAS} , and V_{CC} .

Terminal RF_{IN} is coupled to the input impedance match section 120 for receiving an RF input signal, terminal RF_{OUT} is coupled to the output bias and impedance match section 130 for outputting an amplified RF signal, terminal V_{CC} is for connecting to a primary power supply voltage used to power the amplifier network 100 and is coupled to the output 102 of the power-amplifying device Q_{RF} through the output bias and impedance matching section 130, and terminals V_{REF} and V_{BIAS} are input terminals for the bias circuit 110. The input terminals V_{REF} and V_{BIAS} connect with a reference voltage V_{REF} and a bias voltage V_{BIAS} , respectively.

The reference voltage V_{REF} is a well-controlled voltage used to set a quiescent bias condition for the amplifier network 100. It is commonly derived from a specialized bandgap reference circuit to provide a consistent DC voltage over a specified range of current at the output 102 of power-amplifying device Q_{RF} and over a specified range of ambient temperature. The bias voltage V_{BIAS} is from a DC voltage source with sufficient voltage amplitude and current capacity to power the bias circuit 110 and feed the input 101 of the power-amplifying device Q_{RF} . In common battery-powered applications, V_{BIAS} is often connected to V_{CC} through a suitable network of decoupling capacitors. In high voltage applications associated with high power amplifiers, the bias voltage V_{BIAS} is often derived from V_{CC} but is regulated down to a lower voltage.

In one embodiment, the power-amplifying device Q_{RF} includes a bipolar junction transistor (BJT) having a base and a collector coupled to the input 101 and output 102 of

5

the power-amplifying device Q_{RF} , respectively, and the amplifier **100** further includes impedance elements Z_1 , Z_2 , and Z_3 . Impedance element Z_1 is coupled between the bias circuit **110** and the blocking capacitor C_{IN} , impedance element Z_2 is coupled between the impedance element Z_1 and the base of the BJT, and impedance element Z_3 is coupled between an emitter of the BJT and a circuit ground. Power-amplifying device Q_{RF} may include a single transistor device, such as a BJT, or a transistor cell array, as discussed below. In a non-limiting example, the BJT is suitable for connecting to a high voltage (e.g., 10-28 V) V_{CC} power supply can sustain a collector current I_{CC} up to a maximum collector current I_{Cmax} of about 100 mA.

Under small-signal operation, the bias circuit **110** provides the necessary bias current, I_{BIAS} , to the base of the power-amplifying device Q_{RF} , so as to set the quiescent bias condition of the amplifier network **100** while offering minimal RF loading to the base of the power-amplifying device Q_{RF} . The quiescent operating point is associated with a set of direct current (DC) operating conditions without any applied RF stimulus or input signal. In the case that the power-amplifying device Q_{RF} includes a BJT, the set of DC operating conditions include such parameters as quiescent collector current I_{CC} , quiescent base current I_B , which comes from the bias current I_{BIAS} , quiescent base-emitter voltage V_{BE} , etc., for the power-amplifying BJT device. In addition to providing the bias current I_{BIAS} for the power-amplifying device, the bias circuit **110** also controls the current and temperature compensation for the power-amplifying device.

The input impedance matching section **120** is designed to transform the impedance presented by the series combination of the blocking capacitor C_{IN} and the input **101** of the power-amplifying device Q_{RF} to the conjugate of a source impedance Z_S . The output bias and match section **130** serves two functions: one is to provide a low-loss path for a DC current to supply power to the power-amplifying device Q_{RF} . The second function is to provide a low-loss impedance transformation to maximize the transfer of RF power from the output **102** of the power-amplifying device Q_{RF} to a load impedance Z_L .

Under large-signal class-AB operation, the RF input power in the RF input signal received at the input terminal RF_{IN} reacts with the base-emitter junction of a BJT in the power-amplifying device Q_{RF} to draw additional current from the bias circuit **110**. This in turn draws more collector current I_{CC} at the collector of the BJT from the V_{CC} power supply. For a typical class-AB BJT in the power-amplifying device Q_{RF} , FIG. 2 demonstrates how the power gain $GAIN$, base-emitter voltage V_{BE} , and collector current I_{CC} of the BJT responds to varying levels of continuous-wave RF input power (P_{IN}) and includes $GAIN(P_{IN})$ curves **212**, **214**, and **216**, $V_{BE}(P_{IN})$ curves **222**, **224**, and **226**, and a $I_{CC}(P_{IN})$ curve. As discussed above, with increased RF input power P_{IN} , the interaction of the base-emitter junction of the BJT with the RF input power causes the bias circuit **110** to supply additional base current I_B . As the bias circuit **110** responds to supply more base current I_B , the finite resistance of bias circuit **110**, along with Z_1 , Z_2 , and Z_3 , causes the voltage V_{BE} across the base-emitter junction of the power-amplifying device Q_{RF} to decrease with increasing amount of RF input power P_{IN} , as shown by $V_{BE}(P_{IN})$ curves **222**, **224**, and **226**.

