# An NBTI model for efficient transient simulation of analogue circuits

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## Abstract

To date the only model for the NBTI and PBTI effects that support continuously variable stress input values is based on molecular processes involving diffusing hydrogen. This model and its abstractions are either too imprecise or computationally inapplicable for long time analyses. We present a different approach to BTI models for analogue circuits on transistor level. Our model is based on and backwards compatible with a proven digital-only model, which uses loading and unloading capacitors. We have developed a mechanism that allows error controlled and efficient simulation of transients for transistor circuits over long time scales. This mechanism, based on a predictor-corrector method, exploits the structure of our new BTI model.

## 1 Introduction

Many attempts have been made to handle the problem of (negative and positive) **b**ias **t**emperature **i**nstability (referenced as just BTI in the following) caused threshold voltage shift. Just about all of them come up with a simple model that can be fit against stress curves resulting from (often interrupted) *constant* voltage stress. Unfortunately this is of no use when running circuits in analogue settings. In the generic case no two transistors of an analogue circuit are stressed in the same way. In particular, when simulating monitoring or compensating mechanisms (like [GDW08, QS08]), the stress level may change significantly during a circuits lifetime. This potentially happens to all of the devices, while any of them might be mission-critical.
At first glance it seems necessary to simulate the physical and chemical processes during transient analysis. This would take into account the actual voltages and the present state of the device. Thus, such practice promises most accurate results. We will sum up reason why this could be misleading in Section 2.2. The impact of stress caused in analogue circuits could be handled if all sorts of input signals are classified, and for all classes the effects of BTI are already known. We discuss the intractability of this (*binning*) approach in the BTI case in section 2.1.
Our proposal is based on a semi-abstract model introduced originally by [SGRG10] and based on physical observations presented in [GKG+]. We have extended this model to be able to process analogue stress during transient analysis. While we skip the dependency on temperature, we introduce an analogue model that copes with arbitrary input signals. Our model extends the switching DC model – in a sense that it reproduces its behaviour when exposed to switching stress (see Section 3.1).
At constant stress the BTI model allows for direct extrapolation of the device aging state, i. e. by evaluation of a closed expression. This is possible due to the separation of stress time constants as proposed in [SGRG10] for switching DC stress modelling. Similar closed expressions can be used to extrapolate aging information if our model is subject to arbitrary analogue stress. In this case more caution and an error estimating mechanism is needed.
So far the aging simulator has been implemented into a modern post-SPICE circuit simulator [Dav03], which is part of the gEDA project. With a predictor-corrector mechanism we achieve quite fast simulation results. In particular, we are able to evaluate the BTI model, see Section 4. Although the simulation can be controlled using SPICE-syntax, all the aging models are still hardwired in C++.

## 2 BTI-stress modelling

### 2.1 Binning and existing methods

Classification of stress input signals (of sensible duration) by its BTI impact allows preprocessing of the stress mechanism. This is particularly desirable, since the physical model is computationally intensive. If this can be done, a transient simulator could measure a stress voltage function during simulation and look up the associated change of device parameters and state in a table. To make use out of this, the number of possible stress features needs to be reduced to keep the table small.
An optimist would wish that an arbitrary stress signal (like a sine period with a certain offset) could be abstracted to an equivalent DC stress value and that the

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device state could be represented by a few numbers. In that case, a lookup table could possibly speed up simulation. Such an approach requires the verification of the abstraction in question. In order to do this, in addition to the method itself, a vast number of data is required, which must be measured or be simulated. We thus refrain from premature simplifications of the BTI mechanism, at least unless a fast simulation method has gained enough credibility.

Approaches to circuit aging like [TRL+91] use a sort of binning and preprocessing mechanism. These to the best knowledge of the authors don't support analogue stress BTI modelling. Other commercial tools are found to even mispredict the switching stress BTI effect. An example for such a case can be found in [SGRG10, page 1], where the simulator in question remains anonymous.

### 2.2 Diffusion reaction model

The diffusion-reaction model, as introduced 1977 in [JS77], originally elucidated the BTI effect. It has been widely accepted and verified, that hydrogen vanishing from the interface between gate oxide and the polysilicon channel accounts for the observed threshold voltage shift. The hydrogen placed and located at the so called *defect sites* leaves behind unbound electrons, which act as p-doping. Jeppson and Svensson observed how the vanishing rate is dependent on the applied voltage, which controls the reaction between bound and unbound hydrogen, and dependent on the speed at which the gas diffuses through the oxide. We will briefly sum up reasons, why the R-D model and its abstractions are probably not suitable for long-term analogue transient simulations.

