A New Method for Determining the Reverse Transit Time in Bipolar Transistors

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***Abstract*—The reverse transit time is an important parameter for determining the delay of bipolar transistor in saturation. A new method is proposed here to extract the reverse transit time of bipolar transistors. The technique is based on ac short-circuit current gain measurements using a network analyzer. The method is very simple and is useful for on-wafer measurements.**

I. Introduction

The REVERSE transit time (τ_R), which accounts for the charge storage in the forward-biased base–collector junction of bipolar transistors, is one of the important parameters of the bipolar transistor in saturation. It should be precisely determined for the simulation of the transient response of bipolar and BiCMOS integrated circuits vulnerable to saturation. For example, the switching speed of the common BiCMOS driver is limited by the saturation of bipolar transistors and inaccurately extracted τ_R misguides the circuit design as illustrated in Section II.

The pulse measurement technique [1], [2] is commonly used to determine τ_R. In this method, the saturation delay is measured from the pulse response and τ_R is calculated by

$$\tau_R = \frac{t_s}{\ln\left[(I_{BF} + I_{BR})/(I_{CF}/\beta_F + I_{BR})\right]} \cdot \left(\frac{1 - \alpha_F\alpha_R}{\alpha_R}\right) - \left(\frac{\alpha_F}{\alpha_R}\right)\tau_F$$

[2]. But this method requires that the forward transit time (τ_F) be known and that measurements of dc operating currents be made; so errors in τ_F and dc operating currents can lead to inaccurate τ_R extractions. It is also a difficult procedure for on-wafer evaluation due to the parasitics of the contact pads and measurement setup. If the reverse dc gain (β_R) is much greater than 1, τ_R can be determined in a similar manner to the on-wafer ac extraction method for τ_F but with the transistor biased in the reverse active mode. But for the modern diffused bipolar transistor with asymmetrical structures, β_R is less than or just greater than unity and thus the ac τ_F extraction method can not be directly applicable to the extraction of τ_R.

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This paper proposes a simple and accurate method for measuring τ_R for bipolar transistors used in integrated circuits. It is based on the ac short-circuit current gain (h_{21}) measurements in the reverse active mode. The s-parameters are measured using a network analyzer and microwave probes and then converted into the h-parameters. The merit of this method is that it is very simple and does not require measuring dc bias currents or gain. Another advantage is the easy elimination of probe parasitics by compensation, so on-wafer measurement of τ_R is possible.

In Section II, we investigate the importance of τ_R in the BiCMOS driver. The proposed method and measurements are presented in Section III.

II. Impact of τ_R on the Circuit Performance

BiCMOS technology has demonstrated a speed performance advantage compared to CMOS technology, but the speed of BiCMOS circuits is limited by the saturation of bipolar transistors. The commonly used BiCMOS driver is shown in Fig. 1. As V_{IN} goes to V_{CC} (= 5.0 V), MOSFET's *MN*2 and *MN*3 are turned on. Consequently, the pullup bipolar transistor *Q*1 is turned off and the pulldown bipolar transistor *Q*2 is turned on. Then I_{C2} discharges the load capacitor C_L and V_{OUT} goes to zero. But the base voltage of *Q*2 stays around 1.0 V for a short period of time so that both the base–emitter and base–collector junctions of *Q*2 are forward-biased, i.e., *Q*2 is in saturation. When V_{IN} goes to zero, *MP*1 is turned on and thus *Q*1 is also turned on. So, I_{C1} charges C_L and V_{OUT} goes to V_{CC} minus one diode voltage drop. But if the pulsewidth of V_{IN} is narrow, *Q*1 is turned on while *Q*2 undergoes the reverse recovery. Charges stored in the base and collector of *Q*2 are slowly removed by *MN*3 and *MN*4. Therefore, both transistors *Q*1 and *Q*2 are conducting currents undesirably, as shown in Fig. 2, which is called "the crowbar." Figs. 2 and 3 show the simulated transient collector currents of *Q*1 and *Q*2. Accurately extracted and optimized MOSFET and BJT parameters are used in the simulations. If τ_R = 1 ns is used instead of 5 ns, the crowbar problem is hardly observed (see Fig. 3). Thus if the τ_R value used in simulations is underestimated, circuit designers could fail to simulate the crowbar problem. These simulations indicate the importance of accurate extraction of τ_R for predicting the bipolar transistor operating in the saturation region.

