Design Methodology Based on Carbon Nanotube Field Effect Transistor (CNFET)

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by 
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Abstract

This thesis investigates design issues of high speed and low power circuit design using CNTFET Technology. In this thesis modeling and performance benchmarking for nanoscale devices and circuits have been performed for both nanoscale CMOS and carbon nanotube field effect transistor (CNFETs) technologies. Carbon nanotubes with their superior transport properties, excellent thermal conductivities, and high current drivability turned out to be a potential alternative device to the bulk CMOS technology. However, the CNFET technology has new parameters and characteristics which determine the performances such as current driving capability, speed, power consumption and area of circuits. As a result, new design methodology is needed to optimize performances.

This research presents a development of systematic design method to optimize circuit speed and power consumption. The optimization methods are different from traditional CMOS circuit design and characteristics of circuits. In this thesis, as a demand for these circumstances, three optimization methods are proposed and some traditional CMOS circuits are modified for CNFET and CNT interconnect technologies. The optimization methods explored in this thesis include digital circuit design, memory circuit design and high speed on chip I/O circuits.

In order to test the effectiveness of the design method, CNFET and CNT interconnect models have been developed and extensive HSPICE simulations have been performed in realistic environments considering screening effects, various noises and PVT variation. The simulation results show that proposed methodologies and modified circuits performed high speed and consumed low power compared to non-optimized and traditional circuits.
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First of all, I must thank Dr. Yong-Bin Kim, my academic and research advisor, not only for his invaluable advice and encouragement leading me in the process of fulfilling my master research, but also for his great guidance and help during my four year PhD study. I would also like to thank the members of committee.

Young Bok Kim
Boston, MA
For my parents
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Chapter 1

Introduction

Ever since the 0.35\(\mu\)m node, the gate length of MOSFET has entered the deep-submicron region. 65 nm technology becomes the mainstream since 2006, and 45 nm technology has been announced in 2007. As CMOS continues to scale deeper into the nanoscale, various device non-idealities cause the I-V characteristics to be substantially different from well-tempered MOSFETs. It becomes more difficult to further improve device/circuit performance by reducing the physical gate length. The discrepancy between the fabricated physical gate length and the ITRS [1] projected gate length becomes larger as the technology advances, as shown in Fig 1.1. On the other hand, as the major driving force for the semiconductor industry, the device contacted gate pitch (Lpitch) is scaled down by a factor of 0.7 every technology node.

The last few years witnessed a dramatic increase in nanotechnology research,
especially the nanoelectronics. These technologies vary in their maturity. The exciting opportunity is to design complex electronic circuits using the cutting-edge silicon technology and/or the novel nanometer-scale transistors in the future. Carbon nanotubes (CNTs) are at the forefront of these new materials because of the unique mechanical and electronic properties. Carbon nanotube field effect transistor (CNFET) is the most promising technology to extend or complement traditional silicon technology due to three reasons: First, the operation principle and the device structure are similar to CMOS devices; we can reuse the established CMOS design infrastructure. Second, we can reuse CMOS fabrication process. And the most important reason is that CNFET has the best experimentally demonstrated device current carrying ability to date [3][4][5].
Technology boosters such as strain have helped the continuation of CMOS historic performance trend up to 45 nm node. As device physical gate length is reduced to below 25 nm at/beyond 65 nm technology node, various leakage currents and device parameter variation become the most important considerations for device optimization. In fact, it can be argued that reduction of gate length below 25 nm may not offer the same advantage as short-gate devices had provided historically in terms of power and performance at the system level [8]. The major detractors are: the lack of a thin equivalent gate oxide (with low leakage current) for effective short channel effect control, the increasing contribution of the fringing parasitic capacitance to the total gate capacitance, and the rising contribution of the source/drain resistance to the total device on-resistance.

1.1 Nanoscale MOSFET

As CMOS continues to scale deeper into the nanoscale, various device non-idealities cause the I-V characteristics to be substantially different from well-tempered MOSFETs. For example, the source/drain series resistance is now a significant component of the total on-resistance. Proposals of metal contacted (Schottky) source/drain UTB SOI FET [6] also alter the I-V characteristics significantly. Novel non-Si devices such as the carbon nanotube FETs (CNFETs) operate with completely different device physics with quasiballistic transport in the channel [3] and Schottky barriers at the source/drain contacts [7].
1.2 Carbon Nanotube Field-Effect Transistors

As one of the promising new devices, CNFET avoid most of the fundamental limitations for traditional silicon devices. All the carbon atoms in CNT are bonded to each other with sp2 hybridization and there is no dangling bond which enables the integration with high-k dielectric materials. In the next section, we will introduce the basic properties of CNFET.

1.2.1 Carbon nanotube

A single-walled carbon nanotube (SWCNT) can be visualized as a sheet of graphite which is rolled up and joined together along a wrapping vector \( Ch = n_1\vec{a}_1 + n_2\vec{a}_2 \), where \([\vec{a}_1, \vec{a}_2]\) are lattice unit vectors as shown by Fig 1.2, and the indices \((n_1, n_2)\) are positive integers that specify the chirality of the tube [9]. The length of \( Ch \) is thus the circumference of the CNT, which is given by,
1.2 Carbon Nanotube Field-Effect Transistors

\[ C_h = a \sqrt{n_1^2 + n_2^2 + n_1 n_2} \]  

Single-walled CNTs are classified into one of their groups (Figure 1.5(a)), depends on the chiral number \((n_1, n_2)\): (1) armchair \((n_1 = n_2)\), (2) zigzag \((n_1 = 0 \text{ or } n_2 = 0)\), and (3) chiral (all other indices). The diameter of the CNT is given by the formula \(D_{CNT} = Ch/\pi\). The electrons in CNT are confined within the atomic plane of graphene. Due to the quasi-1D structure of CNT, the motion of the electrons in the nanotubes is strictly restricted. Electrons may only move freely along the tube axis direction. As a result, all wide angle scatterings are prohibited. Only forward scattering and backscattering due to electron-phonon interactions are possible for the carriers in nanotubes. The experimentally observed ultra long elastic scattering mean-free-path (MFP) \((\sim 1\mu m)\) [3][4][5][10] implies ballistic or near-ballistic carrier transport. High mobility, typical in the range of \(10^3 \sim 10^4 cm^2/V\cdot s\) which are derived from conductance experiments in transistors, has been reported by a variety of studies [11][12]. Theoretical study also predicts a mobility of \(\sim 10^4 cm^2/V\cdot s\) for semiconducting CNTs [13]. The current carrying capacity of multi-walled CNTs are demonstrated to be more than \(10^9 A/cm^2\) about 3 orders higher than the maximum current carrying capacity of copper which is limited by the electron migration effect, without performance degradation during operation well above room temperature [14]. The superior carrier transport and conduction characteristic
makes CNTs desirable for nanoelectronics applications, e.g. interconnect and nanoscale devices.

1.2.2 CNFET

The operation principle of carbon nanotube field-effect transistor (CNFET) is similar to that of traditional silicon devices. This three (or four) terminal device consists of a semiconducting nanotube, acting as conducting channel, bridging the source and drain contacts. The device is turned on or off electrostatically via the gate. The quasi-1D device structure provides better gate electrostatic control over the channel region than 3D device (e.g. bulk CMOS) and 2D device (e.g. fully depleted SOI) structures [15]. In terms of the device operation mechanism, CNFET can be categorized as either Schottky Barrier (SB) controlled FET (SB-CNFT) or MOSFET-like FET[3][4][16]. The conductivity of SB-CNFT is governed by the majority carriers tunneling through the SBs at the end contacts.

The on-current and thereby device performance of SB-CNFT is determined by the contact resistance due to the presence of tunneling barriers at both or one of the source and drain contacts, instead of the channel conductance, as shown by Fig. 1.3(a). The SBs at source/drain contacts are due to the Fermi-level alignment at the metal-semiconductor interface. Both the height and the width of the SBs, and therefore the conductivity, are modulated by the gate electrostatically. SB-CNFT shows ambipolar transport behavior [18]. The
work function induced barriers at the end contacts can be made to enhance either electron or hole transport. Thus both the device polarity (n-type FET or p-type FET) and the device bias point can be adjusted by choosing the appropriate work function of source/drain contacts [17]. On the other hand, MOSFET-like CNFET exhibits unipolar behavior by suppressing either electron (pFET) or hole (nFET) transport with heavily doped source/drain. The non-tunneling potential barrier in the channel region, and thereby the conductivity, is modulated by the gate-source bias (Fig 1.3(b)).

**1.3 Thesis Outline**

Significant efforts have been made in recent years to model and simulate CNT based devices such as the CNFET [2] and a CNT interconnect [21]. However, these efforts have been concentrated mostly at a device-level. However, the dynamic performance of a circuit made of multiple CNFETs and an interconnect
has not been properly addressed in the technical literature. Therefore, it is valuable to propose circuit level design methodology for CNFET technology. This thesis focuses on low power and high speed circuit design issues for CNFET technology and evaluation of nanoscale devices and circuits including the realistic device structures, device/circuit non-idealities and process related imperfections.

This thesis is organized as follows;

In chapter 2, a detailed performance analysis and comparison between CNFETs and bulk nano CMOS technologies are undertaken at circuit-level for high performance and low power dissipation. Logic gates and benchmark circuits are investigated for performance, energy efficiency, and leakage current under different operational conditions by considering process, power supply voltage, and temperature (PVT) variations. After performance comparison with conventional CMOS.

In chapter 3, The characteristics of CNFET is analyzed and novel circuit design methodology for CNFET based combinational circuits is proposed; this methodology considers channel capacitance and current variations to determine optimum PDP(power delay product) value, circuit speed and area.

In chapter 4, As an application of memory circuit, 8T SRAM structure which strengthens the advantages of CNFET is proposed. This proposed design method improves SNM, and lowers the power.

Chapter 5 presents low power and high speed transceiver design method wit
1.3 Thesis Outline

CNFET and CNT bundle wire. The explanation for capacitive driven wire and PWAM (pulse width amplitude modulation) scheme is in chapter 5. Combining CNFET and CNT technologies to PWAM and capacitive driven wire scheme hight speed and low power transceiver is implemented. The implementation includes new circuits which consist of PWAM modulation blocks. In order to evaluate in realistic environment, various noise modeling and bit error rate (BER) estimation is taken.

In chapter 6, this thesis conclusion are addressed.
Chapter 2

Analysis of CNFET Technology

2.1 CNFET Technology

CNTs are sheets of graphene rolled into tubes; depending on the chirality (i.e., the direction in which the graphene sheet is rolled), a single-walled CNT can be either metallic or semiconducting. Semiconducting nanotubes have attracted widespread attention of device/circuit designers as an alternative possible channel implementation for high-performance transistors [22]. A typical structure of a MOSFET-like CNFET device is illustrated in Fig. 2.1. The CNT channel region is undoped, while the other regions are heavily doped, thus acting as the source/drain extended region and/or interconnects between two adjacent devices. Carbon nanotubes are high-aspect-ratio cylinders of carbon atoms. The
2.1 CNFET Technology

electrical properties of a single wall carbon nanotube (SWNT) offer the potential for molecular-scale electronics; a typical semiconducting single-wall carbon nanotube is 1.4nm in diameter with a 0.6eV bandgap (the bandgap is inversely proportional to the diameter). Recent carbon nanotube field effect transistors (CNFETs) have a metal carbide source/drain contact [25] and a top gated structure (Fig 2.1) with thin gate dielectrics[26].