Let $\sigma$ denote the number of defect sites and $\rho_x$ the concentration of free hydrogen at distance $x$ from the interface. Following [JS77], for fixed temperature and voltage, there are constants $A$ and $B$ such that the evaporation of hydrogen at time $t$ complies to

$$\frac{\mathrm{d}}{\mathrm{d}t}\sigma(t) = -A \cdot \sigma(t) + B \cdot \rho_0(t) \cdot (\sigma(t) - \sigma(0)), \quad (1)$$

while hydrogen moves through the gate satisfying the heat equation,

$$\frac{\partial}{\partial t}\rho = \alpha \cdot \Delta\rho, \quad (2)$$

where the diffusivity $\alpha$ is subject to the temperature. The original approach included the simplifying assumption of an infinitely thick gate oxide. To adapt the model to real transistors, it is necessary to find sensible conditions for the opposite boundary. Several approaches, which we wont discuss here, exist [OS95, KCS, AM05].

Simulating the hydrogen diffusion reaction caused by an electrical field implies keeping track of the concentration of hydrogen inside the gate and at the interface. In [GET+06] a simulation of the switching case is realised. The number of grid points inside the oxide is numbered above 30, although it seems more likely, that above $100$ are needed (see [GET+06, Fig. 5]). The main reason for numerical instability is the interface region. Here, at constant voltage, a numerical solver approaches the quasi-stable state after a few iterations. This completely fails when processing analogue signals. Experiments carried out, give rise for the need of a grid much finer and timesteps much shorter than in the switching case.

Long term simulation of aging effect heavily relies on a predictor mechanism. A predictor for the state of the hydrogen distribution is a difficult problem. Consider the hydrogen state after a transient run containing one sine period. There is no way known to the authors on how to predict a later state from this data All in all, there is enough evidence that it makes sense to discard a hydrogen simulation or any of its simplifications when doing circuit level simulations.

# 3 A statistical defect site model

The idea of viewing the NBTI defect state as the sum of defects seems very natural with regard to the original reason of the observed threshold voltage shift – defects on the interface.

A rather new observation is that every defect site has its own characteristic emission and capture behaviour, both for the emission (at a fixed electrical field) and for the recapture of hydrogen Fortunately, the contribution of the defects to the threshold voltage can be assumed additive, such that its total amounts to

$$S_{\mathrm{tot}} = \sum_{d \in \{\text{defect sites}\}} S_{\mathrm{d}} \cdot w_d,$$

where $S_{\mathrm{d}}$ indicates if the hydrogen is still there ($S_{\mathrm{d}} = 0$) or gone ($S_{\mathrm{d}} = 1$), and $w_d$ denotes the voltage shift caused by that particular defect site. This perspective has evolved from a series of physical observations [GKG+, KGMM+09, SGRG10].

Interestingly, no hydrogen diffusion is needed to justify the workings of the defect states. In particular this newer defect site state model contradicts the original diffusion reaction model. A correspondence can to our knowledge only be established by fitting them onto each other. A reason, why the latter approach works so well might be the decreasing thickness of the gate oxide layer.

The model for switching DC-stress proposed in [SGRG10] can be explained by interpretation of the $S_{\mathrm{d}}$ as random variables. Here, exponential decay processes are used to model the emission and recapture. The exponential decay, which e. g. is known from radioactive decay is a positive real valued random variable in one parameter $\mu$. For us, the real value is time and $\mu$ is the expected value for an event to happen. An exponentially distributed random variable with parameter $\mu$ has

