Thermal Impedance Extraction for Bipolar Transistors

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***Abstract*— This paper describes a method for extracting the thermal impedance of bipolar transistors. The measurement is a two-step process: first the fractional temperature coefficients are calibrated at dc and then a transient step response is measured to extract the thermal spreading impedance. Measurement configurations and an example measurement cycle are shown. The measurement results can be fitted to multiple-pole models for use in compact circuit modeling in SPICE.**

I. Introduction

THERMAL effects must be accounted for if bipolar transistors (BJT's) are to be accurately modeled [1], [2]. This applies to all BJT's, to heterojunction BJT's where band-gap engineering leads to large and complex temperature dependences of the current gain, and to SOI and dielectrically isolated transistors where the low thermal conductivity of isolating layers enhances self-heating [3], [4]. Two components must be characterized for accurate modeling: the thermal spreading impedance (temperature response to a change in dissipated power) and the temperature dependence of the current-voltage relationships of the transistor. At first glance measuring the temperature dependence of the currents appears easy; the ambient temperature is varied by a known amount and the resulting changes in the terminal currents are measured. In practice, the situation is complicated by the rise of the actual device temperature above the ambient by an unknown amount because of self-heating. This unknown temperature change can be significant compared to ambient variation especially in high current regions. This paper shows a way to overcome this problem, and how the results can be used to extract the thermal impedance from a transient response. If properly accounted for, the errors created by self-heating can be reduced in precision circuit design as well as in the extraction of the device model itself [5], [6].

The base current is treated as the thermometer in the techniques demonstrated here. The technique is based on the assumption that modulation of the base recombination current by the collector-base voltage is negligible compared to the effect of self-heating on the total base current. This assumption is typically valid for narrow-base devices. If operation is in the forward-active region without significant impact-ionization, any change in base current resulting from a change in collector-base voltage with constant base-emitter voltage is dominated by self-heating. Characterization of the thermal impedance requires two steps. The first step involves finding the fractional temperature coefficients: the fractional change in base or collector current for a unit change in the transistor temperature with fixed base-emitter voltage. This information can be found by sweeping the collector voltage at different ambient temperatures and then cancelling the self-heating. The second step is to find the thermal impedance from the base current response to a collector voltage step. Using the temperature coefficients from the first step, the base current response is converted to a temperature response. After the transient temperature response is normalized by the dissipated power, a thermal impedance model can be fitted to the waveform. The program TIPP (Thermal-Impedance Pre-Processor) can be used to perform this fitting.

Manuscript received April 3, 1995; revised August 28, 1995. The review of this paper was arranged by Associate Editor D. P. Verret. This work was supported by the Seminconductor Research Corporation, Contract 91-SP-087.

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Publisher Item Identifier S 0018-9383(96)01090-8.

II. Temperature Coefficient Measurement

In this part of the measurement cycle the goal is to calibrate the base-current thermometer by measuring the change in current caused by a change in temperature with fixed base voltage. When the temperature is increased the collector current and the power increase, causing the local temperature to increase considerably because of self-heating. At moderate to high currents an error arises when self-heating contributes significantly to the device temperature. Transient measurements, to be discussed in the next section, are typically made in these regions, requiring characterization of the base current dependance on temperature in the presence of self-heating. The total temperature change of a BJT including self-heating is

$$\Delta T = R_{\mathrm{TH}}\Delta P + \Delta T_A = R_{\mathrm{TH}}(I_C \Delta V_C + \Delta I_C V_C) + \Delta T_A \tag{1}$$

where ΔP is the change in power dissipation, ΔT is the total effective temperature change at the emitter, ΔT_A is the ambient temperature change, and R_{TH} is the thermal resistance. Separately measuring the change in base current caused by a change in ambient temperature and then by a change in power gives two equations that allow cancellation of the self-heating from the temperature coefficients. The power is controlled by varying the collector voltage, V_C, while keeping the device in the forward-active region and avoiding significant impact-ionization.