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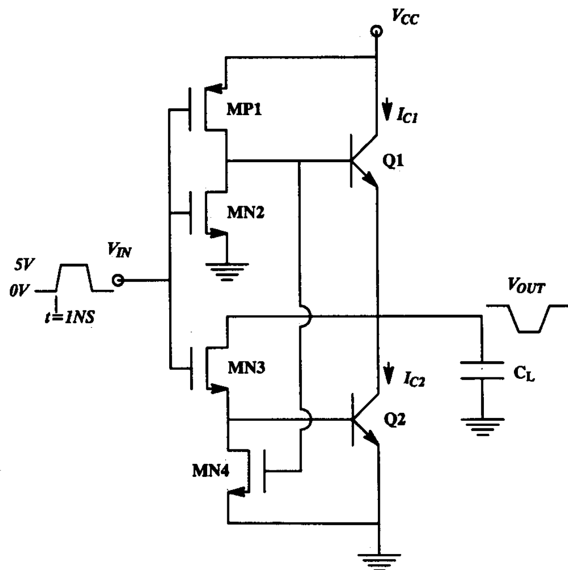

Fig. 1. The BiCMOS inverter circuit. $V_{CC} = 5.0$ V.

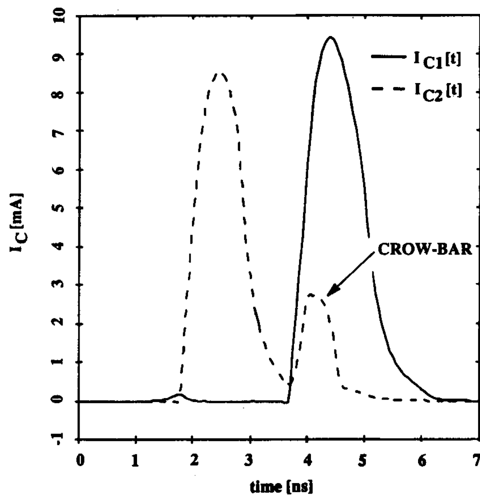

Fig. 2. The MTIME simulated transient collector currents for pulldown ($Q2$) and pullup ($Q1$) transistors in Fig. 1 with $\tau_R = 5.0$ ns. The initial delay, rise time, pulsewidth, and fall time of the input pulse are all 1.0 ns.

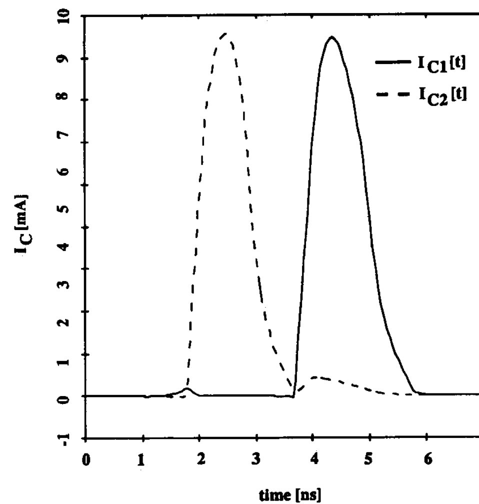

Fig. 3. The MTIME simulated transient collector currents for pulldown ($Q2$) and pullup ($Q1$) transistors in Fig. 1 with $\tau_R = 1.0$ ns. The initial delay, rise time, pulsewidth, and fall time of the input pulse are all 1.0 ns.

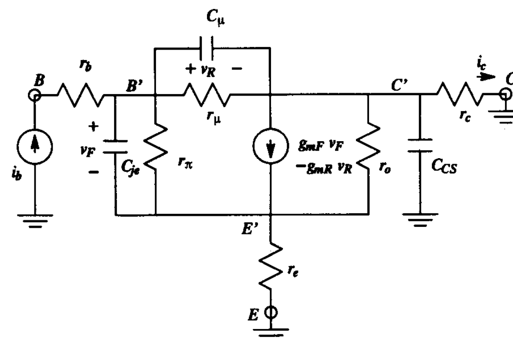

Fig. 4. The small-signal hybrid-π equivalent circuit for both normal-active and reverse-active modes of bipolar transistor in the C-E configuration.

III. Proposed Method and Measurements

The proposed method for extracting τ_R is based on h_{21} measurement in the reverse active mode in the common-emitter (C-E) configuration. The small signal hybrid-π equivalent circuit for the C-E configuration is shown in Fig. 4. We use a generalized hybrid-π representation for both normal-active and reverse-active modes of operation.

