Parameter-Extraction Method for Heterojunction Bipolar Transistors

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Abstract— **A method for determining the equivalent-circuit element values of a small-signal heterojunction bipolar transistor (HBT) is described. Most important is the ability to separate the parasitic emitter resistance from the junction resistance. No use of "cold" measurements or test patterns is required. The method may also be applicable to homojunction bipolars.**

I. Introduction

In the analysis of circuits using HBT's, it is often necessary to have an accurate linear equivalent circuit of the device. Even in nonlinear analysis, the linear equivalent-circuit elements are part of the nonlinear equivalent circuit. If Volterra methods are used for nonlinear analysis, the accuracy of the analysis depends directly upon the accuracy of an initial, linear analysis. Many of the linear-circuit elements have a strong effect on the performance of an HBT, and thus it is important to determine their values accurately. This is especially true of the emitter resistance R_{ee} and the linear part of the current gain $\alpha(\omega)$.

Most existing methods (e.g., [1], [2]) for determining this parameter are not accurate enough for IM or noise calculations. We have developed a new method that is quite accurate and is based on the largely unrecognized fact that the HBT's impedance parameter Z_{12} is real and constant with frequency, and is related simply to Ree. Many of the other equivalent circuit elements can be determined accurately from S-parameter measurements without the use of optimization techniques.

In fitting an equivalent circuit to a set of measured S-parameters, one should determine as many of the elements as possible before resorting to optimization techniques. A common method for determining the element values is to use test patterns fabricated on the same wafer as the device [3], [4]. These patterns, however, may not represent the device adequately, and in any case are not always available to the circuit designer. Other methods use unbiased or "cold device" measurements [5]; however, the equivalent-circuit elements in the "cold" device are often different from those of the "hot" device. Here we describe a method of parameter extraction that is quite accurate and does not require the measurement of test patterns or the unbiased device.

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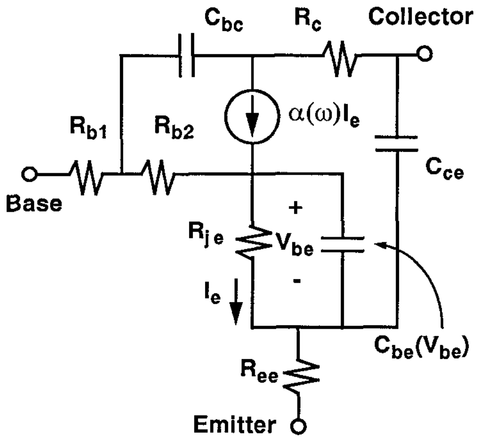

Fig. 1. Small-signal HBT equivalent circuit.

II. Parameter Extraction

The small-signal equivalent circuit of the HBT is shown in Fig. 1. Although this is a linear circuit, by treating the relevant elements as nonlinear, one can use the same topology for nonlinear analysis [6].

The philosophy behind the method is to determine as many resistive elements as possible without resorting to S-parameter fitting. This reduces the number of variables in the fitting process, and makes it more reliable. Indeed, most of the resistive parasitics in the model can be determined in this manner, and others can be estimated closely. The resistive elements are found from low-frequency S-parameter measurements at a number of bias points. All S-parameter measurements are made on-wafer. The calibration technique used for the network analyzer sets the reference plane at the edge of the device, thereby removing the HBT's contact pads from the measurement.

The HBT's emitter current is given by the well-known expression

$$I_e = I_s\left(e^{\delta V_{be}} - 1\right). \tag{1}$$

Because HBT's have high thermal resistance and their junction parameter I_s is very sensitive to temperature, most dc methods (especially those involving the use of Gummel plots) do not produce sufficiently accurate values of R_{ee} or δ. Instead, we use measurements of Z_{12}. In a bipolar transistor (heterojunction or homojunction), in common-emitter configuration,

$$Z_{12} = R_{ee} + R_{je}. \tag{2}$$

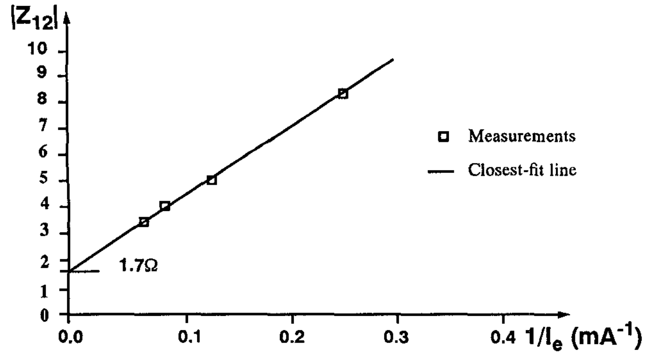

Fig. 2. Plot of Z_{12} vs. $1/I_e$, measured at approximately 1 GHz. Extrapolated point where $1/I_e = 0$ is $Z_{12} = R_{ee}$.