When changing the ambient temperature with V_C constant, (1) gives a net temperature change of

$$\Delta T|_{\Delta V_C=0} \cong \Delta T_A + R_{\mathrm{TH}} V_C \Delta I_C \tag{2}$$

where $\Delta I_C = \frac{\partial I_C}{\partial T}\Delta T = T_{CF}(I_C) I_C \Delta T$ is the change in collector current due to a change in temperature, and $\mathrm{TC}_F(I_C) = \frac{1}{I_C}\frac{\partial I_C}{\partial T}$ is the fractional temperature coefficient of collector current. Substituting into (2) and solving for ΔT gives

$$\Delta T|_{\Delta V_C=0} = \frac{\Delta T_A}{1 - \mathrm{TC}_F(I_C) R_{\mathrm{TH}} I_C V_C}. \tag{3}$$

When T_A is held constant and V_C is varied, (1) reduces to

$$\Delta T|_{\Delta T_A=0} = R_{\mathrm{TH}}(I_C \Delta V_C + \Delta I_C V_C) \tag{4}$$

where $\Delta I_C = \frac{\partial I_C}{\partial T}\Delta T + \frac{\partial I_C}{\partial V_C}\Delta V_C = \mathrm{TC}_F(I_C) I_C \Delta T + \frac{\Delta V_C}{V_A} I_C$ and V_A is the Early voltage. Substituting into (4) and solving for ΔT gives

$$\Delta T|_{\Delta T_A=0} = \frac{I_C R_{\mathrm{TH}}\left(1 + \frac{V_C}{V_A}\right)\Delta V_C}{1 - \mathrm{TC}_F(I_C) R_{\mathrm{TH}} I_C V_C}. \tag{5}$$

Notice that the denominators of (3) and (5) are the same. Combining (3), (5) and the base-current dependance given by

$$\begin{aligned}\Delta I_B &= \frac{\partial I_B}{\partial T}\Delta T + \frac{\partial I_B}{\partial V_C}\Delta V_C \\ &\approx \mathrm{TC}_F(I_B) I_B(\Delta T_A + R_{\mathrm{TH}} V_C \Delta I_C)\end{aligned} \tag{6}$$

where the second term is the base-collector conductance (assumed to be negligible), results in

$$\frac{(\Delta I_B/\Delta V_C)|_{\Delta T_A=0}}{(\Delta I_B/\Delta T_A)|_{\Delta V_C=0}} = I_C R_{\mathrm{TH}}(1 + V_C/V_A). \tag{7}$$

This can be solved for R_{TH} if V_A is known or is much greater than V_C. Note that R_{TH} is an *effective* value, representing the steady-state thermal response, and thus includes the thermal resistances of the die and package in addition to the thermal spreading resistance. Once R_{TH} is known, (6) can be solved for the fractional temperature coefficient of base current

$$T_{\mathrm{CF}}(I_B) = \left.\frac{\Delta I_B/I_B}{\Delta T_A + R_{\mathrm{TH}} V_C \Delta I_C}\right|_{\Delta V_C=0}. \tag{8}$$

Note that this rather complicated procedure is not needed at currents low enough that self-heating can be neglected.

To make the measurements the transistor is connected in common-emitter configuration in an oven or on a probe-station thermal chuck. It is important to have stable, accurate ambient temperature control during the measurement, because the voltage sweep takes a long time and errors in the ambient temperature significantly affect the accuracy of the results.

The algorithm for the measurement is simple although time consuming. First, the ambient temperature is allowed to reach steady state, then a base voltage appropriate to the current range of interest is applied, and the collector voltage is swept to produce a significant, linear increase in I_B. (Impact ionization causes a *reduction* of I_B, and must be avoided.) After each power step, self-heating must be allowed to reach steady state before measurement begins. A long delay (typically 10–20 s) is recommended. (The HP 4145 Semiconductor Parameter Analyzer requires control by an external computer to provide delays longer than about 5 s.) For each V_{BE} in the range of interest, V_{CE} is stepped and the resulting steady-state currents are recorded. Next, the temperature is changed and the preceding procedure is repeated. A temperature step of 5–10°C is typically required to balance errors caused by the strongly nonlinear dependance of current on temperature against the effects of measurement uncertainty. Mean values of the currents are used in (7) and (8) to compute the fractional temperature coefficients at each base voltage value. Geometric means should be used to form the averages over temperature to help correct for the nonlinearities.

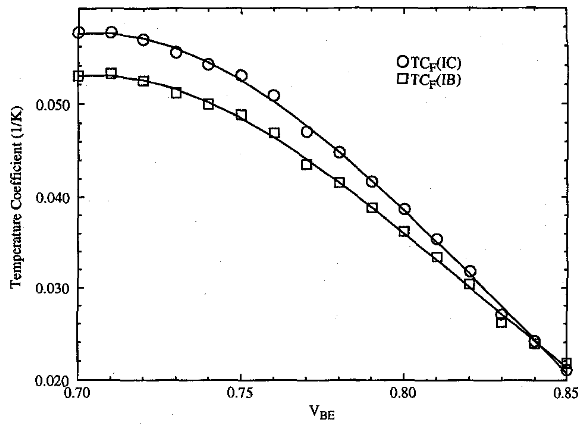

Fig. 1. Plot of the fractional temperature coefficients of base and collector current versus fixed V_{BE} for a $3\times50\ \mu\mathrm{m}^2$ Harris UHF bipolar NPN transistor. The squares and circles show measured data and the lines are least-squares fits.