The characteristics of the $V_{BE}(P_{IN})$ curves, along with the initial quiescent and tuning conditions, would determine the nature of the amplifier's gain response $GAIN(P_{IN})$. In general, a $V_{BE}(P_{IN})$ function with decreasing V_{BE} at low input power, as shown by the curve **226**, would tend to exhibit a $GAIN(P_{IN})$ characteristic with early gain compression, as

6

shown by the $GAIN(P_{IN})$ curve **216**. A $V_{BE}(P_{IN})$ function that reasonably maintains V_{BE} over a wider input power range, as shown by curve **222**, would tend to exhibit flat gain or even gain expansion, as shown by curve **212**. Curves **224** and **214** demonstrate normal situations. The detailed nature of the $GAIN(P_{IN})$ characteristic and the dynamics of how this function responds to a modulated RF signal is closely related to the odd-order intermodulation distortion generated by the amplifier.

Under modulated RF drive conditions, a modulated RF signal is supplied to the input terminal RF_{IN} . A graphical depiction of a modulated RF signal **140** is also shown in FIG. 1 at the RF_{IN} terminal and indicates the time-varying nature of an RF envelope **142** along with a magnified view of a single RF cycle **144** in an RF carrier. Since the frequency of the modulation is small compared to the frequency of the RF carrier, the time scale for changes to the envelope is long compared to the RF cycle time. Given the relatively slowly varying envelope, a single RF cycle will exhibit a positive half cycle **146** that closely matches its negative half cycle **148**.

For a class-AB BJT in the power-amplifying device Q_{RF} under large-signal operations, the RF input signal **140** can be viewed as a sequence of positive-half-cycle and negative-half-cycle perturbations to the quiescent bias condition of the BJT. During the positive half-cycle, the RF input signal is sufficient to turn ON the base-emitter junction of the BJT and transfer charge from the blocking capacitor C_{IN} to the BJT with the flow of a first current component i_{POS} . During the negative half-cycle, the RF input signal is sufficient to turn OFF the base-emitter junction of the BJT, and in the absence of this conduction path, charge is transferred to the blocking capacitor C_{IN} from the bias circuit **110** through impedance element Z_1 with the flow of a second current component i_{NEG} . The time integration of these two current components creates an increasing DC base current $I_{B,TOT}$ with increasing RF input power and leads directly to the upward $I_{CC}(P_{IN})$ curve shown in FIG. 2. These two current components also interact with Z_1 , Z_2 , Z_3 and the bias circuit **110** to affect other characteristics such $V_{BE}(P_{IN})$, $GAIN(P_{IN})$ and the odd-order IMD of the amplifier circuit **100**.

Impedance elements such as Z_2 and/or Z_3 are often required for BJT power-amplifying devices because BJT operation at high power (e.g., 2-10 W) and high voltage (e.g., >10 V) requires substantial thermal ballasting in either or both of the base and emitter terminals of the power-amplifying device Q_{RF} . Unlike its FET counterpart, a BJT RF power-amplifying device often requires proper base impedance tuning at both the second harmonic frequency as well as the modulation frequency to achieve optimum linear performance. A simple input matching circuit **110** outside the bias current path is unable to satisfy these requirements for RF frequencies near 1 GHz and above. So, when a BJT is used as power-amplifying device Q_{RF} , the inclusion of Z_1 and/or Z_2 is often necessary to provide a combination of emitter and base ballasting with the equivalent effect of:

$$R_{ballast} \approx Z_{3,DC} + (Z_{1,DC} + Z_{2,DC})/\beta$$

where $Z_{1,DC}$, $Z_{2,DC}$, and $Z_{3,DC}$ are the DC resistance associated with the impedance elements of Z_1 , Z_2 , and Z_3 , respectively.

Having included the three impedance elements Z_1 , Z_2 , Z_3 to meet the stated electro-thermal requirements, it is found that these same elements also hinder the flow of i_{NEG} and i_{POS} under large signal operation and cause $V_{BE}(P_{IN})$ to drop prematurely with increasing input power P_{IN} . This leads to

poor large-signal RF power and linearity performance as depicted by curves **216** and **226** in FIG. 2. The present invention seeks to remedy this problem by constructing the bias circuit **110** that interacts cooperatively with the blocking capacitor C_{IN} , the power-amplifying device Q_{RF} , and the three impedance elements Z_1 , Z_2 , and Z_3 to simultaneously satisfy the electrical and thermal requirements for stable, efficient, and linear class-AB BJT amplifier operation.