![Figure 2.1: CNFET Structure.](image)

The contact resistance and the subthreshold slope of a CNFET are comparable to those of a silicon MOSFET. While a silicon FETs current drive is typically measured in current per unit device width (e.g. μA/μm), the CNFETs current is measured in current per tube (as reflecting the structure of the CNFET as an array of equal carbon nanotubes with constant spacing and fixed diameter).
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

The CV/I performance of an intrinsic CNFET is 13 times better than the CV/I performance of a bulk n-type MOSFET because the CNFETs effective gate capacitance of one CNT per gate is about 4% compared to bulk CMOS and the driving current ability of each CNT is about 50% of a bulk n-type MOSFET with minimum gate width (48 nm) at a 32 nm node (due to the ballistic transport nature of a CNT). Moreover due to the similar behavior and the current driving capability of a pFET compared to those of a nFET, the performance improvement of a pFET over a PMOS is better than the one of a nFET over a NMOS. Even though a CNFET has a leakage current in the off-state, this leakage current is controlled by the full band gap of the CNTs and the band to band tunneling; this is less than for a MOSFET [26][27].

The expected (optimistic) performance advantage of a CNFET (as described previously) is unlikely to be achievable in a real device and will be significantly degraded for the CV/I (6 times for a nFET and 14 times for a pFET) due to device/circuit non-ideal conditions. These non-ideal conditions include the series resistance of the doped source/drain region, the Schottky barrier (SB) resistance at the metal/CNT interface, the gate outer-fringe capacitance and the interconnect wiring capacitance. However, the need for low power consumption
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

and high operating frequency has resulted in geometry and supply scaling with a significant increase in operating temperature for a device. With these scaling features, the effects of systematic and random variations in process, supply voltage, and temperature (PVT) may cause an inconsistent delay and increase in leakage to appear even in low power circuits, thus becoming one of the major challenges in nano scale devices. Therefore, to measure the actual performance of a CNFET compared to a MOSFET, it is necessary to compare performance at a circuit level under PVT variations. In this paper, logic gates and benchmark circuits are designed at 32nm for both CNFET and CMOS technologies; delay, power, power delay product (PDP), leakage current and frequency response are simulated and compared [26] [28].

2.2.1 Deciding ratio between PMOS and NMOS

For comparing the performance of CNFETs with MOSFETs at circuit level, the inverter as a fundamental logic gate is considered first; the inverter is designed with minimal width and a number of tubes in 32nm technology. For Si CMOS, a PMOS/NMOS ratio between 2 and 3 is used for compensating the difference in mobility between PMOS and NMOS. In this paper, a 3 to 1 (PMOS:NMOS) ratio is used when designing the inverter because the Voltage Transfer Characteristic (VTC) of the MOSFET inverter shows a more symmetrical shape in the center of the logic threshold voltage (VDD/2) for a ratio of 3 in 32nm technology (as
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

shown in Fig 2.2). However for the CNFET case, a 1 to 1 (pFET:nFET) ratio is used because the nFET and the pFET have almost the same current driving capability with same transistor geometry [26]. Fig 2.2 shows that the VTC of the CNFET also has a symmetrical shape at a 1 to 1 (pFET:nFET) ratio. Even though the current in a CNT is smaller than a minimum sized MOSFET (at 32nm technology), a CNFET has a steeper curve in the transition region due to the higher gain. This contributes to a 22.5% improvement in Noise Margin (NM), and this progressed performance is preserved under a decrease in power supply voltage as shown in Fig 2.3. In this paper, ratio value for the inverter is also utilized when designing more complex logic gates and benchmark circuits and used to determine the width and number of tubes of the CNFET.

To determine the PMOS/NMOS ratio for CMOS, the transistor width (W) of a MOSFET is modified; however in a CNFET, the number of CNTs is changed because a CNFET uses CNTs for the conducting channel between the source and drain. Therefore, when the width of the CNFET is changed, by implication the number of tube is also changed.

2.2.2 Logic gates performance analysis

In this section different metrics are utilized to compared CNFET and CMOS logic gates at 32nm features size.
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

Figure 2.2: Voltage Transfer Characteristic (VTC) for 32nm MOSFET and CNFET inverters at 0.9V supply.

Figure 2.3: Voltage Transfer Characteristic (VTC) of 32nm CNFET and MOSFET inverters at different power supplies.
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

**Power Delay Product (PDP)**

Due to the increased demand for high-speed, high-throughput computation, and complex functionality in mobile environments, reduction of delay and power consumption is very challenging. MOSFET and CNFET can be compared using the Power Delay Product (PDP) as metric. Table 2.1 shows the delay, power, and Power Delay Product (PDP) of logic gates in 32nm MOSFET and 32nm CNFET technologies; the PDP of the 32nm MOSFET is about 100 times higher than that of the 32nm CNFET.

**Leakage**

As process dimensions shrink further into the nanometer ranges, traditional methods for dynamic power reduction are becoming less effective due to the increased impact of static power [29]. In general, leakage power is different depending on the applied input vector. Fig 2.4 shows the maximum and minimum leakage power for 32nm MOSFET and CNFET-based logic gates. The maximum leakage power of the MOSFET-based gates is 75 times larger than for CNFET gates. The minimum leakage power of the MOSFET is about three times larger than for CNFET. Fig 2.4 also shows that the maximum leakage power shows a similar trend for both CNFET and MOSFET-based gates, while the minimum leakage power shows somewhat different trends, because the stack effect is reduced in CNFET circuits.
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

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<th>CNFET</th>
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<tr>
<td></td>
<td>Delay(sec)</td>
<td>Power(watt)</td>
</tr>
<tr>
<td>Inverter</td>
<td>1.77E-11</td>
<td>1.39E-06</td>
</tr>
<tr>
<td>NAND2</td>
<td>2.26E-11</td>
<td>1.96E-06</td>
</tr>
<tr>
<td>NAND3</td>
<td>2.99E-11</td>
<td>2.77E-06</td>
</tr>
<tr>
<td>NOR2</td>
<td>3.97E-11</td>
<td>2.58E-06</td>
</tr>
<tr>
<td>NOR3</td>
<td>6.97E-11</td>
<td>4.04E-06</td>
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<td>1.11E-07</td>
</tr>
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<td>1.89E-07</td>
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<td>NOR3</td>
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<td>2.73E-07</td>
</tr>
</tbody>
</table>

Table 2.1: Delay, Power, and PDP for 32nm MOSFET and 32nm CNFET logic gates.

**Frequency Response**

For establishing the frequency response, AC simulation has been performed for both the MOSFET and CNFET inverters. The results are given in Fig 2.5, the CNFET inverter shows nearly 3dB more voltage gain and 3 times higher 3dB frequency (f3dB) than the MOSFET inverter, thus confirming its superiority in terms of this metric.

**2.2.3 Logic gates PVT variations**

With technology scaling, the effects of systematic and random variations in process, supply voltage, and temperature (PVT) have led to inconsistent delay and leakage in low power circuits, thus becoming a major obstacle for device scaling. Technology scaling beyond 90 nm is affected by significant levels of process variations, and they are changing the design environment from a deterministic
to a probabilistic one. Moreover, the requirement of low power relies on supply voltage scaling, making voltage variations a significant challenge. The quest for increase in higher operating frequencies has resulted in significantly high junction temperature and within-die temperature variation [28][29]. Therefore, the possible performance degradation due to PVT variations has become a major criterion in assessing the performance of a new technology.

**Process Variations**

When investigating physical process variations, among them channel length and width are considered for a MOSFET transistor in CMOS. However, CMOS and CNFET have different characteristics as evidenced in Fig 2.6. The current change in a MOSFET is about ±30%(±13%) for a ±10% change in length (width) at a gate voltage of 0.9V while the current change in a CNFET is below ±0.5%. However when the diameter of the CNFET is changed by ±10%, the current change in a CNFET is about ±17% (as shown in Fig 2.7). Therefore for a CNFET, the diameter variation is more important because a CNFET is more sensitive to diameter variation than length and width variations. Based on this observation, the PDP and leakage of a CNFET are computed and shown in Fig 2.8 and Fig 2.9, respectively. When the diameter of a CNFET is changed, then the PDP changes too. Fig 2.9 shows that the maximum leakage power increases when the diameter is increased. Also note that the threshold voltage
and diameter of a CNFET are determined based on the chirality of the CNTs used in this type of transistor.

Figure 2.4: Maximum and minimum leakage power for 32nm MOSFET and CNFET logic gates.
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

Voltage Variation

The reduction in power consumption due to voltage scaling is also confronted with the increased sensitivity to voltage variations; this is a major concern to assess the performance of a new technology such as CNFETs. Fig 2.10 and 2.11 show the Power Delay Product (PDP) for 32nm MOSFET and CNFET logic gates, respectively when the supply voltage is decreased until the gate stops functioning. These figures show that the inverter and the other logic gates operate until the supply voltage decreases to 0.5V and 0.6V, respectively. Even though the PDP is changed depending on the supply voltage, this change is more or less the same as the change in PDP for the MOSFET logic gates; hence, the overall PDP of the CNFET-based gates is significantly lower than for

![Graph showing frequency response for 32nm MOSFET and CNFET inverters.]

Figure 2.5: Frequency response for 32nm MOSFET and CNFET inverters.
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

Figure 2.6: $I_{DS}$ vs. $V_{GS}$ curve with 10% change of gate length and width for the 32nm MOSFET and CNFET.
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

Figure 2.7: $I_{DS}$ vs. $V_{GS}$ with 10% change of carbon nanotube diameter (chirality) for the 32nm CNFET.

Figure 2.8: Power delay product (PDP) of 32nm CNFET logic gates vs. diameter (chirality) of carbon nanotube.
Figure 2.9: Maximum leakage power of the 32nm CNFET logic gates vs. diameter (chirality) of carbon nanotube.
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

the MOSFET-based gates. Fig 2.12 and Fig 2.13 plot the maximum leakage power for 32nm MOSFET and CNFET based gates, respectively versus the supply voltage. They show that the maximum leakage power of the MOSFET (CNFET) gates decreases exponentially (linearly), however the overall leakage power of the MOSFET gates is greater than that for the CNFET gates.

![Figure 2.10: Power delay product (PDP) of 32nm MOSFET logic gates vs. supply voltage.](image)

**Temperature Variation**

With an increase in circuit speed, a larger power consumption is often encountered, thus resulting in more heat at chip level. Circuits with an excessive power dissipation are more susceptible to run-time failures and account for serious reliability problems [28][29]. Fig 2.14 and 2.15 show that the PDP of the
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

Figure 2.11: Power delay product (PDP) of 32nm CNFET logic gates vs. supply voltage.

Figure 2.12: Maximum leakage power of 32nm MOSFET logic gates vs. supply voltage.
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

Figure 2.13: Maximum Leakage Power of 32nm CNFET logic gates vs. supply voltage.

Figure 2.14: Power delay product (PDP) of 32nm MOSFET logic gates vs. temperature.
2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

Figure 2.15: Power delay product (PDP) of 32nm CNFET logic gates versus temperature.

Figure 2.16: Maximum Leakage Power of 32nm MOSFET logic gates vs. temperature.
MOSFET gates increases with temperature; however, the PDP of the CNFET logic gates is constant. Moreover, the maximum leakage power of the MOSFET gates increases linearly with temperature, while for the CNFET-based gates this increase is exponential (as shown in Fig. 2.16 and 2.17, respectively).

