

Analysis and Mitigation of Multiple Radiation Induced Errors in Modern Circuits

by

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A Dissertation

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# DISSERTATION APPROVAL

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in the field of Electrical and Computer Engineering

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## AN ABSTRACT OF THE DISSERTATION OF

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Due to technology scaling, the probability of a high energy radiation particle striking multiple transistors has continued to increase. This, in turn has created a need for new circuit designs that are can tolerate multiple simultaneous errors. A common type of error in memory elements is the double node upset (DNU) which has continued to become more common. All existing DNU tolerant designs either suffer from high area and performance overhead, may lose the data stored in the element or are vulnerable to an error after a DNU occurs which makes the devices unsuitable for clock gating. In this dissertation, a novel latch design is proposed in which all nodes are capable for fully recovering their correct value after a single or double node upset which is referred to as DNU robust. The proposed latch offers lower delay, power consumption and area requirements compared to existing DNU robust designs.

Multiple simultaneous radiation induced errors are a current problem that must be studied in combinational logic. Typically, simulators are used early in the design phase which use a netlist and rudimentary information of the process parameters to determine the error rate of a circuit. Existing simulators are able to accurately determine the effects when the problem space is limited to one simultaneous error. However, existing methods do not provide accurate information when multiple concurrent errors occur due to inaccurate approximation of the glitch shape when multiple errors meet at a gate. To improve existing

error simulation, a novel analytical methodology to accurately determine the pulse shape when two simultaneous errors occur is proposed. Through extensive simulations, it was shown that the proposed methodology matches closely with HSPICE while providing a speedup of 15X.

The analysis of the soft error rate of a circuit has continued to be a difficult problem due to the calculation of the logical effect on a pulse generated by a radiation particle. The most common existing methods to determine logical effects use either exhaustive input pattern simulation or binary decision diagrams. The problem with both approaches is that simulation of the circuit can be intractably time consuming. To solve this issue, a simulation tool is proposed which employs partitioning to reduce the simulation time and space overheads associated with the use of binary decision diagrams. In addition, the tool integrates an accurate electrical masking model. Compared to existing simulation tools, the proposed tool can simulate larger circuits faster.

## DEDICATION

This dissertation is dedicated to my parents Daniel and Debra Burris and my wife Yijing Zhang Watkins for their support during the creation of this work.

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# TABLE OF CONTENTS

Abstract . . . . .	ii
Dedication . . . . .	iv
Acknowledgments . . . . .	v
List of Tables . . . . .	vii
List of Figures . . . . .	viii
Introduction . . . . .	1
1 Robust Latch Designs for Multiple Node Upsets . . . . .	7
1.1 Background on Existing DNU Tolerant Latches . . . . .	8
1.2 Proposed HRDNUT Latch Design . . . . .	10
1.3 Simulation Results . . . . .	18
1.4 Filtering of Transients Arriving on the Latch Input . . . . .	20
1.5 Analysis of a Triple Node Upset Tolerant Latch . . . . .	22
1.6 Conclusion . . . . .	26
2 An Analytical Pulse Approximation Algorithm for Multiple Event Transients . .	28
2.1 Pulse Approximation Model . . . . .	29
2.2 Experimental Results . . . . .	35
2.3 Conclusion . . . . .	40
3 Accurate Soft Error Simulation Using Partitioning . . . . .	41
3.1 Preliminaries . . . . .	44
3.2 Description of the FAST_MET Simulator . . . . .	50
3.3 FAST_MET Simulation Flow . . . . .	54
3.4 Results . . . . .	60
3.5 Conclusion . . . . .	61
Appendix . . . . .	67
Vita . . . . .	73

## LIST OF TABLES

1.1	SPICE Simulations of Existing Latches using the 1.05V 32nm PTM library . . . . .	20
2.1	Proposed and [33] compared to HSPICE for 2400 points. . . . .	38
2.2	Execution Time vs Accuracy ( $\times 10^{-3}$ ) . . . . .	38
3.1	Functions for Transient Pulse Generation . . . . .	46
3.2	Functions for a MET Arriving on the Inputs of a NAND Gate . . . . .	48
3.3	Comparision of FAST_MET vs Monte Carlo . . . . .	60
3.4	FAST_MET Performance vs Number of Partitions . . . . .	61



## LIST OF FIGURES

1	A energetic particle creating electron-hole pairs in a transistor. . . . .	1
2	A particle striking two node concurrently causing a double node upset. . . . .	2
3	An example of a radiation induced transient pulse. . . . .	4
4	Two pulses arriving at a gate input. . . . .	5
1.1	HSMUF latch [35] with a weak keeper on the output. . . . .	10
1.2	DONUT latch as proposed in [6]. . . . .	11
1.3	Modified low-power DONUT latch. . . . .	11
1.4	Basic data storage loop block. . . . .	12
1.5	Schematic of the block-based latch. . . . .	13
1.6	Waveforms of the HRDNUT latch during normal operation. . . . .	15
1.7	Schematic of the HRDNUT latch. . . . .	16
1.8	The pulse filtering circuit provided in [10]. . . . .	21
1.9	The enhanced pulse filtering circuit that can tolerate an error. . . . .	21
1.10	Block based latch that forms the basis for a TNU tolerant design. . . . .	23
1.11	Schematic of the TNU tolerant latch. . . . .	25
2.1	(a) The transistor model used for a NMOS. (b) The transistor model for a PMOS.	30
2.2	(a) The transistor model used for a NMOS. (b) The transistor model for a PMOS.	31
2.3	A generalized representation of the gate model. . . . .	32
2.4	(a) Three stacked NMOS transistors with two intermediate nodes $V_{N1}$ and $V_{N2}$ . (b) The transistors converted to the proposed circuit model. . . . .	35
2.5	(a) Output of NAND2F with 300 points. (b) Output of NAND2F with 900 points . . . . .	39
2.6	(a) Output of NAND3F with 300 points. (b) Output of NAND3F with 900 points . . . . .	39
3.1	A low-high-low transient pulse being generated by a strike on the PMOS. . . .	45

3.2	A transient pulse being masking by a controlling value on the off-input. . . . .	47
3.3	(a) A transient pulse arriving on input a and masked on b. (b) A pulse arriving on input b and masked on a. (c) Two pulses arriving on the inputs simultaneously.	48
3.4	The truth table and corresponding BDD for an AND gate. . . . .	51
3.5	An example of the BDD creation for a rising and falling pulse. . . . .	53
3.6	An example on c17 showing how the pulses are propagated between partitions.	57
3.7	Waveforms for CLK and D. . . . .	68
3.8	Node pair n1 and n2 upset and recovery. . . . .	68
3.9	Node pair n2 and out upset and recovery. . . . .	69
3.10	Node pair n1 and n5 upset and recovery. . . . .	69
3.11	Node pair n3 and n4 upset and recovery. . . . .	70
3.12	Node pair n4 and out upset and recovery. . . . .	70
3.13	Node pair n1 and n3 upset and recovery. . . . .	71
3.14	Node pair n1 and n6 upset and recovery. . . . .	71
3.15	Node pair n5 and out upset and recovery. . . . .	72
3.16	Node pair n3 and out upset and recovery. . . . .	72

## INTRODUCTION

The continued reduction of the transistor feature size has led to the increase in the vulnerability of modern circuits to error. This effect, in turn, has made modern circuits more susceptible to radiation induced errors in space and terrestrial environments. A radiation induced error, commonly referred to as a soft error, occurs when a high energy particle from space or packaging strikes a transistor. Shown in Fig. 1, the particle arrives externally and deposits energy in the active volume generating electron-hole pairs. This creates a new diffusion region that could allow a non-conducting device to temporarily conduct current. The mechanism causes a temporary voltage pulse, referred to as a single event transient (SET) to occur at the device. If the SET occurs in combinational logic it may propagate to a memory element causing an error in the system. Furthermore, the error can occur in the memory element causing the stored data to be corrupted.

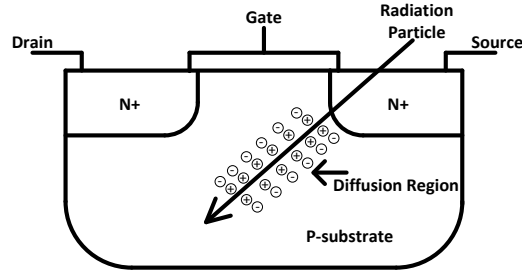


Figure 1. A energetic particle creating electron-hole pairs in a transistor.

In the radiation effect on memory element will be discussed. If a memory element loses its value due to a strike by a single particle, this is referred to as a single event upset (SEU). Additionally, due to the constant scaling down of the feature size, a single particle can also strike two transistors simultaneously which is referred to as a double node upset (DNU). Fig. 2 gives a diagram of a SEU tolerant latch to demonstrate the DNU phenomenon. In the diagram, radiation is denoted as a high energy particle which passes through two transistors on a DICE latch [3]. In the case of a SEU, the DICE latch would normally be

able to tolerate the error. As can be observed, the particle passes through two transistors causing the respective nodes to switch from "1" on the first node to "0" and from "0" to "1" on the second node. The upset of both nodes drives the remaining two nodes to an erroneous value. This observance is alarming since it signals that current designs are not sufficient for future processes. In turn, this provides a need for new memory element designs that can tolerate DNUs.

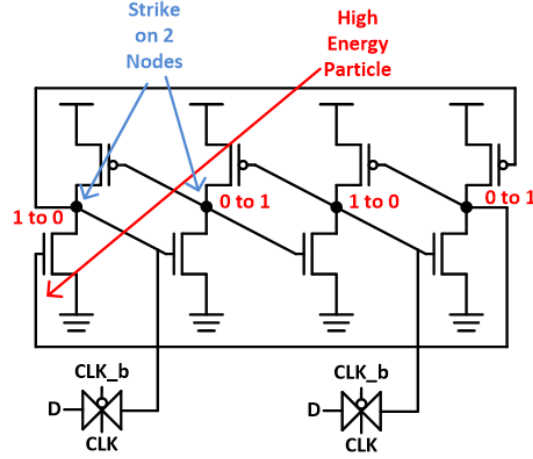


Figure 2. A particle striking two node concurrently causing a double node upset.

In addition, many modern circuit designs employ a technique referred to as clock gating. Clock gating is defined as setting the clock to constant value thus reducing the power consumed. Use of clock gating may lead to cases where the current state held by the memory element must be held for many clock cycles. This phenomenon leads to a much longer than normal vulnerable window in which an improperly hardened device may lose the stored value. While there are existing DNU tolerant designs [15, 16, 35], these designs move to a high impedance state after a DNU. If a DNU occurs while the latch is gated, the voltages within the latch may degrade thus losing the stored value. In Chapter 1, the aim is to alleviate this problem by providing an efficient design that is capable of recovering all nodes after a DNU occurs. This, in effect, guarantees that the data will be held even if the latch is struck by a DNU while gated.

In addition to the effects on memory elements, accurate approximation of the error rate

of a combinational circuit is also important. The goal of combinational circuit simulation is the accurate estimation of the soft error rate (SER). In Chapters 2 and 3 of this work, the focus is on the evaluation of the SER using near exact estimation methods on combinational circuits. Typically, this type of simulation entails the estimation of the resulting pulse shape when a particle strikes a transistor and during pulse propagation, the determination of the Boolean functions that allow the pulse to propagate and the estimation of the latching characteristics at an output flip-flop. Through accurate computation of the aforementioned effects, the probability of the pulse reaching an output flip-flop is determined. Assume that the probability of an error reaching and being latched in a flip-flop at output  $i$  of a circuit is represented as  $P(O_i)$ , the particle hit rate in a particular area is given as  $R_{PH}$ , the fraction of particle hits resulting in charge generation as  $R_{eff}$  and  $A_{cir}$  gives the area of the circuit. The equation for the SER at output  $O_i$  is given below [21].

$$SER_{O_i} = P(O_i) * R_{eff} * R_{PH} * A_{cir} \quad (1)$$

The focus of all SER estimation simulators is the calculation of the term  $P(O_i)$ . This is a difficult problem due to the presence of the following three masking factors: electrical masking, logical masking and temporal masking. Electrical masking deals with the calculation of the pulse shape as it propagates through the circuit. Logical masking is the estimation of the logical 1's and 0's and how they may mask the pulse. For example, if a NAND gate has a value of "0" on an input, the pulse will not propagate since the output will be held to a "1". Lastly, temporal masking involves the latching characteristics, specifically the set-up and hold times and how they relate to the pulse width and clock period, of the output flip flop. Accurate consideration of all three masking effects is crucial to accurate SER estimation.

Accurate consideration of the electrical masking effect is an important but often simplified component of SER estimation. The most common method to calculate the pulse

shape, proposed in [27] uses a linear ramp to approximate the rising and falling transitions and a straight line for the peak giving a trapezoidal shape. While this method executes quickly and is easy to implement, it does not accurately represent the realistic shape of the pulse. As can be observed in Fig. 3, a transient pulse does not take a trapezoidal shape in a real case. For this reason, accurate consideration of the pulse shape must consider the non-linear aspects of the pulse. In [33], the authors proposed an enhanced pulse approximation method which is accurate within 5% of HSPICE. However, their method has the drawback that it can only consider a single pulse arriving at a gate. This is problematic in modern circuit simulation due to the continued scaling of the feature size has increased the likelihood of a radiation particle generating two or more pulses simultaneously, referred to as a MET. In the case of a MET, the number of pulses within the circuit substantially increase leading to a higher likelihood of multiple pulses arriving a gate simultaneously as shown in Fig. 4. This trend implies that future soft error simulators must be able to accurately consider multiple pulses. In Chapter 2 an enhanced analytical pulse approximation algorithm which can consider any number of concurrent input pulses is proposed.

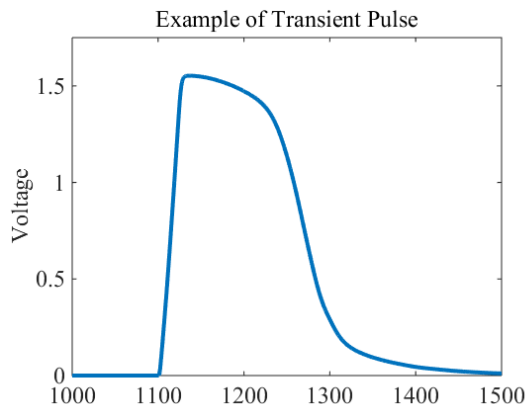


Figure 3. An example of a radiation induced transient pulse.

In addition to the electrical masking effect, the logical masking effect must also be accurately considered. Many existing methods use Monte Carlo based simulation which consists of either exhaustively apply all patterns or randomly applying a subset of patterns [33, 31, 24, 38, 30]. While these methods are able to simulate with a very low memory

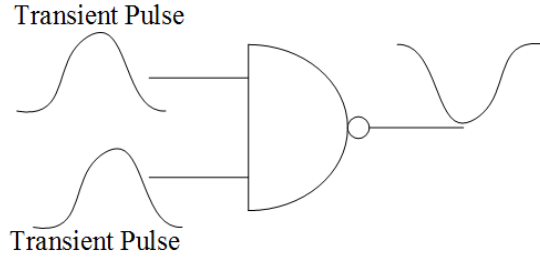


Figure 4. Two pulses arriving at a gate input.

overhead, they have an intractable execution time for circuits with more than 30 inputs. This problem is further compounded by the fact that due to the nearly infinite number of possible pulse durations and strike locations. To alleviate this issue, the authors in [39] propose the use of probabilistic transfer matrices [PTM]. While the matrices do provide exact calculation of the logical masking effect, their memory usage blows up even for relatively small circuits. As an improvement to the use of PTMs, the use of decision diagrams have become more common [36, 20]. Decision diagrams offer an improved approach since they balance the cost of execution time and memory consumption. However, decision diagrams are not a perfect solutions since, like PTMs, blow up if the circuit is sufficiently large. To solve this issue, the authors in [36] proposed the use of partitioning to limit the size of the decision diagrams. The drawback with this approach is that each time the circuit is partitioned, additional error is added to the result. To ensure that the effect on the result is low, the number of partitions should be minimized based on the system used and amount of time allowed.

In this dissertation, novel approaches to the aforementioned problems are proposed. In Chapter 1, a latch design is proposed which is capable of recovering all nodes after a DNU. This design has specific applications in circuits that employ clock gating. Chapter 2 provides an enhanced analytical pulse approximation algorithm which can determine the output pulse shape when two pulses arrive at a gate input. The method is implemented and compared to existing algorithms and HSPICE. Chapter 3 discusses a soft error simulator

which employs partitioning and the pulse approximation mode in Chapter 2. Results are provided which compare the effect of partitioning to Monte Carlo simulation for the calculation of the SER.



# CHAPTER 1

## ROBUST LATCH DESIGNS FOR MULTIPLE NODE UPSETS

The focus of this chapter is on the design of soft error hardened latches. There has been extensive research in the field of hardening latches against single even upsets (SEU). The simplest design for safety critical applications is the triple modular redundancy (TMR) latch. This design consists of 3 standard latches connected to a 3-input majority voting circuit. While this design is tolerant against errors, it has the drawback of high area, delay and power consumption. For this reason there have been many other designs proposed that offer high SEU reliability with lower area, delay and power consumption. The first and most common cell is the DICE cell proposed in [3]. The DICE cell consists of eight cross-coupled PMOS and NMOS transistors connected in series which forms four nodes. Due to the relatively high delay and power consumption of the DICE latch, there have been many other SEU tolerant latch designs proposed that provide reliability using blocking Muller C-elements [23], redundancy or delay in the feedback path [28, 10, 14, 29, 18, 37, 25].

