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A Cascadeable Pipelined Fast Fourier Transform Switch With Built-In Self-Test

by

Clifford Po Shek Yeung, B.Sc.

**A thesis submitted to the Faculty of Graduate
Studies and Research in partial fulfillment
of the requirements for the degree of**

Master of Engineering

**Ottawa -Carleton Institute for Electrical Engineering
Department of Electronics
Faculty of Engineering
Carleton University
Ottawa, Canada
January, 1992**

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**"A CASCADEABLE PIPELINED FAST FOURIER TRANSFORM
SWITCH WITH BUILT-IN SELF TEST"**

**submitted by Clifford Po Shek Yeung, B.Sc.
in partial fulfilment of the requirements for
the degree of Master of Engineering**

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January, 1992**

Abstract

The single pipelined fast Fourier transform (FFT) is a one dimensional systolic array composed of two building blocks, namely an FFT Switch and a Complex Butterfly. By partitioning the computation requirements among a number of processing elements, the pipelined architecture offers a speed improvement over the traditional Von-Neumann architecture. A design for the radix-2 pipelined FFT Switch is described in this thesis. The generic design of the proposed architecture enables the FFT Switch to support a number of different transform sizes for a radix-2 FFT processor of up to 8192 points. Additional circuitries are incorporated into the FFT Switch in order to facilitate off-line testing. The Built-in self-test (BIST) is fast becoming an essential technique for verification of complex integrated circuits. The pipelined FFT processor is part of the multi-carrier demodulator intended for on-board digital signal processing for satellite communications.

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Table Of Contents

Acceptance.....	i
Abstract	ii
Acknowledgements	iii
Table Of Contents	iv
List Of Figures	vii
List Of Tables.....	x
List Of Abbreviations	xi

Chapter 1 Introduction

1.1 Thesis Motivation and Objectives	1
1.2 Thesis Contributions	5
1.3 Thesis Format	5

Chapter 2 Fast Fourier Transform Implementations and Fourier Transform Alternatives

2.1 Introduction	6
2.2 The Discrete Fourier Transform and Fast Fourier Transform Algorithms	7
2.3 FFT Implementations Overview	11
2.3.1 Introduction	11
2.3.2 General Purpose Digital Systems	12
2.3.3 General-purpose Digital Signal Processor	13
2.3.4 Dedicated Processors for DFT and FFT	15
2.3.4.1 Introduction	15
2.3.4.2 Pipelined and Parallel Pipelined FFT Architecture	15
2.3.4.3 Single Chip Approach	20
2.3.4.4 Multichip Approach	21
2.4 Summary	23

Chapter 3 Architectural developments for the Pipelined Fast Fourier Transform Chip Set

3.1	Introduction	24
3.2	Problems Definition	25
3.3	Architectural Considerations for the FFT Switch	26
3.3.1	Introduction	26
3.3.2	Cascadeability	27
3.3.3	Switch Control	30
3.3.4	Testability Issues	32
3.4	Architectural Considerations for the Complex Butterfly	34
3.5	System Control	35
3.6	Functional Simulation of the Pipelined FFT	36
3.6.1	Introduction	36
3.6.2	Models for the FFT Switch and the Complex Butterfly	37
3.6.3	Simulation Results	39
3.7	Summary	39

Chapter 4 Circuit Description of the FFT Switch

4.1	Introduction	45
4.2	Computer Aided Design Tools	46
4.2.1	Introduction	46
4.2.2	Schematic Capture Tools	46
4.2.3	Simulation Tools	47
4.3	Design Implementation	48
4.3.1	Design of the Delay Circuit and Commutator blocks ...	48
4.3.2	Control Unit Design	53
4.3.2.1	Introduction	53
4.3.2.2	The Clock Cycles Generator	53
4.3.2.3	The Sync Pulse Logic (SPL)	55
4.3.2.4	The Cascadeability Control	56
4.3.2.5	Information Latch	57
4.3.3	Input and Output	59
4.3.4	Design for Testability	59

4.3.4.1 Built-in Self-test	59
4.3.4.2 Interconnectivity Test	64
4.4 Summary	65
Chapter 5 Simulation and Test results	
5.1 Introduction	66
5.2 Simulations of the FFT Switch	67
5.2.1 Stand-alone module	67
5.2.1.1 Introduction	67
5.2.1.2 Functional Test	67
5.2.1.3 Non-functional Simulations	77
5.2.2 Cascaded Structure Simulations	82
5.3 Prototype for the FFT Switch	87
5.4 Prototype Tests	87
5.5 Summary	90
Chapter 6 Conclusions	
6.1 Summary	92
6.2 Future Works	94
Reference.....	95
Appendices.....	102
Appendix A	
Appendix B	
Appendix C	
Appendix D	
Appendix E	