### 2.2.4 Performance analysis of benchmark circuits

Few combinational circuits (such as a 4 stage inverter chain, a 2:4 decoder, a 4:16 decoder, the ISCAS-85 Benchmark Circuit C17, a 1-bit full adder, a 3-bit Ripple Carry Adder, and the ISCAS-85 Benchmark Circuit 74182) have also been evaluated. Table 2.2 shows the simulation results of circuits designed with 32nm CNFET and CMOS technologies. As shown in Table 2.2, the average

![Figure 2.17: Maximum Leakage Power of 32nm CNFET logic gates vs. temperature.](image-url)
## 2.2 Performance Analysis between CNFET and CMOS Under PVT Variation

<table>
<thead>
<tr>
<th></th>
<th>Delay(s)</th>
<th>Power(w)</th>
<th>PDP(joule)</th>
</tr>
</thead>
<tbody>
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<td></td>
<td></td>
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<tr>
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<tr>
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<tr>
<td>C17</td>
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<td>2.66E-16</td>
</tr>
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<tr>
<td>74182</td>
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<td>9.22E-06</td>
<td>5.90E-16</td>
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<td></td>
</tr>
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</tbody>
</table>

Table 2.2: Delay, Power, and PDP for 32nm MOSFET and CNFET circuits.

The delay and power consumption of the 32nm MOSFET circuits is about 10 times and 100 times higher than for the 32nm CNFET circuits respectively; these results confirm the findings found previously in Section B for the CNFET and MOSFET logic gates. Moreover, it shows that indeed CNFET-based designs offer significant improvements over MOSFET-based designs.
Chapter 3

Optimization Method for

Combinational CNFET Circuit

In this section the optimal combinational circuit design method for CNTFET is shown. In section 3.1, a brief review focusing on a CNTFET inclusive of device characteristics and physical features is provided. In section 3.2, a performance analysis is pursued for comparison with nano scale CMOS technology. Section 3.3 describes a novel circuit design methodology for CNTFET based circuits; this methodology considers channel capacitance and current variations to determine the best pitch, circuit speed and area. Simulation results are also presented to compare the results obtained by using the proposed methodology with the ones in which circuits were not optimized (such as for the case of either a too large or small screening effect due to an incorrect selection of the pitch).
3.1 Analysis of Technology Parameters

In this section, a design approach for CNTFET-based circuits is presented with emphasis on the differences with CMOS. For CMOS technology, the gate capacitance is proportional to WL (where W is the channel width and L is the channel length). In CNTFET technology, the gate capacitance depends on the number of tubes and the pitch (where the pitch is defined as the distance between the centers of two adjacent CNTs in the same device[30]). As the pitch decreases, the gate capacitance is also reduced due to the potential between adjacent CNTs (affecting the total gate capacitance). Moreover, the pitch also affects the current in the CNTFETs. Due to the screening effect [30], the total current of a CNTFET decreases as the pitch decreases. By analyzing the device characteristics of a CNTFET, performance metrics such as high speed, low power and low area overhead can be achieved when designing circuits using this technology. This aspect will be analyzed further in the next sections.

3.1.1 Channel and gate capacitance vs pitch

To calculate the coupling capacitance $C_{01}$ consider Fig. 3.1; the effects of two tubes around tube 1 can be lumped into the equivalent capacities of tube 2 and tube 3 (denoted as $C_{02}$ and $C_{03}$ respectively [13]). By applying the same voltage $V_1$ between tubes 1, 2 and 3 and the electrode, the charges $Q_1$, $\eta_1 Q_1$, and $\eta_2 Q_1$ are induced on these tubes due to the different coupling capacitances $C_{01}$, $C_{02}$,
3.1 Analysis of Technology Parameters

Figure 3.1: The CNTs in array and electrode.

and $C_{03}$. $C_{0i}$ is the equivalent coupling capacitance between the electrode and tube $i$. $\eta_1$ and $\eta_2$ are defined as the ratio of $C_{02}$ and $C_{03}$, respectively, over $C_{01}$ i.e. [13]:

$$\eta_1 = \frac{C_{02}}{C_{01}} \quad \eta_2 = \frac{C_{03}}{C_{02}}$$  \hspace{1cm} (3.1)

The charges on tube 2 and tube 3 affect the electric field and the electrostatic potential profile between the electrode and tube 1. The charge redistribution in the circumferential direction can be ignored for a one-dimensional device geometry. Using the superposition principle, the capacitance C01 can be expressed as [13]:

$$C_{01} = \frac{Q_1}{V_1} = \frac{Q_1}{V_0 + V_{adj}}$$  \hspace{1cm} (3.2)

$V_o$ and $V_{adj}$ are the potential differences between the electrode and tube 1 that are caused by the charges on tube 1 and the adjacent CNTs (i.e. tubes 2 and
3.1 Analysis of Technology Parameters

3.2 Figure 3.2: Three CNTs in parallel to the gate.

3, respectively), acting as independent electrodes. As the pitch decreases, \( V_{adj} \) increases. So, the capacitance \( C_{01} \) located in the middle of the CNTs decreases.

Fig 3.2 shows the gate capacitance of the CNTFET and its parameters. In this paper, \( h \) and \( H_{\text{gate}} \) are fixed to 4nm and 64nm, respectively, as default values of the HSPICE library [14]. \( C_{gc,e} \) denotes the capacitance of the CNT located at the boundary edge of the CNFET and \( C_{gc,m} \) denotes the capacitance of the CNT located in the middle of the CNTFET. The following capacitances are modeled: the gate to channel capacitance, the outer fringe capacitance, the gate to gate, or gate to source/drain coupling capacitance. By varying the pitch and the number of tubes, the channel capacitance changes significantly, while the other capacitances are barely affected. Fig 3.3 and Fig 3.4 show the total potential due to the adjacent CNT and its image charge as well as the capacitances \( C_{gc,e} \) and \( C_{gc,m} \) for different values of pitch, respectively. As mentioned previously, the pitch of a CNT affects the current (the current for different values of pitch and number of tubes is plotted in Fig 3.4). If there are more than \( N \) (\( N \geq 2 \)) tubes in the CNTFET, then the total channel capacitance
3.1 Analysis of Technology Parameters

Figure 3.3: The total potential due to the adjacent CNT and its image charge.

Figure 3.4: The gate to middle and edge CNT channel capacitances.
3.1 Analysis of Technology Parameters

<table>
<thead>
<tr>
<th>pitch (nm)</th>
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<th>4</th>
<th>6</th>
<th>8</th>
<th>10</th>
<th>12</th>
<th>14</th>
<th>16</th>
<th>18</th>
<th>20</th>
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<td>η</td>
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<td>0.59</td>
<td>0.73</td>
<td>0.82</td>
<td>0.88</td>
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<td>0.93</td>
<td>0.95</td>
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<td>0.96</td>
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</table>

Table 3.1: η versus pitch ($V_{DD} = 0.6V$, diameter = 2nm, channel length = 32nm).

is given by $(N - 2)C_{ge,m} + 2C_{ge,e}$; this can be expressed as $(\eta(N - 2) + 2)C_{ge,e}$ (in terms of $C_{ge,e}$). The total channel capacitance $C_{ch}$ is therefore given by;

\[
C_{ch} = \begin{cases} 
C_{ge,1} & : N = 1 \\
2C_{ge,e} & : N = 2 \\
(\eta(N - 2) + 2)C_{ge,e} & : N > 2 
\end{cases} \tag{3.3}
\]

\[
C_{gc,tot,CNT} \approx N \cdot C_{gc,CNT} \cdot L_g = \{\eta(N - 2) + 2\} \cdot C_{ge,e,CNT} \tag{3.4}
\]

where $N$ is the number of tubes, $C_{ge,1}$is the channel capacitance of the CNT-FET with only one tube, and $\eta$ is the ratio between $C_{ge,e}$ and $C_{gc,m}$, i.e. $\eta = (C_{ge,e}/C_{gc,m})$. $\eta$ is determined based on the pitch, supply voltage, channel length, and the diameter of the CNT. $\eta$ is shown in Table 3.1 (found based on Hspice simulation). The best (optimal) number of tubes and pitch can be found by considering the different parameters inclusive of delay, power dissipation and area overhead.

The gate capacitance affects the speed and power consumption, and is determined based on device size. In a CNTFET, the device size is proportional to the gate width (that is a function of the number of tubes) and the pitch.
The CNTFET gate capacitance ($C_{\text{gate,CNT}}$) consists of three components [13]: the gate to channel capacitance ($C_{g\text{c, tot,CNT}}$), the gate outer fringe capacitance ($C_{fr, tot,CNT}$), and the coupling capacitance between the gate and the adjacent contacts ($C_{gtg, tot,CNT}$). These components are approximated as follows:

$$C_{fr, tot,CNT} \approx C_{fr, CNT} \cdot L_g$$ (3.5)

$$C_{gtg, tot,CNT} = C_{gtg} \cdot W_g$$ (3.6)

$$C_{\text{gate,CNT}} \approx \{\eta(N - 2) + 2\} \cdot C_{g\text{c, x,CNT}} + C_{gtg} \cdot W_g$$ (3.7)

### 3.1.2 Size of CNFET

The total size of the CNFET is determined by the width of the gate (as for the device structure of Fig 2.1). The gate width can be determined by the pitch. By setting the minimum gate width $W_{\text{min}}$ and the number of tubes $N$, the gate width can be approximated as

$$W_g = \text{Max}(W_{\text{min}}, N \cdot \text{pitch})$$ (3.8)

### 3.1.3 CNFET threshold voltage

The drain current of the CNTFET is dependent on pitch value, which determine the amount of screening effect. Therefore the current is not linearly proportional
3.1 Analysis of Technology Parameters

Figure 3.5: Drain current with different values of pitch and number of tubes.

to the number of carbon nanotube. The figure 3.5 shows drain current with different values of pitch and number of tubes. To design a circuit with best performance based on an average power consumption and speed, it is very important to determine the threshold voltage because this affects the switching speed, the current and leakage power. A single wall carbon nanotube (SWCNT) with charities \((n_1, n_2)\), can be metallic if \(n_1 - n_2\) is an integer multiple of 3 \((mod(n_1 - n_{2,3}) = 0)\), else, it is semiconductor [14]. For a SWCNT with charities \((n_1, n_2)\), the diameter \((DCNT)\) is given by [14]

\[
D_{CNT} = a\sqrt{\frac{n_1^2 + n_1n_2 + n_2^2}{\pi}}
\]  

(3.9)
3.1 Analysis of Technology Parameters

<table>
<thead>
<tr>
<th>Diameter</th>
<th>V\text{th}\text{(V)}</th>
<th>Delay(sec)</th>
<th>Power(W)</th>
<th>PDP(J)</th>
</tr>
</thead>
<tbody>
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<tr>
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<td>5.79E-06</td>
<td>3.12E-17</td>
</tr>
</tbody>
</table>

Table 3.2: Delay, Power and PDP of the FO4 inverter for various V\text{th}.

where \( a = 2.49\text{Å} \) is the lattice constant. Since the bandgap of semiconducting CNTs is proportional to the diameter, then the threshold voltage of the intrinsic CNT channel can be approximated to the first order as the half bandgap (which is inversely proportional to the diameter). By adjusting the diameter, the threshold voltage can be controlled and is given by [14]

\[
V_{\text{th}} \approx \frac{E_g}{2e} = \frac{\sqrt{3}}{3} \frac{aV_\pi}{eD_{\text{CNT}}}
\]  

(3.10)

where \( V_\pi (3.033\text{eV}) \) is the carbon \( \pi - \pi \) bond energy in the tight bonding model [18] and \( e \) is the unit electron charge.