In the following descriptions, like reference characters and symbols designate like or corresponding parts in the various embodiments. For example, the circuit element such as C_{IN} or Q_{RF} performs similar functions in the illustrated embodiments. It is understood that the exemplary illustrations are for the purpose of describing the preferred embodiments of the present invention and are not intended to limit the invention thereto.

In one embodiment of the present invention, as illustrated in FIG. 3, the bias circuit **110** in the amplifier network **100** includes bias transistors Q_1 and Q_2 , two emitter degeneration resistors R_{E1} and R_{E2} associated with the bias transistors Q_1 and Q_2 , respectively, a reference resistor R_{REF} , a buffer transistor Q_4 , a detecting transistor Q_{DET} , and a bypass capacitor C_{DET} associated with the detecting transistor Q_{DET} . The bias circuit **110**, the power-amplifying device Q_{RF} , and the impedance elements Z_1 , Z_2 , and Z_3 together form a current mirror with the bases of transistor Q_4 and Q_{DET} connected to a common connected-base node **310**, transistor Q_1 coupled with transistor Q_2 in a Darlington configuration with the emitter of transistor Q_2 coupled to the base of transistor Q_1 , and power-amplifying device Q_{RF} coupled with transistor Q_{DET} also in a Darlington configuration with the emitter of transistor Q_{DET} coupled to the input **101** of power-amplifying device Q_{RF} . Thus, DC conducting circuit elements in the amplifier network **100** to the left of the connected-base node **310**, such as transistors Q_1 and Q_2 , and resistors R_{E1} and R_{E2} , form a reference branch **330** of the current mirror, and DC conducting elements to the right of the base node **310**, such as transistors Q_{RF} and Q_{DET} , and DC conducting elements in the impedance elements Z_1 , Z_2 , Z_3 , form an amplifier branch **340** of the current mirror.

Transistor Q_4 is a buffer transistor for the current mirror and has its base coupled to the V_{REF} input via the reference resistor R_{REF} , its emitter coupled to the connected-base node **310**, and its collector coupled to the V_{BIAS} input. Transistor Q_{DET} has its collector coupled to the V_{BIAS} input and its emitter coupled to the input **101** of the power amplifying device Q_{RF} through impedance elements Z_1 and Z_2 . Transistor Q_2 has its collector coupled to the V_{BIAS} input and its emitter coupled to the base of transistor Q_1 via the resistor R_{E2} . Transistor Q_1 has its emitter coupled to the circuit ground through resistor R_{E1} and its collector coupled to the V_{REF} input through resistor R_{REF} . Bypass capacitor C_{DET} is coupled between the base of transistor Q_{DET} and an RF ground **312**, which can be a circuit ground, or a DC voltage terminal, such as the V_{REF} , V_{BIAS} , or V_{CC} terminal.

In one embodiment, each of the four transistors Q_1 , Q_2 , Q_4 , and Q_{DET} is a BJT formed on a same semiconductor die with the power-amplifying device Q_{RF} , and shares a common DC current gain factor, $\beta \gg 1$ with the BJT in the power-amplifying device Q_{RF} . FIG. 4 is a simplified block diagram of a cross-sectional view of a BJT **400** that can be used as the transistor Q_1 , Q_2 , Q_4 , Q_{DET} , or the BJT in the power-amplifying device Q_{RF} . As shown in FIG. 4, BJT **400** includes an emitter **410** formed over an emitter layer **412**, which is formed over a base layer **422**, on which is also formed a base **420**. The base layer **422** is formed over a collector layer **432**, which is formed over a subcollector

layer **434**, on which is also formed a collector **430**. The subcollector layer is formed near a top surface **452** of a substrate **450**.

The BJT **400** can be fabricated using any of the BJT technologies in the art. In one embodiment of the present invention, the BJT **400** is fabricated using a Gallium Arsenide HBT semiconductor processing technology, and the emitter **410** includes a N+ InGaAs contact over a N+ InGaAs graded cap layer over a N+ GaAs cap contact layer. The emitter layer **412** includes a N-type AlGaAs or InGaP layer, the base layer **422** includes a P+ GaAs layer, the collector layer **432** includes a N- GaAs layer, the subcollector layer **434** includes a N+ GaAs layer, and the substrate **450** is a GaAs substrate. The BJT **400** may also be fabricated using an Indium Phosphide (InP) semiconductor process technology with an InP layer in the emitter layer **412**, the base layer **422**, and/or the collector layer **432**. In another embodiment, the BJT **400** is fabricated using a SiGe semiconductor processing technology with SiGe in the emitter layer **412**, the base layer **422**, and/or the collector layer **432**.