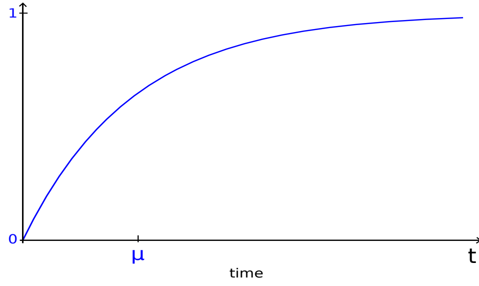


**Figure 1** Exponential decay (CDF) with parameter $\mu$. Probability for an even to happen before $t$.

the distribution $\mathrm{Exp}_\mu \colon t \mapsto 1 - \exp(-t/\mu)$ (see Fig. 1). This, in our model at $t$, evaluates to the probability, that an event occurs before time $t$. So for any time $t$ we may say that the event has happened with probability $\mathrm{Exp}_\lambda(t)$ before $t$.
In combination with an input signal it makes sense to evaluate $\mathrm{E}_t(S_{\text{tot}})$, the expected value of $S_{\text{tot}}$ at time $t$. The big number of defect sites allows the interpretation of $\mathrm{E}_t(S_{\text{tot}})$ as threshold voltage shift. Lets fix a reference voltage $u_{\text{ref}}$ which the model is supposed to work at. For clarity, we have chosen this voltage to always be the absolute value of the stress voltage, such that a bigger $u_{\text{ref}}$ always corresponds to more stress.
Say $f$ is a step function in time mapping to the set $\{0, u_{\text{ref}}\}$, changing its value at $0 = t_0 < t_1 < t_2 \cdots$. So formally we have $f(t) = ((-1)^i + 1)u_{\text{ref}}/2$ whenever $t_i \leq t < t_{i+1}$ (see Fig. 2). We will now calculate the expected value $\mathrm{E}_t(S_{\text{d}})$ for a fixed defect site $d$ at time $t$.
At voltage $u_{\text{ref}}$ let $\tau_{\mathsf{e}}^\bullet$ denote the expected emission time of d and let $\tau_{\mathsf{c}}^\bullet$ denote the corresponding capture time (at voltage zero). Trivially, we have $\mathrm{E}_0(S_{\text{d}}) = 0$. For $0 < t < t_1$ the probability for $S_{\text{d}} = 1$ at time $t$ is the probability for an emission until time $t$, which in our model is $1 - \exp(-\tau_{\mathsf{e}}^\bullet \cdot t)$. This also is $\mathrm{E}_t(S_{\text{d}})$, the expected value for $S_{\text{d}}$ at time $t$, since the complementary event maps to 0. In general, if $t_i < t \leq t_{i+1}$ we have

$$\mathrm{E}_t(S_{\text{d}}) = \left(\mathrm{E}_{t_i}(S_{\text{d}}) - \frac{(-1)^i + 1}{2}\right) \cdot \exp\left(\tau_i^\bullet \cdot (t - t_i)\right) + \frac{(-1)^i + 1}{2}, \qquad (3)$$

where in the index of $\tau^\bullet$ we identify even numbers with e and odd ones with c.
PROOF: In the case $0\text{s} \leq t \leq t_2$ the expression evaluates to our previously deduced expected values. Now let $i$ be odd, then $\mathrm{E}_t(S_{\text{d}})$ is the probability of $S_{\text{d}} = 1$ at $t = t_i$ plus the probability of $S_{\text{d}}$ to emit between $t$ and $t_i$ given $S_{\text{d}} = 0$ at $t$. One verifies that this is the sum $\mathrm{E}_{t_i} + (1 - \exp(\tau_i \cdot (t - t_i))) \cdot (1 - \mathrm{E}_{t_i})$ and complies with our claim. Otherwise $i$ is even. Then the expected value at time $t$ is the probability for $S_{\text{d}} = 1$ at $t$ multiplied by the probability of *no* capture between $t_i$ and $t$ – all other events map to zero. A capture happens with probability $1 - \exp(\tau_i^\bullet \cdot t - t_i)$, so the resulting product is $\mathrm{E}_{t_i}(S_{\text{d}}) \cdot \exp(\tau_i^\bullet \cdot (t - t_i))$. □
Parametrizing the switching model for BTI means figuring out the time constants and the impact of each defect site. These parameters can be found approximatively with the following consideration: Assume we know the time constants of our (finitely many) defect sites, let $\mathcal{E}, \mathcal{C}$ denote the set of occurring emit time constants and capture times, respectively, For simplicity we may assume that the pairs $(e, c) \in \mathcal{E} \times \mathcal{C}$ correspond to the defect sites. For $e \in \mathcal{E}$ denote $f_e$ for the function that maps $t$ to $u_{\text{ref}}$ if $t < e$, and to $0\,\text{V}$ otherwise. For fixed parameters $w_{\text{d}}$ (fixed for all d), we get expected values for $S_{\text{tot}}$ at all combinations $(e, c) \in \mathcal{E} \times \mathcal{C}$. Precisely, we get a *linear* map $\Lambda$ taking $|\mathcal{E}| \cdot |\mathcal{C}|$ weights and returning the same number of values for $S_{\text{tot}}$, namely the values

$$\mathrm{E}_{c+e}^{f_e}(S_{\text{tot}}) = \sum_{d \in \{\text{defect sites}\}} \mathrm{E}_{c+e}^{f_e}(S_{\text{d}}) \cdot w_d$$

