On the Safe Operating Area of Bipolar Cascode Amplifiers

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Abstract—This paper presents an investigation of the safe-operating-area boundary of bipolar cascode amplifiers in GaAs and SiGe technologies. A simple relation is derived to predict the instability onset due to electrothermal and avalanche effects. Circuit simulations and experiments performed on GaAs test structures are employed to analyze the influence of thermal coupling between transistors and the beneficial impact of base ballasting.

Keywords—bipolar transistor; cascode amplifier; common-emitter amplifier; impact ionization; safe operating area; thermal resistance

I. INTRODUCTION

The bipolar cascode amplifier is a two-transistor configuration composed by the series-connection of a common-emitter (CE) stage followed by a common-base (CB) current-buffer stage, as shown in Fig. 1. Compared to the standard single-device CE amplifier, the cascode exhibits improved performance due to the higher output resistance, increased output swing, and larger bandwidth resulting from smaller Miller effect; these benefits come at the cost of (i) a second device – which is not dramatic in ICs – and (ii) a high supply voltage due to the double collector-emitter drop. Nevertheless, bipolar cascode stages and variants are attractive for use in a large variety of applications, e.g., bandgap and reference circuits, high-linearity RF and instrumentation amplifiers, LNAs, as well as high-speed mixed-signal circuitry and drivers [1]–[3].

In spite of the popularity of this circuit topology, the investigation of its operating limits has received little attention in literature, the exceptions being represented by [4], [5]. However, it is well-known that state-of-the-art bipolar technologies suffer from exacerbated positive-feedback mechanisms, i.e., (i) electrothermal (ET) effects, steadily growing due to transistor scaling, poor material thermal conductivity (GaAs), as well as use of insulating oxide-based schemes (SiGe), and (ii) impact-ionization (II) mechanisms arising at low output voltages (SiGe).

This work is aimed to analyze the influence of ET and II effects on the safe operating area (SOA) of both GaAs and SiGe cascode amplifiers. A simple formulation is derived to predict the instability onset locus and its dependence on all physical parameters playing a role. Calibrated circuit simulations and experiments performed on GaAs test structures are used to explore the influence of thermal coupling between the lower and upper transistor, as well as the impact of base ballasting. Furthermore, the SOA boundary of cascode amplifiers is compared to that of their single-stage CE counterparts.

Fig. 1. Single-device (left) and cascode (right) bipolar amplifiers.

II. FLYBACK LOCUS

This section is devoted to the derivation of an analytical formulation for the SOA boundary of a cascode amplifier, intended as the collection of flyback points in the V_B -constant output I - V curves, at which a negative differential resistance branch takes place due to the interplay between ET and II effects.

With reference to Fig. 1, the collector current I_{C1} of transistor Q_1 can be expressed as follows:

$$I_{C1} = M_1 \cdot A_{E1} \cdot J_S \cdot \left(1 + \frac{V_{CE1}}{V_A} \right) \cdot \exp \left(\frac{V_{BEj1} + \phi \cdot \Delta T_{j1}}{\eta \cdot V_{T0}} \right) \quad (1)$$

where A_{E1} is the effective emitter area [μm^2], J_S is the reverse saturation current density [$\text{A}/\mu\text{m}^2$], M_1 is the V_{CB1} -dependent avalanche multiplication factor, V_A is the Early voltage [V], V_{BEj1} is the internal (junction) base-emitter voltage [V], ϕ is the absolute value of the temperature coefficient of V_{BEj1} [V/K], $\Delta T_{j1} = T_{j1} - T_{AMB}$ is the temperature rise above ambient of the base-emitter junction [K], η is the ideality factor, and V_{T0} is the thermal voltage at ambient temperature [V].