Because of the current mirror configuration in the bias circuit **110**, as shown in FIG. 3, the bias current $I_{B,TOT}$ drawn by the power-amplifying device Q_{RF} from the bias circuit **110** is insensitive to variations in the current gain β of transistor Q_1 and current gain β_{RF} of power-amplifying device Q_{RF} , so long as β and β_{RF} track one another, a simple requirement that is not difficult to achieve for transistor devices manufactured on a same semiconductor die. The quiescent collector current of transistor Q_{RF} can be designed as a scaled ratio of the quiescent collector current of Q_1 by adjusting an effective emitter area ratio associated with the two transistors, again a simple requirement for devices on a same semiconductor die. The quiescent base-emitter voltage V_{BE1} of transistor Q_1 sets the quiescent base-emitter voltage V_{BE} of transistor Q_{RF} and can provide a nearly constant V_{BE} over increasing RF_{IN} power up to a certain point, as illustrated by the curves **222**, **224**, and **226** in FIG. 2.

The buffer transistor Q_4 amplifies and buffers the reference current through the resistor R_{REF} and effectively decouples the bias current from any limitations of V_{REF} and R_{REF} while maintaining the benefits of the current mirror configuration, as described above.

In one embodiment, transistors Q_1 and Q_2 in the reference branch **330** of the current-mirror serve as a scaled-down version of transistors Q_{RF} and Q_{DET} , respectively, in the amplifier-branch **340** of the current mirror. In one embodiment, the effective emitter areas of transistors Q_{RF} and Q_{DET} are N-times as large as the effective emitter areas of transistors Q_1 and Q_2 , respectively, where N is a scaling factor. The scaling factor N is generally chosen to be large such that the current drawn through transistor Q_1 is small compared to the quiescent collector current of the power-amplifying device Q_{RF} . If K is the factor by which the collection current I_{CC} of the power-amplifying device Q_{RF} would increase over the quiescent collector current I_{CQ} to support a required RF power range, then N should preferably satisfy the inequality, $NK \ll \beta^2$, to ensure that the reference branch of the current mirror is not starved for current during peak RF envelope power excursions. In one embodiment, the emitter area of buffer transistor Q_4 is chosen so as not to limit the current into the base of transistor Q_{DET} during peak RF operating conditions.

Resistor R_{REF} is designed to set the quiescent collector current of transistor Q_{RF} , which is approximately given by,

$$I_{CQ} \cong \left(\frac{V_{REF} - V_{BE1} - V_{BE2} - V_{BE4}}{R_{REF} + R_{E1}} \right) N$$

where V_{BE1} , V_{BE2} , and V_{BE4} are the forward voltage drops across the base-emitter junctions of Q_1 , Q_2 and Q_4 , respectively, and the voltage drop across R_{E2} is assumed to be small. For a given BJT process technology, the magnitude of the voltage V_{BE} across the base-emitter junction of power-amplifying device Q_{RF} is relatively insensitive to the forward current and is typically in the range of 0.9-1.0V when power-amplifying device Q_{RF} includes a Si-based BJT, or in the range of 1.3-1.4V when power-amplifying device Q_{RF} includes a GaAs-based HBT.

Resistors R_{E1} and R_{E2} are generally chosen to mirror, in a scaled form, the DC resistance of Z_3 and (Z_1+Z_2) , respectively, such that

$$R_{E1} \cong N(Z_{3,DC}), \text{ and}$$

$$R_{E2} \cong N(Z_{1,DC} + Z_{2,DC}).$$

The design constraints pertaining to impedance elements Z_1 , Z_2 , and Z_3 are numerous. To ensure thermal stability for the transistor or transistor cell array comprised in Q_{RF} , an impedance sum of $Z_{Ballast} = (Z_3 + Z_2/\beta)$ should be sufficiently large at DC and up to a frequency roughly equal to $1/\tau_{th}$, where τ_{th} is a dominant thermal time constant for the power-amplifying device Q_{RF} . In the situation power-amplifying device Q_{RF} comprises a transistor array, the impedance sum $Z_{Ballast}$ may also include a contribution of Z_1/β from impedance element Z_1 , depending on the exact layout of the Q_{RF} transistor cell array. The sum of $Z_{Ballast} = (Z_3 + Z_2/\beta + Z_1/\beta)$ at DC is generally designed to be a minimum value required for thermal stability, as any excess resistance in the amplifier network **100** places increasing demands on the bias circuit **110** to compensate for the voltage drops along the bias current path through the input **101** of the power amplifying device Q_{RF} . At the RF carrier frequency, Z_1 , Z_2 , and Z_3 should cooperate with the input match **120** to provide low reflection at the RF_{IN} terminal. At the modulation frequency and the second harmonic frequency, Z_1 , Z_2 , and Z_3 are generally designed to provide low IMD distortion when the input signal power is relatively small.