(Here we use the symbol $\mathrm{E}_t^f$ for the expected value at $t$, using $f$ as stress input). This map $\Lambda$ can be seen to be bijective. So conversely, we may measure the voltage shift at all time pairs $(e, c)$. For these values, which we take for the corresponding expected values, we compute the inverse image under $\Lambda$ and get weights $w_d$ for every site (i. e. for all time pairs). These weights exactly model the measured expected values. This method completely ignores the sensible choice of the time constants and lacks theoretical assessment in terms of bias investigation. A probably similar approach based on the same data is presented in [SGRG10].

## 3.1 An analogue model

By now, we have worked out that the statistical model inherently makes sense for the switching case. A useful generalization which is capable of analogue stress input processing, thus should reproduce the behaviour described in the last section, when treated with a switching DC-signal of any amplitude. Also an analogue model should be parameterized in a way that allows to reuse the time constants already known for switching DC-stress (at a fixed voltage $u_{\text{ref}}$).
Our new approach again models the expected value for the state of a defect site. And again a defect site state is a random variable, mapping to $\{0, 1\}$. Since the random process time constant is voltage dependent, we choose a voltage dependent time constant $\tau_{\mathsf{e}}$ that is monotonously falling. So, if the hydrogen at the defect site is present, the probability for it to be gone after an infinitesimal time $\mathrm{d}t$ is $1 - \mathrm{d}t/\tau_{\mathsf{e}}(u)$ if we apply $u$. Monotonicity simply assures that a higher voltage causes a faster defect rate. Here, due to the continuous nature of $u$ it doesn't make sense to switch this random process on or off. We rather introduce a second comple-

mentary process, the recapture of gone hydrogen. After infinitesimal time $\mathrm{d}t$ the hydrogen recaptures with probability $1 - \mathrm{d}t/\tau_\mathsf{c}(u)$. The map $\tau_\mathsf{c}$, a monotonously rising function, controls its speed.
Let us evaluate the model for a switching input voltage. For this, assume, that at time 0, the defect site is defect with probability $p_0$, that is the probability of the hydrogen to be there is $1 - p_0$. Now we apply a fixed voltage $u_\text{on}$ and observe that the probability of the hydrogen to be gone at infinitesimal time $\mathrm{d}t$ equals

$$p(t + \mathrm{d}t) = p(t) \cdot (1 - \mathrm{d}t/\tau_\mathsf{c}(u_\text{on})) + (1 - p(t)) \cdot \mathrm{d}t/\tau_\mathsf{e}(u_{on}).$$

That is the sum of the probability, that the hydrogen is gone and not coming back, and the probability of it to be there and about to leave. We immediately derive the differential equation for $p$,

$$\frac{\mathrm{d}p}{\mathrm{d}t}(t) = \lim \frac{p(t + \mathrm{d}t) - p(t)}{\mathrm{d}t} = -p(t)(1/\tau_\mathsf{c} + 1/\tau_\mathsf{e}) + 1/\tau_\mathsf{e}. \quad (4)$$

The solution for $p$ is an exponential function $t \mapsto \ell \pm exp(-t/\tau)$, where $\ell(u) = \tau_\mathsf{c}(u)/(\tau_\mathsf{c}(u) + \tau_\mathsf{e}(u))$ is the limit for $t \to \infty$ and

$$\tau(u) := \tau_\mathsf{e}(u)\tau_\mathsf{c}(u)/(\tau_\mathsf{e}(u) + \tau_\mathsf{c}(u)) \quad (5)$$

is the **resulting time constant**, depending on $u$.
Independent of the choice of $\tau_\mathsf{e}$ and $\tau_\mathsf{c}$ we get back the behaviour of a switching DC-model as follows. Starting with the voltage 0, we find that the expected value for the defect site state converges to $\ell(0)$, not 0 (as before). So we will have to start at $p(0) = \ell(0)$, for a fresh device. Similarly the $\ell(u_\text{ref})$ is not 1 anymore. But still these inputs make the modeled probability converge with the time constants $\tau(0)$ and $\tau(u_\text{ref})$, respectively. The time constants don't change, if we apply an affine transformation, namely