Fig. 1 shows a plot of the temperature coefficients versus base voltage for an NPN Harris UHF transistor with a $3 \times 50\ \mu\mathrm{m}^2$ emitter. The UHF process yields high-frequency, dielectrically isolated, vertical transistors. (Such devices are especially prone to self-heating.) The collector voltage was swept from 2–4 V while the base voltage was varied from 700–850 mV in 5 mV intervals. The ambient temperature was controlled to within ±0.5°C by an oven; temperatures used were 25 and 33°C. This measurement produces errors of approximately 10% that are dominated by oven-temperature accuracy, and nonlinearity of the data.

III. Thermal Impedance Measurement

The objective of this part of the measurement cycle is to measure the normalized thermal response to a step in power. This is done by converting the base current response into a temperature response using the temperature coefficients extracted above. Solving the fractional temperature coefficient definition for the change in temperature gives

$$\Delta T(t) = \frac{I_B(t) - I_B(0)}{I_B T_{\mathrm{CF}}(I_B)} \tag{9}$$

where I_B is the median value of the base current for the transient. $\Delta T(t)$ can be normalized by dividing by the power

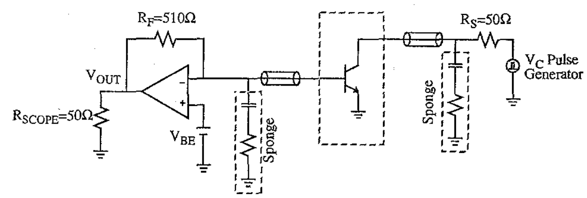

Fig. 2. Setup for thermal transient response measurement of a bipolar transistor using a 50 Ω RF wafer probe station. Each sponge, composed of a 800 pF capacitor and 50 Ω resistor, absorbs reflected power due to mismatches from 45 MHz to 20 GHz.

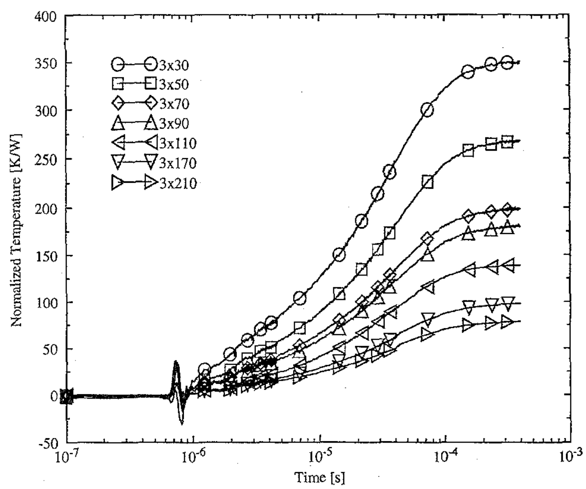

Fig. 3. Normalized temperature step-response versus time for different emitter size UHF NPN transistors.

of the step:

$$\Delta T_{\text{norm}}(t) = \frac{\Delta T(t)}{\Delta I_C V_C + \Delta V_C I_C} \tag{10}$$

where the denominator represents the magnitude of the power step.

A suitable measurement setup is shown in Fig. 2, where the BJT in the center is the device under test. The current-to-voltage converter transforms the base current into a voltage waveform for the oscilloscope. With V_{BE} fixed, the base current is

$$I_B(t) = (V_{\text{out}} - V_{BE})/R_F. \tag{11}$$

where R_F is the feedback resistance. Unfortunately, the intrinsic base-emitter voltage is not constant because of the parasitic base and emitter resistances. Since the V_{BE} variation is small compared to the thermal voltage, a linear correction is applicable

$$I_B(t) = \frac{V_{\text{OUT}} - V_{BE}}{R_F - R_B - (1+\beta)R_E} \tag{12}$$

where R_B and R_E are estimated base and emitter resistances and β is the large-signal current gain.