In more recent times, the further reduction of the transistor feature size has increase the likelihood of a single event causing a transient pulse on multiple nodes simultaneously, commonly referred to as a single event multiple upset (SEMU). This trend necessitates the development of new latch designs that are tolerant to multiple node strikes to guarantee reliability in current and future technologies. As in the SEU case, the goal of these designs are to minimize the power, delay and area overheads. However, contrary to the SEU case, the latches are designed to tolerate two simultaneous errors, commonly referred to as a double node upset (DNU). Currently there are many existing latch designs that are tolerant to DNUs which are discussed in Section 1.1.

Many modern circuit designs employ a technique commonly referred to as clock gating to further reduce the power consumption. Clock gating consists of setting the clock to a stable value or "gating" the clock. If clock gating is used with a latch, it may need to

hold the current state for many clock cycles. In the presence of DNU's, this increases the likelihood of multiple errors occurring during the hold phase. In many existing DNU tolerant designs, a DNU puts the latch to a high impedance state in which the correct value could be lost if the latch experiences a further SEU or DNU before the transparent mode. Additionally, if the latch is gated for a sufficient number of cycles, the data held in the latch may degrade. For this reason, there is a need for new designs that are capable of holding the correct output value after a DNU for any number of clock cycles. In this chapter, we classify all DNU tolerant designs as either DNU robust or DNU non-robust. A DNU robust design is defined as being capable of resisting further errors and by not allowing any high impedance states after a DNU occurs. A DNU non-robust design is a latch that does not meet the all of stated criteria. In this chapter, a novel latch design referred to as the HRDNUT (Highly Robust Double Node Upset Tolerant) is proposed.

## 1.1 BACKGROUND ON EXISTING DNU TOLERANT LATCHES

Currently, there are a few existing DNU tolerant designs. The first proposed design found in [16], referred to as the DNCS latch, consists of two DICE cells connected to an output Muller C-element. This design tolerates DNU's since each DICE element requires a DNU to flip its state. Since the assumption is that only two errors can occur at once, in the worst case only one DICE element flips its state. Due to the C-element, the latch output does not change value. This design has been shown to be very resilient to DNU's at a very high cost of area, delay and power. The authors in [15] propose an enhanced design compared to [16]. Their latch design consists of six 2 input C-elements connected in series which are then fed into a 3 input C-element. Like the DNCS latch, this design offers high resiliency to DNU's, however the power consumption and area overheads are still very high.

More recently, a highly area and power efficient design has been proposed in [35] and is referred to as the HSMUF latch. Fig. 1.1 provides the design. The HSMUF uses the TP-DICE [1] structure which consists of 6 cross-coupled elements. In the case of a DNU, if

the error is on an adjacent node (such as a strike on  $n1$  and  $n2$ ), the TP-DICE element will fully recover the previous state. However, if the strike occurs on two non-adjacent nodes, the TP-DICE will not fully recover leaving one output node with an erroneous value, one node at high impedance and the remaining output node held at the error-free value. To provide reliability, the three nodes are connected to a C-element, as in Fig. 1.1, which allows the correct value to be held at the latch output.

While all of the previously discussed designs do provide high DNU reliability, none of them are classified as DNU robust since a DNU will result in high impedance states on the internal and output nodes. If an error occurs after a DNU, these latch designs will flip their held value. A popular remedy to this issue is to place a weak keeper on the latch output as in Fig. 1.1. However, adding a weak keeper greatly increases the power, area and delay overheads since the output C-element must be re-sized so that their driving strength exceeds that of the keeper. According to the simulations in Section 1.3, the addition of the keeper to the HSMUF latch nearly triples the power consumption and delay. Additionally, the latch is still vulnerable to error after a DNU since the TP-DICE is in a high impedance state.

To the author's knowledge, the existing most efficient DNU robust design capable of recovering all nodes after a DNU is the DONUT latch [6] in Fig 1.2. The design, as proposed in their paper, uses only 36 transistors but has a much higher power consumption compared to the HSMUF (See Section 1.3). The reason for the high power consumption is due to contention on the input lines during the transparent mode. For example, if the node  $n2$  in Fig. 1.2 is observed during the transparent mode, the node is driven by three cross-coupled elements. This contention will increase the amount of time required to change the node thus drastically increasing the dynamic power consumption. To optimize their design, the 48 transistor DONUT-M latch was created in which each component connected to an input node is modified, as shown in Fig. 1.3 so that the line is at high impedance for the whole duration of the transparent mode. This, in effect, removes the data contention

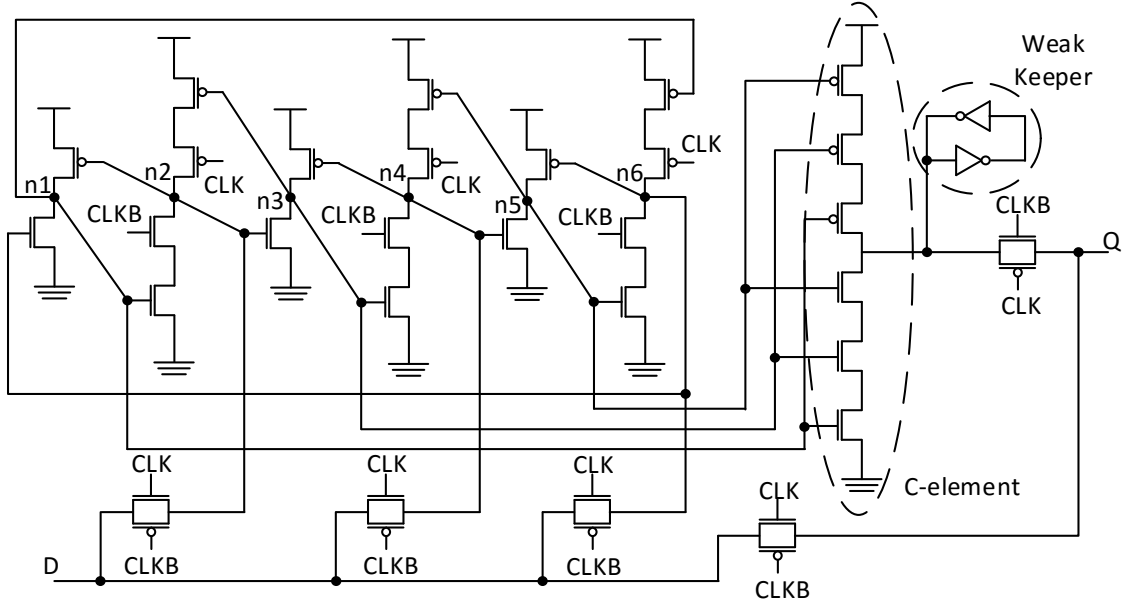


Figure 1.1. HSMUF latch [35] with a weak keeper on the output.

problem thus reducing the overall dynamic power and delay.

## 1.2 PROPOSED HRDNUT LATCH DESIGN

In this section the proposed DNU robust latch is discussed. The latch implementation is based on three cross connected storage loops connected to three C-elements. The basic design of the storage loop is given in Fig. 1.4. The data loop is based on the standard latch design with a 3-input C-element inserted to replace one of the inverters. The purpose of the C-element is to separate the feedback loop so that an error will not be held. Additionally, a PMOS is connected to the positive clock signal (CLK) and a NMOS is connected to the negative clock signal (CLKB) to remove contention when data is loaded to the latch. As in the modified DONUT latch, the addition of these transistors drastically reduces the delay and power consumption.

Using the basic storage block, the block based latch is constructed as in Fig. 1.5. The latch was designed with the goal of ensuring that none of the nodes directly drove itself.

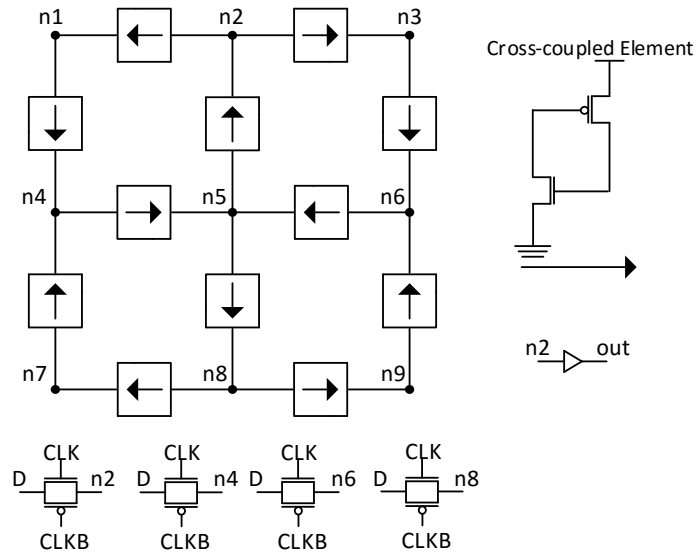


Figure 1.2. DONUT latch as proposed in [6].

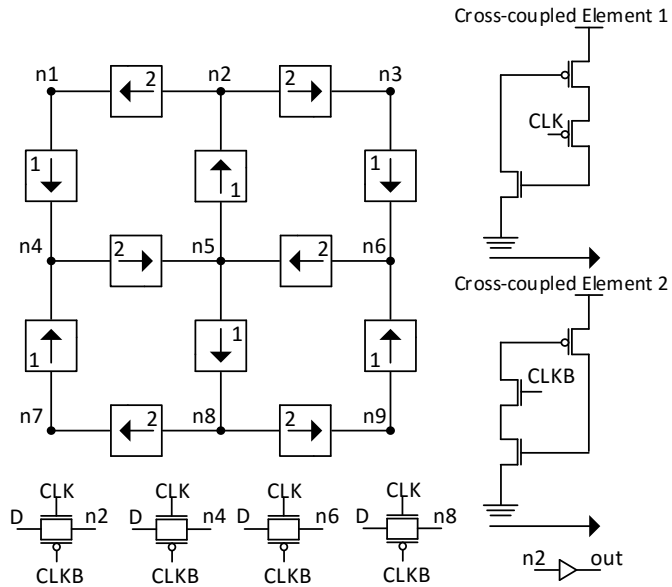


Figure 1.3. Modified low-power DONUT latch.

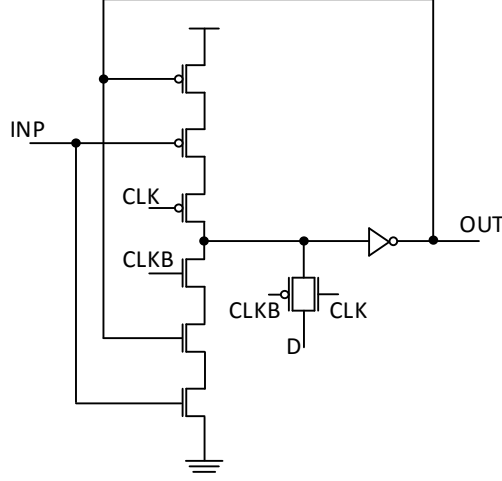


Figure 1.4. Basic data storage loop block.

For example, it can be seen that in Fig. 1.4 the node *out* is fed into the input of the 3-input C-element. If an error strikes node *out*, the cell will never be able to recover its previous state since one of the C-element inputs will be held to an erroneous value by its output. To prevent this issue, the design is based on cross-connecting three of the storage loop blocks so that the C-element is driven by three separate block outputs. In Fig. 1.5 a basic latch design using this idea is provided. If a single error occurs on any node in this design, the circuit is capable of fully recovering the previous data.

To demonstrate this, consider a strike on node *n2*. When the strike occurs, the erroneous value will be propagated to the C-elements driving nodes *n1* and *n3*. However, since there is no change on *n1* or *n3*, the C-elements *C1* and *C3* will hold their previous value thus preventing the error from propagating to the output. Additionally, since node *n2* is driven by nodes *n1* and *n3*, *n2* will completely recover the correct state.

A problem, however, with the latch in Fig. 1.5 is that it is not capable of tolerating DNU's. For example, if an error occurs on nodes *n1* and *n2* the erroneous values will propagate to the inputs of C-element *C3* and flip the value of *n3* thus changing the output value. However, since the latch has recovery capability for SEUs, we modify it so it can

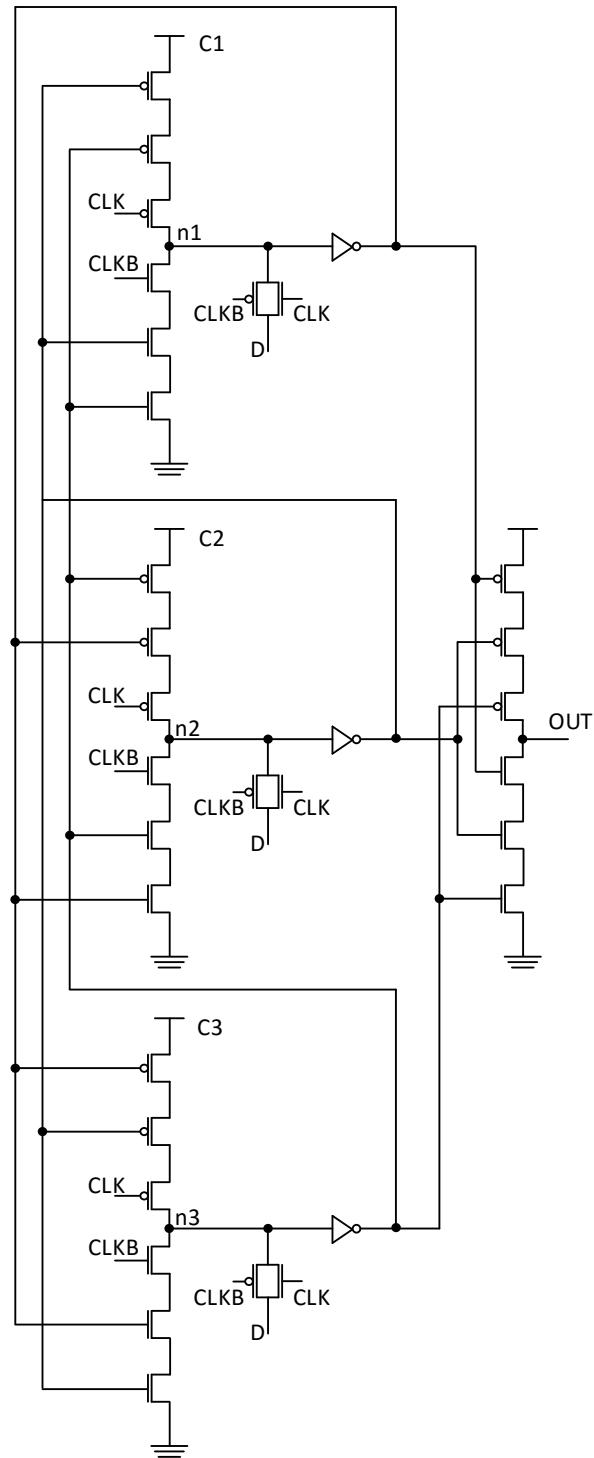


Figure 1.5. Schematic of the block-based latch.

tolerate DNUs and recover all nodes to the previous state. In Fig. 1.7 the schematic of the proposed HRDNUT latch is provided. The design uses the block-based latch in Fig. 1.5 as a base and adds additional C-elements to prevent errors from being held by the data loop.

Initially, the HRDNUT latch is evaluated during normal operation. When the positive clock signal (CLK) has a high value and the negative clock signal (CLKB) has a low value, the latch is in transparent mode. At this stage, the transistors connected to the clock signal in C-element *C1* deactivates the PMOS and NMOS stacks thus causing the node *n1* to be in a high impedance state. This, in effect, reduces data contention thus reducing delay and dynamic power consumption. Next, the data is loaded through the pass gates connected to nodes *n1*, *n2* and *out*. Since the output node *out* is loaded directly, the propagation delay is minimized and all nodes are set to their respective error free values. When CLK changes to a low value and CLKB to a high value, the latch moves into the hold mode. In this stage, the pass gates are deactivated and the state of the latch is held since each node is driven to the correct value using a C-element. Fig. 1.6 provides the waveforms of the CLK, D and OUT nodes for both the transparent and hold modes of operation.

In the case of an SEU, the HRDNUT retains the excellent resiliency of the block based latch and the ability to recover every node after an error. In the case of any internal node being struck by an error, the latch will not change value due to all internal C-elements requiring at least 2 identical input values to change the output. In the case of an error hitting the output node *out*, the latch fully recovers since *out* does not directly drive C-element *C7*.