$$\lambda\colon p \mapsto (p - \ell(0))/(\ell(u_\text{ref}) - \ell(0)), \quad (6)$$

nor does the exponential shape of the time evolution change. Consequently, interpretation of the value of $\lambda$, weighted by some $w$ as the voltage shift contribution, yields the same as before (Fig. 2).
Note that here, the initial expected value for a defect site state is allowed to be positive. This makes sense, since all we know is some empirical data about reaction rates, and it might well be true, that at any time or voltage the real state is a sort of equilibrium.
After all we consider our model to be semi-abstract based on the following opinion. The modeling of actual happenings with dopants at defect sites together with the physical view on the impact of single defects make our concept look fairly concrete. On the other hand, we abuse the expected values of a defect site state random

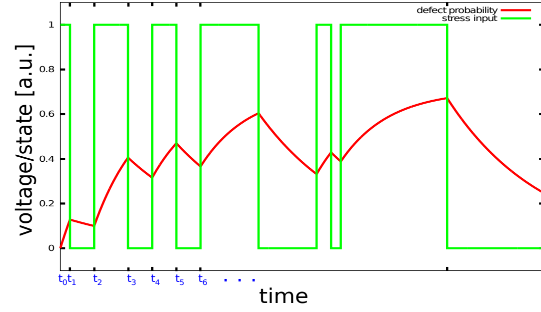


**Figure 2** Switching input signal (dotted) and resulting probability for defect.

variable for the real share in voltage shift. Also, we use two competing hypothetical stochastic processes and only compute the resulting expected behaviour. The partition of defect sites into fixed bins of equal time constants – which we adapted from the switching stress model – further abstracts the concept.

## 3.2 The time constants

The definite voltage dependency of $\tau_\mathsf{e}$ and $\tau_\mathsf{c}$ can hardly be measured directly, apart from lack of instruments. Note that from our perspective, prior measurements, eġ. in [RGGS10], rather measure what we call the *resulting* time constant (5).
Our goal is simulation and analysis of circuitry. For this we present a conceptual ready-to-use time constant model reusing and reproducing the time constants of a given switching stress model.
Lets take a switching model working at $u_\text{ref}$ with emission time constant $\tau_\mathsf{e}^\bullet$, capture time constant $\tau_\mathsf{c}^\bullet$ and weight $w$. For simplicity, we set $u_\text{ref} = 1\text{V}$, since we want to present working a concept, rather than involved formulas. Also, as announced we skip temperature dependence of the time constants, since at present it is not totally clear, how temperature affects both the emission and recapture rates.
Using the parameters $\alpha$ and $v_0$, we choose the voltage dependency model to be

$$\tau_\mathsf{c}(u) = \tau_\mathsf{c}^\bullet \cdot \exp(\alpha \cdot u) \text{ and} \quad (7)$$

$$\tau_\mathsf{e}(u) = v_0/u, \infty \text{ for } u \leq 0 \quad (8)$$

for the time constants. The exponential dependency of $\tau_\mathsf{c}$ on $u$ is motivated by the reaction model, where the reaction speed is exponential in the electric field. Nevertheless, this is an abstract model and it requires further investigation to map these functions to real physical processes. For now, to set $\alpha$ and $v_0$, we choose a fitting parameter $s$ and set

$$\alpha = \ln(s \cdot \tau_\mathsf{e}^\bullet/\tau_\mathsf{c}^\bullet + 1) \text{ and}$$
$$v_0 = (\tau_\mathsf{c}^\bullet * \tau_\mathsf{e}^\bullet + s \cdot \tau_\mathsf{e}^{\bullet 2})/(((s-1) * \tau_\mathsf{e}^\bullet + \tau_\mathsf{c}^\bullet)).$$

This ensures that $\tau(0) = \tau_\mathsf{c}^\bullet$ and $\tau(u_\text{ref}) = \tau_\mathsf{e}^\bullet$, as the reader may verify using (5).