Feedback resistance R_F must be chosen carefully to optimize the gain-bandwidth product and measurement error. Gain must be large because the signals are typically very small, yet a high bandwidth (typically at least 100 MHz) is needed to resolve the earliest part of the response. In our measurements, an amplifier with unity gain-bandwidth of 1 GHz was used with a loop gain of about 10 for a bandwidth of 100 MHz. The "sponges" shown in Fig. 2 are high-frequency sinks that absorb the energy at frequencies beyond those of interest in the thermal measurement, damping reflections and possible oscillations.

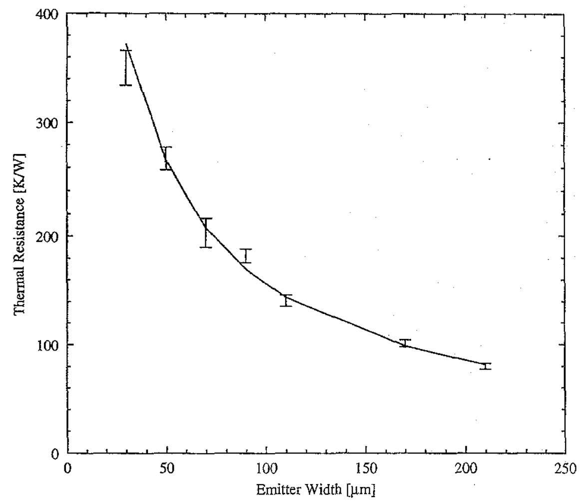

Fig. 4. Plot of thermal resistance versus emitter width for the UHF transistor showing the errors for the measured point. The line shows the least-squares fit.

To make the step response measurements, V_{BE} is first chosen to set a current level that leads to a measurable base current response to a collector voltage square wave. The best frequency for the collector voltage waveform depends on the transistor and its packaging. The frequency must be low enough for the response to settle before the end of each half cycle, yet high enough to null the heating time constants of the chip and package. A frequency of 1.25 kHz was chosen for our measurements, resulting in a 400 μs pulse width. Because of the distributed nature of the response, op-amp output is recorded using three overlapping time windows of increasing size to capture with high resolution the fast beginning of the response while keeping the number of data points to a minimum for the long tail. Next, an ammeter was connected in series with the collector to measure the current at the pulse maximum and minimum values. The op-amp output is also measured at this time to compute the base current using (12). From this data the β of the transistor is determined, so that the transient base current can be converted into a collector current for computation of the power step magnitude. The base current minimum and maximum in this dc measurement will differ from the values in the transient because of the difference in mean power being dissipated. Using (9) and (10), the normalized temperature response can be computed. Fig. 3 shows the normalized temperature responses for transistors from the same process as those in Fig. 1.

Fig. 4 shows the thermal resistance extracted from the same transistors versus emitter width. Estimated measurement errors arise primarily from errors in the extracted fractional temper-

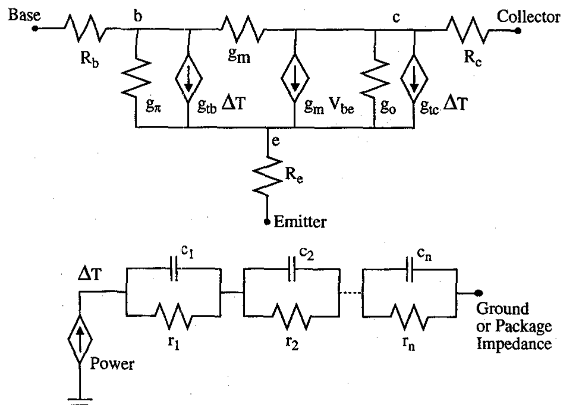

Fig. 5. Model for modified SPICE thermal analysis. The thermal impedance is connected to the ΔT terminal supplied with a power source to generate a voltage proportional to the temperature change. Sources G_{TB} and G_{TC} are the base and collector thermal transconductances.

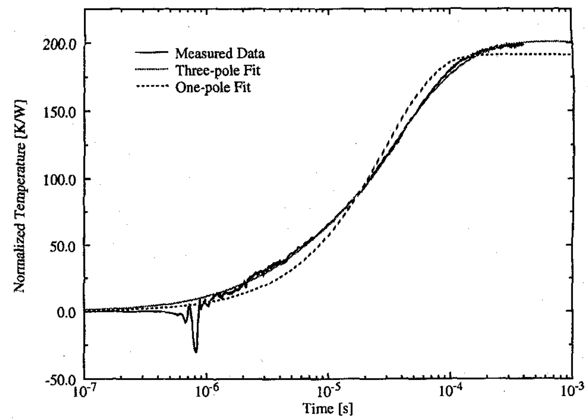

Fig. 6. Normalized temperature response of a $3 \times 70\ \mu m^2$ UHF NPN transistor; measured and modeled with three-pole and single-pole RC ladders generated by TIPP.

ature coefficient, the variance of β over the range of collector voltage, and the limited resolution of the oscilloscope.