Lastly, the latch will be evaluated in the case of a DNU. Note that unless otherwise stated, it is assumed that the analysis applies to both when D=0 and D=1. For the analysis, the possible DNU strike combinations are categorized into 9 distinct cases based on their effect in the HRDNUT latch. Waveforms for each category can be found in Appendix 3.5. The categories are discussed in detail below:

1. Consider strikes at nodes *n1* and *n2*. In this case, the error at *n1* will propagate



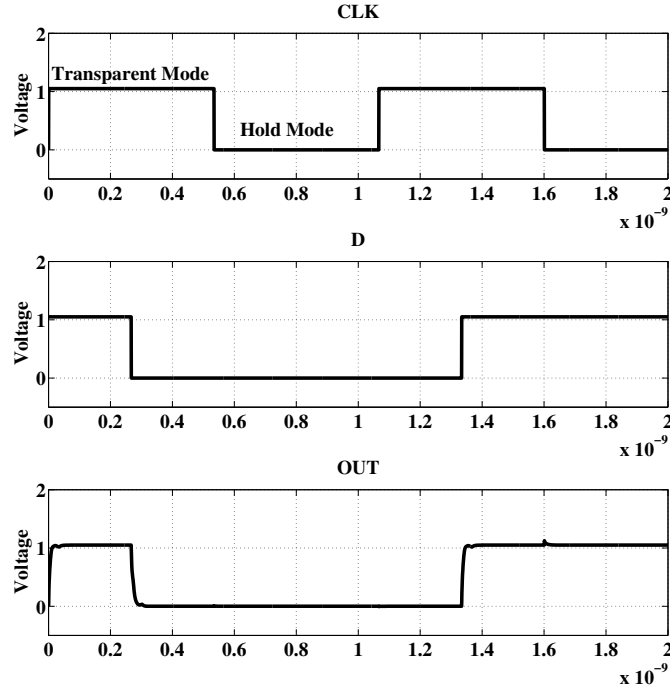


Figure 1.6. Waveforms of the HRDNUT latch during normal operation.

to C-elements  $C5$  and  $C7$  but will not cause a flip since the error at  $n2$  will be blocked by C-element  $C4$ . Additionally, since the inputs of C-elements  $C1$  and  $C2$  are unchanged, the nodes will recover their initial values. This analysis can be applied to node combinations containing node  $n2$  except for the combination with node  $out$  since the error will be blocked by C-element  $C4$ .

2. In the case of a DNU upsetting nodes  $n2$  and  $out$ , the error at  $n2$  will propagate through C-element  $C4$ . However, C-elements  $C1$  and  $C3$  will block the error and nodes  $n1$ ,  $n3$ ,  $n5$  and  $n6$  will hold their values thus driving node  $out$  to the correct state.
3. Consider when a DNU strikes nodes  $n1$  and  $n5$ . In this case, the error at  $n1$  hits the output of C-element  $C1$  which is propagated to  $C7$ . The error on  $n5$  is also propagated to C-element  $C7$ . Since node  $n3$  and the inputs of C-elements  $C1$  and  $C5$

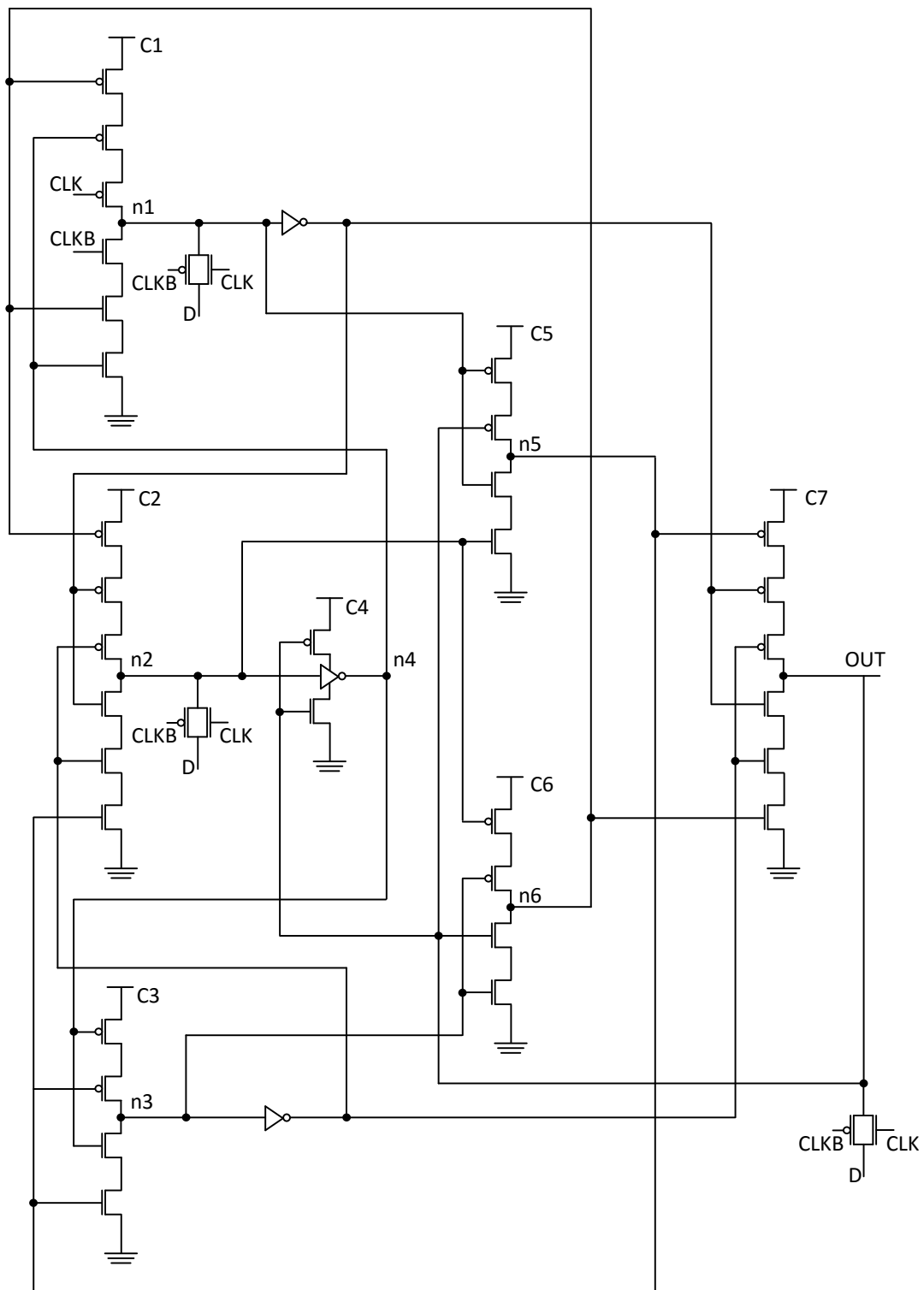


Figure 1.7. Schematic of the HRDNUT latch.

are unaffected by an error, the output retains the error-free value and the latch fully recovers the previous state. The above analysis also applies to the node combination  $(n3, n6)$ .

4. In the case of a DNU hitting nodes  $n3$  and  $n4$ , the error at  $n4$  is propagated to C-element  $C3$  and the error at  $n3$  is propagated to  $C7$  and  $C6$ . After the error on  $n3$  subsides,  $C4$  will drive node  $n4$  and, due to the connection at  $C3$ , node  $n3$  back to the error-free value. The node combination  $(n1, n1)$  can be analyzed similarly. For the node combinations of  $(n4, n5)$  and  $(n4, n6)$ , the latch will also recover the previous result since the inputs to  $C4$  are unchanged. This implies that after the error occurs at  $n4$ , the node will be driven back to the correct value thus also driving the nodes  $n5$  or  $n6$  back to the correct value.
5. When a DNU upsets the combination of  $n4$  and  $out$ , the error at  $out$  is propagated to  $C4$ ,  $C5$  and  $C6$  and the error at  $n4$  to  $C1$  and  $C3$ . Since none of the inputs to  $C7$  are changed by the error,  $out$  is flipped back to its error-free value which drives  $n4$  through  $C4$  back to its previous state.
6. Consider when a DNU strikes nodes  $n1$  and  $n3$  being struck. In this case, the errors are propagated to C-elements  $C2$ ,  $C5$ ,  $C6$  and  $C7$ . However, since the errors do not manifest into an error on any other node, the latch fully recovers from the error.
7. When a DNU strikes the nodes  $n1$  and  $n6$ . The error at node  $n6$  propagates to  $C1$  and  $C7$  while the error at  $n1$  also propagates to  $C7$ . Due to the error-free node  $n3$  driving  $C7$ , the previous value is held at the output by  $C7$ . Additionally,  $n3$  will drive  $C6$  back to its previous value thus driving  $C1$  back to the error free state. This analysis can be applied similarly to the node combination of  $(n3, n5)$ .
8. In the case where a DNU strikes nodes  $n5$  and  $out$  the error at  $n5$  propagates to  $C7$ ,  $C2$  and  $C3$  and the error at  $out$  goes to  $C4$ , a PMOS in  $C5$  and a NMOS in

*C6*. When the error-free value at *out* is 1, the value at *n5* is 0. The error at the nodes change the values to 0 and 1 respectively and the erroneous value at *out* is propagated to the PMOS at *C5* and the NMOS at *C6*. This, in effect, causes the PMOS at *C5* to be activated and the NMOS at *C6* to be deactivated. However, since nodes *n1* and *n2* remain error-free, the NMOS stack of *C5* will drive *n5* back to the correct value. This, in turn, forces *C7* to also drive *out* back to the error-free value. In the case where *out* has an ideal value of 0, the error will be fully recovered since the NMOS stack will be entirely driven by fault-free nodes. The above analysis can be applied to the node combination of (*n6*, *out*).

9. Lastly, the node combinations (*n1*, *out*), (*n3*, *out*) and (*n5*, *n6*) are analyzed. In these cases the errors do not cause a change on the inputs of any C-elements driving the node thus the previous value will always be recovered.

### 1.3 SIMULATION RESULTS

The proposed HRDNUT latch has been implement using the 1.05V 32nm PTM library [39] and simulated in HSPICE. All transistors were set to the minimum size with the PMOS widths set to W=80nm and the NMOS widths set to W=40nm. To evaluate the DNU reliability of the design, current pulses were injected for every possible error combination. The injection current was calculated using the equation found in [40]. The equation is given below with  $\tau$  as the technology dependent constant,  $Q_o$  as the injection current value and  $t$  as the variable for time.

$$I(t) = \frac{2Q_o}{\tau\sqrt{\pi}} \sqrt{\frac{t}{\tau}} e^{-\frac{t}{\tau}} \quad (1.1)$$

Using equation (1.1)  $\tau$  was set to  $32 \times 10^{-12}$  and  $Q_o$  was set to 5fC. In all simulations, the latch was operated at a frequency of 1Ghz. In Appendix I Figs. 3.7-3.16, the waveforms for each case discussed in Section 1.2 are presented and show that the HRDNUT is fully

capable of recovering all nodes in the presence of a DNU.

Next, the HRDNUT is compared to existing SEU and DNU tolerant methods. As in the HRDNUT latch, all latches were designed using the 32nm PTM library and operated at 1Ghz. For the analysis, the following SEU tolerant latches were compared: DICE [3], FERST [10] and HIPER [28]. Additionally, the following DNU tolerant designs we also analyzed: DNCS [16], Interception [15], HSMUF [35] and DONUT [6]. All transistors for the implemented latches were set to minimum width and length except for the designs that use a C-element with a weak keeper. In these designs the C-element's PMOS width was set to  $W=320\text{nm}$  and the NMOS width was set to  $W=160\text{nm}$ . The weak keeper was sized to be at minimum width. The C-element was sized so that the output driving strength did not allow the keeper to drive an erroneous value in the event of an error.

To provide a fair comparison, the propagation delay, average power consumption and area of all designs is measured and each design is categorized based on whether they can tolerate a DNU and if they are robust from error after a DNU occurs. The delay was measured as the time between when a transition occurs on input  $D$  to when a transition was observed on the output. The average power was computed using the error-free operation for each latch for a duration of 200ns. To compare the area overhead, the unit size transistor (UST) metric was adopted as in [16] which represents the number of unit sized (minimum width is  $W=40\text{nm}$  in this case) transistors required for the design. Table 1.1 provides the results of these simulations.

According to Table 1.1 the only DNU robust designs are the two DONUT latch implementations and the HRDNUT. Compared to the modified DONUT latch, the HRDNUT provides DNU robustness while reducing the power consumption and the number of transistors by 11.3% and 8.33% respectively while also reducing the delay by 72.5%. For the above reasons, the HRDNUT is the best design for clock gating applications due to its high robustness, even after a DNU occurs, and lower power, delay and area overheads.

Table 1.1. SPICE Simulations of Existing Latches using the 1.05V 32nm PTM library

Latch	DNU Immune	DNU Robust	Power ( $\mu$ W)	Delay (ps)	Area (UST)
DICE	No	No	1.332	8.145	16
FERST	No	No	3.178	31.648	60
HIPER	No	No	1.292	2.221	27
DNCS	Yes	No	4.948	22.486	61
[15]	Yes	No	5.606	79.168	89
HSMUF	Yes	No	1.871	1.0626	51
HSMUF (Keeper)	Yes	No	3.787	3.945	78
DONUT [6]	Yes	Yes	4.021	14.722	54
DONUT-M (Section 1.1)	Yes	Yes	2.760	8.421	72
HRDNUT (Proposed)	Yes	Yes	2.450	2.310	66

#### 1.4 FILTERING OF TRANSIENTS ARRIVING ON THE LATCH INPUT

In addition to radiation particles hitting the internal nodes in a latch, transient voltage pulses may arrive on the inputs of a combinational circuit. In a typical digital system design, combinational logic is placed between two sets of flip-flops, one set driving the input and the other being driven by the output of the combinational circuit. While there are masking factors in combinational circuits, they may still allow a pulse to arrive on the input of a flip flop. A novel circuit to filter incoming pulses was proposed in [10] and is given in 1.8. Their design consisted of routing the pulse through two paths, one with no delay and one with sufficient delay that the two pulses will not overlap. The general design is given below.

An issue with the approach in Fig. 1.8 is that in a radioactive environment, a high energy particle can hit the delay element causing a voltage pulse on the node. If this occurs while a pulse is being propagated through the circuit, it could create a secondary pulse at the delay element causing an erroneous value to be latched in the flip flop. A simple way to mitigate this effect is to add an additional delay element with twice the delay. This, in effect, will allow the device to function properly even if an error occurs during the filtering

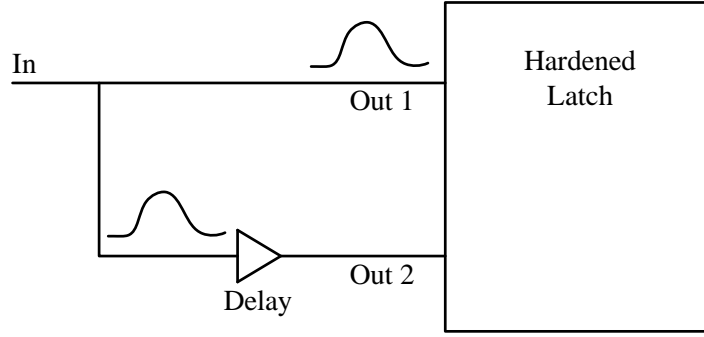


Figure 1.8. The pulse filtering circuit provided in [10].

stage. The design of the device is given in Fig. 1.9. Due to the use of three outputs, the device can be used with the HRDNUT latch without modification.

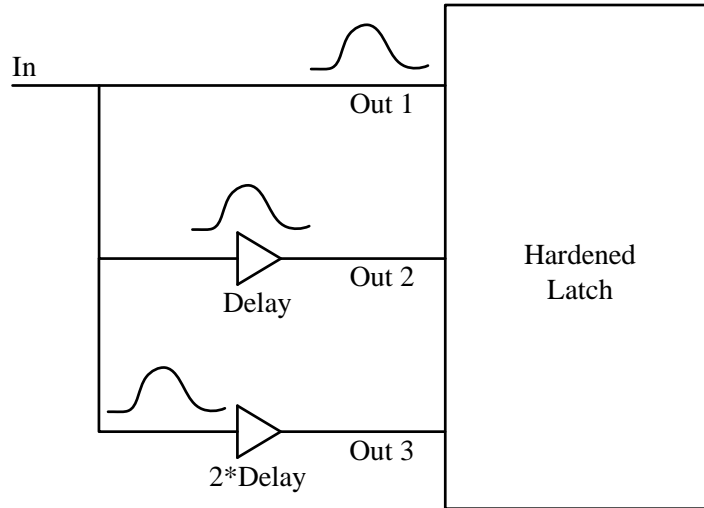


Figure 1.9. The enhanced pulse filtering circuit that can tolerate an error.

Next the filtering circuit is evaluated. If a pulse arrives on node  $In$ , it is propagated to the three delay elements. The pulse arriving on  $OUT1$  will propagate to the latch with no increased delay. The pulse on  $OUT2$  will arrive at the latch with a delay of one unit and the pulse on  $OUT3$  will arrive with a delay of 2 units. If an error occurs on the buffer at  $OUT3$  while a pulse is being propagated, an additional pulse will be generated at the buffer causing two node to simultaneously have the same data value. However, since node  $OUT2$

is error free during this stage, the circuit will still filter the pulse allowing the correct data to be latched. This analysis can be applied similarly to the case where a strike occurs on the buffer connected to *OUT2*.

To determine the power overhead of the filtering circuit, the device was implemented in the 32nm PTM library and operated at 1Ghz. The circuit was designed to filter pulses with a width of 20ps. This was accomplished using a chain of 4 buffers for the short delay and 8 buffers for the long delay which had a 5ps propagation delay each. Simulations were completed in HSPICE which show that the addition of the filtering circuit increases the power consumption by 3.15 ps and has a minimum delay of 40ps.