As discussed above, the bias circuit **110** sets a bias current and controls the quiescent condition and temperature compensation for the power-amplifying device Q_{RF} . As the RF input power increases, the base-emitter junction of the power-amplifying device Q_{RF} begins to rectify as does the base-emitter junction of transistor Q_{DET} , and this sets up the sequence of positive-half-cycle and negative-half-cycle perturbations to the quiescent bias condition described above with reference to FIG. 1. In one embodiment of the present invention, a detector function as described below is provided by the detecting transistor Q_{DET} and bypass capacitor C_{DET} . The base-emitter junction of the detecting transistor Q_{DET} in the bias circuit **110** provides a PN junction diode **320** having a P-terminal coupled to the RF ground **312** through the bypass capacitor C_{DET} and an N-terminal coupled to the terminal RF_{IN} through the impedance element Z_1 , the blocking capacitor C_{IN} , and the input match section **120**. Because the P-terminal is held to a DC voltage through the bypass capacitor, the PN junction diode **320** is allowed to respond to the RF modulation envelope in the input RF signal by rectifying each negative half cycle in the input RF signal.

The charge accumulated in each negative half cycle is deposited in the bypass capacitor C_{DET} and a properly amplified and delayed version of this charge is returned back through the emitter of the detecting transistor Q_{DET} . The series combination of bypass capacitor C_{DET} , the base-emitter junction of the detecting transistor Q_{DET} , impedance element Z_1 , and blocking capacitor C_{IN} , as shown in FIG. 3, facilitates a charge-pump action to help drive this charge into the base of Q_{RF} , resulting, on an average, an effective bias circuit gain depending on the RF input power in the dynamic bias current $I_{B,TOT}$.

The effective bias circuit gain results in increasing voltage at the base of the power-amplifying transistor Q_{RF} with increasing RF power level in the RF input signal. The increasing voltage at the base of the power-amplifying transistor Q_{RF} offsets the increasing voltage drops across Z_1 , Z_2 , and Z_3 as the RF power level increases. Thus, the bias circuit **110** dynamically modifies, on the time scale of the modulation in the RF input signal, the base-to-emitter voltage of the RF amplifying transistor Q_{RF} . When this dynamic bias action is adjusted correctly through proper design of the three impedance elements, the embodiments of the present invention achieve a much improved $V_{BE}(P_{IN})$ curve and, as a result, a substantial reduction in odd-order intermodulation distortion over a useful input power range (approximately 6-10 dB), while at the same time providing the necessary bias isolation, matching, power/phase distribution, and ballasting for the power-amplifying device Q_{RF} . In this way, the higher efficiency of reduced conduction angle operation is realized while the linearity requirements of the communication signal is satisfied.

The charge-pump action described above should be slow enough to smooth the individual RF cycle peaks but fast enough to follow the modulation envelope. These requirements can be used to set upper and lower bounds for the values of C_{DET} . As a non-limiting example, the upper and lower bounds of C_{DET} are set to be about

$$\frac{1}{2\pi Z_{BASE} F_{MOD}} \text{ and } \frac{1}{2\pi Z_{BASE} F_{RF}},$$

respectively, where F_{MOD} is the highest modulation frequency in the RF input signal **140**, F_{RF} is the RF carrier frequency in the RF input signal **140**, and Z_{BASE} is the magnitude of the impedance at the base node of detecting transistor Q_{DET} when detecting capacitor C_{DET} is removed from the circuit.

FIGS. 5(a)-5(f) are a schematic diagrams illustrating exemplary embodiments of the impedance elements Z_1 , Z_2 , and Z_3 as two-terminal networks for the sake of illustration. At frequencies near 1 GHz and above, these elements are more accurately described as two-port impedance networks interacting with a circuit ground plane. Elements Z_2 and Z_3 should be placed in close physical proximity with transistors Q_{RF} and Q_{DET} and are typically integrated within the same semiconductor die as the transistors. As such, they may be constrained by the integrated circuit fabrication technology and limited to such networks depicted in FIGS. 5(a), 5(b), and 5(c) with the capacitance C typically less than 100 pF, or limited to such networks depicted in 5(d), 5(e), and 5(f) for modest values of inductance L typically less than 5 nH. Element Z_1 is generally less constrained and can take any of the forms shown in FIGS. 5(a) through 5(f), or another form as needed to satisfy the electrical and thermal requirements described above with reference to FIG. 3. Z_1 can be realized

11

within the same semiconductor die as the transistors, outside the semiconductor die, or as some combination thereof.