For the measurement of τ_R we bias the bipolar transistor into the reverse-active mode ($V_{BC} > 0$ V and $V_{BE} = 0$ V for n-p-n transistor). In the reverse region of operation, the resistance r_π, r_0, and forward transconductance g_{mF} can be neglected. The input ac base current i_b in the resulting equivalent circuit is given by

$$i_b = \frac{1 + j\omega r_\mu C_\mu}{r_\mu} v_R + i_{je} \tag{1}$$

where i_{je} is the current fed through C_{je}

$$\begin{aligned} i_{je} &= j\omega C_{je} v_F \\ &= j\omega C_{je}[v_R + r_c i_c - r_e(i_{je} - g_{mR} v_R)]. \end{aligned} \tag{2}$$

Summation of currents at the internal collector (C') gives

$$-i_c - j\omega C_{cs} r_c i_c + \frac{1 + j\omega r_\mu C_\mu}{r_\mu} v_R + g_{mR} v_R = 0. \tag{3}$$

Assuming $\omega C_{cs} r_c << 1$, (3) yields

$$i_c \simeq \left(\frac{1 + j\omega r_\mu C_\mu}{r_\mu} + g_{mR} \right) v_R. \tag{4}$$

Substituting (4) in (2) and assuming $\omega C_{je} r_e << 1$

$$i_{je} \simeq j\omega C_{je}[1 + (r_c + r_e) g_{mR} + \frac{r_c}{r_\mu}(1 + j\omega r_\mu C_\mu)] v_R. \tag{5}$$

Using (1), (4), and (5),

$$h_{21}(\omega) \equiv \frac{i_c}{i_b}(v_c = 0) = \frac{1 + \frac{g_{mR} r_\mu}{1 + j\omega r_\mu C_\mu}}{1 + \frac{j\omega C_{je} r_\mu}{1 + j\omega r_\mu C_\mu}[1 + (r_c + r_e) g_{mR} + \frac{r_c}{r_\mu}(1 + j\omega r_\mu C_\mu)]}. \tag{6}$$

The right-hand side of the denominator in (6) can be neglected for the forward-biased base-collector junction ($C_\mu >> C_{je}$) and thus

$$h_{21}(\omega) = 1 + \frac{g_{mR} r_\mu}{1 + j\omega r_\mu C_\mu}$$

$$= (1 + g_{mR} r_\mu)\left(\frac{1 + \frac{j\omega r_\mu C_\mu}{1 + g_{mR} r_\mu}}{1 + j\omega r_\mu C_\mu}\right). \tag{7}$$

The ac short-circuit current gain in (7) has one pole (ω_p) and one zero (ω_z) with ω_z at higher frequency than ω_p as verified by measurements at end of this section.

The reverse transconductance g_{mR}, which accounts for the carrier collection by the reverse-biased base-emitter junction, is given by

$$g_{mR} \equiv \frac{dI_{EC}}{dV_{B'C'}}$$

$$= \frac{I_{EC}}{V_T} \tag{8}$$

where I_{EC} is the reverse transport current [2] and V_T is the thermal voltage (kT/q). The resistance r_μ is given by

$$r_\mu \equiv \frac{dV_{B'C'}}{dI_B}$$

$$= \frac{V_T \beta_R}{I_{EC}} \tag{9}$$

where β_R is the reverse-current gain (I_E/I_B). The base-collector junction diffusion capacitance, which controls the switching speed of the transistor in the reverse-active mode, is expressed by

$$C_\mu \equiv \frac{d(\tau_R I_{EC})}{dV_{B'C'}}$$

$$= \frac{\tau_R I_{EC}}{V_T}. \tag{10}$$

Substituting (8)–(10) in (7) gives

$$h_{21}(\omega) = (1 + \beta_R)\frac{1 + j\omega \frac{\beta_R \tau_R}{1 + \beta_R}}{1 + j\omega \beta_R \tau_R}$$

$$= h_{21}(0)\frac{1 + j\omega/\omega_z}{1 + j\omega/\omega_p} \tag{11}$$

where

$$h_{21}(0) = 1 + \beta_R \tag{12}$$

$$\omega_z = \frac{1 + \beta_R}{\beta_R \tau_R} \tag{13}$$

$$\omega_p = \frac{1}{\beta_R \tau_R}. \tag{14}$$

Note that ω_z is greater than ω_p by a factor of $1 + \beta_R$.