## 1.5 ANALYSIS OF A TRIPLE NODE UPSET TOLERANT LATCH

An inevitable issue with the development of more complex latches to tolerate multiple errors is the increase of the number of transistors and area. This issue leads to the possibility of new latch designs being vulnerable to more simultaneous error such as a triple node upset (TNU). The goal of this section is to estimate the area and device overhead and to evaluate the feasibility of a TNU tolerant latch. To begin the analysis, the design methodology used in the development of the HRDNUT is applied. The basic element used in this design is the block based latch given in Fig. 1.5. In the case of the HRDNUT, 3 blocks were needed to harden the latch against 2 errors. Based on this observance, it is estimated that hardening against TNUs will require 4 blocks. Additionally, to hold to the principle discussed earlier in this paper that no element drives itself, each c-element will have an additional transistor added to accompany the extra block. The block based latch for TNUs is given in Fig. 1.10 which has 56 total transistors. Since the block based latch for the HRDNUT has 36 transistors, it can be extrapolated that TNU hardening requires a 35% transistor overhead.

To further estimate the power, delay and area overheads of a TNU latch, a design, provided in Fig. 1.11, was created which shows promise of being TNU tolerant and robust. It is based on the ideas used for the HRDNUT and derived from the TNU block latch.



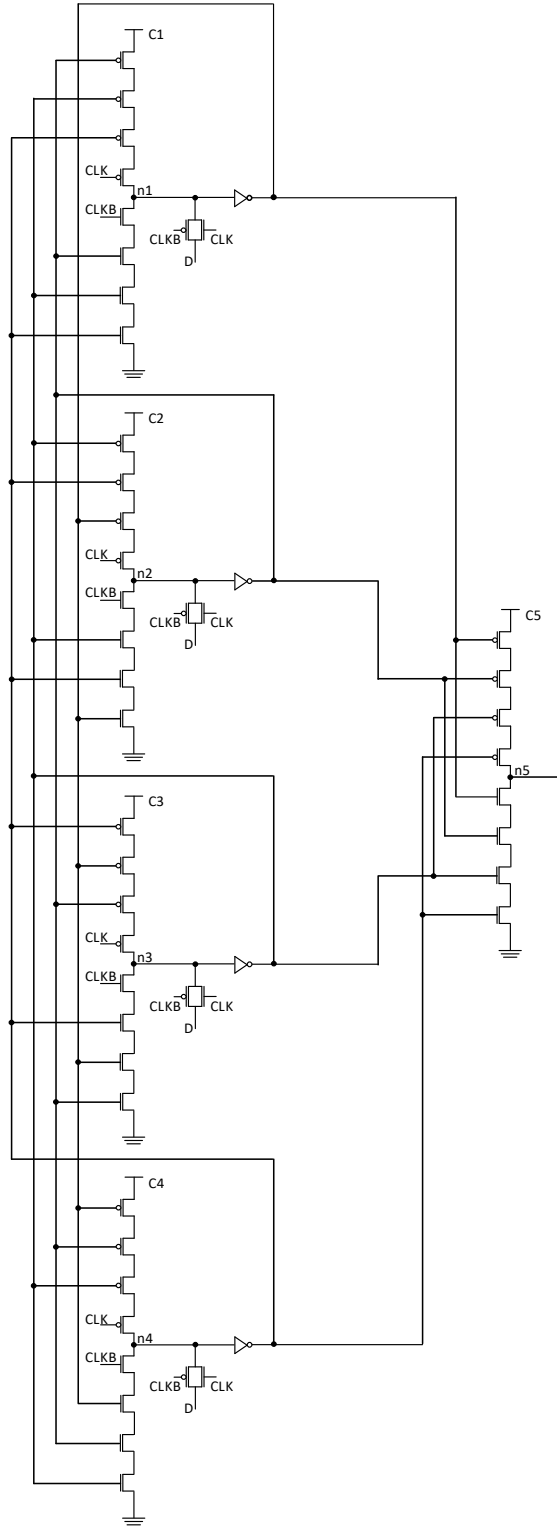


Figure 1.10. Block based latch that forms the basis for a TNU tolerant design.

Due to it being based on the HRDNUT, it can be assured that it is both SEU and DNU tolerant. However, it cannot be guaranteed that the design is totally TNU tolerant due to the complexity of testing all 165 possible cases. For the sake of discussion, three examples of how the latch tolerates a TNU are given below.

1. Consider a TNU on nodes *out*, *n5* and *n6*. The error at *n5* will effect C-elements *C1*, *C3*, *C4* and *C9*. The transient at *n6* will effect nodes *C1*, *C2*, *C4* and *C8*. Lastly, the error at *out* will arrive at C-elements *C5* and *C6*. In this case, the output node will directly be driven back to the correct value due to no change on any of the inputs on C-element *C11*. Since *out* is driven back to the correct value, all of the inputs on C-elements *C5* and *C6* are returned back to the error free value. This will, in turn, drive nodes *n5* and *n6* back to the correct value thus fully recovering all nodes.
2. Assume a TNU strikes nodes *n7*, *n8* and *out*. Node *n7* has a transient that propagates to C-element *C11*. Similarly, node *n8* also has a transient propagates to *C11*. The transient pulse on node *out* arrives on the inputs of C-elements *C5* and *C6*. In this case, all transient pulses are blocked by the respective C-element since not each element has at least one error free input. This implies that all nodes will fully recover to their respective value.
3. Consider a TNU which generates pulses on nodes *n9*, *n2* and *out*. In this case, the errors on the nodes will flip the output of C-element *C5* thus also causing a transient on node *n5*. The error on *n5* will propagate to C-elements *C1*, *C3*, *C4* and *C10*. However, the pulse will be blocked at all elements. Additionally, since *n2*, *n9* and *out* all restored to their error free value, *n5* will also return to the correct value.

The design was implemented in HSPICE using the 32nm PTM library with all transistors set to minimum width (PMOS=80nm, NMOS=40nm). The clock frequency was set to 1Ghz and  $V_{dd}$  to 1.05V. The latch was operated with no injected transient pulses to

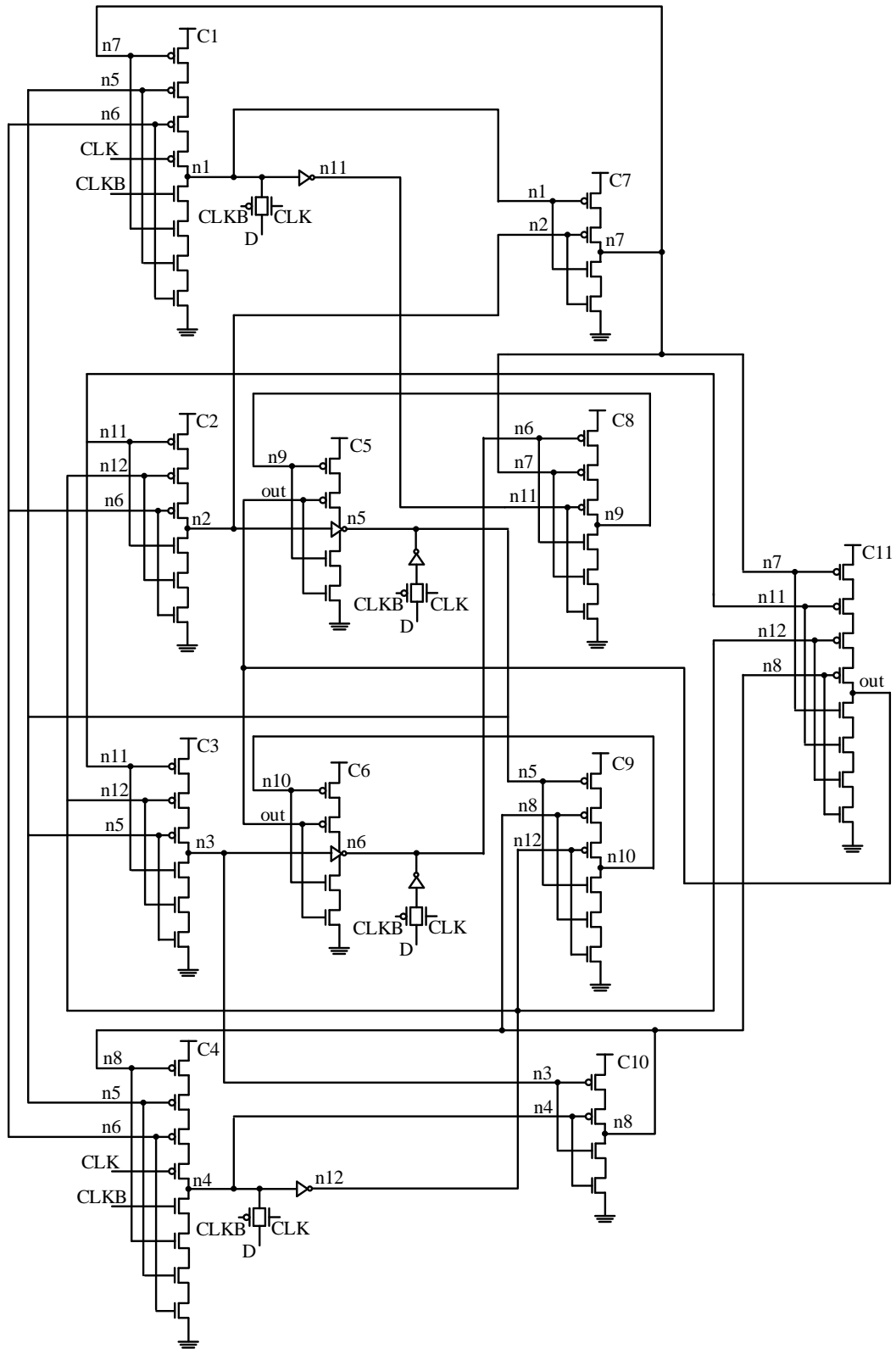


Figure 1.11. Schematic of the TNU tolerant latch.

measure the average power. It was found that the TNU latch consumes  $4.09 \mu\text{W}$  which is 60% more compared to the HRDNUT while having an UST area of 129 and delay of 2.77 ps which is double the HRDNUT.

Based on the design, there are few similarities to the HRDNUT that have been observed that can be used to design latches that tolerate 4 or more errors. The similarities are summed up in the following list:

1. Assuming the latch will tolerate a  $n$  number of errors, the block based latch will have  $n + 1$  number of elements.
2. Each storage block in the block based latch will have at least one C-element with a  $n$  number of inputs.
3. At least one of the blocks will have two C-elements with a  $n$  number of inputs.
4. Each block should be wired such that no wire drives itself. This is to ensure robustness.
5. An intermediary level of C-elements must be inserted between the blocks and the output device to ensure robustness.
6. The output must be looped back to drive a C-element.
7. To load data to the latch, a  $n + 1$  number of nodes must be driven by the data line.
8. At least one C-element will be set to high impedance during the transparent mode to increase performance.

## 1.6 CONCLUSION

In this section the HRDNUT latch was proposed which is suited for clock gating schemes. Since clock gating may require the latch to remain in a hold state for many clock cycles, the susceptibility of error increases. In many existing designs, a DNU may either

change the state of the latch or push the latch into a state where the output may discharge over time due to a high impedance state. A common method to solve this problem is the addition of a weak keeper on the output. As shown in this paper, the addition of the keeper causes much higher power consumption. Since the HRDNUT latch does not stay in a high impedance state after a DNU, the HRDNUT provides high reliability during the whole duration of the hold mode while providing the lowest delay, power and area compared to other latches suitable for clock gating. Simulation results show that the HRDNUT is 11.3% more power efficient while requiring 8% less transistors and 72.5% less delay compared to the highly robust DONUT latch.

Additionally, an enhanced circuit to filter incoming transient pulses was proposed. The new circuit is able to tolerate a single error during the filtering phase. The area of the new filter circuit is dependent on the size of the expected transient but overall is low. Lastly, an analysis of the TNU was conducted and the block based circuit was given. It was determined that a TNU tolerant circuit would have a 35% greater overhead compared to the HRDNUT. Based on this finding, it is recommended that a TNU tolerant latch is used in extreme radiation environments where system stability is crucial.

## CHAPTER 2

### AN ANALYTICAL PULSE APPROXIMATION ALGORITHM FOR MULTIPLE EVENT TRANSIENTS

As the transistor size has continued to decrease, the probability of a SEU occurring has continued to greatly increase. This, in effect, has created a need for accurate and efficient SEU simulators. The SEU problem has been investigated deeply. [20, 38, 36] However, due to the close proximity of transistors and the increase in transistor error sensitivity, the probability of a single or multiple radiation strikes producing multiple voltage pulses has become a concern. The authors in [13] investigated a 65 nm 3x3 inverter matrix and found that 40% of all single radiation strikes resulted in the generation multiple significant voltage pulses. Considering the possibility that a modern circuit is also vulnerable to multiple simultaneous strike, there is a significant concern for multiple transients. This creates the need for either the enhancement of existing SEU tools or the development of new tools which can consider the single event multiple transient (SEMT) and multiple event multiple transient (MEMT) effect.

Previous work in SEUs have shown that combinational circuits are vulnerable to 3 types of error masking effects: logical masking, electrical masking and temporal masking. Accurate estimation of the soft error rate requires consideration of all masking factors for not only SEUs but also the SEMT and MEMT cases. In this paper, the focus is on the consideration of the electrical masking effect. There have been many approaches to the electrical masking problem for the SEU case [27, 33, 34]. However, a major drawback with these approaches is that they do not accurately consider the effect of multiple pulses occurring on the gate input (referred to as convergent pulses). Since, in the SEMT and MEMT cases, there are many pulses generated, the occurrence of convergent pulses is much higher. To model these cases, an accurate electrical masking model must consider convergent pulses.

There has been a few attempts at modeling the effect multiple transient pulses, however all approaches suffer from inaccurate estimation of electrical masking. The authors in [21, 8] both proposed a simulation tool to determine the soft error rate (SER) of a combinational circuit in the presence of multiple transients, however, they calculate the output transient of a gate with a convergent input through the simple superposition of the pulse widths. As stated in [33], this is insufficient since accurate electrical masking modeling requires the inclusion of the pulse amplitude and non-linear portions of the glitch.

An additional possible solution to the pulse convergence problem is the simple extension of [33] for convergent pulses. In [33] an accurate iterative model for the single pulse case is proposed. Their model can easily be extended by generating larger look-up tables. However, since they characterize stacked transistors using look-up tables, gates with more than 2-inputs require tables with 125,000 entries for 3-input gates and over 7 million for 4-input gates. Thus our model provides an enhancement to [33] by allowing the characterization of stacked transistors using a look-up tables for the transistor and the Miller capacitance between the gate and drain terminals.

Based on the above discussion, there is a lack of accurate and fast methods to consider convergent pulses. This provides the motivation of this work. To the authors knowledge, the proposed model is the first masking model that accurately and efficiently approximates the entire output pulse shape for convergent input pulses. Through extensive experimentation, the proposed model has been shown to be robust for various input shapes and process technologies while providing a 15X speed-up over HSPICE.

## 2.1 PULSE APPROXIMATION MODEL

As in [33], an important aspect of the proposed model is the accurate approximation of the transistor drain to source current ( $I_{DS}$ ). To consider the  $I_{DS}$  current, a look-up table for each transistor is created by sweeping both the gate-source voltage ( $V_{GS}$ ) and the drain-source voltage ( $V_{DS}$ ).

Additionally, as in [33], the effects of pulse overshoot and undershoot are considered through the use of a Miller capacitor. In contrast to [33], however, we consider the MOS parasitic effects on each individual transistor through the inclusion of a Miller capacitance between the gate and drain terminals. Fig. 2.1 demonstrates the proposed model for a NMOS and a PMOS transistor.

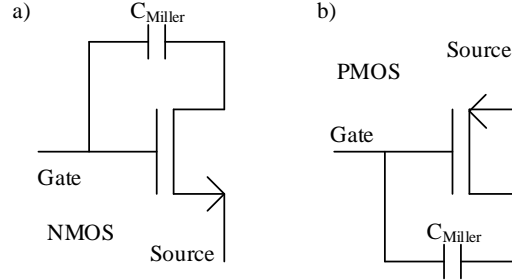


Figure 2.1. (a) The transistor model used for a NMOS. (b) The transistor model for a PMOS.

As stated in [33], it was observed that the value of the Miller capacitance varied over the range of drain-source ( $V_{DS}$ ) and gate-source ( $V_{GS}$ ) voltages. To consider this effect, it is proposed to generate a look-up table with each entry pertaining to the Miller capacitance at each specific  $V_{GS}$  and  $V_{DS}$  bias. The Miller capacitance look-up tables can be characterized using HSPICE by sweeping the  $V_{GS}$  and  $V_{DS}$  voltages and measuring the current flowing through the gate node of the transistor ( $i_m$ ). Using the current  $i_m$ , the derivative of the difference between the gate and drain voltages and equation 2.1, the capacitance  $C_{Miller}$  can be calculated.

$$C_{Miller} = \frac{i_m}{\frac{d(V_{DS}-V_{GS})}{dt}} \quad (2.1)$$

As stated before, the  $I_{DS}$  current and the Miller capacitance for both the PMOS and NMOS transistors are characterized. Using this data, each transistor is modeled as a current source and its Miller capacitor. Any given gate can then be represented by replacing



each transistor with a current source and a Miller capacitor connected between the drain and gate terminals. At the output of the gate, a load capacitance is added to consider the gate loading effects and to calibrate the model. Fig. 2.2 demonstrates the conversion to the proposed model on a 2-input NAND gate.