Voltage V_{BEj1} can be expressed as [6]

$$V_{BEj1} = V_B - R_{B1} \cdot I_{B1} - R_{E1} \cdot I_{E1} = V_B - R_{EBeq1} \cdot I_{C1} \quad (2)$$

where

$$R_{EBeq1} = R_{B1} \cdot \left(\frac{1}{M_1 \cdot \alpha_{F1}} - 1 \right) + \frac{R_{E1}}{M_1 \cdot \alpha_{F1}} \quad (3)$$

α_{F1} being the common-base current gain. The temperature rise ΔT_{j1} is given by

$$\Delta T_{j1} = R_{TH1} \cdot P_{D1} + R_M \cdot P_{D2} \quad (4)$$

where R_{TH1} and R_M are the self-heating thermal resistance of device Q_1 and the mutual thermal resistance between Q_1 and Q_2 , respectively, while $P_{D1} \approx V_{CE1} \cdot I_{C1}$ and $P_{D2} \approx V_{CE2} \cdot I_{OUT}$ are the

powers dissipated by Q_1 and Q_2 . The output current can be expressed as

$$I_{OUT} = M_2 \cdot \alpha_{F2} \cdot I_{E2} = M_2 \cdot \alpha_{F2} \cdot I_{C1} \quad (5)$$

By making use of (2)-(5), (1) can be recast as

$$I_{C1} = M_1 \cdot A_{E1} \cdot J_S \cdot \left(1 + \frac{V_{CE1}}{V_A}\right) \cdot \exp\left(\frac{V_B}{\eta \cdot V_{T0}}\right) \cdot \exp\left[I_{C1} \cdot \frac{\phi \cdot (R_{TH1} \cdot V_{CE1} + R_M \cdot V_{CE2} \cdot M_2 \cdot \alpha_{F2}) - R_{EBeq1}}{\eta \cdot V_{T0}}\right] \quad (6)$$

By assuming that V_{BE2} is constant, also $V_{CE1} = V_{B2} - V_{BE2}$ and $V_{CB1} = V_{CE1} - V_{B1}$ are constant, and the RHS can be reviewed as a function $f(I_{C1}, V_{CE2})$. By differentiating (6), it is obtained that

$$\frac{dI_{C1}}{dV_{CE2}} = \frac{\partial f}{\partial V_{CE2}} \bigg|_{I_{C1}} \cdot \left(1 - \frac{\partial f}{\partial I_{C1}} \bigg|_{V_{CE2}}\right)^{-1} \quad (7)$$

Consequently, the flyback condition in the (V_{CE2}, I_{C1}) plane is

$$\frac{dI_{C1}}{dV_{CE2}} \rightarrow \infty \Rightarrow \frac{\partial f}{\partial I_{C1}} \bigg|_{V_{CE2}} = 1 \quad (8)$$

After simple algebra, the flyback locus can be derived as

$$I_{C1} = (1/I_{C1SH} + 1/I_{C1M})^{-1} \quad (9)$$

where

$$I_{C1SH} = \frac{\eta \cdot V_{T0}}{\phi \cdot R_{TH1} \cdot V_{CE1} - R_{EBeq1}} \quad I_{C1M} = \frac{\eta \cdot V_{T0}}{\phi \cdot R_M \cdot V_{CE2} \cdot M_2 \cdot \alpha_{F2}} \quad (10)$$

I_{C1SH} and I_{C1M} being related to the self-heating of device Q_1 and to the thermal interaction between Q_1 and Q_2 , respectively. Equations (9) and (10) represent the flyback locus (i.e., the SOA boundary) of Q_1 in the (V_{CE2}, I_{C1}) plane (V_{CE1} is assumed constant). It is worth noting that the SOA shrinks by increasing R_M , that is, the thermal coupling has a deleterious impact, contrarily to what happens in other basic two-transistor circuits, like simple current mirrors [7] and differential pairs [8]. The flyback locus in the (V_{OUT}, I_{OUT}) plane can be obtained from (9)-(10) by replacing V_{CE2} with $V_{OUT} - V_{CE1}$ and multiplying both sides by $M_2 \cdot \alpha_{F2}$, where factor M_2 depends upon $V_{CB2} = V_{OUT} - V_{CE1} - V_{BE2}$. Lastly, it must be remarked that, in the absence of II effects and assuming $\alpha_{F2} = 1$, in the extreme case of $R_M \rightarrow R_{TH1}$ the flyback locus would coincide with that of a single device induced by self-heating only.