Now the reverse transit time τ_R can be determined from the frequency response of $h_{21}(\omega)$. First β_R can be extracted from $h_{21}(0)$ using (12). Then we calculate the magnitude of $h_{21}(\omega = \omega_P)$

$$|h_{21}(\omega_p)| = \gamma \frac{h_{21}(0)}{\sqrt{2}} \tag{15}$$

where

$$\gamma = \sqrt{1 + (\omega_p/\omega_z)^2} = \sqrt{1 + (1 + \beta_R)^{-2}}$$

is a factor determined by the distance between ω_p and ω_z. For β_R less or slightly greater than unity, as in most modern diffused bipolar transistors, ω_z is close to ω_p and $\gamma > 1$. For β_R much greater than unity, $\omega_z >> \omega_p$ (the device has a dominant pole ω_p) and $\gamma = 1$. Now ω_p is graphically determined from the frequency response plot $h_{21}(\omega)$ using (15). Finally, τ_R can be obtained from (14).

To verify the proposed method, measurements were performed on a polysilicon emitter n-p-n transistor with an emitter size of 0.8 × 20 μm^2. It was fabricated by a 0.5-μm BiCMOS triple-polysilicon technology for 4-Mb fast SRAM's [3]. The s-parameter measurements were taken using an HP8753 network analyzer and Cascade Microtech Probes. DC biasing (using an HP4145), s-parameter calibration, and data acquisition and analysis were handled by HP IC-CAP software [4]. The network analyzer, cables, probes, and bond pad parasitics were calibrated using calibration standards on silicon wafers [5]. The transformation of the measured s-parameters to h-parameters was performed with IC-CAP. The measured $|h_{21}|$ as a function of frequency in the reverse-active mode of operation is shown in Fig. 5. The low-frequency $|h_{21}|$

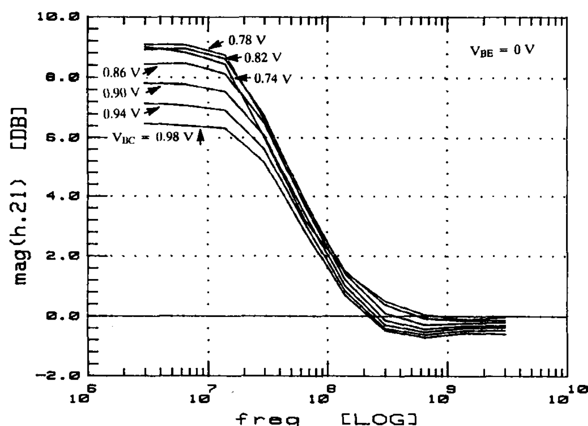

Fig. 5. Measured h_{21} versus frequency characteristics. $V_{BE} = 0$ V, $V_{BC} =$ 0.74 to 0.98 V, and $\Delta V_{BC} = 0.04$ V. Note that the y-axis is $20 \log_{10} |h_{21}|$.

corresponds to $1 + \beta_R$ taken from dc measurements. When $\omega_p < \omega < \omega_z$, $|h_{21}|$ decreases but the slope is not 20 dB/decade due to the adjacent zero ω_z. When $\omega > \omega_z$, $|h_{21}|$ is equal to unity as predicted in (11). Small deviations of $|h_{21}|$ for $\omega > \omega_z$ from unity in Fig. 5 are believed to be caused by the distributed circuit elements at very high frequencies which are represented by the lumped elements in the hybrid-π equivalent circuits. Consider, as a sample calculation of τ_R, the case where $V_{BC} = 0.78$ V. Then from Fig. 5, $20 \log_{10} |h_{21} (\omega = 0)| = 9.2$ or $|h_{21}(\omega = 0)| = 1 + \beta_R = 2.88$. Thus $\sqrt{2}/\gamma$ in (15) is 1.336 and from Fig. 5, $\omega_p = (2\pi)\ 3 \times 10^7 = 1/(1.88\tau_R)$ which results in $\tau_R = 2.8$ ns. The reverse transit time τ_R changes slightly with V_{BC}, but for the constant τ_R of the Gummel–Poon model used by most circuit simulators, the extraction of τ_R is recommended near the operating bias point. The constant τ_R model is usually used for large-scale integrated circuit simulations which require fast computational time without sacrificing the accuracy.

IV. Conclusions

The importance of τ_R of bipolar transistors in the BiCMOS driver has been investigated and a new method for determining τ_R has been presented and demonstrated. It is based on h_{21} measurements in the reverse-active mode of operation.

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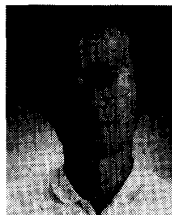

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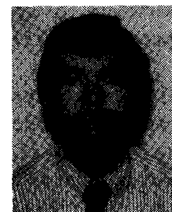

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