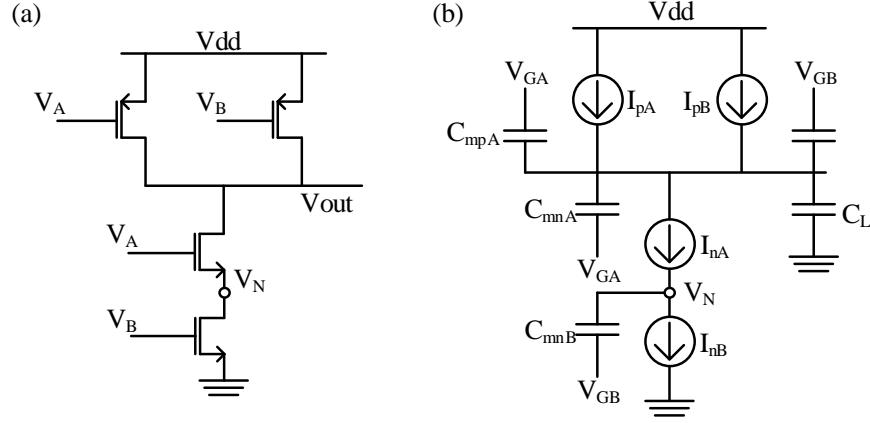


Figure 2.2. (a) The transistor model used for a NMOS. (b) The transistor model for a PMOS.

Before beginning the discussion on the equation derivation it is important to note that this model is general for all types of gates but for the sake of this discussion, the derivation will focus on CMOS NAND and NOR gates.

The proposed gate modeling for a given gate, as in Fig 2.2b, can be generalized for an  $i$  number of inputs as in Fig. 2.3 where a single current source represents the current for the pull-up network and a second current source represents the pull-down network. This is based on the assumption that the gate is a CMOS gate in which one network consists of parallel current sources and the equivalent current is equal to the sum of each parallel transistor current. The other current source represents the current flowing through series connected transistors which ideally have a single constant current. However, in practice, due to parasitic effects, the current in each stacked transistor varies. It was found through many simulations that the best approximation for this current was by taking the average

between each transistor current in the stack. For example, if we have the circuit in Fig. 2.2b, the current on the stacked transistors can be calculated using the following equation:

$$I_{pd} = \frac{I_{nA} + I_{nB}}{2} \quad (2.2)$$

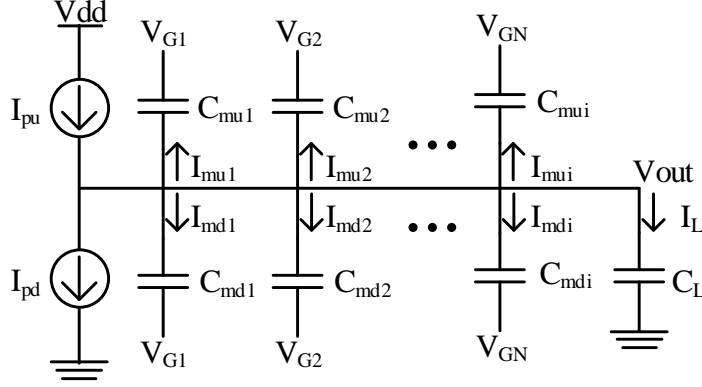


Figure 2.3. A generalized representation of the gate model.

To derive the output equation for the node  $V_{out}$  in Fig. 2.3, we first use Kirchoffs Current Law (KCL) at the node  $V_{out}$ . Assuming that  $I_{mu_i}$  and  $I_{md_i}$  are the currents through the  $i$ -th Miller capacitor,  $I_L$  is the current through the load capacitor and  $I_{pu}$  and  $I_{pd}$  represent the pull-up and pull-down current respectively, KCL will result in the following:

$$I_{pu} = I_{pd} + I_L + \sum_{i=1}^N I_{mu_i} + \sum_{i=1}^N I_{md_i} \quad (2.3)$$

For our model, we discretize the time into time steps with the variable  $t_{step}$ . Additionally, we represent each variable in the current time instant with the subscript  $i$  and each variable in the previous time instant with the subscript  $i - 1$ . Lastly, we represent the change in a value between two time instances using  $\Delta$  (i.e.  $V_i - V_{i-1} = \Delta V$ ).

According to the charge conservation law, we get equation 2.4 for each Miller capacitor

$C_{m_i}$  connected between  $V_{out}$  and the input voltage  $V_{G_i}$ .

$$I_{m_i} = \frac{[\Delta V_{out} - \Delta V_{G_i}]C_{m_i}}{t_{step}} \quad (2.4)$$

Next, consideration of the load capacitance  $C_L$  gives the following equation:

$$I_L = \frac{(\Delta V_{out}C_L)}{t_{step}} \quad (2.5)$$

By substituting into equation 2.3 the current  $I_{m_i}$  from 2.4 for the currents  $I_{mu_i}$  and  $I_{md_i}$  and replacing the current  $I_L$  with equation 2.5, we get the following for a  $N_u$  number of Miller capacitors connected between  $V_{out}$  and the  $i$ -th input in the pull-up network (denoted by  $C_{mu_i}$  and a  $N_d$  number of Miller capacitors connected to  $V_{out}$  and the  $i$ -th input on the pull-down network (denoted by  $C_{md_i}$ ):

$$I_{pu} = I_{pd} + \frac{(\Delta V_{out}C_L)}{t_{step}} + \sum_{i=1}^{N_u} \frac{[\Delta V_{out} - \Delta V_{G_i}]C_{mu_i}}{t_{step}} + \sum_{i=1}^{N_d} \frac{[\Delta V_{out} - \Delta V_{G_i}]C_{md_i}}{t_{step}} \quad (2.6)$$

Rearranging 2.6 and solving for  $\Delta V_{out}$  gives the following:

$$\Delta V_{out} = \frac{(I_{pu} - I_{pd})t_{step} + \sum_{i=1}^{N_u} \Delta V_{G_i} C_{mu_i} + \sum_{i=1}^{N_d} \Delta V_{G_i} C_{md_i}}{C_L + \sum_{i=1}^{N_u} C_{mu_i} + \sum_{i=1}^{N_d} C_{md_i}} \quad (2.7)$$

Assuming that  $\Delta V_{out_i}$  equals  $V_{out_i} - V_{out_{(i-1)}}$  results in:

$$V_{out_i} = \Delta V_{out} + V_{out_{i-1}} \quad (2.8)$$

As can be observed in equations 2.7 and 2.8 an important aspect of the proposed model is the accurate consideration of both the pull-up and pull-down currents. As stated before, cases where the pull-up or pull-down transistors are composed of parallel connected transistors the current can simply be found through the sum of the drain to source currents.

However, in cases where the transistors are in series, the determination of the equivalent current is more complicated.

In Fig. 2.2a the pull-down branch consists of series connected NMOS transistors which will be used to demonstrate the derivation. Note that using the same method, the equations for stacked PMOS devices can be easily derived. Within the stack there is an intermediate node voltage  $V_n$  which is crucial for the determination of the drain to source currents in the transistors. Through many simulations using the transistor drain to source current look-up tables, it was found that accurate approximation of the node voltages provided accurate current values for each transistor in the stack.

To calculate the intermediate voltage, it is proposed to first convert each transistor to the representations in Fig. 2.1a and 2.1b. which results in the circuit in Fig. 2.2b. In Fig. 2.2b. the intermediate node voltage  $V_n$  is connected to a static capacitor representing the loading effect of the subsequent transistor stack. Next, the node voltage equation will be derived.

Similar to equation 2.7, the node voltage  $V_n$  is modeled using discrete time steps. Using KCL at  $V_n$  in Fig. 2.2b we get the following equation:

$$I_{D1} = I_{D2} + I_m \quad (2.9)$$

As in equation 2.4, the current through the static capacitor can be determined using the charge conservation law resulting in the following for the current  $I_m$ .

$$I_m = \frac{C_{m1}(V_{n_i} - V_{n_{i-1}})}{t_{step}} \quad (2.10)$$

By substituting 2.10 into 2.9 and solving for  $V_{n_i}$ , the equation for the node voltage  $V_n$  at the time instant  $i$  is derived.

$$V_{n_i} = \frac{(I_{D1} - I_{D2})t_{step}}{C_{m1}} + V_{n_{i-1}} \quad (2.11)$$

In cases where there are more than 2 stacked transistors, the described model is directly extendable for each intermediate node. Fig. 2.4a and b provides the conversion for 3 stacked NMOS transistors to our model. The node voltages  $V_{N1}$  and  $V_{N2}$  can be found in any order using equation 2.11 and by considering the currents flowing into the node. For example, to solve for the node voltage  $V_{N2}$  the currents  $I_{D2}$  and  $I_{D3}$  would be used in equation 2.11 since they are directly connected to the node. Additionally, the order for solving the node equations does not matter since the node voltages from the previous step  $i - 1$  are used.

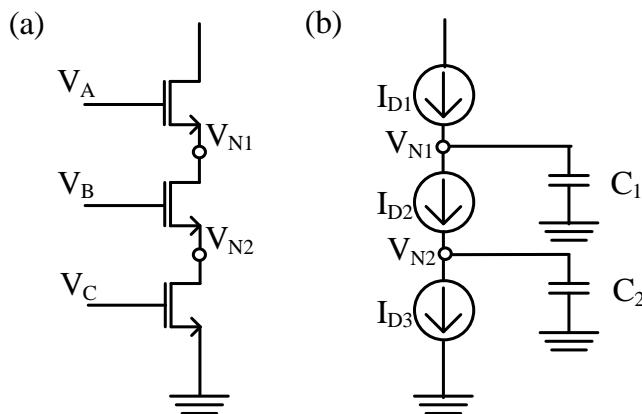


Figure 2.4. (a) Three stacked NMOS transistors with two intermediate nodes  $V_{N1}$  and  $V_{N2}$ . (b) The transistors converted to the proposed circuit model.

## 2.2 EXPERIMENTAL RESULTS

The proposed algorithm was implemented in Matlab and characterized using the 32 nm and 45 nm PTM library [39] and the 65 nm IBM library. To test the proposed model, 3 benchmark circuits for each process library were created to test two convergent pulses on a 2 and 3 input NAND gate and a 2 input NOR gate. The first two benchmark circuits denoted by NAND2R and NAND2F in column 1 of Table 2.1, consists of two NAND gate

chains with each off-input held to a non-controlling value. The chains contain two gates and the R and F in the benchmark names represent either a rising (R) or falling (F) pulse on the convergent gates inputs. Each chain is then connected to an input of the NAND gate. The benchmark circuits NOR2R and NOR2F are constructed similarly but the chains consist of NOR gates and are connected to the input of a 2 input NOR gate. The last benchmark circuits constructed, NAND3R and NAND3F, were created with 2 2 gate chains connected to the first and second input of the 3-input NAND gate. The remaining off-input was held to a non-controlling value.

Next, to accurately model the injected pulse shape, the injection current is represented as a current source located at the output of the first gate in the chain. We then used the function for the pulse shape given in [40] to relate the strike current to a given charge. Equation 1.1 was used to model the current generated by a pulse strike.

Using equation 1.1,  $Q_o$  was varied over a range of charge values that results in an observable error. For our test cases, we found that a charge ranging from 3-16 fC for a of 90x10-12 provided an observable output shape for all processes tested. We then propagated the pulses through the gate chains to gate of interest and compared the generated outputs using the proposed model and the model in [33] to HSPICE. Table 2.1 details the average error for 30 cases per gate and transition type for the 32, 45 and 65 nm processes using 2400 data points over a range of 0 to 1.2 pico-seconds. The first column gives the benchmark name. All subsequent columns provide the MSE error for each test case for the proposed method and the method in [33]. Assuming that  $N$  represents the number of calculated points,  $V_i^H$  represents the pulse voltage at point  $i$  on the signal from HSPICE and  $V_i^C$  is the voltage calculated by the proposed model, the MSE was calculated using the following equation:

$$MSE = \frac{\sum_{i=0}^N (V_i^H - V_i^C)^2}{N} \quad (2.12)$$

As can be seen from Table 2.1, the proposed model is much more accurate over the model proposed in [33]. In all tested cases, the method in [33] predicted there would be an output pulse. However, they only propagate the pulse with the largest width which fails to consider the effect of multiple pulses and provides a pulse with a much smaller width. For this reason, their method tends to underestimate the severity of the pulse in the presence of convergent pulses. According to Table 2.1, our method closely predicts the output pulse shape.

Next, we run a simulation to observe how the number of points selected relates to the speed and accuracy. In this simulation, we used the NAND2F and NAND3F benchmark and applied a pulse to each input of the terminating NAND gate. We then varied the number of output data points for the terminating NAND gate in both HSPICE and the proposed model. Table 2.2 provides the results of this simulation.

The first column provides the number of simulation points used to calculate the result. Columns 2 and 3 provide the error of the proposed model and [33] for the given number of points. Columns 4, 5 and 6 gives the execution time of the proposed method, the method in [33] and HSPICE respectively. As can be seen from Table 2.2, the reduction in simulation time scales linearly to the reduction in simulation points.

In Fig. 5, we provide the output waveform of the simulation in Table 2 for 900 and 300 points respectively. As can be observed in Fig. 5a, the result for the proposed model closely matches the result obtained in HSPICE when 900 simulation points are used. If a higher accuracy is desired, the proposed model provides a near perfect match with 2400 points. For both the 900 and 300 point cases, it can be observed that the method in [33] greatly underestimates the resulting pulse thus providing a high MSE. The proposed model provides an accurate result with an MSE of  $7.40 \times 10^{-3}$  while still providing a speed-up of 15X compared to HSPICE for this benchmark. In general we have observed a speed-up of 15X for any 2-input gate using 900 points while maintaining a high accuracy.

In Fig. 6a and 6b, we provide the output waveforms for the NAND3F simulation

for 900 and 300 points respectively. Similar to the NAND2F benchmark, a simulation with 900 points provides a close waveform approximation. However, if a lower accuracy is tolerable, the number of points can be reduced further providing a faster simulation time as demonstrated in Table 2.2 and Fig. 6b. As can be observed in Table 2, proposed model provides a speed-up of 12X for the 900 point case. In general for any 3-input gate, we have observed a speed-up of 12X using 900 points while maintaining a high accuracy

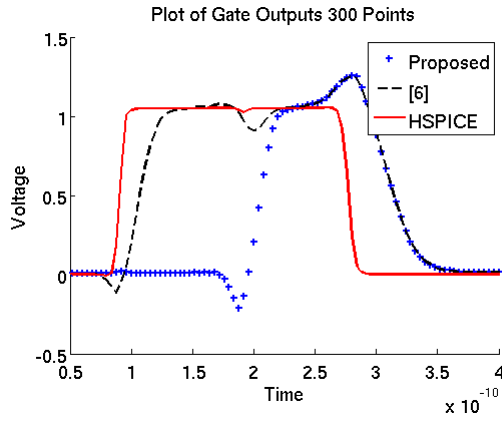
Table 2.1. Proposed and [33] compared to HSPICE for 2400 points.

Circuit	MSE [65nm] ( $\times 10^{-3}$ )		MSE [45nm] ( $\times 10^{-3}$ )		MSE [32nm] ( $\times 10^{-3}$ )	
	Proposed	[33]	Proposed	[33]	Proposed	[33]
NAND2R	0.760	108	0.453	35.6	0.190	45.6
NAND2F	1.000	110	0.550	67.8	0.467	50.9
NOR2R	0.220	23.0	0.345	133.0	0.405	146.0
NOR2F	0.986	103	0.678	42.5	0.493	33.9
NAND3R	0.887	115	0.785	55.6	0.352	39.2
NAND3F	0.998	98.9	0.467	39.8	0.465	30.8

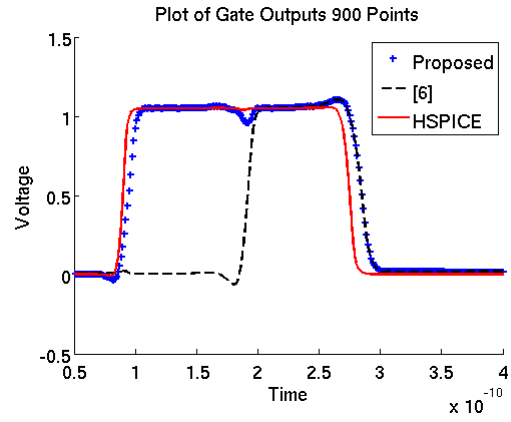
Table 2.2. Execution Time vs Accuracy ( $\times 10^{-3}$ )

Points	Error Proposed	Error [33]	Execution Time Proposed	Execution Time [33]	Execution Time HSPICE
NAND2F					
2400	0.486	85.7	8.39	8.03	100
1200	3.90	92.0	4.34	4.01	60
900	7.40	97.8	3.43	3.05	50
300	41.1	115.9	1.23	1.18	20
NAND3F					
2400	0.529	37.5	12.1	11.4	120
1200	6.40	38.7	6.40	5.90	70
900	13.0	42.8	4.90	4.50	60
300	60.9	85.0	1.91	1.50	30



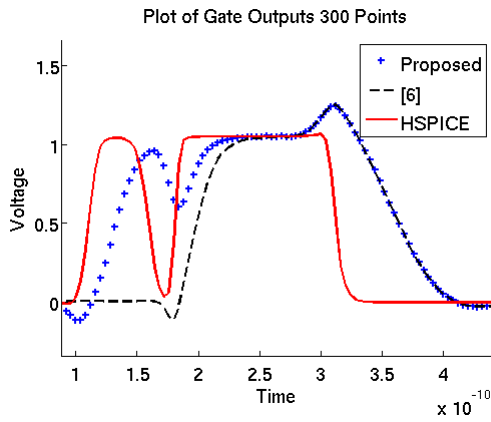


(a)

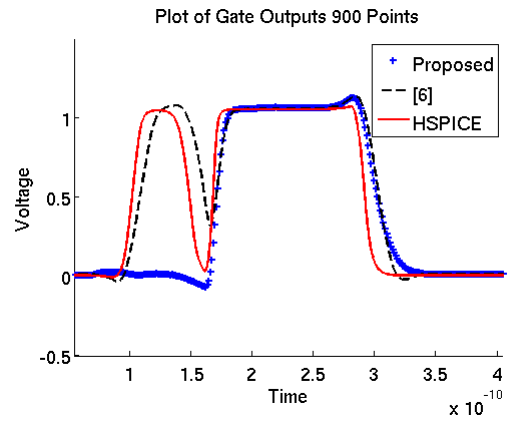


(b)

Figure 2.5. (a) Output of NAND2F with 300 points. (b) Output of NAND2F with 900 points



(a)



(b)

Figure 2.6. (a) Output of NAND3F with 300 points. (b) Output of NAND3F with 900 points

## 2.3 CONCLUSION

In this section, a model was proposed that improves SEMT and MEMT error simulation by calculating the output pulse shape in the presence of convergence with an MSE of less than  $7.40 \times 10^{-3}$  for 2-input gates and  $13 \times 10^{-3}$  for 3-input gates using 900 points and a speed-up of 15X and 12X respectively compared to HSPICE. Future work with our model includes the study of the number of points effect the simulation result on large benchmarks and the study of SEMT and MEMT on large standard benchmark circuits using the proposed model.