The sensitivity of the PN junction diode **320**, and thus the sensitivity of the effective bias circuit gain, to the changes in the RF input power, can be adjusted by adjusting the amount of RF impedance in the impedance elements Z_1 . For a given Z_1 , the rectifying effect of the PN junction diode **320** in the detecting transistor Q_{DET} is about the same, whether Z_1 is lumped or distributed among the transistor cells in the power-amplifying device Q_{RF} , if power-amplifying device Q_{RF} includes an array of transistor cells.

In a non-limiting example, Z_2 is not required, Z_1 takes the form shown in FIG. 5(a) with R in the range of about 15Ω to about 100Ω, and Z_3 also takes the form shown in FIG. 5(a) with R in the range of about 1Ω to about 6Ω. In another example, Z_2 is not required, Z_3 takes the form shown in FIG. 5(a) with R in the range of about 0.5Ω to about 3Ω, and Z_1 takes the form shown in FIG. 5(c) with R_1 in the range of about 15Ω to about 100Ω, R_2 in the range of about 50Ω to about 300Ω, and C around 10 pF. In a further example, Z_1 takes the form shown in FIG. 5(a) with R in the range of about 15Ω to about 100Ω, Z_2 takes the form shown in FIG. 5(b) with R in the range of about 50Ω to about 300Ω and C around 10 pF, and Z_3 takes the form shown in FIG. 5(a) with R in the range of about 0.5Ω to about 3 Ω.

The circuit topology and design guidelines described above should provide a dynamic bias circuit gain sufficient to overcome, to a first order, the voltage drops across Z_1 , Z_2 , and Z_3 thus leading to a desirable $V_{BE}(P_{IN})$ characteristics and linear power gain characteristic, as depicted by the GAIN(P_{IN}) curves **222** and **212** in FIG. 2. Computer simulation can be used to provide a more detailed circuit analysis and to further refine the design of the amplifier network **100** to satisfy the linearity requirements of a particular communication signal or system.

In an alternative embodiment, the bias circuit **110** lacks the buffer transistor Q_4 , as shown in FIG. 6, and is generally suited to applications requiring a smaller amplifying transistor Q_{RF} , or applications having a lower reference voltage, V_{REF} , which is unable to support the voltage drops across the base-emitter junctions of all of the three transistors Q_4 , Q_{DET} , and Q_{RF} , as shown in FIG. 3. Without the buffer transistor Q_4 , the base and collector of transistor Q_2 are tied and are coupled to the V_{REF} terminal through the resistor R_{REF} . FIG. 6 also illustrates an example of a negative feedback network **610** including a capacitor C_F and a resistor R_F serially coupled with each other between the V_{BIAS} terminal and the connected bases of transistors Q_2 and Q_{DET} , and is designed to suppress electrical oscillations within the bias circuit **110**.

In yet another embodiment, the bias circuit **110** includes a transistor-resistor tier **710** including a resistor R_{E1} , a transistor Q_1 , a resistor R_{E2} , a transistor Q_2 , a resistor R_{E3} , and a transistor Q_3 serially connected with each other between R_{REF} and a circuit ground. Transistors Q_1 , Q_2 , and Q_3 each has its base and collector tied. The bias circuit **110** in FIG. 7 includes an additional transistor Q_5 coupled with the transistor Q_3 in a current mirror arrangement with their bases connected to a common connected-base node **720**. The buffer transistor Q_4 is absent and the negative feedback network **610** is provided between the base and collector of transistor Q_5 . The DC conducting circuit elements in FIG. 7 to the left of the connected-base node **720** form a reference branch **722** of the current mirror, while DC conducting elements to the right of the base node **720** form an amplifier branch **724** of the current mirror. In one embodiment, the reference branch **722** includes transistors Q_1 , Q_2 , and Q_3 ,

12

and the amplifier branch **724** includes transistors Q_{RF} , Q_{DET} , and Q_5 . As in the previously described embodiments, transistors Q_1 , Q_2 , and Q_3 in the reference branch **722** of the current-mirror serve as a scaled-down version of transistors Q_{RF} , Q_{DET} , and Q_5 , respectively, in the amplifier-branch **724** of the current mirror.