For the fan-out-of-4 (FO4) inverter, Table 3.2 shows the average power consumption, the propagation delay, and the PDP (Power Delay Product) for different threshold voltages. When \( V_{\text{th}} = 0.2\text{V} \), the smallest value of PDP is attained; therefore in this paper, 0.2V is chosen for \( V_{\text{th}} \).
3.2 Performance Optimization Methods

3.2.1 Optimization for CNFET-based circuits

The pitch, diameter and number of CNTs, determine the gate capacitance and drain current, threshold voltage. In this section, the optimization methods will be proposed with given technology parameters. Fig 3.6 shows the optimization flow overview.

Determination of optimum $V_{DD}$

The energy-delay product (EDP) defined as

$$EDP = PDP \times t_p = \frac{C_L V_{DD}^2}{2} r_p$$  \hspace{1cm} (3.11)
3.2 Performance Optimization Methods

The $\tau_p$ can be approximated as below [19]

\[
\tau_p = \frac{C_{gg,CNFET}V_{DD}}{I_{CNFET}}
= \eta_{CNT} \cdot \frac{C_{gg,CNFET}L_gV_{DD}}{g_{CNT}(V_{DD} - V_{th,CNT} - V_{DSAT}/2)}
\] (3.12)

Where $I_{CNFET}$ is given by [14]

\[
I_{CNFET} = \frac{n \cdot g_{CNT}(V_{DD} - V_{th,CNT} - V_{DSAT}/2)}{1 + g_{CNT}L_s\rho_s}
\] (3.13)

Where $g_{CNT}$ is the transconductance per CNT, and $L_s$ is the source length (doped CNT region), $\rho_s$ is the source resistance per unit length of doped CNT, $\eta_{CNT}$ is the pre-defined technology factor, $C_{gg,CNFET}$ is gate capacitance. $V_{DSAT}$ is the voltage which start to saturate the drain. Put $\tau_p$ in 3.12 into 3.11, the EDP as follow

\[
EDP = PDP \times t_p = \frac{\alpha C_L^2 V_{DD}^3}{2(V_{DD} - V_{th} - (V_{DSAT}/2)^2)}
\] (3.14)

Taking the derivative equation 3.14 with respect to $V_{DD}$ and equating the result to 0. The result is

\[
V_{DD} = \frac{3}{2}(V_{th} + V_{DSAT}/2)
\] (3.15)

In order to find $V_{DSAT}$, drain current equation of CNT will be used. The drain current equation is derived from the Landauer formula [18], which describes ballistic transport with ideal contacts. Its expression represents the sum of the
3.2 Performance Optimization Methods

energy sub-band contributions of two terms where $\Delta p$ is the minima of the $p$th energy sub-band, $e$ is the electron charge, $k_B$ the Boltzman constant, $h$ the Planck constant, and $T$ the temperature.

$$I_D = \frac{4ek_BT}{h} \sum_{p=1}^{\infty} \left[ \ln \left( 1 + \exp \left( -\frac{\Delta_p + V_{CNT}}{k_B T} \right) \right) - \ln \left( 1 + \exp \left( -\frac{V_{DS} - \Delta_p + V_{CNT}}{k_B T} \right) \right) \right]$$

(3.16)

Since it is not possible to obtain an analytical closed-form expression to calculate the number of carriers in the channel, $\eta_{CNT}$, the following empirical relationship has been proposed in [18].

$$V_{CNT} = V_{GS} \quad \text{for} \quad V_{GS} < \Delta_1$$

$$= V_{GS} - \alpha(V_{GS} - \Delta_1) \quad \text{for} \quad V_{GS} \geq \Delta_1 \quad (3.17)$$

Where $\Delta p$ is the energy level for the first sub-band, and

$$V_{CNT} = V_{GS} \quad \text{for} \quad V_{GS} < \Delta_1$$

$$= V_{GS} - \alpha(V_{GS} - \Delta_1) \quad \text{for} \quad V_{GS} \geq \Delta_1 \quad (3.18)$$

$\alpha 0, 1, 2$ are fitting parameters depending on the gate-oxide capacitance, CNT diameter. Taking the derivative equation 3.14 with respect to $V_{DS}$ equating the result to approximately 0 , the $V_{DS}$ which saturate the drain current can be found. With given technology parameters in this work, this $V_{DSAT}$ is almost 0.4
3.2 Performance Optimization Methods

V. Therefore, the optimal $V_{DD}$ in this work which obtained from equation 3.15 is about 0.6V.

**Fan-out factor for best delay**

The process of sizing up an inverter reduces the delay; however, it also increases its input capacitance. For determining the input loading effect, the relationship between the input gate capacitance $C_g$ and the intrinsic output capacitance $C_{int}$ of the inverter must be established. Both are proportional to gate sizing; the following well known relationship holds in this case [15]

$$C_{int} = \gamma C_g \quad (3.19)$$

where $\gamma$ is a proportionality factor that is only a function of technology and is determined by the pitch and the gate width, (that also determines the intrinsic capacitance). Its optimum value can be found by differentiating the minimum delay expression [15] by the number of stages and setting the result to 0, i.e.

$$\gamma + \sqrt[3]{F} - \frac{\sqrt[3]{F} \ln F}{N} = \gamma + f - f \ln f = 0 \quad (3.20)$$

where $F$ denotes the overall effective fan-out of the circuit and $f$ denotes the effective fan-out factor. For CNTFETs, the value of $\gamma$ can be found by HSPICE through simulation (and it is reported in Table 3.3). It is a function of the pitch.
3.2 Performance Optimization Methods

<table>
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<tr>
<td></td>
<td>1.31</td>
<td>1.38</td>
</tr>
</tbody>
</table>

Table 3.3: γ Values for 4nm, 10nm pitches with different number of tubes.

and the number of tubes.

Using the simulated value of γ, the best (optimum) fan-out factor f is given by 4; note that this value is different from the optimum fan-out of CMOS circuits.

**Optimum number of CNTs for minimum delay**

For driving a load in CMOS technology, the ratio between the load and drivers gate capacitances is often used. In CMOS, this ratio is almost same as the ratio \( (W_{\text{load}}/L_{\text{load}})/(W_{\text{driver}}/L_{\text{driver}}) \). However for CNTFET, as shown in a previous section, the gate capacitance is not proportional to the number of carbon nanotubes. This is caused by the screening effect that affects the gate capacitance and the current. Therefore, it is important to choose a suitable number of tubes in each driver and load. As an example, a buffer chain is considered; in this circuit, the ratio between the gate capacitance of the \( i-th \) inverter and the next stage is

\[
\{\eta(N_i - 2) + 2\}C_{\text{gc,R}} \approx f\{\eta(N_{i-1} - 2) + 2\}C_{\text{gc,R}}
\]

(3.21)
3.2 Performance Optimization Methods

<table>
<thead>
<tr>
<th>CMOS(32nm)</th>
<th>Number of inputs</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>n</th>
</tr>
</thead>
<tbody>
<tr>
<td>Logical effort</td>
<td>inverter</td>
<td>1</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>NAND</td>
<td>5/4</td>
<td>6/4</td>
<td>7/4</td>
<td>(n+3)/4</td>
<td></td>
</tr>
<tr>
<td>NOR</td>
<td>7/4</td>
<td>10/4</td>
<td>13/4</td>
<td>(3n+1)/2</td>
<td></td>
</tr>
</tbody>
</table>

Table 3.4: Logical Effort for CNTFET.

Solving 3.11 for \( N_i \)

\[
N_i = f \cdot N_{i-1} + 2 \left( 1 - f + \frac{f}{\eta} - \frac{1}{\eta} \right)
\]

\[
= f \cdot N_{i-1} + K(f, \eta)
\]  

(3.22)

where \( K(f, \eta) = 2 \left( 1 - f + \frac{f}{\eta} - \frac{1}{\eta} \right) \) \( N_i \) denotes the number of carbon nanotubes of the \( i \) th stage, and \( f \) is the effective fan-out factor. \( K(f, \eta) \) is a function of \( f \) and \( \eta \); the values are determined by the pitch and the number of tubes in the CNTFET. Based on \( \gamma \) as reported in Table 3.3, in most cases the optimum value of \( f \) is 4 and the value of \( \eta \) can be found from Table 3.1.

Logical effort for combinational logic

The so-called logical effort of a gate denotes the additional input capacitance of a gate to deliver the same output current as an inverter. The logical effort is an important figure of merit, because it can be used to reduce the path delay of a circuit [16].

In a CNTFET, the P-type and N-type have almost the same carrier mobility.
3.2 Performance Optimization Methods

<table>
<thead>
<tr>
<th>CMOS(32nm)</th>
<th>Number of inputs</th>
<th>Logical effort</th>
</tr>
</thead>
<tbody>
<tr>
<td>inverter</td>
<td>1</td>
<td>2 3 4 n</td>
</tr>
<tr>
<td>NAND</td>
<td>5/4 6/4 7/4 (n+3)/4</td>
<td></td>
</tr>
<tr>
<td>NOR</td>
<td>7/4 10/4 13/4 (3n+1)/2</td>
<td></td>
</tr>
</tbody>
</table>

Table 3.5: Logical Effort for 32nm CMOS.

![Example circuit for optimization method.](image)

 unlike CMOS technology (which is 3:1). Therefore, the sizes of the sizes of the P-type and N-type CNTFETs can be the same in an inverter. Tables 3.4 and 3.5 show the logical efforts for CNTFET and 32nm CMOS gates. Based on the CNTFET logical effort, different measures (such as the path logical effort, the path electrical effort, the path effort, the path branching effort, the path delay and the path effort delay) can be found [16]. As an example, consider the circuit in Fig. 3.7; the logical effort can be used to reduce the delay of this CNTFET based circuit. The path logical effort is given by $G = (1) \cdot (4/2) \cdot (3/2) \cdot (1) = 12/4$. The path electrical effort is $H = 5$, the path branching effort is $B = 3 \times 2 = 6$, the path effort is $F = GBH = 152$ and $f = \sqrt[3]{F} = 5.3$.

As the parasitic delay is $P = 2 + 3 + 2 = 7$, then the minimum path delay is $D = 3 \cdot 5.3 + 7 = 23$. The gate capacitance are computed as $y = 45 \cdot (3/2)/5.3 = 12.7$ and $x = (12.7 + 12.7) \cdot 4/2/5.3 = 9.6$.
3.2 Performance Optimization Methods

3.2.2 Simulations results for combinational circuits

In this paper simulation utilizes a 32nm CNFET HSPICE model that includes non-idealities of the CNTFET [22]. Simulation was performed to verify that the propose design methodology for finding the optimum pitch and number of tubes for best gate capacitance and logical effort. The technology parameters for the CNTFETs are as follows: Physical channel length = 32.0nm, Length of doped CNT drain/source extended region = 32.0nm, Fermi level of the doped S/D tube = 0.6 eV, Thickness of high-k top gate dielectric material = 4.0nm, chirality of tube = (17,2), Vfbe, Vfbp =0 (Flatband voltage for n-CNFET and p-CNFET), Physical gate length = 32.0nm, Mean free path of intrinsic CNT = 200.0nm, Length of doped CNT source/drain extension region = 32.0nm, Mean free path in p+/n+doped CNT = 15.0nm, Work function of Source/Drain metal contact = 4.6eV, and CNT work function = 4.5eV.