III. EXPERIMENTAL MATERIAL AND SIMULATION APPROACH

A. Experimental material

A mixed-signal InGaP/GaAs HBT process was made available by Skyworks Solutions; sample dies thinned to 100 μm were affixed to a bulk Si wafer with a thermally conductive epoxy. The technology is optimized for mobile handset applications, and features maximum $f_T/f_{\max} = 30/40$ GHz, BV_{CBO} of about 27 V, minimum emitter size of $2 \times 2 \mu\text{m}^2$, double non-self-aligned base contact, CE current gain β_F of about 100, three layers of interconnect metal, integrated Schottky diodes, and MIM capacitors. All HBTs investigated in this study have two emitter fingers, each with $3.4 \times 17.8 \mu\text{m}^2$ area ($A_E = 121 \mu\text{m}^2$). Various cascode structures with identical transistors Q_1 and Q_2 were manufactured by varying the edge-to-edge spacing between

the HBTs from 6.8 μm to 27.2 μm with the aim of obtaining different levels of thermal coupling; Fig. 2 (top) shows a top-view image of the circuit with closest devices. For each cascode amplifier, a corresponding “reference” structure with separate (i.e., independently accessible) Q_1 and Q_2 was also fabricated, as illustrated in Fig. 2 (bottom), in order to estimate the key transistor parameters and the thermal resistances.

On-wafer measurements were performed by feeding DC voltages through 26 GHz RF probes, the stability being ensured by 50 Ω RF terminated bias tees. In particular, the output voltages (V_{CE} and V_{OUT}) were swept at a constant V_B for both the single HBT and the cascode amplifier by assigning proper limits (raised with increasing V_B) on the output currents (I_C and I_{OUT}) not to damage the parts due to ET/II-induced current runaway.

Fig. 2. Photograph of a cascode amplifier (top) and corresponding reference structure (bottom), along with the related schematics (left and right, respectively).

B. Simulation approach

Simulations were carried out in the environment of the popular PSPICE tool [9] using the *analog behavioral modeling* technique: the standard device is replaced by a subcircuit, which, besides the conventional electrodes, is equipped with thermal (input) and power (output) nodes, and includes a transistor macromodel that (i) allows the variation of all temperature-sensitive parameters (e.g., V_{BEj} and β_F) during the simulation run, (ii) accounts for all relevant effects, like impact ionization (the avalanche multiplication factor is described by the formula presented in [10]) and parasitic resistances, and (iii) relies on simple physics-based formulations, the parameters of which can be easily tailored by comparison with Gummel and output characteristics measured at various thermochuck (baseplate) temperatures T_B . In particular – differently from the first-order transistor model exploited in section II – various negative-feedback mechanisms are enabled, namely, the logarithmic reduction of parameter ϕ by increasing collector current [11]–[13], the negative temperature coefficient of gain β_F , and high-injection effects. The temperatures are determined from the dissipated powers by connecting to the subcircuit(s) a thermal feedback block, which for the cascode amplifier includes the self-heating thermal resistances of Q_1 and Q_2 and the mutual resistance R_M .

The choice of this approach was motivated by the following considerations: (a) PSPICE simulations allow an accurate detection of the SOA boundary of the parts under analysis without the need of potentially destructive experiments; in particular, the flyback behavior – which would correspond to a thermal runaway onset in a voltage-controlled measurement – is determined by sweeping the output current and evaluating the

voltage; (b) the key voltages, currents, and temperatures can be easily monitored, thus favoring an in-depth analysis of the circuit behavior; moreover, simulations are very fast: less than 1 s is required to obtain an I-V curve.