Because the base-to-emitter capacitance in these devices is very small, and the output current is controlled by the current in R_{ee}, Z_{12} is real. It is also remarkably constant at frequencies up to at least 10 GHz in our devices; it varies only a few tenths of an ohm over this range. This property is generally true of bipolar devices when the reactance of the base-to-collector capacitance is large; the small decrease in Z_{12} observed at the higher frequencies is probably caused by this capacitance. This property allows us to establish its value very accurately at the outset. Z_{12} is found by converting S-parameters to Z-parameters. It is given by

$$R_{je} = \left(\frac{\partial I_e}{\partial V_{be}}\right)^{-1} \approx \frac{1}{\delta I_e}. \tag{3}$$

R_{je} is independent of I_s (it is, of course, dependent on δ; however, δ's temperature dependence is much weaker than that of I_s). Since R_{je} is proportional to $1/I_e$, one can find $R_{ee} = Z_{12}$ at the extrapolated point where $1/I_e = 0$ (one can also find d from the slope of this plot). Fig. 2 shows such a plot for a TRW small-signal HBT. Similarly, one can obtain an estimate of $R_{b1} + R_{b2}$ by calculating Z_{12} in common-base configuration, and an estimate of R_c from Z_{12} calculated in common-collector configuration. However, the S-parameters of the device are not very sensitive to these resistances; specifically, changing $R_{b1} + R_{b2}$ does not change S_{11} very much, nor does R_c change S_{22}. Consequently, these quantities are not as accurate as the determination of Z_12 from R_{ee} and R_{je}: we have observed that the value of Z_{12} in common-emitter configuration is reliably constant within 0.1 Ω up to several GHz; in the other configurations it often deviates by ±1 Ω or more.

The current gain, $\alpha(\omega)$, is expressed by

$$\alpha(\omega) = \alpha_0 e^{-j\omega\tau}. \tag{4}$$

$\alpha_0$is found from the low-frequency common-emitter current gain H_{21}, which is in turn obtained from an S- to H-parameter conversion:

$$\alpha_0 = \frac{|H_{21}|}{1 + |H_{21}|}. \tag{5}$$

When these quantities are determined, the remaining elements, mostly capacitances, can be found very reliably from conventional S-parameter fitting. During this fitting, R_{je}, R_{ee}, and

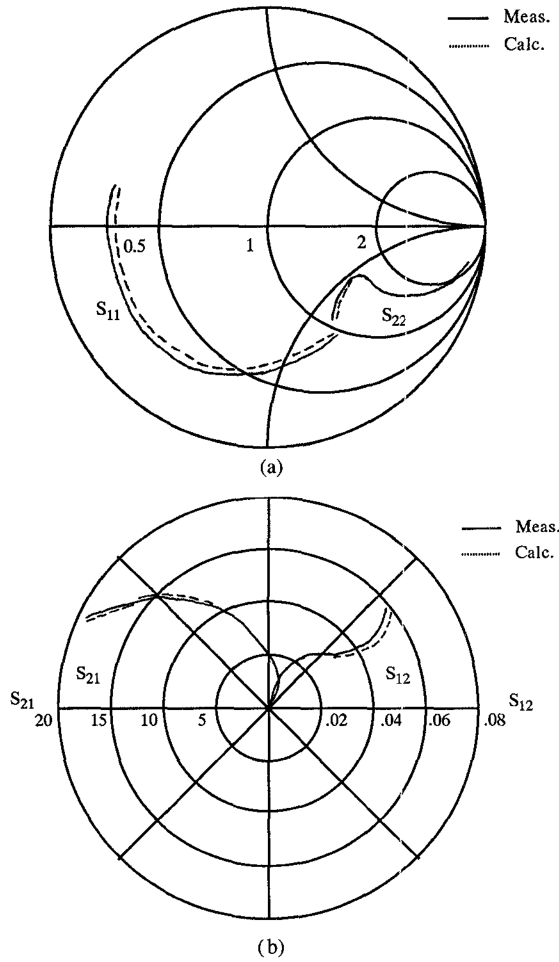

Fig. 3. Comparison of measured and calculated (a) S_{11} and S_{22} and (b) S_{21} and S_{12} of a 2 x 10-micron, quad-emitter TRW HBT from 0.5 to 26 GHz. $V_{ce} = 3.0$ V, $I_c = 16$ mA.

TABLE I
HBT MODEL PARAMETERS

Parameter	Value
R_{ee}	1 70
R_{je}	1.75
R_{b1}	3.7
R_{b2}	3.7
R_c	3.0
C_{bc}	58.7 fF
C_{ce}	15.8 fF
C_{be}	1.85 pF
τ	3.66 psec
α_0	0.9764

TRW 2 x 10-micron, quad-emitter HBT. $V_{ce} = 3.0$ V, $I_c = 16$ mA.

$\alpha(\omega)$ are not modified. Although it is very small, C_{bc} has a remarkably strong effect on S_{22} as well as its expected dominant effect on S_{12}, and therefore is easy to fit. Fig. 3 compares the S-parameters of the model and the TRW device, and Table I shows the element values.

III. CONCLUSION

We have shown that many of the microwave equivalent-circuit elements of a small-signal HBT can be obtained from a straightforward use of S-parameter measurements. When these are determined, the number of remaining elements is reduced considerably and the equivalent circuit can be found with little ambiguity from conventional S-parameter fitting.

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