## CHAPTER 3

### ACCURATE SOFT ERROR SIMULATION USING PARTITIONING

As process technology continues to the scale down, the likelihood of a radiation induced error increases. This trend provides a need for accurate and efficient methods to calculate the soft error rate of a given circuit. Since it is expensive and time consuming to design a circuit and test at a later time, efficient tools that can accurately determine the error rate for a given netlist before fabrication can significantly reduce the design time. However, it has proven to be very difficult to characterize combinational circuits due to three pulse masking factors. The masking factors are as follows:

1. Logical Masking: The pulse is removed due to the boolean inputs on the gate during pulse generation or a controlling value on an off-input during pulse propagation.
2. Electrical Masking: The first order RLC effects within the gate cause the pulse to degrade and fully attenuate.
3. Temporal Masking: The pulse arrives at a flip-flop but does not meet the set-up and hold time parameters.

Considering the above factors, all current soft error simulators consist of modeling a particle at a specific energy which is used in an equation to model the error pulse shape [40]. Once the pulse is generated, it is propagated through the combinational logic to the output. It is then assumed that the output is connected to a flip-flop thus the set-up and hold times are considered. It has been shown in [20, 21] that concurrent estimation of all masking factors is required in order to ensure that the soft error rate is calculated accurately. However, due to limitations on simulation time and available memory, efficient and accurate consideration of all masking factors has proven to be a difficult and ongoing problem.

There have been many approaches to improve the estimation of the three factors. The most common method to calculate the logical masking effect is the use of a Monte Carlo based simulation which consists of randomly applying patterns as in [33, 38, 24, 31, 30]. While this approach is very easy to implement, it suffers from accuracy and run time issues when the circuit has more than 30 inputs. In the case of large circuits, the simulator must resort to generating random input patterns which cover only a small subspace of the possible inputs. This implies that the error in logical masking estimation increases with the circuit size.

To enhance on Monte Carlo based simulation, the authors in [39] propose the use of probabilistic transfer matrices (PTM). This approach consists of representing the Boolean functions within a matrix in which Boolean operations can be applied. While being exact in the estimation of the probability, the use of PTMs are non-ideal since they have high memory and simulation time costs that explode on small circuits. The authors in [12] propose an enhanced alternative to the PTM method which uses stochastic logic to calculate the signal probabilities for logical masking estimation. While their method is fast, it relies on the use of random input patterns that can limit the accuracy of the calculation.

Another common approach found in [4, 17] uses the correlation coefficient method (CCM) proposed in [7]. This method uses basic gate signal probability functions to determine the output. For example, the probability of an AND gate is given as  $P(O) = P(A)P(B)$  where  $P(O)$  is the output probability,  $P(A)$  is the probability of input  $A$  being "1" and  $P(B)$  is the probability of inputs  $B$  being "1". A drawback with using the basic probability functions is that they are unable to consider the correlations between signals. CCM improves on this by adding a correlation factor to estimate the signal probability. While CCM does provide quick estimation of the logical masking effects with virtually no memory overhead, it can only estimate first order correlation. This implies that as the circuit gets larger, the accuracy of the method substantially decreases.

The last approach discussed in this section, which is of most concern in this paper,

is the use of binary decision diagrams (BDDs). BDDs provide a very accurate way to determine the logical masking effects since they are a condensed canonical form of a Boolean function. The main problem with BDDs however, is that they scale exponentially with circuit size thus leading to the possibility of blow up on medium to large circuits. Despite this problem, there have been many proposed simulators that use BDDs [36, 20, 21]. The main advantage to the use of BDDs is that all input patterns can be considered concurrently thus allowing for exact calculation of logical masking probabilities.

To alleviate the inherent blow up problem, the authors in [36] propose the use of partitioning. In their work, they show that the use of partitioning allows for much faster calculation of the soft error with no threat of blow up at a cost of accuracy. The authors in [21] also investigate partitioning but do not give an in depth study on how partitioning effects the output error. In both approaches, the circuit is partitioned before simulation with each partition size being set to a predetermined value. In this section, a new tool is proposed that uses partitioning to reduce the simulation time. Compared to existing approaches, the new tool allows for the integration of an accurate electrical masking model that can calculate the output pulse shape for any input pulse.

In addition to the partitioning approach, the proposed simulator has the capability of considering multiple concurrent pulse strikes commonly referred to as multiple event transients (METs). The consideration of multiple strikes has become of greater concern due to smaller feature sizes, leading to a lower critical charge for upset, and the closer placement of transistors [32]. These type of events can occur in two forms: single event multiple transient (SEMT) and multiple event multiple transient (MEMT). A SEMT is defined as the case where a single radiation particle upsets multiple transistors causing multiple transients pulses. A MEMT is defined as when multiple particles hit different circuit components simultaneously. While the probability of the SEMT vs the MEMT depends strongly on the radiation environment, methods developed for one type of error can easily be modified to consider the other.

There are a few existing methods that include the consideration of multiple event errors. The authors in [21] consider the correlation of METs however their method uses BDDs which tend to blow up and uses the width and amplitude to consider the pulse shape. While they do investigate the use of partitioning, their method is only tested for small circuits using only a few concurrent pulses. An additional simulation tool used for MET analysis found in [9] uses probabilistic functions. This approach has shown to be fast and efficient but is not suited for the use of complicated electrical masking methods. To improve on the existing methods, the simulator FAST\_MET (Fast and Accurate Simulation Tool for Multiple Event Transients) which uses BDDs to calculate the logical masking effect in conjunction with the accurate pulse approximation tool proposed. FAST\_MET improves on the method in [21] by allowing for the integration of accurate electrical masking models and through the extensive use and evaluation of partitioning to avoid BDD blow up.

### 3.1 PRELIMINARIES

A crucial aspect in the estimation of the logical masking effects is accurate creation of the Boolean functions. To ensure full consideration of the logical masking effects, Boolean functions must be created to evaluate the circuit inputs for pulse generation and the off-inputs for pulse propagation.

When a high energy particle strikes a transistor, the magnitude and polarity of the pulse will depend on the configuration of the transistor. In the case of CMOS logic, a transient pulse is generated when a particle hits a blocking transistor. This will, in turn cause the transistor to temporarily conduct current allowing for the generation of the voltage pulse. In Fig. 3.1 a 2 input NAND gate is given to demonstrate this mechanism. In the NAND gate, a particle can hit a single PMOS or a NMOS transistor to cause an error. However, since the transistor must be blocking, the input pattern of the gate has a large effect on where the error will occur and the polarity of the pulse.

For example, in the 2 input NAND gate in Fig. 3.1, if both inputs have a value of

"1", a strike on either of the PMOS transistors will likely cause a transient pulse. In the case of one input having a "0" value and the other a "1" value, a strike on the blocking NMOS transistor will cause an error. In the case of both inputs having a "1" value, the gate will only generate a transient pulse if both NMOS transistors are struck concurrently. Addition, the polarity of the pulse also requires consideration. If the output is has a "1" value, the pulse will start at a high value, go low and back to a high value. When the output has a "0" the pulse will start at a low value, go high and back to a low value. Based on this observation, the function for the pulse generation for the AND, NAND, OR and NOR are given in Table 3.1 for a  $N$  number of inputs with each  $i$ -th input denoted as  $I_i$ .

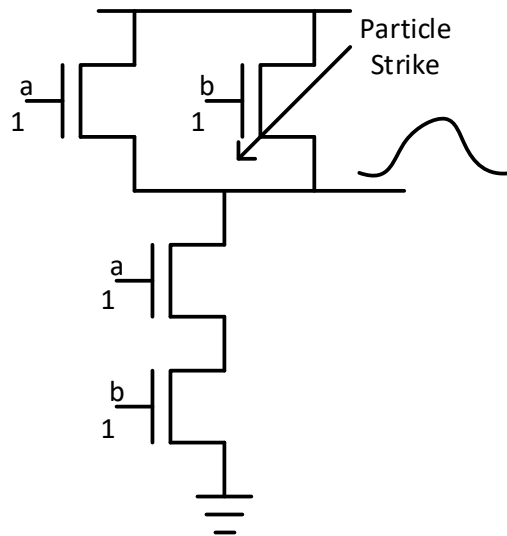


Figure 3.1. A low-high-low transient pulse being generated by a strike on the PMOS.

For the consideration of the generation of METs, the equations in Table 3.1 must be modified such that the joint probability of the error occurring is considered. In the case of pulse generation, it is assumed that a single radiation particle will strike two or more gates causing transient pulses. As stated previously, in order for a pulse with a specific polarity to be generated, the output of the gate must have a specific value. Based on this, the probability of pulses being generated ( $P_{gen}$ ) assuming it strikes  $N$  gates with sufficient

Table 3.1. Functions for Transient Pulse Generation

Gate	Polarity	Generation Function ( $F(G)$ )
AND	Rising	$F(G) = \neg(I_1 \wedge I_2 \wedge \dots I_N)$
AND	Falling	$F(G) = I_1 \wedge I_2 \wedge \dots I_N$
NAND	Rising	$F(G) = I_1 \wedge I_2 \wedge \dots I_N$
NAND	Falling	$F(G) = \neg(I_1 \wedge I_2 \wedge \dots I_N)$
OR	Rising	$F(G) = I_1 \wedge I_2 \wedge \dots I_N$
OR	Falling	$F(G) = \neg(I_1 \wedge I_2 \wedge \dots I_N)$
NOR	Rising	$F(G) = \neg(I_1 \wedge I_2 \wedge \dots I_N)$
NOR	Falling	$F(G) = I_1 \wedge I_2 \wedge \dots I_N$

energy is given in equation 3.1 where  $P(G_{i,p})$  is the probability that gate  $G_i$  is capable of generating a pulse of polarity  $p$ . If the pulse generated is rising,  $G_{i,p}$  pertains to the probability that the output of  $G_{i,p}$  is "0". If the pulse is falling,  $G_{i,p}$  is the probability of the output being "1".

$$P_{gen} = \prod_{i=1}^N P(G_{i,p}) \quad (3.1)$$

A particle hitting two transistors simultaneously allows for multiple cases since the input may be such that the struck gate is not blocking. Based on this, the total number of possible ways a strike can produce a pulse ( $P_{num}$ ) is given in equation 3.2 assuming a  $N$  number of struck gates and  $r$  as the number of gates that produce a transient pulse.

$$P_{num} = 2 * \sum_{r=1}^N \frac{N!}{(N-r)!} \quad (3.2)$$

Equation 3.2 is based on the fact that multiple pulses are not always generated when a particle strikes both transistors. As stated previously, the polarity and existence of the pulse depends on the input patterns. If one assumes that the particle hits a  $N$  number of gates, this creates a  $P_{num}$  number of cases where  $P(C_i)$  represents the probability of case  $i$  calculated using equation 3.1. The total probability of the inputs having the values to generate a SET or MET ( $P_{gen}$ ) is given below:



$$P_{gen} = \sum_{i=1}^{P_{num}} P(C_i) \quad (3.3)$$

In the case of a pulse being propagated to the input of gate, the pulse will be masked if one of the off-inputs has a controlling value. Fig. 3.2 gives a NAND gate with one of the inputs having a pulse and the other input having a controlling "0" value. In this case, the output will be a logical "1" value despite the pulse on the input.

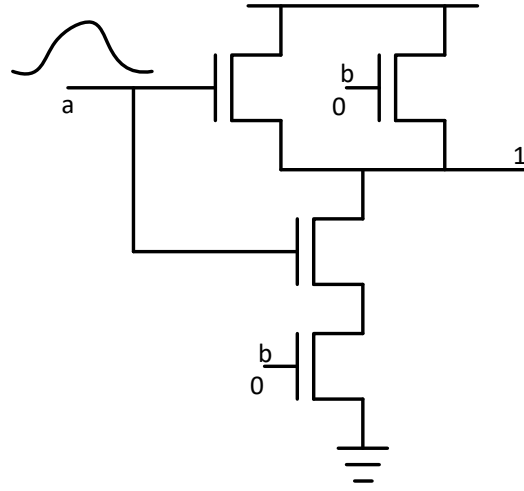


Figure 3.2. A transient pulse being masking by a controlling value on the off-input.

To calculate the logical masking factor during pulse propagation, the functions of the off-inputs are calculated using the logical AND operation. Assume that  $F(OI_i)$  is the Boolean function of the off-input  $OI_i$  representing the input patterns that allow a non-controlling value and there are a  $N$  number of inputs, the function representing the case where the pulse is propagated ( $F(M)$ ) is given below:

$$F(M) = F(OI_1) \wedge F(OI_2) \wedge \dots F(OI_N) \quad (3.4)$$

In equation 3.4 only  $N - 1$  inputs are considered since the input which contains the transient pulse is not included in the calculation.

For the consideration of METs, two or more pulses may arrive at a gate. If the two pulse case is assumed, there are 3 possible ways that the pulses can be propagated: the pulse on  $a$  is propagated and  $b$  is masked, the pulse on  $b$  is propagated and  $a$  is masked and, lastly, both pulses are propagated simultaneously providing an output pulse. Fig. 3.3 provides an illustration of the cases.

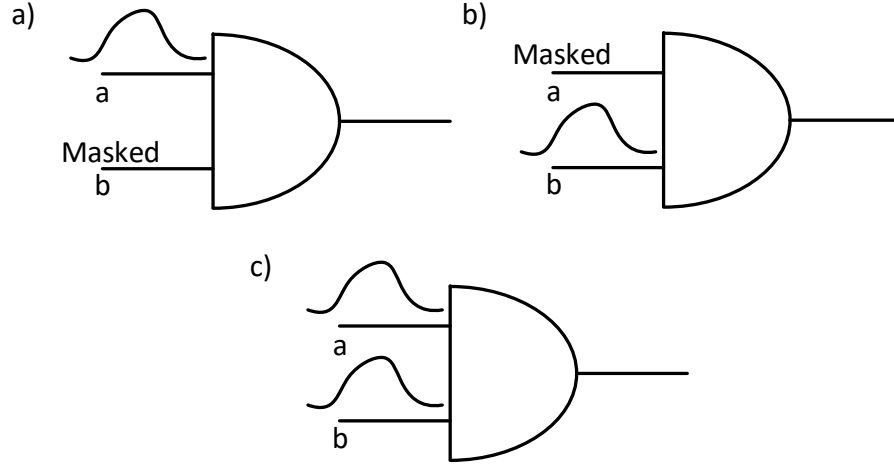


Figure 3.3. (a) A transient pulse arriving on input  $a$  and masked on  $b$ . (b) A pulse arriving on input  $b$  and masked on  $a$ . (c) Two pulses arriving on the inputs simultaneously.

Table 3.2 gives the propagation functions  $F(G)$  for each case with  $F(a_p)$  and  $F(b_p)$  being the functions representing the input values that allow for the pulse to be propagated. While this example is only for a two input gate, larger gates can be considered similarly.