IV. RESULTS AND DISCUSSION

A. GaAs technology

The thermal resistances were determined by measurements performed on the reference structures. In particular, the self-heating resistance R_{TH} ($R_{TH1}=R_{TH2}=R_{TH}$ since the devices are identical) was estimated to be nearly 540 K/W by applying the classic approach proposed in [14] to one of the HBTs, while the mutual resistances R_M associated to each spacing were assessed by exploiting transistor Q_2 as a temperature-sensing diode through the following procedure: (i) Q_2 was biased under open-collector conditions with a low I_{B2} ($=I_{E2}=2 \mu A$) to avoid self-heating ($T_{j2} \approx T_B$); (ii) the temperature coefficient ϕ of this HBT was calibrated by measuring the $V_{BE2}-T_B$ ($\approx V_{BE2}-T_{j2}$) curve; (iii) the power P_{D1} dissipated by device Q_1 was increased and the resulting V_{BE2} reduction was monitored; (iv) the temperature rise ΔT_{j2} was calculated for each P_{D1} as the ratio between $V_{BE2}(P_{D1}=0 \text{ W}) - V_{BE2}(P_{D1})$ and ϕ ; lastly, $R_M = \Delta T_{j2}/P_{D1}$, where the evaluation is made at low P_{D1} so as to safely disregard nonlinear thermal effects. R_M was found to span from 120 K/W for the lowest spacing to 98 K/W for the highest, this narrow range being ascribed to the poor lateral heat spreading through the GaAs substrate. As a consequence, only the topologies with the closest and furthest devices – hereinafter denoted as #1 and #2, respectively – are analyzed in the following. In PSPICE, the thermal conductivity reduction in the GaAs substrate was accounted for by resorting to the Kirchhoff transformation with a power factor $m=1.25$ [15].

Fig. 3a illustrates the comparison between the V_B -constant output I-V characteristics of the single HBT and cascode #1 in which V_{BB} is maintained at 2.6 V. An inspection of the figure reveals that: (i) experimental data – determined for various V_B values – are in a good agreement with PSPICE simulations, which confirms the accuracy of the calibration of the macromodel parameters and extraction of thermal resistances; (ii) the SOA of the cascode is rather larger than the single HBT counterpart; (iii) the analytical model for the flyback locus derived in section II was used by assigning to coefficient ϕ the values corresponding to the currents I_C (for the single HBT) and I_{C1} (for the cascode amplifier) at the lowest- V_B flyback points. This guarantees a good prediction of the “actual” SOA boundary computed by PSPICE in the high-voltage/low-current region. However, the accuracy is reduced at high currents, where the model is found to describe a shrunk SOA compared to PSPICE due to the approximations of (i) negligible negative-feedback effects, and (ii) constant V_{BE2} . Nevertheless, the model can still be suggested as a means for a straightforward determination of a *worst-case* SOA. As far as the analytical flyback locus of the single HBT is concerned, the formula proposed in [6], [16] was adopted; in this case a favorable matching with the “actual” SOA boundary is obtained, since only approximation (i) plays a role, but is somehow compensated by nonlinear thermal effects – enabled in PSPICE – which become relevant at high currents. Fig. 3a shows also the

benefits of base ballasting from the merely thermal viewpoint [17]; it can be seen that the flyback locus of cascode #1 moves rightward by increasing the value of two identical resistors tied to the bases of Q_1 and Q_2 . However, care should be taken when adopting this strategy due to the expected reduction in RF performance. Fig. 3b illustrates the junction temperature rises calculated by PSPICE and associated to the curves reported in Fig. 3a; it is found that the temperature of the cascoded HBT Q_2 at the flyback – highlighted in the figure – is much higher of that of the lower transistor Q_1 , which is almost equal to the one that would be obtained for the HBT of the single-stage amplifier.

Fig. 3. (a) PSPICE I_C-V_{CE} and $I_{OUT}-V_{OUT}$ characteristics of a single HBT and cascode #1 biased with $V_{BB}=2.6 \text{ V}$, respectively, for V_B ranging from 1.26 V to 1.36 V with a 10 mV step (solid lines), along with experimental data obtained for some V_B values (dotted); the flyback points are identified by white circles (single HBT) and squares (cascode #1); also reported are the loci for the cascode variants ballasted with two resistors of 100 (light-grey squares), 200 (grey), and 300 Ω (black) tied to the bases of Q_1 and Q_2 , and the analytical boundaries evaluated with the model derived in section II (dashed lines). (b) PSPICE temperature rises over ambient for the single HBT (ΔT_j , solid lines), and for the cascode transistors (ΔT_{j1} , dashed, and ΔT_{j2} , dot-dashed); the flyback points associated to ΔT_{j2} are identified with hexagons.