Delay

Fig 3.8 shows the inverter delay (for channel length =32nm, $V_{DD}$ = 0.6, $V_{th}$ = 0.2, number of tubes =4) versus pitch. As the pitch changes, the speed of the circuit also changes due to the gate capacitance and current. Fig 3.9 shows the delay of the inverter chain for various values of fan-out factor. The result shows that the minimum (least) delay is obtained when the fan-out factor is 4 for both cases of 4nm and 20nm pitch (as not affected by the screening effect, $\eta$=1). This
Figure 3.8: Delay of inverter (channel length =32nm, VDD = 0.6, Vth = 0.2, number of tubes =4) for different pitches.
3.2 Performance Optimization Methods

The simulation result is the same as the result of section 3.3.3. To verify the proposed design methodology to find the number of tubes for each logic stage, an inverter chain is designed and its delay has been simulated. There are 6 different cases with different numbers of tubes for each stage. Each CNTFET in the inverters has the same pitch (4nm) and fan-out factor ($f = 4$) except for the fourth stage (i.e., the load). The load has 64 tubes and 20nm pitch. The relation between the gate load and the number of tubes in the output gate is not linear (as indicated in 3.8). Different numbers of tubes have been selected to verify 3.12. The number of tubes is chosen near the theoretical value (obtained from 3.12). For example, if the number of tubes in the first stage is 1, then the number of tubes in the

![Graph showing inverter chain delay for different fan-factor and pitch.](image-url)

Figure 3.9: Inverter chain delay for different fan-factor and pitch.
Theoretically, the number of tubes for the second stage must be 8. However, since an approximation was used when deriving 3.12, the exact number of tubes for the second stages can be different from 8. Simulation was performed for different numbers of tubes near the value of 8 for the second stage to find the optimum delay. The optimum number of tubes is found in the same way for the third stage. The number of tubes in each stage is denoted as $(1^{st}, 2^{nd}, 3^{rd}, 4^{th})$.

From case 1 to case 6 the numbers of tubes are given by (1,4,16,64), (1,6,24,64), (1,7,32,64), (1,8,36,64), (1,9,40,64), (1,10,44,64). For Case 1, the screening effect is not considered. For all cases, the first stage inverter has only one tube (corresponding to a minimum sized inverter), the number of tube for the second
3.2 Performance Optimization Methods

stage (i.e. (4,6,7,8,9,10)) was chosen near the theoretical value (i.e. 8), and the number of tubes for the third and the fourth inverters are determined based on 3.12. The delay of the inverter chain is shown in Fig 3.10; these results shows that either case 3 or case 4 provides the best delay and this value is close to the value obtained from 3.12. Therefore, simulation has shown that the proposed methodology is effective in finding the optimum number of CNTs for minimum delay. Fig 3.11 shows the results for the logical effort of the circuit example of Fig 3.7. Two cases are considered. The first case is the non-ideal one with the screening effect (pitch 4nm, $\eta = 0.6$). The other corresponds to the ideal case in which there is no screening effect (pitch 20nm, $\eta = 1$). In both cases, the numbers of tubes is calculated from 3.12. From these results, a 48% difference between ideal and non-ideal cases is found, thus showing that the minimum delay can be achieved by utilizing the optimum number of tubes and the logical effort as in the proposed methodology.

**Power consumption**

Both the gate capacitance and the current are reduced due to the screening effect (at a small pitch).

Both these reductions cause to dissipate a smaller amount of power. Fig 3.12 and 3.13 show the power consumption of the inverter with channel length $=32$nm, $V_{DD} = 0.6$, $V_{th} = 0.2$ number of tubes $=4$. As expected, the small
3.2 Performance Optimization Methods

Figure 3.11: Delay of the circuit example for logical effort in Figure 7 ($\eta=0.6$ and 4nm pitch, $\eta=1$, 20nm pitch for ideal CNTFET case with no screening effect).

Figure 3.12: Power consumption of inverter (channel length =32nm, $V_{DD} = 0.6$, $V_{th} = 0.2$ number of tubes =4) versus pitch.)
3.2 Performance Optimization Methods

Figure 3.13: Power consumption of the test circuit in Figure 23.

pitch consumes less power.

Performance Analysis of Benchmark Circuits

Few combinational circuits (such as a 4 stage inverter chain, a 2:4 decoder, a 4:16 decoder, the ISCAS-85 Benchmark Circuit C17, a 1-bit full adder, a 3-bit Ripple Carry Adder, and the ISCAS-85 Benchmark Circuit 74182) have also been evaluated. Table 3.6 shows the simulation results of circuits designed with 32nm CNTFET and CMOS technologies. As shown in Table 3.6, the average delay and power consumption of the 32nm MOSFET circuits is about 10 times and 100 times higher than for the non-optimized 32nm CNTFET circuits respectively. The optimized circuit show 1.8 2.5 faster than non-optimized at the point of delay and power, which result what the PDP is 211~321% less.
### Table 3.6: Delay, Power and PDP for 32nm MOSFET and CNTFET circuits.

#### 3.3 Summary

The Carbon Nanotube Field Effect Transistor (CNTFET) is one of the most promising devices among emerging technologies to extend and/or complement the traditional Si MOSFET. The quantitative results of this paper have confirmed that the CNTFET technology is a viable solution to replace the conventional bulk MOSFET technology and, in particular, the paper makes it possible to quantitatively estimate the delay, leakage current, and power of CNTFET-based gates. This has been accomplished by proposing a new optimization methodology for CNTFETs. As the characteristics of a CNTFET is different from conventional bulk CMOS, new criteria must be established. By an appropriate selection of the diameter from the chirality, threshold voltages can be determined. As the channel capacitance and current vary as pitch is changed due to the screening effect, the gate capacitance has been established as function
of the number of tubes in the device and the optimum fan-out factor has been found. Using these parameter, the logical effort has been calculated and the minimum delay of a multistage circuit topology has been analyzed. To prove the effectiveness of the proposed gate-level design method, simulation has been performed using HSPICE with the CNTFET library of [22], Results have demonstrated that the proposed design methodology is both effective and practical. To design a CNFET circuit, many parameters must be considered, among them the diameter at certain chirality, pitch and the optimum number of tubes have been shown to be of primary importance.
Chapter 4

Low Power SRAM Design using CNFET

4.1 Introduction

In this Chapter, a new SRAM cell design based on Carbon Nanotube Field-Effect Transistor (CNFET) technology is proposed. The proposed SRAM cell design for CNFET is compared with SRAM cell designs implemented with the conventional CMOS and FinFET in terms of speed, power consumption, stability, and leakage current. The HSPICE simulation and analysis show that the dynamic power consumption of the proposed 8T CNFET SRAM cell’s is reduced about 48% and the SNM is widened up to 56% compared to the conventional CMOS SRAM structure at the expense of 2% leakage power and 3% write delay.
4.1 Introduction

An estimated 70% of the transistors in a billion-transistor superscalar microprocessor are expected to be used for memory arrays, especially for large L2 and L3 SRAM data caches. Therefore, it is essential to develop a low power SRAM design technique for the new device technology such as CNFET.

In this paper, as a circuit level design of CNFET, a novel low power 8T SRAM cell design is proposed and its performance and viability are demonstrated by performing various simulations. The performance, power consumption, stability, and leakage currents of the 6T and 8T SRAM cell based on CNFET are compared with those of the conventional CMOS and FinFET based SRAM cell designs to show the viability of the CNFET based SRAM cell design.

In section II, the characteristics and physical features of CNFET are explained, and the mechanisms of the read and write operations of the proposed 8T CNFET SRAM cell are explained and the schemes for deciding the number of nanotubes of each CNFET are described in section III. In section IV, the simulation results are presented to compare the performance and viability of the CNFET technology with those of other technologies, followed by the conclusion in Section V.
4.2 Low power 8T SRAM Cell

4.2.1 Write/Read operations

In the proposed 8T SRAM, the write and read bits are separated. While bit and bit-bar lines are used for writing data in the traditional 6T SRAM, only the WRITE_BIT in Figure 2 is used in the proposed SRAM cell to write for both "0" and "1" data. The writing operation starts by disconnecting the feedback loop of the two inverters. By setting 'W_bar' signal to "0", the feedback loop is disconnected. The data that is going to be written is determined by the WRITE_BIT voltage. If the feedback connection is disconnected, SRAM cell has just two cascaded inverters. WRITE_BIT transfers the complementary of the input data to Q2, cell data, which drives the other inverter (P2 and N2) to develop Q_bar. WRITE_BIT have to be pre-charged "high" before and right after each write operation. When writing "0" data, negligible writing power is consumed because there is no discharging activity at WRITE_BIT. To write '1' data at Q2,

The WRITE_BIT have to be discharged to ground level, just like 6T SRAM cell. In this case, the dynamic power consumed by the discharging is the same as 6T SRAM. The write circuit does not discharge for every write operation but discharges only when the cell writes "1" data, and the activity factor of the discharging WRITE_BIT is less than "1", which makes the proposed SRAM cell
4.2 Low power 8T SRAM Cell

more power effective during writing operation compared with the conventional ones.

All the Read_Bit lines are pre-charged before any READ operation. During read operation, transistor N5 is turned on by setting W_bar signal high and the READ_ROW(RD) is "high" to turn on N6. When Q2="0", the N4 is off making the READ_BIT voltage not change from the pre-charged value, which means the cell data Q2 holds "0". On the other hand, If Q2= "1", the transistors N4 and N6 are turned on. In this case, due to charge sharing, the READ_BIT voltage will be dropped about 100 200mV, which is enough to be detected in the sense amplifier.

4.2.2 Carbon nanotube configuration

The operation of writing "1" is stable because NMOS transistor N3 can pass "0" faithfully. On the other hand, when writing "0", WRITE_BIT is pre-charged high (VDD) and N5 is turned off. The node voltage at Q1 is less than VDD due to the threshold voltage drop between the gate and source of the transistor N3. To compensate this voltage drop, the transistor N2 and P2 must be designed as a low-skew inverter to ensure Q2 to be solid ground level to represent "0" state. A low-skewed inverter has a weaker PMOS transistor. If the PMOS CNFET have only one tube, the current is minimum. Let’s suppose that the cell stores "0" at Q2 and "1" at Q_bar after WL(Word Line) is deactivated and
W_bar is activated. In this case, the voltage at Q1 is less than VDD due to the threshold voltage drop across the gate and source of the transistor N5. The degraded voltage at Q1 may turn on the transistor P2 slightly causing short circuit current through transistors P2 and N2. To overcome this problem, the low skewed inverter (N2 and P2) mentioned for writing "0" case is justified again and the Vth of the transistor N5 needs to be controlled low to reduce the voltage difference between Q_bar and Q1. To implement a low skewed inverter with transistors N2 and P2, transistor ratio of N2 to P2 should be at least 2 to have a solid ground level at Q2.