Table 3.2. Functions for a MET Arriving on the Inputs of a NAND Gate

Case	Propagation Function ( $F(G)$ )
a	$F(G) = F(a_p) \wedge \neg F(b_p)$
b	$F(G) = \neg F(a_p) \wedge F(b_p)$
c	$F(G) = F(a_p) \wedge F(b_p)$

Based off equation 3.2, the total number of cases can be calculated as the number of combinations assuming  $i = 1, 2, \dots, N$  simultaneous pulses. Assuming that  $P_{num}$  is the

number of possible cases and  $r_i$  represents a  $i$  number of concurrent pulses, the equation for the total number of propagation cases is given below:

$$P_{num} = \sum_{i=1}^N \frac{N!}{(N - r_i)!} \quad (3.5)$$

Based on equation 3.5, the total probability at a gate  $G$  ( $P(G_{tot})$ ) with the assumption that  $P(G_i)$  is the probability of pulse  $i$  arriving at gate  $G$  is given below:

$$P(G_{tot}) = \sum_i^{P_{num}} P(G_i) \quad (3.6)$$

When a SET or MET is generated, it is represented as an event  $E_k$  with  $k$  being the event number. For each event, the generated transients are propagated through gates to the primary outputs. Once the pulse arrive at the gate output, the temporal masking probability can be considered. Based off the equation found in [27], let  $P_{L,i}$  be the probability of the pulse being latch on primary output  $i$ ,  $W$  be the pulse width,  $t_{setup}$  and  $t_{hold}$  being the setup and hold times respectively and  $T_{clk}$  be the clock period, the probability of a pulse being latch on an output flip-flop is calculated in the below equation:

$$P_{L,i} = \frac{W - (t_{setup} + t_{hold})}{T_{clk}} \quad (3.7)$$

Assuming that the pulses from event  $E_k$  propagate to a  $M$  number of primary outputs with each output having a probability of  $P_{L,i}$  the probability of error for event  $E_k$  is calculated as the following:

$$P(E_k) = \sum_{i=1}^M P_{L,i} \quad (3.8)$$

To evaluate the error probability for all events, the mean error susceptibility (MES), defined in [21] is used. The MES represents the average probability of error for all events combined. Assuming that there are a  $n_E$  number of events and  $n_d$  number of input proba-

bility distributions, the MES is found using the following equation:

$$MES = \sum_{k=1}^{n_E} \frac{P(E_k)}{n_E n_d} \quad (3.9)$$

Based off the MES, the soft error rate (*SER*) can be calculated as in below where  $R_{eff}$  represents the effective concentration of particle in the given area,  $P_{eff}$  is the probability of a particle hitting the sensitive region of a transistor and  $A$  as the area of the sensitive volumes.

$$SER = MES * R_{eff} * P_{eff} * A \quad (3.10)$$

### 3.2 DESCRIPTION OF THE FAST\_MET SIMULATOR

FAST\_MET is a fast and accurate tool to estimate the soft error rate. The method aims to provide a good trade-off between simulation time and accurate SER calculation. To accurately determine the logical masking effects, BDDs are used. A BDD is a tree structure which can be used to represent a Boolean function. The first instance of the use of BDDs for function representation was proposed in [2] and has continued to be used regularly in circuit simulation and synthesis since. Fig. 3.4 gives the representation of a AND gate as BDD. Note that the structure consists of nodes which represent variables. For each node, there are two branches, one denoting a "true" value (solid line) and the other a "false" value (dashed line). The BDD graph also consists of multiple paths to two terminating nodes, one with a "true" value denoted by "1" and one with a "false" value denoted by "0".

The main advantage of a BDD is that they provide an efficient canonical form to represent a Boolean function. Specifically, this means that Boolean operations (such as AND, NAND, OR ... etc) can be easily and quickly applied to the graph. In the case of the FAST\_MET simulator, BDD representation of the functions are used to both calculate

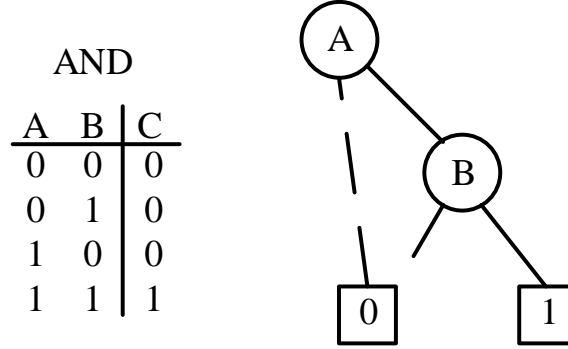


Figure 3.4. The truth table and corresponding BDD for an AND gate.

the signal probability and the probability for pulse propagation.

In FAST\_MET, the goal is to accurately and efficiently represent a sensitization function as a BDD. A sensitization function is defined as the Boolean function that allows for a generated transient pulse to propagate through a circuit. Once the function is derived, it can be evaluated to determine the logical masking probability of a pulse.

In addition to BDDs, the FAST\_MET simulator operates in a topological order. This implies that the primary inputs will be visited first. For each primary input visited, it is represented as a variable in a BDD. Once all inputs are visited, the simulator starts on the gates within the circuit. When a gate is visited, the sensitization function representing the gate output is calculated. The "1" terminal on this BDD represents the Boolean function that makes the output a "1" value. To increase efficiency, the FAST\_MET simulator also generates a rising and falling pulse at each gate using the method proposed in Chapter 2 for a specific energy.

When a pulse is generated, the gate input functions are considered using the input sensitization functions. If the pulse is rising (starting from a low value then going high), the pulse is only generated when the output is low. To consider this, the BDD function is inverted such that a termination to "1" represents the case at which the pulse exists. Similarly, if the pulse is falling, the output must be a high value. Since the function

already denotes a high value on the output, the sensitization function can be used directly. Additionally, as discussed in Section 3.1, in the case of a gate having multiple inputs, the output function must be considered according to Table 3.1. Using the table, the correct generation function can be found by applying the function for the gate specified. Fig. 3.5 provides an example of the creation of the generation functions for a gate.

To consider the MET case, the location and correlation between the pulses must be considered. As discussed in [5], layout information is important in accurately modeling the location of the MET for a real design since gates that are near each other in a netlist may not necessarily be close in the layout. In this section, for the sake of simplicity, METs are generated using close proximity nodes from the netlist. Just for simple comparison and analysis, using the netlist is not problematic.

To determine the logical masking effects of a MET, the correlation between the inputs of both struck gates are found by using the logical AND operation between the generation functions derived as in Table 3.1. The logic behind this operation is that all pulses must be logically sensitized concurrently. Since the functions may share common inputs, they may have dependencies that must be considered.

In the case of pulse propagation, the sensitization functions are represented as BDDs. As discussed in Section 3.1, a pulse will propagate through a gate if all off-inputs have a non-controlling value. As can be observed in equation 3.4, the functions representing the case at which the off-input has a non-controlling value are logically AND'd together. The basis behind this idea is that all off-inputs must be non-controlling in order for the pulse to be propagated.

To consider pulse propagation in the FAST\_MET simulator, the logic functions are first represented in a BDD. A logic function is the Boolean function which provides the variable states that allow a "1" and a "0" value. The sensitization function for the gate is then determined by using the logical AND operation on each off-input BDD. If the non-controlling value for the gate is a "0", such as in a OR gate, the BDDs are inverted so that

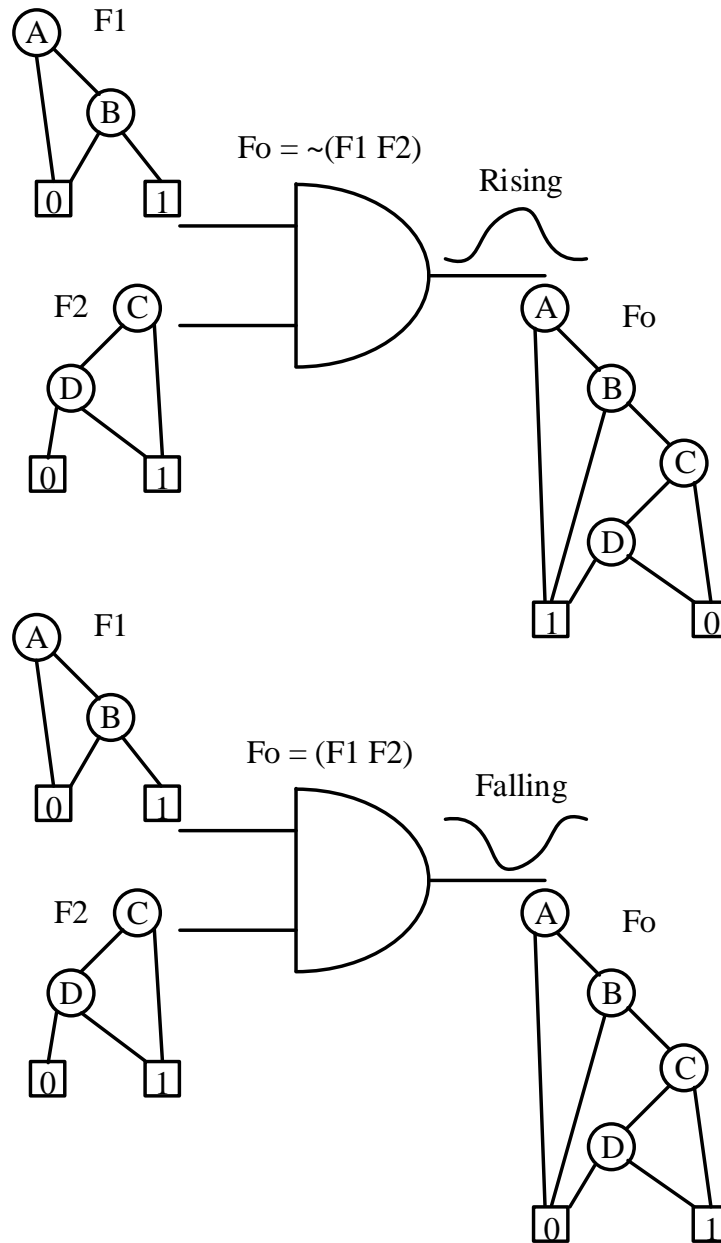


Figure 3.5. An example of the BDD creation for a rising and falling pulse.

the "1" terminal represents the case at which the pulse propagates. If the non-controlling value is a "1" the BDDs are left unchanged. The result of this operation is a single BDD in which the paths that lead to the "1" terminal represent all of the input patterns that allow the pulse to be propagated while the paths that lead to the "0" terminal represent the patterns that will mask the pulse.

In the case of multiple pulses arriving at a gate simultaneously referred to as convergent pulses, the off-inputs are considered as discussed previously. As shown in Table 3.2, a MET can result in multiple output cases which must be considered. The first case is where only one pulse will propagate which is calculated as previously discussed. All remaining cases include the circumstances where multiple pulses will arrive at the gate input simultaneously. In these cases, the logic function is determined by using the AND operation on the input functions which have a pulse. The resulting function from this operation is then AND'd with the non-controlling off-input logic functions. The basis behind this idea is that the pulse sensitization functions give the cases where a pulse will arrive at the gate input. For example if two pulses arrive simultaneously, both pulses must be sensitized concurrently. The logical AND states that both functions must evaluate to true indicating that the pulse arrives at the inputs while all off-inputs must be non-controlling for the pulse to be propagated.

### 3.3 FAST\_MET SIMULATION FLOW

The goal of the FAST\_MET simulator is to calculate the SER of a circuit. To do this, the simulator must inject pulses in each gate. The main difficulty for this type of simulation is that for even relatively small circuits, it may take an intractable amount of simulation time and memory to consider all generation and propagation cases. The FAST\_MET simulator provides a good trade off compared to existing methods since it uses the pulse approximation method in Chapter 2 in conjunction with BDDs to accurately and quickly determine the SER.



The first step for the FAST\_MET simulator is to partition the circuit. As shown in [36, 21], the use of BDD functions can lead to memory blow up. To alleviate this issue, the FAST\_MET partitions the circuit into a user defined number of parts. It is shown in Section 3.4 that the use of partitioning can substantially reduce the simulation time at a cost of accuracy in the SER estimation. For this work, the circuits were partitioned using the Fiduccia and Mattheyses (FM) algorithm [11] which allows for a circuit to be partitioned into two parts in linear time. To achieve a  $k$  part partition, the FM algorithm can be applied recursively a  $k-1$  number of times. This algorithm was chosen based off its linear time complexity and relatively easy integration into the FAST\_MET tool. Through proper integration, one can also use other partitioning tools such as hMetis [19]. Partitioning the circuit improves the performance by reducing the size of the logic and pulse sensitization BDDs. For moderately large circuits or larger, partitioning must be used since the BDD functions increase exponentially in size with respect to the number of inputs.

After partitioning, the FAST\_MET simulator processes the primary inputs of the circuit. At each input node, the signal probability is set and the BDD variables are initialized such that each input is a variable. In this work, it is assumed that there are not pulses on the inputs, however, if the method is used for the consideration of sequential circuits such as in [22], pulses can be injected on the inputs. This is due to the inputs being connected to the output of a flip-flop which drives the circuit.

After the inputs are processed, the internal nodes are visited in a topological order. At each node, at least one pulse is injected for a given charge. The particle effect is modeled as a current pulse using equation 1.1. A current source is placed on the output of the gate to simulate the particle effect and the resulting pulse shape is determined using the method discussed in Chapter 2. Multiple pulses can be injected concurrently in a gate, however, the simulation time and memory resources increase substantially as more pulses are injected.

Additionally, FAST\_MET is able to simulate the effects of a MET. To accommodate this, each injected pulse or MET is given an event number  $k$  to create an event  $E_k$ . This is

important since FAST\_MET injects pulses at each gate concurrently to speed up the SER characterization. Thus, when an MET is generated, a pulse is created on two or more gates which all have the same event number.

Once the pulses are injected and the corresponding BDD functions are determined, FAST\_MET propagates all pulses found on the gate inputs. Each fanin gate is checked individually for pulses. If a pulse is found the propagation function is determined as in Section 3.2. If the function evaluates to a false value, implying that there are no input patterns that allow the pulse to be propagated, the pulse is not considered further. However, if the function is not false, the output pulse shape is calculated. This operation continues for each pulse found on the inputs.

When a node is on the edge of a partition circuit, the values calculated on the previous partition must be considered. To do this, each node on the output edge of a partition is evaluated to determine their probability. This value is stored with the pulse which is propagated between the partitions. Since a partition may be separated from the primary inputs that drive it, virtual inputs are created which contain all pulses with their attached probabilities and the probabilities from the logic functions of the gate that was connected to the input on the previous partition. Using these values, the probabilities are multiplied by the evaluated probability of the function to determine the overall probability. This is demonstrated in Fig. 3.6.

An issue that was found through numerous simulations using this method is that the functions will not evaluate to a false value even if the pulse does not propagate since a single partition does not consider the whole circuit. This, in effect, substantially reduces the number of pulses that are attenuated which can offset any improvement increases gained through partitioning. To solve this issue, it was found that setting a minimum probability threshold ( $P_{thr}$ ) in which only pulses with a probability greater than this number are considered substantially saved simulation time with a slight increase in estimation error.

Once each individual pulse is propagated, a routine searches the input pulses and

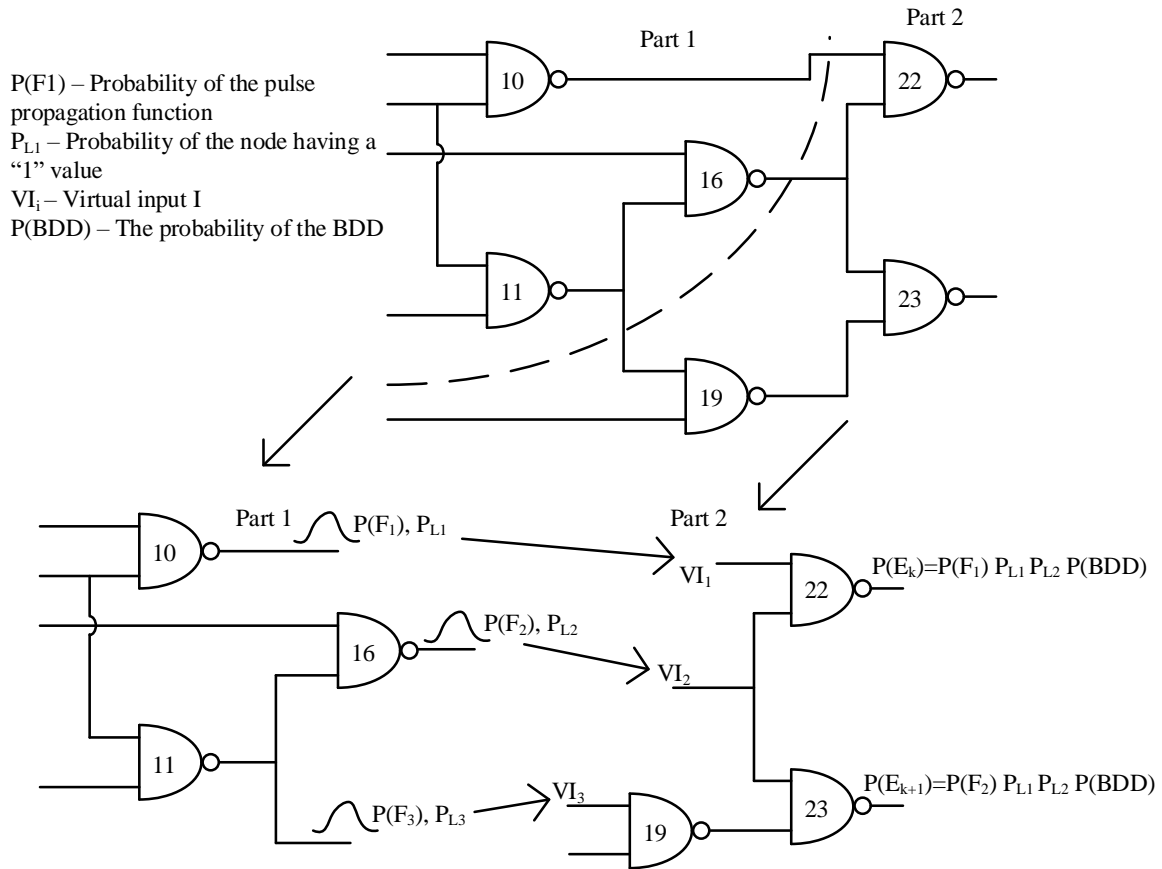


Figure 3.6. An example on c17 showing how the pulses are propagated between partitions.

organizes them into a hash table based on their event number. For each event  $E_k$ , the pulses are checked for overlap. If the pulses do overlap, they are treated as convergent pulses and are considered as previously discussed. One issue that arises during this step is that there may be many possible cases in which the pulses overlap. For example, there may be 2, 3 or more concurrent overlapping pulses. In this work, to reduce the simulation time, only two pulses are considered. While the electrical masking model can accurately consider the effects of 3 or more convergent pulses, it is assumed that the likelihood of the pulses propagating is low.