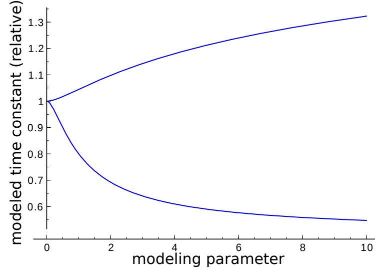


**Figure 3** Dependency of $\tau$ on the model parameter $s$, we chose $\tau_{\mathrm{c}}^{\bullet} = \tau_{\mathrm{e}}^{\bullet} = 1$, and $u = u_{\mathrm{ref}}/2$ (upper plot) and $u = 2u_{\mathrm{ref}}$ (lower plot)

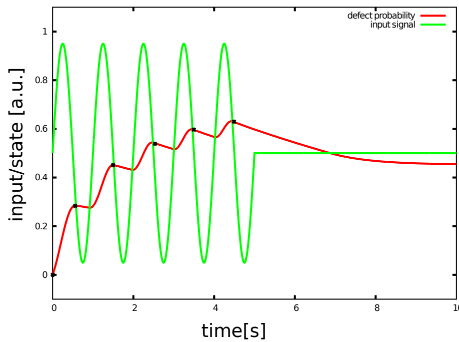


**Figure 4** Sine input signal (dotted) and resulting probability for defect.

We see that this example is particularly simple, as it only allows positive input $u$. It also satisfies $\ell(0) = 0$. At $s = 1$ we have $\ell(u_{\mathrm{ref}}) = \tau_{\mathrm{c}}^{\bullet}/(\tau_{\mathrm{c}}^{\bullet} + \tau_{\mathrm{e}}^{\bullet})$. The parameter $s$ controls the voltage dependency outside the points $0, u_{\mathrm{ref}}$. For example a bigger value for $s$ increases the time constant $\tau(u)$ for $0 < u < u_{\mathrm{ref}}$ and decreases $\tau(u)$ for $u_{\mathrm{ref}} < u$ (see Fig. 3).
Put into a transient simulator, this model yields the expected value for the defect after arbitrary stress, see Fig. 4 for an example. Unlike in Fig. 2, this time, there is no simple closed expression for the modeled value. That's why we have developed a way to speed up simulation.

# 4 The simulator

A degradation based on transient stress evaluation is worthless without a simulator. We introduce a simulation method that speeds up transient simulation of arbitrary aging effects, given that the model supports a specific interface.
Consider a periodic input signal, where one period has length $\Delta t$. We want to simulate the transient until Time $T \gg \Delta t$, taking into account the aging that happens during this time interval.
The simulator is derived from a usual transient simulator, around which we have implemented a loop (Fig. 5). The simulator does a `transient` simulation until $\Delta t$, during which the aging state (for us, the probabilities

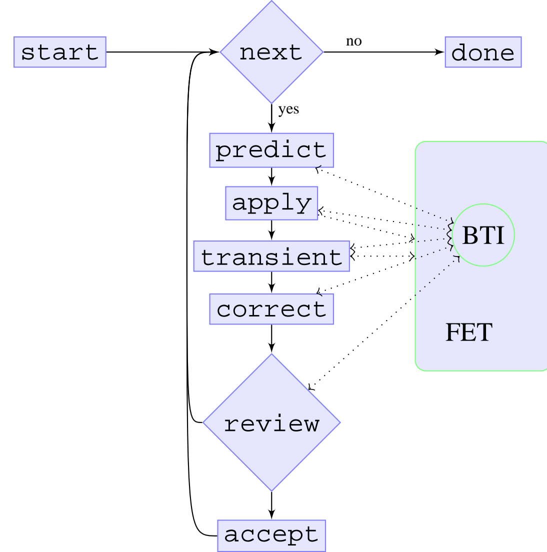


**Figure 5** Simplified flow and interfaces for the simulator

at the defect sites) at $\Delta t$ is computed. Some algorithm chooses a `next` time $t_1 \geq \Delta t$ and the `predictor` extrapolates the aging state to $t_1$ using the model and the history. Then, with the aging `applied` to the devices, another transient simulation is done until $t_1 + \Delta t$. After this simulation, the `corrector`, now knowing the transient values of the second simulation, ensures the consistency of the simulated values with the device model. If this is possible, the simulator `accepts` and does another simulation at $t_2 > t_1 + \Delta t$. Otherwise the simulation data at $t_1$ will be discarded and another simulation will be performed at a smaller $t_1$. The boxes on the right stand for the implementation of the BTI-device as a subdevice of the transistor. The dotted arrows roughly indicate the communication between the simulator core and the devices.