The impact of the thermal interaction between Q_1 and Q_2 was analyzed by comparing the flyback loci determined by PSPICE for amplifiers #1 and #2, as well as for (non-fabricated) circuits with $R_M=30 \text{ K/W}$ and 160 K/W . Results are shown in Fig. 4; as stated in section II, the SOA reduces for structures with tighter coupling. Similar conclusions were drawn in [4], where it was shown that a more rugged six-transistor cascode amplifier can be obtained by thermally insulating the CE stage from the CB one.

Fig. 4. PSPICE I_{OUT} - V_{OUT} characteristics of cascode #2 for $V_{BB}=2.6$ V and V_B spanning from 1.26 V to 1.36 V with a 10 mV step (solid lines), along with experimental data for some V_B values (dotted); the flyback locus computed by PSPICE (rhombi) is compared to the one determined from the model presented in section II (dashed line). Also depicted are the PSPICE locus associated to cascode #1 (squares) and those evaluated by setting R_M to 30 (triangles) and 160 K/W (stars), as well as the one determined for the single HBT (circles).

B. SiGe technology

A variant of the PSPICE tool was created to determine the SOA boundary of bipolar cascode amplifiers in state-of-the-art SiGe technology. This is of particular interest since the aggressive scaling aimed at performance improvement and the trench-based insulating schemes are pushing the thermal resistances of the devices into the thousands of K/W [13], [18]. Since only individual transistors were available for experimental characterization, the analysis was conducted by making use only of PSPICE simulations. The parameters of the macromodel were preliminarily calibrated so as to ensure good agreement with DC data measured on HBTs featuring peak $f_T/f_{max}=230/310$ GHz, $\beta_F=1000$, and BV_{CBO} equal to about 5.5 V; the thermal resistances R_{TH} were experimentally extracted by resorting to a common-base variant [13], [18] of the approach in [14]. In particular, the simulated cascode amplifier was composed by identical devices sharing $A_E=0.10 \times 2.69 \mu m^2$ and $R_{TH}=7200$ K/W, while the mutual resistance R_M was set to 1500 K/W. The comparison between the single HBT and the cascode biased with $V_{BB}=1.8$ V is reported in Fig. 5. It can be inferred that in this case the analytical locus may lead to a significant underestimation of the “actual” cascode boundary predicted by PSPICE for high V_B values. This discrepancy can be mainly attributed to the approximation of a constant V_{BE2} , which plays a major role in comparison to the GaAs technology due to the (much) lower BV_{CEO} ; as V_{CE2} exceeds this voltage, current I_{B2} exiting the base terminal rapidly grows, resulting in a large drop across the parasitic resistance $R_{B'1}I_{B2}$, and consequently leading to a significant V_{BE2} reduction.

V. CONCLUSIONS

In this paper, an extensive analysis of the SOA of GaAs and SiGe cascode amplifiers has been carried out with the support of measurements performed on GaAs circuits and calibrated PSPICE simulations. Based on a simple theoretical basis, an analytical model of the flyback locus has been developed, which can be employed at the design stage to identify a *worst-case* SOA

boundary and thus prevent damages induced by electrothermal and avalanche effects. The deleterious impact of thermal coupling has been quantified; in particular, it is shown that the SOA of cascode structures with close transistors tends to reduce to that corresponding to a standard single-stage CE amplifier.

Fig. 5. PSPICE I_C - V_{CE} and I_{OUT} - V_{OUT} curves of the single HBT and the cascode amplifier biased with $V_{BB}=1.8$ V in SiGe technology, for V_B ranging from 0.7 V to 0.85 V with a 10 mV step (solid lines); the flyback points determined by PSPICE (symbols) are compared to those calculated by the model derived in section II (dashed lines).

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