![Figure 4.1: The 8T SRAM CNFET structure.](image)

However, by increasing the number of tubes, the P2 and N2 area sizes can be same. That is, if P2 has only one tube and N2 has 2 tubes, then the current
ratio N2/P2 can be more than 2. This means that the inverter transistor sizes N2/P2 can be smaller than 2 by controlling the number of tubes. Transistor ratio N3/P2 of 1.3, N1/P2 of 3, and low Vth of the transistor N5 guarantees a stable READ operation when Q_bar stores ”0”. However, if the similar approach to N2/P2 sizing is used to optimize transistor ratios among N1, N3 and P2, the transistor sizes can be further reduced. If N3 has only one tube, N1 has two tubes and P2 has one tube, the transistor N1 needs to be only 1.5 times larger than transistor P2 to satisfy the relationships among N1, N3, and P2. Combining the threshold voltage controllability F of the CNFET varying the diameter of tubes and transistor sizing techniques, the proposed 8T SRAM cell can accomplish low power consumption due to smaller node capacitance and tuning Vth at the minimal cost of the area overhead.

4.3 SIMULATION RESULTS

In order to compare performances, total 8 different SRAM cells are designed; Each 6T SRAM and 8T SRAM cells are designed using tied FinFETS (front and back gates of the FinFETS are tied together), independent double gates Fin-FETS (front and back gates are independently controlled), CMOS and CNFET. The 6T independent gate FinFET SRAM (6T-Ind) is implemented by using independent gate control which connects the back gates of the NMOS (PMOS) transistors to GND (VDD) to reduce the leakage current. An independent-gate
4.3 SIMULATION RESULTS

FinFET operates in the dual-gate mode (DGM) when both gates are biased to induce channel inversion. Alternatively, an independent-gate n-FinFET (p-FinFET) operates in the single-gate mode when one of the gates is deactivated by connecting the gate to ground (VDD). Disabling one of the gates in the single-gate mode (SGM) increases the absolute value of the threshold voltage compared to DGM. Therefore, it is possible to modulate the threshold voltage of the FinFET by biasing the two gates independently [36]. And the proposed 8T SRAM (8T-Ind) configuration is that back gate of PMOS connected to the VDD and front and back gates of N1, N2, N6 are tied together and back gate of N3, N4 are connected together.

4.3.1 Simulation setup

The technology parameters for the FinFETs are; channel length (L) = 32nm, fin height (Hfin) = 32nm, fin thickness (tsi) = 8nm, oxide thickness (tox) = 1.6nm, channel doping = 2 x 1020 cm-3, source/drain doping = 2 x 1020 cm-3, work functions (N-FinFET) = 4.5eV, work functions (P-FinFET) = 4.9eV. The technology parameters for the CNFETs are: physical channel length = 32.0nm, 10nm (this value used in simulations for performances with VDD variations), the length of doped CNT drain-side/source-side extension region = 32.0nm, fermi level of the doped S/D tube. = 0.6 eV, the thickness of high-k top gate dielectric material = 4.0nm, chirality of tube = (19,0), Pitch = 10nm, Vfbn,
4.3 SIMULATION RESULTS

$V_{fbp}$ (flatband voltage for n-CNFET and p-CNFET) = 0.0eV, 0.0eV, physical gate length = 32.0nm, The mean free path in intrinsic CNT = 200.0nm, the length of doped CNT source/drain extension region. = 32.0nm, the mean free path in p+/n+ doped CNT = 15.0nm, the work function of Source/Drain metal contact = 4.6eV, CNT work function = 4.5eV. The minimum transistor sizes used for those technologies are $W=48\text{nm}$ and $L = 32\text{nm}$ for bulk CMOS, $H_{\text{fin}}=32\text{nm}$ and $L = 32\text{nm}$ for FinFET, and $L=32\text{nm}$ and the number of tubes =1 for CNFET. The HSPICE using the Predictive Technology Model (PTM) model and Stanford University CNFET model is used to simulate the performance of the proposed 8T SRAM and the conventional 6T SRAM cells designed with CMOS, FinFET, and CNFET transistors.

4.3.2 Dynamic power consumption

The proposed 8T SRAM achieves 48% writing power saving while maintaining the cell performance, read/write delay, and stability of the conventional cell. The power saving comes from the fact that the cell keeps WRITE_BIT ”high” instead of discharging when it writes ”0”, which reduces the activity factor of the WRITE_BIT.

While conventional 6T SRAM always discharges one of the bit lines to write a data into the cell, the proposed 8T SRAM discharge the WRITE_BIT only when it writes ”1”. As the probability of writing ’0’ gets higher, the power dissipation
due to the bit line discharging is reduced comparing to the conventional case. CNFET shows about 5 times less power consumption compared to CMOS.

**Figure 4.2: The Dynamic Power consumption with VDD variation**

Fig 4.2 shows the dynamic power consumption of the CNFET 8T SRAM cell for different VDD. As shown in the Fig. 3, the power saving of the 8T SRAM on CNFET becomes greater as VDD increases since the dynamic power difference between the 6T SRAM and the proposed 8T SRAM increases exponentially as VDD increases.

### 4.3.3 Leakage power consumption

Fig 4.3 show the leakage power of the 6T and 8T CNFET SRAM cell. The proposed 8T SRAM cell design shows slightly higher leakage power because it has one more leakage current path than the conventional SRAM cell. The
4.3 SIMULATION RESULTS

READ_BIT, N4, N6 constitutes an additional leakage current path. However, N4, N6 has a stack effect that reduces the sub-threshold leakage current a bit. As a result, the leakage current through the READ_BIT, N4, and N6 path is relatively small. The difference of the leakage current in all of the four configurations is less than 2%.

![Leakage Power](image)

Figure 4.3: The Leakage Power consumption with VDD variation

4.3.4 Cell stability

Fig. 4.4 shows the Static Noise Margin (SNM) difference between the conventional 6T SRAM and the proposed 8T SRAM. Static Noise Margin is the standard metric to measure the stability in SRAM bit-cells[37]. The static noise margin of SRAM cell is defined as the minimum DC noise voltage necessary to flip the state of the cell. The voltage transfer curves (VTCs) of the back-to-back
inverters in a bit-cell are used to measure SNM [37]. Separating the Read and Write bit offers wider SNM during read operation as shown in Figure 4. When reading the stored data, only READ_BIT affects inv1 (N1/P1) output voltage. Consequently, this fact makes the cell hard to flip. And Table 1 shows CNFET have the highest SNM because of the relatively higher Vth and lower leakage current than CMOS and FinFET based SRAM cells.

![SNM Graph](image)

**Figure 4.4:** SNM for 6T and 8T CNFET cell with VDD variation

The READ access time at the cell level is determined by the time taken for the bitlines to develop a potential difference of at least 100mV. The read time depends on the READ path’s transistors’ sizes. The proposed 8T SRAM cell’s READ delay is almost same as the conventional cells since the transistor sizes are very similar. For write operation, the write delay is defined as the time from the 50% activation of the WL to the time when Q_bar becomes 90% of
4.4 Summary

The new SRAM cell cuts off the feedback connection between the two back-to-back inverters in the SRAM cell when data is written and separates the write and read port with 8 transistors. The proposed technique saves dynamic power by reducing discharging frequency during write operation. Compared to 6T SRAM structure, the proposed 8T SRAM saves power up to 48% and obtains 56% wider SNM during read operation at the minimal cost of 2% leakage power and 3% delay increase. As the cells are more frequently accessed, the dynamic power saving is linearly increased. This paper also compares CMOS, FinFET and CNT 6T and 8T SRAM cells using HSPICE simulations. The result demonstrates from 3 to 7 times less dynamic power consumption, from 11 to 17 times less
leakage power consumption, from 5 to 6 times faster read and write operations, and 1.6 wider SNM than the conventional designs.
Chapter 5

High Speed and Low Power

Transceiver Design for CNFET

and CNT Interconnect

5.1 Introduction

On-chip communication has become an active research area in the past few years. This not only because on-chip interconnects are becoming a speed, power, and reliability bottleneck [38], but also because systems on chips (SoCs) start to become so complex that they require new interconnection approaches [39], [40]. Therefore, high-speed, low power wire-line communication has been required
5.1 Introduction

by the modern computer systems, and the demand for more sophisticated input/output (I/O) transceiver design has been increased accordingly.

However, the environment of the communication channel and the semiconductor technology (transistor size, material and wire material) limit the maximum symbol rate of the serial data transmission and power is consumed by data transmission. Therefore, many researches have been done to save power and increase the data rate while keeping the same symbol rate and sustaining the signal integrity. Among them, increasing the data rate methods are the pulse-amplitude modulation (PAM) [41], pulse-width modulation (PWM) [42] and using carbon nano tube (CNT) bundle [45] by reducing wire capacitance.

[40] introduced a transceiver that uses both PWM and PAM modulation for the first time to boost the data transmission rate. However, its application is limited to a relatively low-speed data transmission (250M symbol/s) and it utilizes the conventional PWM and PAM symbol representations. It also needs complex circuit components for PWM, PAM modulation and demodulation. Due to the switched capacitances of interconnect, on-chip wires also present an increasing energy problem. A CMOS wire driver running at an effective frequency must switch a total wire capacitance $C_w$ through the voltage, leading to a power cost proportional to $CV^2f$. Under technology scaling $C_w$, remains largely constant, voltage scales down slowly, and frequency (f) scales up while leading to nearly constant power per wire. However, as chip’s device integration level increases
continuously, this constant power per wire gets multiplied by an ever-increasing number of wires. In order to save power consumption which is caused by wires, $C_w$ and $V^2$ have to be reduced. These can be realized by using low swing voltage as a wire symbol signal and changing wire material with CNT bundle. Circuits using optimized low-voltage swings as a wire signal have shown 10 times energy savings[4] because reducing voltage level save the power which is proportional to $CV^2f$. But such low-swing systems typically require an expensive second power supply, which breaks up power distribution planes on the chip and in the package requiring additional voltage regulator modules at low target impedances. To overcome these drawback of the low swing techniques, a coupling capacitor in line with the long wire which pre-emphasizes transitions is proposed in [46][47] to reduce wire delay and to reduce the load seen by the driver. It also lowers the wire’s voltage swing without additional power supply. A number of on-chip transceiver improvements have been proposed in the past, but they usually reduce either the power consumed in the interconnect [48], [49] or improve the data-rate achievable over the interconnect [50],[51]. In this paper, the proposed transceiver will adapted PWAM modulation techniques and capacitive transmitting scheme to achieve high data rate and save power with relatively simple modulation circuit and only one power supply voltage. Furthermore, the proposed transceiver is implemented with CNFET and CNT bundle technologies and it will show superior performance compared to CMOS and Cu wire.
Its performance will be proven by various simulation including PVT variations, random jitter, crosstalk noise and bit error rate (BER)

5.2 CNT interconnect technologies

Carbon nanotubes have recently been proposed as a possible candidate to replace the metal interconnects in future technologies [45].

Because of their extremely desirable properties of high mechanical and thermal stability, high thermal conductivity and large current carrying capacity [4], CNTs have aroused a lot of research interest in their applicability as VLSI interconnects of the future. However, the high resistance associated with an isolated CNT (greater than 6.45 KΩ) [45] necessitates the use of a bundle (rope) of CNTs.