## 4.1 The aging model interface

The aging model interfaces to the transient simulator just as a usual device. For our BTI model, this means that in every transient time step the probabilities for the device states are updated corresponding to the voltage applied to the FET gate. It furthermore needs to provide functions doing the predictor/corrector and review routines.
In the example of the BTI model, the predictor fits the values for the state at the boundaries of the transient runs to a meta-model (in contrast to simple extrapolation). This meta model is inspired by the behaviour of the switching DC-model. Consider a periodic switching signal and its effect on the modeled expected value. It can be seen that this value, sampled only after each period, lays on an exponential curve. This exponential curve can be written as a function $t \mapsto \ell \pm \exp(-\alpha t)$. We also use this as envelope for the analogue stress

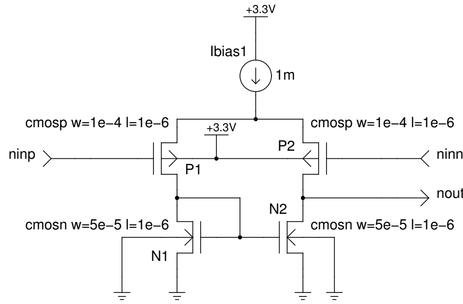


**Figure 6** A CMOS comparator

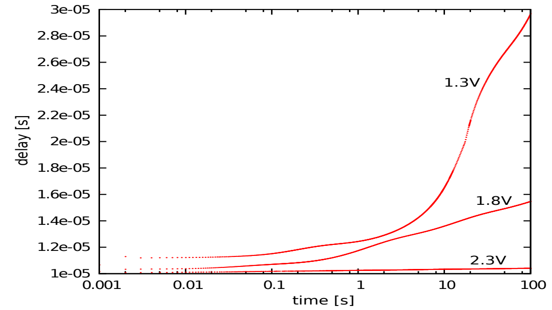


**Figure 7** Delay over time of the crossing of $v_{\mathrm{dd}}/2$ at $u_0 = 1.3, 1.8, 2.3\mathrm{V}$

model, since we expect the states in question to behave like this (see the crosses in Fig. 4).
Fitting to the proposed meta-model roughly works like this: After we have computed a transient for one period (i. e. only until to the second cross from the left), we extrapolate and go on doing transient at a later time $t$ (this way skipping the transient simulation of several periods). So we predict the probability state at $t$ and do another transient run starting from $t$. Knowing the difference in probability for two periods starting at two different states we can derive the parameters for the meta-model, and correct the device state at $t$ retroactively. This way, depending on the situation, we can gap some 100 to 1000 times the period length of a transient run with high accuracy. To control the size of the gap, the difference between the expected and the afterwards modeled state has proven to be a good measure.

# 5 Examples

Threshold voltage shift can be observed easily in a comparator circuit. A comparator (Fig. 6) takes two input voltages (at ninn and ninp) and outputs a high or low voltage (at nout) depending on which of the input voltages is bigger. If the transistors are affected by different $V_{\mathrm{th}}$ shift, the comparison trivially is biased by the difference of these shifts. To demonstrate the simulator we apply a constant input $u_0$ to ninp and a sine wave (amplitude $0.2\mathrm{V}$ at offset $u_0$ and 10ms period) to ninn. We monitor the aging indirectly by measuring the time of the output crossing $\mathrm{vdd}/2$ for the first time. We use the BTI model DC parameters from [SGRG10, Fig. 13] to calibrate the analogue model.
We observe that with our model parameters, the threshold voltage for the pFET exposed to the sine is bigger than the one exposed to DC stress. Also the effect crucially depends on $u_0$. Note that the bumps originate from the discretization of the time constants, as similarly observed in in [SGRG10, Fig. 10].

# 6 Conclusion and future plans

The paper presents a method for modeling and efficiently simulating NBTI and PBTI effects for analog circuits with high accuracy. We have shown that our model can handle analog signals, degradation and recovery in a reasonable way. It can efficiently be evaluated in a circuit simulator, which on the other side has been extended to handle the different time scales of degradation and analog signal processing. The experimental results show that analog excitation leads to individual signal dependent degradation of MOS transistors noticeably changing the overall circuits' performance.
Next steps are further development of simulator control enabling larger degradation time scales and investigations in identifying NBTI-sensitive analog circuits. With the ability of accurate simulating degradations effects, like parameter drifts (including annealing) of analogue blocks, we further work on incorporating these effects into behavioral models of these blocks.

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