For small transistors that operate at low currents it is hard to get changes in I_B large enough to overwhelm system noise and avoid round-off error. Unfortunately thermal effects in these transistors are not insignificant. For example, in our measurements errors were dominant for collector currents below about 1 mA. For a typical transistor biased at 1 mA undergoing a 5 mW power step with an R_{TH} of 500 K/W, an approximately 15% increase in I_B occurs. However, this change results in only a 1 mV signal at the op-amp output. This signal is hard to resolve with a oscilloscope with a 1 mV/division minimum resolution. Even with averaging, the typical 2 mV of noise will tend to degrade the signal even further. One way to ameliorate this problem is to wire several of these small BJT's in parallel electrically but decoupled thermally. The resulting larger signal would then enhance the dynamic range.

IV. Thermal Impedance Model and SPICE

The above results must be compressed into a manageable model for use in circuit simulation. Using a version of SPICE modified to include a temperature (ΔT) terminal [8], shown in Fig. 5, the thermal impedance can be incorporated into precision circuit simulation of dc, ac and transient. This has shown to be important in many applications, especially in analog circuits [5]. In this version of SPICE the thermal impedance is modeled using an RC ladder subcircuit connected between the ΔT terminal and ground. The ΔT terminal, sourced by a current proportional to the power dissipation of the BJT, generates a voltage through the RC ladder equal to the change in local temperature of the emitter. Then, through the thermal dependencies shown as G_{TB} and G_{TC} in the figure, this ΔT is used to modify the operating-point, ac and transient behavior.

The program TIPP [9] can be used to derive the thermal resistance and thermal capacitance values for a thermal equivalent circuit from the measured step-response data. TIPP uses the following algorithm to fit the thermal step response. The thermal step response magnitude is "scanned" for specific points (one point for the one-pole model, two points for the two-pole model, etc.) which correspond to certain predetermined percentages of the step maximum. The times at which these percentages are attained are assumed to approximate the thermal time constants. A linear, least-squares fit is then used to approximate the thermal resistance components. The thermal capacitances are then computed from the time constants and the thermal resistances. This simple algorithm successfully approximates thermal responses for a wide variety of device structures. Fig. 6 shows the normalized temperature response for a $3 \times 70\ \mu m^2$ transistor and the resulting three-pole and single-pole fits. The three-pole fit is very good, and generally is the best trade-off between accuracy and complexity for precision applications.

V. Conclusion

This technique for characterizing BJT thermal impedance parameters can be used for narrow-base BJT's and HBT's (which tend to be especially temperature sensitive). Measurement equipment is constantly improving, resulting in diminishing errors for this technique and allowing characterization of smaller, lower current devices. The method requires linearization of the current versus temperature characteristics, which induces some error. To improve the accuracy of this technique, nonlinear optimization could be used for the temperature coefficient extraction if a circuit model including thermal effects is linked to the optimizer. Ideally the thermal impedance should be included in the extraction process of all of the Gummel-Poon parameters to reduce errors caused by self-heating at moderate to high currents.

Because of the very long thermal time constants associated with self-heating, the initial calibration step required in this method is inherently very slow. The time needed to extract temperature sensitivities for all device geometries in a given technology would be excessive, so an alternative methodology is desirable. A simple empirical model for interpolating between temperature-coefficient measurements would help, but a physics-based model would be even more useful. Such a model could predict the thermal parameters for a given technology, even for difficult to measure small geometries, with minimal need for tuning through measurements.

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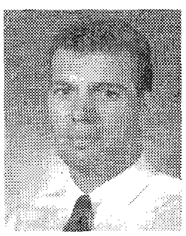

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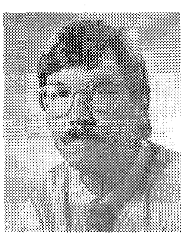

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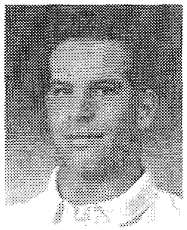

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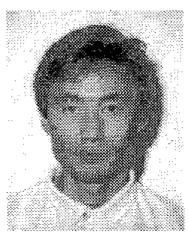

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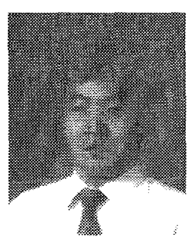

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