Figure 5.1: CNT bundle interconnect structure.
conducting current in parallel to form an interconnection [21] as shown in Fig. 5.2. Moreover, due to the lack of control on chirality, any bundle of CNTs consists of metallic as well as semi-conducting nanotubes. CNTs are also classified into single-walled and multi-walled nanotubes. Single-walled CNTs (SWCNTs) have electron mean free paths of the order of a micron [45]. Therefore, in the domain of interconnects, metallic SWCNTs are the preferred candidates, and the single-walled metallic CNT bundle structure is used for the channel media between the transmitter and receiver in this paper. Noise and bit error rate of the transceiver is estimated and analyzed based on the interconnect structure.

5.3 PWAM SCHEME

PWAM-(MxN) scheme combines PAM-M (M is the number of different voltage levels) and PWM-N (N is the number of different pulse widths) together. With M different pulse amplitudes and N different pulse widths, there exist M×N different symbol representations. M×N symbols represent $\log_2 M + \log_2 N$ bit binary information. Therefore, the bit rate of PWAM-(M×N) is $\log_2 M + \log_2 N$ times the symbol rate. PWAM-(4×4) is illustrated in Fig 5.2(c) as an example to show the symbol representations of PWAM. One PWAM-(4×4) symbol denotes 4-bit data. The 4-bit data have 16 different values, thus 16 PWAM-(4×4) symbols are used to stand for 16 different values (hereafter, 4-bit data are represented by hex 0, 1, ..., a, b, c, d, e, f). These 16 symbols are divided into four groups
Figure 5.2: The PWAM Signaling (a) Conventional Symbols (b) Conventional PAM-4 Symbols (c) PWAM-(4 x 4) Symbols.
as shown in Fig 5.2(c): Group W, X, Y, Z consists of symbol 0 1 2 3, symbol 4 5 6 7, symbol c d e f, and symbol 8 9 a b, respectively.

5.4 Driving a Wire with a Capacitor

In previous section, 4-PWAM scheme was explained as a method to increase data rate. In this section, the method of driving wire with a capacitor will be described to make low swing signal. Using this method a lot of power can be saved. Driving a long wire of capacitance $C_w$ through a capacitor $C_c$ reduces the signal swing on the wire through capacitive voltage divider [46]. As shown in Fig 5.3, including parasitic capacitance on the right side of the coupling capacitor and including the final load capacitance gives a final wire voltage swing of $V_{swing} = V_{DC}/(C_c+C_w+C_{p2}+C_l)$. Typically, $C_{p2}$ and $C_l$ are both small compared to $C_w$.

![Figure 5.3: The capacitively driven wire](symbols)

The node immediately after the coupling capacitance will initially overshoot and then settle to $V_{swing}$, while the end of the wire will show a rapid rise to the final $V_{swing}$ voltage. Because it acts as a high-frequency short, the capacitor
increases the 3-dB bandwidth of the long wire, allowing shorter cycle times and decreasing latency. Long wires have a low-pass frequency response that limits their operating speed. Capacitive coupling to a long wire creates a pole-zero pair, giving high-frequency emphasis that mitigates the low-pass wire response and increases performance.

Therefore, the capacitive transmitter increases the bandwidth of the interconnect, as it emphasizes each transition with an overshoot. Compared to the low-swing transmitters that switch between supplies, the capacitive transmitter also is much less sensitive to supply noise, as the capacitor divider also attenuates this noise.

5.5 Energy Consumption Of Transceiver

The energy-cost for a rising edge with swing V equals the well-known $E = CV^2$. Half of this energy is dissipated during charging. The other half is stored in the interconnect and dissipated at a later time when the interconnect is discharged (the resistance of the interconnect prevents efficient charge-recycling techniques). Therefore, to reduce the link power, it is necessary to reduce the swing. If only a single supply voltage is available and active circuits are used to reduce the swing, there is no quadratic but linearly relation with the swing. When a dedicated supply voltage is available to generate the low-swing signal, then the power is again quadratically dependent on the swing. Therefore, using dedicated
supply voltage is effect to generate low swing signal as a result, many low-swing techniques with a dedicated supply voltage (either generated on- or off-chip) for the transmitter have been introduced in the past [48],[49],[46], [52].

However, in this work the capacitive transmitter uses a series capacitance (Cc) to drive the interconnect with only single supply voltage. It also lowers the energy consumption by using a simple structure at the driver circuits.

5.6 4-PWAM Circuit Design

In this section block diagram and circuits of the proposed 4-PWAM transceiver will be described. In the transmitter circuits, the phase controller circuit, PWM modulation circuit and adaptive capacitor circuit are proposed and the receiver circuits are also newly designed.

5.6.1 Transmitter circuit

The PWAM transmitter is shown in detail in Fig 5.4. It contains three main building blocks: a PWM modulator, a PAM modulator, and 8 phase PLL. An 8 Phase PLL provides evenly spaced clock phases that are used to produce PWM-encoded signals. After processing Tx-bit0 and Tx-bit1 using the PWM technique, PAM signaling is used to modulate the information from Tx-bit2 and Tx-bit3. In addition to the PWAM modulator, the PWAM modulator, capacitively driven wire act as a pre-emphasis block. Pre-emphasis
compensates for the limited bandwidth of the package leads and channel medium [46].

2-PWM modulator

The pulse width modulation can be implemented by the 8 phase PLL and phase controller. 6 phase signals and control data $D0 \sim D3$ are used as inputs. Based on their combinations, the pulse width is determined. The phase controller and its timing diagram are shown in Fig. 5.5 and Fig. 5.6, respectively.

2-PAM Modulator with capacitive pulse amplitude scheme

In order to realize low power PAM modulator, the low swing capacitive pulse amplitude modulation scheme is proposed in this chapter. The capacitive PAM
Figure 5.5: Phase controller.

Figure 5.6: Timing diagram of Phase controller.
driver can be implemented using adaptive capacitance which is shown in Fig 5.7. The bit3, bit4 determine the output of the Demux D1, D2, D3, D4 which consequently decide the capacitance. By changing the capacitance, multiple Vswing can be obtained, and the level of the voltage swing Vswing is expressed as Equation (5.1)

\[ V_{\text{swing}} = V_{DD}(\sum D_k C_{ck} + C_W + C_{p2} + C_l) \] (5.1)

Where Dk is 0,1 and Cck is a capacitance element of adaptive capacitance network. Cck can be expressed as (5.2);

\[ C_{ck} = N_k \cdot C_{c1} + C_{gate,k} \] (5.2)

Where Cc1 is the capacitance of one carbon nanobue in P1 ~ 4, Nk is the number of carbon nanotubes and Cgate,k is the capacitance due to the gate size in CNFET in adaptive capacitance network. The Cck is nearly proportion to Nk. The N1 ~ 4 are N1 = 120, N2 = 220, N3 = 410, N4 = 790 in this paper.

5.6.2 Receiver circuits

Fig 5.8 shows the block diagram of the receiver. The receiver consists of 4 differential amplifier, 4 comparators, one PLL, the pulse width demodulation circuits, and supplementary circuits. Bit2 and bit3 are recovered by PAM demodulator
5.6 4-PWAM Circuit Design

and bit1 and bit0 are generated from pulse width demodulation circuits. As shown in Fig 5.8. The architecture of the PWM demodulator is quite simple. It is composed of three D-latch and a 3bit to 2bit decoder. The PLL in the receiver end provides 8-phase clock required by the pulse width demodulation circuit. The timing diagram of the PWM architecture is shown in Fig 5.9.

Because the voltage swing of the signal received from wire is small, it is amplified by differential amplifier ($\Delta V_{\text{swing}} = 100\text{mv}$). The different amplitude level of the received signal after amplification is $200\text{mV}$. Accordingly, the reference voltages for comparators are $V_{\text{ref1}} = 1.5\text{V}$, $V_{\text{ref2}} = 1.3\text{V}$, $V_{\text{ref3}} = 1.1\text{V}$, and $V_{\text{ref4}} = 0.9\text{V}$. This multiple voltage references can be obtained from [10]. After amplification, the signal is fed into the comparator which determines digital

![Figure 5.7: Adaptive Capacitance.](image-url)
5.6 4-PWAM Circuit Design

Figure 5.8: Receiver circuits.

Figure 5.9: Timing diagram of PWM Demodulation.
5.7 Noise models and Bit Error Rate

5.7.1 Noise models

In order to estimate the bit error rate of the transceiver, three different noise models (flicker noise, thermal noise, and shot noise) are implemented using HSPICE noise model reference [53] and the wave form of the combined noise is shown in Fig 5.10. As shown in Fig 5.10, the noise voltage amplitude is less than 0.1mV. It is too small compared with PAM voltage swing, and it is

<table>
<thead>
<tr>
<th>Comparator out:1,2,3,4</th>
<th>Bit2</th>
<th>Bit3</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 1 1 1</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>0 1 1 1</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>0 0 1 1</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>0 0 0 0</td>
<td>1</td>
<td>1</td>
</tr>
</tbody>
</table>

Table 5.1: Decoded bit2, bit 3 from comparator out.

output 0 or VDD. The final bit2 and bit 3 are recovered by the decoder based on comparator output. Table 5.1 shows the comparator out mapping with bit2 and bit3.
negligible for bit error.

![Wire model for crosstalk noise.](image)

Figure 5.11: Wire model for crosstalk noise.

The major portion of the noise that causes a bit error is crosstalk noise. Crosstalk noise model is shown in Fig 5.11 [54]. The low swing signal can be victimized by large aggressor signal severely. If the large aggressor signal affects the small victim signal, shielding technique [54] is needed between small and large swing signals. Fig 5.12 and Fig 5.13 show the difference between the shielded and unshielded victim signals. Even though the shield technique is used, it is evident that there is crosstalk noise between two small signals.

### 5.7.2 Bit error rate (BER) estimation

The probabilities of bit error in this transceiver system can be expressed as the following:

\[
P_{err-pwm} = P_{err-pwm} + P_{err-pam}
\]  

(5.3)
5.7 Noise models and Bit Error Rate

Figure 5.12: Crosstalk noise caused by large aggressor signal victim signal without shield technique.

Figure 5.13: Crosstalk noise caused by large aggressor signal victim signal with shield technique.
5.8 Simulation Results

\[ P_{err-pwm} = P_{err-Djitter} + P_{err-DIjitter} \]  

(5.4)

\[ P_{err-pam} = P_{err-process} + P_{err-voltage} + P_{error-crosstalk} \]  

(5.5)

Where \( P_{err-pwm} \), \( P_{err-pam} \), \( P_{err-Djitter} \), \( P_{err-DIjitter} \), \( P_{err-process} \), \( P_{err-voltage} \), and \( P_{error-crosstalk} \) are probabilities of error caused by PWM, PAM, DD(data dependent) jitter, DI (data independent) jitter, and process/voltage variation including crosstalk, respectively. It can be assumed that all probabilities are independent so there is no correlation between them. In this paper, all error events are assumed having Gaussian distribution except \( P_{error-crosstalk} \). Therefore, all probabilities can be found by Monte Carlo simulation parameter and PRBS-11 test bit patterns that cause crosstalk noise.