The detector function in FIG. 7 is provided by the detecting transistor Q_{DET} , which is configured in the form of a level-shifting diode with its collector and base tied and connected to the emitter of transistor Q_5 . As in all the previously described embodiments, the bypass capacitor C_{DET} is coupled between the base of the detecting transistor Q_{DET} and the circuit ground. The embodiment shown in FIG. 7 offers more flexibility in terms of temperature compensation via transistors Q_1 and Q_2 and their respective emitter resistors R_{E1} and R_{E2} . Charge drawn from the bypass capacitor C_{DET} down through the detecting transistor Q_{DET} in the amplifier network **100** shown in FIG. 7, however, lacks the current gain exhibited by the embodiments shown in FIGS. 3 and 6. But the basic operating principles of the bias circuit **110** are the same in all of the embodiments described above.

FIG. 8 is a schematic diagram illustrating an amplifying network **800** according to yet another embodiment of the present invention. As shown in FIG. 8, amplifying network comprises an array of n rows by m columns of transistor cells with a total of (m)(n) transistor cells each labeled as Q_{RF} . Such transistor cell arrays are common in modern semiconductor RF power devices. The transistor cell array may be large enough to require the three impedance elements Z_1 , Z_2 , Z_3 and blocking capacitor C_{IN} to be distributed among the transistor cells to satisfy the RF phase and thermal stability requirements. The detecting transistor Q_{DET} and its associated bypass capacitor C_{DET} are generally required to be placed in close proximity with the amplifying transistor cells, so these elements are distributed across the m columns in the transistor array but can be lumped within each of the m columns, as shown in FIG. 8. A resistor R_{STAB} is often included in the bias circuit and is coupled between the emitters of two adjacent detecting transistors Q_{DET} to suppress odd-mode oscillations therein. The values of the distribution of parallel elements shown in FIG. 8 are scaled from the respective lumped circuit element values described in FIG. 6, as listed in Table 1.

TABLE 1

(I) Lumped Circuit Element value in FIG. 6	(II) Equivalent Distributed Circuit Element Value in FIG. 8	Change from (I) to (II)
R_{REF}	R_{REF}	Unchanged
R_{E1}	R_{E1}	Unchanged
R_{E2}	R_{E2}	Unchanged
(Q_1 Area)	(Q_1 Area)	Unchanged
(Q_2 Area)	(Q_2 Area)	Unchanged
(Q_{DET} Area)	(Q_{DET} Area)/m	Scale Down by m
C_{DET}	(C_{DET})/m	Scale Down by m
C_{IN}	$C_{IN}/(mn)$	Scale Down by mn
Z_1	$Z_1(mn)$	Scale Up by mn
Z_2	$Z_2(mn)$	Scale Up by mn
Z_3	$Z_3(mn)$	Scale Up by mn
(Q_{RF} Area)	(Q_{RF} Area)/(mn)	Scale Down by mn

The input matching section **220** and output bias and matching section **230** need not be changed in going from the lumped circuit depicted in FIG. 6 to the distributed circuit depicted in FIG. 8. It is good RF design practice, however, to match the input and output phase delays to the individual transistor cells Q_{RF} to maximize the gain and power com-

13

binning efficiency of the transistor cell array. Such standard layout practices are common and familiar to those skilled in the art of integrated-circuit power amplifiers and are not explicitly shown in FIG. 8.

FIG. 9 is a schematic diagram illustrating another embodiment of a distributed amplifier network **900** wherein the transistor column length n is short enough and the impedance element Z_1 is low enough to allow Z_1 to be lumped within each transistor column like the Q_{DET} and C_{DET} elements. Thus, amplifier network **900** shown in FIG. 9 is preferred where possible to amplifier network **800** shown in FIG. 8 because a simpler layout is provided in the amplifier network **900**. The scaling rules for scaling the values of relevant circuit elements in network **900** from the values of the corresponding lumped circuit depicted in FIG. 6 are similar those for scaling the relevant circuit elements in network **800**, as listed in Table 1, except that Z_1 is scaled up by a factor of m instead of mn , and the capacitance C_{IN} is unchanged relative to that in FIG. 6.

FIG. 10 is a schematic diagram illustrating yet another embodiment of a distributed amplifier network **1000** wherein the transistor column length, n , is short enough and the impedance elements Z and Z_2 are low enough to allow both Z_1 and Z_2 to be lumped within each column like the Q_{DET} and C_{DET} elements. Thus, amplifier network **1000** shown in FIG. 10 is preferred where possible to amplifier network **800** or **900** shown in FIG. 8 or 9, respectively, because a simpler layout is provided in the amplifier network **1000**. The scaling rules for scaling the values of relevant circuit elements in network **1000** from the values of the corresponding lumped circuit depicted in FIG. 6 are similar those for scaling the relevant circuit elements in network **800**, as listed in Table 1, except that both Z_1 and Z_2 are scaled up by a factor of m instead of mn , and the capacitance C_{IN} is unchanged relative to that in FIG. 6. With both Z_1 and Z_2 lumped, however, Z_3 needs to provide the full amount of ballasting.