Lastly, when the simulator processes a primary or partition output, the BDD functions for each pulse are solved to determine the probability. If the probability is solved on a primary output, the MES is determined as discussed in 3.1. An overview of the whole simulation flow is given in Algorithm 1.

---

**Algorithm 1** FAST\_MET

---

```

1: Set Process Parameters
2: Parse Netlist
3: Organize Circuit Topologically
4: Parse Circuit Into  $k$  Parts
5: for Each Partition do
6:   Extract Partition
7:   Set Virtual Inputs
8:   for Each Gate do
9:     Generate Pulse
10:    Generate Sensitization Function
11:    Propagate Pulse
12:    Calculate Convergence
13:    if Primary Output then
14:      Calculate  $P_{L,i}$ 
15:    else if Edge of Partition then
16:      Solve BDD Probability
17:    end if
18:  end for
19: end for

```

---

---

**Algorithm 2** Generate Pulse

---

```
1: F = Calculate Generation Function
2: if F != false then
3:   Calculate Pulse Shape
4:   Add To Pulse List
5: end if
```

---

---

**Algorithm 3** Propagate Pulse

---

```
1:  $F_{sens}$  = Sensitization Function
2: for Each Input Pulse  $i$  do
3:    $F_{sens} = F_{sens}(i) \wedge F_{sens}$ 
4: end for
5: if ( $F_{sens} \neq \text{false}$ )  $\wedge$  (Solve BDD( $F_{sens}$ ) >  $P_{thr}$ ) then
6:   Calculate Pulse Shape
7:   Add To Pulse List
8: end if
```

---

---

**Algorithm 4** Calculate Convergence

---

```
1: Load Hash Table For Each Event
2: for Each Event do
3:   I = Check For Overlap
4:   if I == true then
5:     F = Calculate Convergence Function
6:     if (F != false)  $\wedge$  (Solve BDD(F) >  $P_{thr}$ ) then
7:       Calculate Pulse Shape
8:       Add To Pulse List
9:     end if
10:   end if
11: end for
```

---

### 3.4 RESULTS

The FAST\_MET simulator was test on the ISCAS 85 combinational benchmark circuits. FAST\_MET was implemented in C++ on a machine with a quad core i7 and 16GB of RAM. The transistor look-up tables were characterized in HSPICE using the 32nm Predictive Technology Model (PTM) [39] where  $V_{dd}$  was set to 1.05V. The unit gate capacitance was set to a constant capacitance of 2 fF to simulate the loading effects of the gates and interconnects. To calculate the MES, a single pulse with an energy of 15 fC was applied to each gate using equation 1.1 with  $\tau$  being set to  $32 \times 10^{-15}$ . It was assumed that each output is connected to a flip-flop implemented in the same process library with a setup time ( $t_{setup}$ ) of 22 ps and a hold time ( $t_{hold}$ ) of -7 ps in accordance to [26]. For all simulations, the probability of MET was set to 10% and the error threshold for simulation  $P_{thr}$  was set to 0.025 when partitioning was used. When an MET occurred, it was injected to a neighboring gate based on the netlist.

First, the accuracy of FAST\_MET without partitioning is compared to Monte Carlo simulation on c17 and c880 to ensure that the probabilistic functions provided here exactly compute the logical masking effect. To ensure that both simulators use the same pulses, the pulse injection routine in FAST\_MET was moved to before the main loop. Larger circuits were not tested since they lead to long simulation times and lead to memory blow up for FAST\_MET without the use of partitioning. As can be observed in the Table 3.3, FAST\_MET provides exact estimation of the output error with an over 5x speedup.

Table 3.3. Comparision of FAST\_MET vs Monte Carlo

Circuit	Simulator	MES	Run Time
c17	Monte Carlo	$3.92 \times 10^{-4}$	0.429
c880	Monte Carlo	$4.15 \times 10^{-4}$	9110.35
c17	FAST_MET	$3.92 \times 10^{-4}$	0.0974
c880	FAST_MET	$4.15 \times 10^{-4}$	1909.37

The FAST\_MET simulator can provide an additional speedup through the use of partitioning. Table 3.4 provides the MES and the simulation time for various ISCAS 85 circuits

while changing the partition size. According to the results, it shows that partitioning can provide an upward of 20x reduction in simulation time at a cost of accuracy compared to not using partitioning. Based off the results, it is shown that increasing the number of partitions follows the law of diminishing returns. For example in c880, the use of 2 partitions reduces the simulation time by 13x while the use of 4 partitions reduces the simulation time by 18x. While this is still a large decrease in simulation time, the error is increased by 4x. Additionally, it can be seen that in c17 the use of partitioning actually increases the simulation time. This is due to the time to simulate the circuit is less than the overhead incurred by the partitioning algorithm. Based on this result, it is suggested that the partition size is carefully selected to ensure that the error is minimized.

Table 3.4. FAST\_MET Performance vs Number of Partitions

Circuit	Parts	MES	Run Time
c17	2	$3.83 \times 10^{-4}$	0.1377
c880	2	$3.13 \times 10^{-4}$	139.54
c880	4	$8.37 \times 10^{-4}$	107.72
c1355	2	$1.96 \times 10^{-4}$	459.00
c1355	4	$2.85 \times 10^{-4}$	242.60
c1355	8	$3.67 \times 10^{-4}$	167.961

### 3.5 CONCLUSION

In this chapter, the FAST\_MET simulator is proposed. First, probabilistic equations were provided which allow for accurate consideration of all propagation effects. These equations were then represented as BDDs and used to determine the logical masking effect. In the results it was shown that the tool provides a good trade-off between simulation time and accuracy due to the use of partitioning and an accurate masking model. Additionally, it is shown that the number of partitions reduces the simulation time up 18x compared to not using partitioning and 90x compared to Monte Carlo simulation.

An open question that may occur is the consideration of process variations. In a

realistic case, the transistor parameters given in the standard cells may vary considerably. For this reason, it is prudent to simulate circuits in which the gate process parameters vary by some amount. The electrical masking model proposed in this dissertation is able to consider process variations by characterizing look up tables for transistors that have differing parameters. Base of this, it is recommended to characterize at least 3 transistors look up tables, one with the lowest possible width, one with the highest width and one with an ideal value. Using these tables, multiple copies of the circuit can be created in which gates build randomly with the different tables are injected. For each circuit copied, the SER can be computed and analyzed.



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## APPENDICES

# APPENDIX I

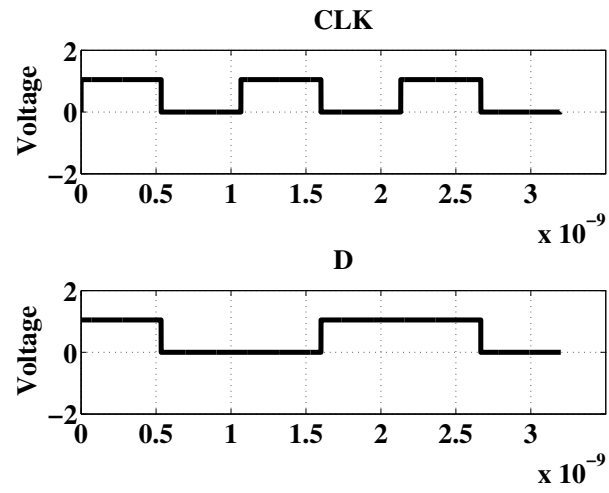


Figure 3.7. Waveforms for CLK and D.

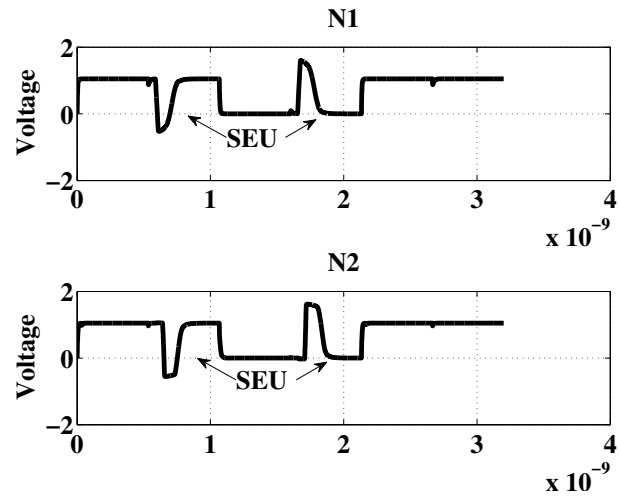


Figure 3.8. Node pair n1 and n2 upset and recovery.

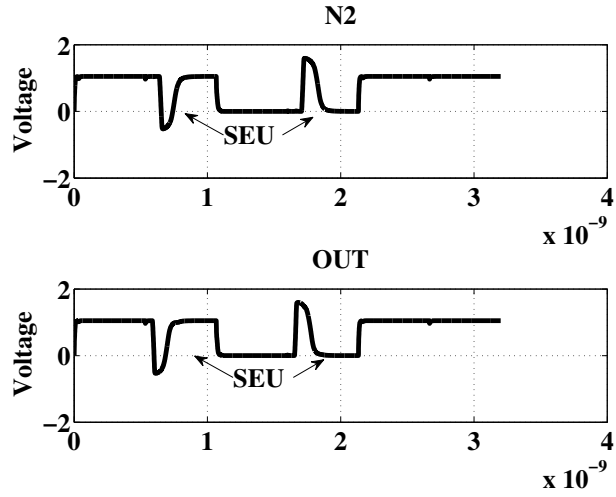


Figure 3.9. Node pair n2 and out upset and recovery.

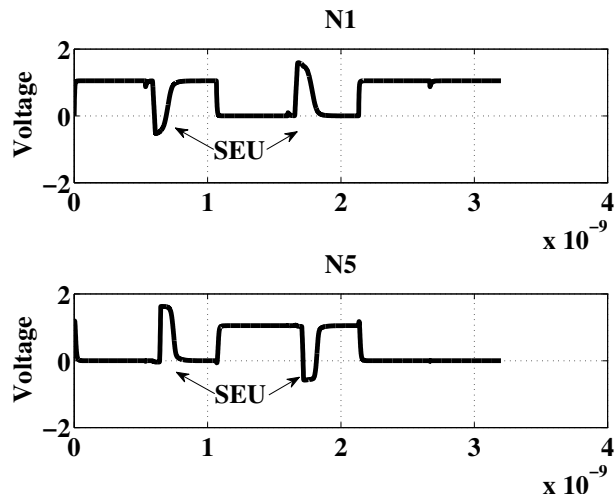


Figure 3.10. Node pair n1 and n5 upset and recovery.



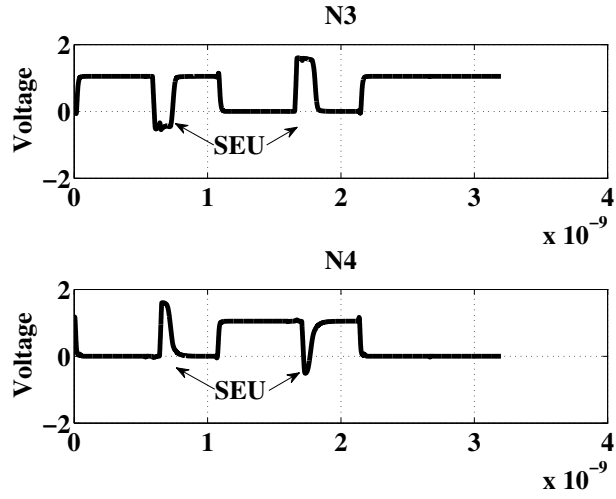


Figure 3.11. Node pair n3 and n4 upset and recovery.

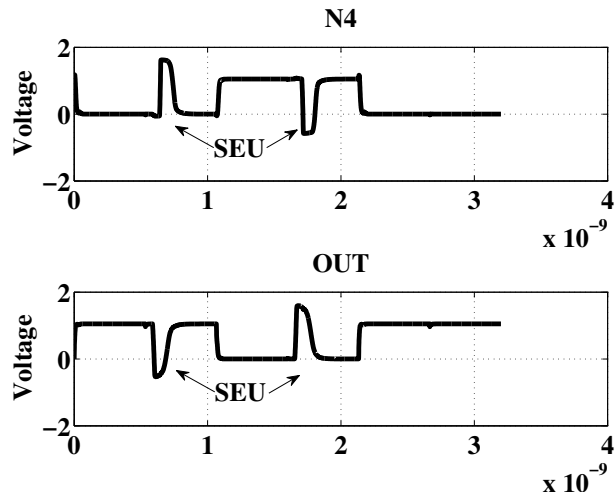


Figure 3.12. Node pair n4 and out upset and recovery.

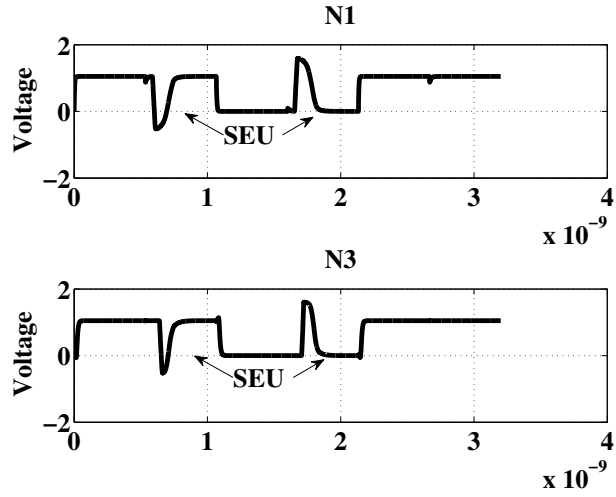


Figure 3.13. Node pair n1 and n3 upset and recovery.

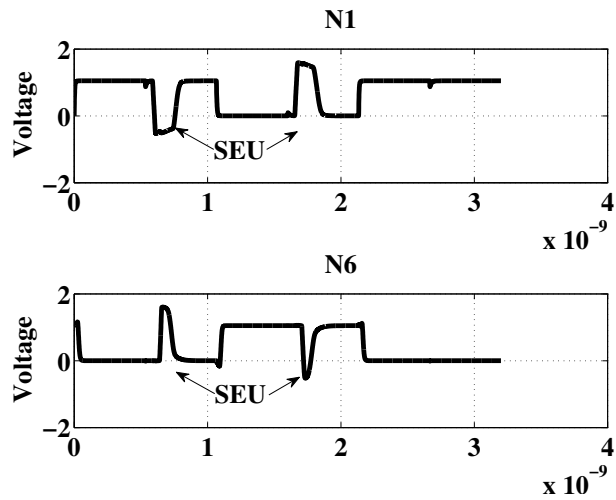


Figure 3.14. Node pair n1 and n6 upset and recovery.

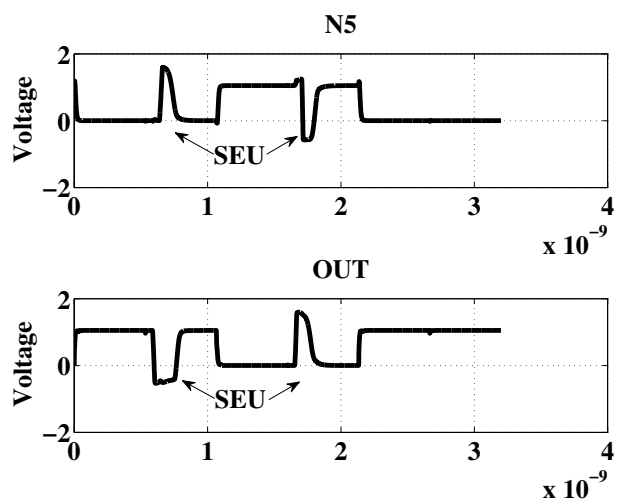


Figure 3.15. Node pair n5 and out upset and recovery.

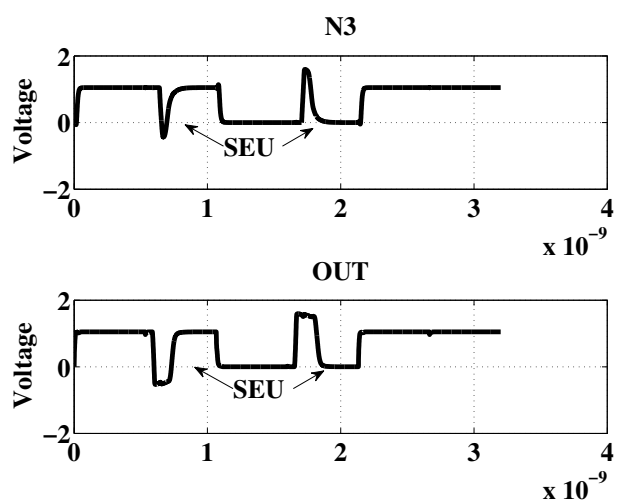


Figure 3.16. Node pair n3 and out upset and recovery.

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Publications:

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