5.8 Simulation Results

5.8.1 Simulation setup

The technology parameters for the CNFETs are: Physical channel length = 32.0nm, The length of doped CNT drain-side/source-side extension region = 32.0nm, Fermi level of the doped S/D tube = 0.6 eV, The thickness of high-k top gate dielectric material = 4.0nm, chirality of tube = (19,0), Pitch = 10nm,
5.8 Simulation Results

Vfn, Vfp (Flatband voltage for n-CNFET and p-CNFET) = 0.0eV, 0.0eV,
Physical gate length = 32.0nm, The mean free path in intrinsic CNT = 200.0nm,
The length of doped CNT source/drain extension region = 32.0nm, The mean
free path in p+/n+ doped CNT = 15.0nm, The work function of Source/Drain
metal contact = 4.6eV, CNT work function = 4.5eV. The wire model and its
parameters were obtained from [45] (wire length = 2 mm, Cbundle= 140fF,
Rwire = 500Ω), and the supply voltage 1.8V is used in this simulation. In order
to test the proposed 4-PWAM transceiver, the PRBS-11 signal is used. The test
was done with various environments with Monte Carlo method with 1000 times
iteration.

5.8.2 Eye diagram analysis

Fig 5.14 shows eye diagrams of the 4 different pulse widths and 4 different levels
of voltage amplitudes. The voltage difference of each level is around 100mv in
the proposed design.

The first test was accomplished without any random jitter and PVT vari-
ations. The eye diagram for this was shown in Fig 5.14a. The proposed
transceiver works without any error (BER=0) and there is only data depen-
dent jitter which is $\Delta T_{rms} = 5.72ps$ and $\Delta T_{pk} - pk = 20.1ps$.

The second simulation was done by adding random jitter which has Gaussian
distribution with nominal value = 50ps and variation = 0.1, $\sigma = 3$. The eye
5.8 Simulation Results

diagram of this environment is shown in Fig 5.14b. As shown in the Fig 5.14b, the eye is almost 90% opened. As a result, the bit errors are too small to measure. The third simulation is done by putting process variation with Gaussian distribution which of 0.1 channel length and width variation and 3. Its eye diagram is shown in Fig 5.13c. The eye is almost closed between 1.29 ∼ 1.49. As the eye is getting closed, the bit error happens because of misreading in comparator voltage level. In this case the bit error rate is expected to be less than 10-10. The fourth simulation is combining the random jitter and process variation used in second and third simulation. As shown in Fig 5.13d, the only difference between the third and the fourth simulation result is the random jitter increase. Because the random jitter in this case is not affecting the bit error rate, the fourth simulation case has same bit error rate with the third one.

In order to see the effect of the supply voltage variation, another simulation was done. The simulation of supply voltage variation was performed with Gaussian distribution which has 0.01 ∼ 0.1 variations and σ = 3 for voltage. By changing the voltage variation, it is observed that the eye opening is changing drastically. Fig 5.12e shows eye-diagram for 0.01 variation and Fig 5.12f shows eye-diagram for 0.5 variation. At 0.01 variation the eye is almost 60 ∼ 70% open and as the variation goes over 0.5, the eye is completely closed.
As a result, the bit error rate is increased about 10^{-8} which can be expected from Gaussian distribution function. The Fig 5.13 shows the expected bit error rate with supply voltage variation. From the eye-diagram analysis, it is concluded that the proposed transceiver is sensitive to voltage variations which come from process and supply voltage variation but not from random jitter.

Because of the PWAM scheme and capacitively driven method, the data rate in this work is higher than any other scheme. It is impossible to make the pulse width 20\% of clock cycle over 1.5GHz, consequently the maximum frequency of this work is 1.5GHz. However, in the [9], it does not need to separate the clock cycle and it can increase the clock up to 5GHz with 50\% eye opening. In case of capacitive driven wire, the data dependant jitter is less than other method. The voltage swing determines data dependant jitter because large voltage swing generally takes more ring and falling time. Consequently the data dependant jitter of this is less than [7] but larger than [9].

The bit rate was summarized in Table 5.2 in the worst case of each simulation environments. To compare the performance of this work to others, [44] and [47] were implemented and its simulation results are summarized in Table 5.3. The proposed transceiver consumes more power than [9] but shows less power consumption than [47]. The results prove that the voltage swing is dominant for power consumption.
5.8 Simulation Results

(a) without random jitter, PVT variations

(b) with random jitter
5.8 Simulation Results

(c) without random jitter, PVT variations

(d) with random jitter
Figure 5.12: Eye-diagram of received PWAM signals (a) without random jitter, PVT variations (b) with random jitter (c) with process variation (d) with process variation and random jitter (e) supply voltage variation (variation = 0.01) (f) supply voltage variation (variation = 0.05).
5.8 Simulation Results

Figure 5.13: Receiver circuits.

<table>
<thead>
<tr>
<th></th>
<th>Random Jitter</th>
<th>Process Variation</th>
<th>Voltage Variation</th>
<th>Temperature Variation</th>
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<tr>
<td>BER</td>
<td>0</td>
<td>$&lt; 10^{-10}$</td>
<td>$&lt; 10^{-8}$</td>
<td>0</td>
</tr>
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Table 5.2: Bit Error rate with Random jitter and PVT variation.

<table>
<thead>
<tr>
<th>Reference</th>
<th>44</th>
<th>47</th>
<th>This work</th>
</tr>
</thead>
<tbody>
<tr>
<td>Technology</td>
<td>0.18</td>
<td>90nm</td>
<td>0.18</td>
</tr>
<tr>
<td>Supply</td>
<td>1.8V</td>
<td>1.2V</td>
<td>1.8V</td>
</tr>
<tr>
<td>Max. Voltage Swing</td>
<td>900mV</td>
<td>120mV</td>
<td>400mV</td>
</tr>
<tr>
<td>Power dissipation</td>
<td>Tx : 86mW</td>
<td>Tx : 14.4mW</td>
<td>Tx : 36.5 mW</td>
</tr>
<tr>
<td></td>
<td>Rx : 79mW</td>
<td>Rx : 18.4 mW</td>
<td>Rx : 45.2 mW</td>
</tr>
<tr>
<td>Data rate</td>
<td>2Gb/s</td>
<td>5Gb/s</td>
<td>8Gb/s</td>
</tr>
<tr>
<td>Jitter</td>
<td>Trms = 49.9 ps</td>
<td>Trms = 9.8 ps</td>
<td>Trms = 11.2 ps</td>
</tr>
<tr>
<td></td>
<td>Tpk-pk = 180 ps</td>
<td>Tpk-pk = 58.2 ps</td>
<td>Tpk-pk = 70.1 ps</td>
</tr>
</tbody>
</table>

Table 5.3: Summary of the performances and comparison.
5.9 Summary

In this chapter, capacitive 4-PWAM transmitter architecture is proposed and its performances are analyzed with random jitter and PVT variation comparing with other works. To implement 4-PWAM transmitter new phase controller and adaptive capacitance network is designed. At the receiver side, new architecture for PWM and PAM demodulation is proposed. The 4-PWAM transceiver in this work has low data dependant jitter which does not affect the bit error rate. But it is very sensitive to supply voltage variation. It is because voltage difference between amplitude modulation is relatively low (100mv) which causes low noise margin. But if the supply voltage variation is lower than 5%, the bit error rate will be less than $10^{-10}$.

The proposed capacitive 4-PWAM transmitter with low swing voltage can save a lot of power consumption comparing with the full swing one and other PAM scheme. Furthermore it can increase data rate with PWAM modulation and pre-emphasis. In this chapter, the proposed transceiver demonstrates that almost $1.74 \sim 2.4$ times power is saved compared to [44] and 4 times higher data rate than [47] is accomplished.
Chapter 6

Conclusion

6.1 Conclusion

The Carbon Nanotube Field Effect Transistor (CNTFET) is one of the most promising devices among emerging technologies to extend and/or complement the traditional Si MOSFET. As the characteristics of a CNTFET is different from conventional bulk CMOS, new design method must be established. As demand of new design method, this thesis analyze the characteristics of CNFET, CNT interconnect technologies and propose new methodologies to design circuits such as digital, memory and I/O circuit.

The quantitative results of chapter 2 have confirmed that the CNTFET technology is a viable solution to replace the conventional bulk MOSFET technology
and, in particular, this work makes it possible to quantitatively estimate the delay, leakage current, and power of CNTFET-based gates.

This has been accomplished by proposing a new optimization methodology for CNTFETs in chapter 3. The parameters such as threshold voltage, gate capacitance, drain current, optimum fan-out factor can be determined by pitch, chirality and the number of carbon nano tubes. Using these parameter, the logical effort has been calculated and the minimum delay of a multistage circuit topology has been analyzed. To prove the effectiveness of the proposed gate-level design method, simulation has been performed using HSPICE with the CNTFET library of [22][17], Results have demonstrated that the proposed design methodology is both effective and practical. To design a CNFET circuit, many parameters must be considered, among them the diameter at certain chirality, pitch and the optimum number of tubes have been shown to be of primary importance.

In chapter 4, new SRAM circuit was proposed. The new SRAM cell cuts off the feedback connection between the two back-to-back inverters in the SRAM cell when data is written and separates the write and read port with 8 transistors. The proposed technique saves dynamic power by reducing discharging frequency during write operation. Compared to 6T SRAM structure, the proposed 8T SRAM for CNFET saves power up to 48% and obtains 56% wider SNM during read operation at the minimal cost of 2% leakage power and 3% delay increase.
As the cells are more frequently accessed, the dynamic power saving is linearly increased. This work also compares CMOS, FinFET and CNT 6T and 8T SRAM cells using HSPICE simulations. The result demonstrates from 3 to 7 times less dynamic power consumption, from 11 to 17 times less leakage power consumption, from 5 to 6 times faster read and write operations, and 1.6 wider SNM than the conventional CMOS designs.

In chapter 5, capacitive 4-PWAM transceiver architecture for CNFET and CNT interconnect technologies is proposed and its performances are analyzed with random jitter and PVT variation comparing with other works as an I/O circuits. The characteristics such as wire capacitance, resistance and inductance of CNT bundle and gate capacitance, analog parameters of CNFET was analyzed to design transceiver. To implement 4-PWAM transmitter new phase controller and adaptive capacitance network is designed. At the receiver side, new architecture for PWM and PAM demodulation is proposed. The 4-PWAM transceiver in this work has low data dependant jitter which does not affect the bit error rate. But it is very sensitive to supply voltage variation. It is because voltage difference between amplitude modulation is relatively low (100mv) which causes low noise margin. But if the supply voltage variation is lower than 5%, the bit error rate will be less than $10^{-10}$. The proposed CNFET capacitive 4-PWAM transceiver with low swing voltage can save a lot of power consumption comparing with the full swing one and other PAM scheme. Furthermore it can
increase data rate with PWAM modulation and pre-emphasis [5]. In this work, the proposed transceiver demonstrates that almost $1.74 \sim 2.4$ times power is saved compared to [7] and 4 times higher data rate than [9] is accomplished.

This research presents a development of systematic design method to optimize circuit speed and power consumption. In this thesis, three optimization methods are proposed and some traditional CMOS circuits are modified for CNFET and CNT interconnect technologies. The optimization methods explored in this thesis include digital circuit design, memory circuit design and high speed on chip I/O circuits. In order to test the effectiveness of the design method, CNFET and CNT interconnect models have been developed and extensive HSPICE simulations have been performed in realistic environments considering screening effects, various noises and PVT variation. The simulation results show that proposed methodologies and modified circuits performed high speed and consumed low power compared to non-optimized and traditional circuits.
Bibliography


[32] Ivan E. Sutherland, Robert F. Sproull, and David F. Harris "Logical Effort: Designing Fast CMOS Circuits" Margan Kaufmann