The foregoing discussion and supporting illustrations are for exemplary purposes only and should not be construed as limiting the present invention. Substantial variations and combinations of the exemplary bias circuits may be practiced without departing from the spirit and scope of the present invention. For example, the embodiments described herein may be modified to include additional circuit elements to affect the stability and frequency response or to adjust the temperature or voltage compensation. Those skilled in the art will readily appreciate that these and other similar modifications are well within the scope of the present invention.

We claim:

1. A radio frequency (RF) amplifier circuit having an input terminal for receiving a RF input signal, comprising:

- a RF power-amplifying transistor having a base coupled to the input terminal;
- a bias circuit coupled to the input terminal and to the RF power-amplifying transistor, the bias circuit including a bypass capacitor and a PN junction diode having first and second terminals, the first terminal being coupled to the base of the RF amplifying transistor through at least one impedance element, the second terminal is coupled to a RF ground in the RF amplifier circuit through the bypass capacitor

wherein the bias circuit is configured to raise a voltage at the base of the RF power-amplifying transistor in response to an increase in an RF power level in the RF input signal; and

14

the PN junction diode is part of a detecting transistor having a base and a collector, the RF amplifier circuit further comprising:

- a reference voltage terminal for connecting to a reference voltage supply;
- a bias voltage terminal for connecting to a bias voltage supply;
- a first bias transistor having a collector coupled to the reference voltage supply through a reference resistor, an emitter coupled to a circuit ground terminal through a first emitter degeneration resistor, and a base; and
- a second bias transistor coupled with the detecting transistor in a current mirror configuration, the second bias transistor having a base connected to the base of the detecting transistor at a common connected base node, an emitter coupled to the base of the first bias transistor through a second emitter degeneration resistor, and a collector coupled to the bias voltage terminal.

2. The RF amplifier circuit of claim 1, wherein each of the at least one impedance element includes a two-port impedance network.

3. The RF amplifier circuit of claim 1, wherein the PN junction diode is part of a detecting transistor having a base and a collector that are tied to each other.

4. The RF amplifier circuit of claim 1, wherein the bypass capacitor is coupled between the common connected base node and the circuit ground.

5. The RF amplifier circuit of claim 1, further comprising a buffer transistor having a base coupled to the reference voltage terminal through the reference resistor, an emitter coupled to the common connected base node, and a collector coupled to the bias voltage terminal.

6. The RF amplifier circuit of claim 1, wherein the PN junction diode is part of a detecting transistor having a base and a collector, the RF amplifier circuit further comprising:

- a reference voltage terminal for connecting to a reference voltage supply;
- a first bias transistor having an emitter coupled to a circuit ground terminal through a first emitter degeneration resistor, and a base and a collector tied with each other; and
- a second bias transistor coupled with the detecting transistor in a current mirror configuration, the second bias transistor having a base connected to the base of the detecting transistor at a common connected base node, an emitter coupled to the tied base and connector of the first bias transistor through a second emitter degeneration resistor, and a collector tied to the base of the second bias transistor and coupled to the reference voltage terminal via a reference resistor.

7. The RF amplifier circuit of claim 6, further comprising a bias voltage terminal for connecting to a bias voltage supply, wherein the detecting transistor has a collector coupled to the bias voltage terminal; and

a negative feedback network coupled between the collector of the detecting transistor and the common connected base node, the negative feedback network including a resistor and a capacitor serially connected with each other.

8. The RF amplifier circuit of claim 1, further comprising: a reference voltage terminal for connecting to a reference voltage supply;

- a tier of serially connected level-shifting diodes coupled between the reference voltage terminal and a circuit ground;
- a bias transistor having an emitter coupled to the PN junction diode, a base coupled to the tier of serially

15

connected level-shifting diodes, and a collector coupled to the base of the bias transistor through a negative feedback network.

9. The RF amplifier circuit of claim 8, wherein the tier of serially connected level-shifting diodes includes first, second and third bias transistors and first, second, and third

16

emitter degeneration resistors, each of the first, second and third bias transistors having a base and a collector tied with each other and an emitter coupled to respective ones of the first, second, and third emitter degeneration resistors.

* * * * *