List Of Figures

Figure 1.1 A Multi-channel Satellite Communication System	3
Figure 1.2 Polyphase Filters cascade with FFT Processor	4
Figure 2.1 A partly decomposed 16 point signal flow diagram	10
Figure 2.2 A fully decomposed flow diagram of a 16-point FFT.	12
Figure 2.3 A Generic Pipelined Architecture.	16
Figure 2.4 A Block Diagram of a Single pipelined Fast Fourier Transform.	17
Figure 2.5 A 32-point radix-2 pipelined fast Fourier transform	18
Figure 2.6 A 32-point Radix-2 parallel pipelined fast Fourier transform	19
Figure 2.7 A timing diagram of the single pipelined FFT.	20
Figure 2.8 A timing diagram of the parallel pipelined FFT.	21
Figure 3.1 A block diagram of the FFT Switch	26
Figure 3.2 Five modes of operations	28
Figure 3.3 A block diagram of a 128 point pipelined FFT processor	29
Figure 3.4 Four cascaded Switches	30
Figure 3.5 A diagram to illustrate the operation of the Sync pulse.	31
Figure 3.6 A built-in self-test arrangement.	33
Figure 3.7 A Block Diagram of a Complex Butterfly	34
Figure 3.8 A model for FFT Switch.	37
Figure 3.9 A model for the Complex Butterfly.	38
Figure 3.10 Operations of the commutator	38
Figure 3.11 Model code for the Delay Element	39
Figure 3.12 Model code for the Commutator	40
Figure 3.13 Model code for the Complex Butterfly	41
Figure 3.14 Model code for the Angle Block	42
Figure 3.15 The simulated Splice output with a 2 KHz input source	43
Figure 3.16 The simulated Splice output with a 10 KHz input source	43
Figure 4.1 A block diagram of a FFT Switch	48

Figure 4.2 Data channels of Left Switch mode	49
Figure 4.3 A two bits delay line structure	50
Figure 4.4 A Commutator with four sets of multiplexers	52
Figure 4.5 A Clock Prescaler	54
Figure 4.6 Delay elements of a 32 point pipelined FFT processor	55
Figure 4.7 Operation of the 32 point pipelined FFT	58
Figure 4.8 A Linear Feedback Shift Register	62
Figure 4.9 A Multiple Input Pseudorandom Compacter	63
Figure 4.10 An input buffer with a NAND gate for a parametric NAND tree.	65
Figure 5.1 A 32 point pipelined FFT processor	68
Figure 5.2 Simulation result of Switch A of Example 1	69
Figure 5.3 Simulation result of Switch D of Example 1	71
Figure 5.4 A two Switches cascaded structure	72
Figure 5.5 Simulation result of End switch mode in Example 2	73
Figure 5.6 Simulation result of Output switch mode in Example 2	74
Figure 5.7 A four Switches cascaded structure	75
Figure 5.8 Simulation result of End switch mode in Example 3	76
Figure 5.9 Simulation result of Left switch mode in Example 4	78
Figure 5.10 Simulation result of Right switch mode in Example 4	79
Figure 5.11 Simulation result of Output switch mode in Example 4	80
Figure 5.12 Simulated response of the GO_NOGO output	81
Figure 5.13 Three two modules combinations include a Switch	83
Figure 5.14 Simulation result of Example 6	85
Figure 5.15 Simulation result of Example 7	86
Figure 5.16 A die photo of the FFT Switch	88
Figure A.1 The simulation result of Switch B in Example 2.....	101
Figure A.2 The simulation result of Switch C in Example 2.....	102
Figure B.1 The simulation result of the Left Switch mode in Example 3	104
Figure B.2 The simulation result of the Right Switch mode in Example 3	105
Figure B.3 The simulation result of the Output Switch mode in Example 3	106

Figure C.1 Operation of a parametric NAND tree.	109
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List of Tables

Table 5.1 A table of all possibilities for “Switch-Switch” combination	84
Table 5.2 “Switch-Butterfly” and “Butterfly-Switch” combinations	87
Table 5.3 Prototype results at self-test mode for 5.0V power supply	89
Table 5.4 Prototype results at self-test mode for 3.3V power supply	90

List of Abbreviations

ADC	Analog to Digital Converter
ASIC	Application Specific Integrated Circuit
BIST	Built-in Self-test
CAD	Computer Aided Design
CFT	Continuous Fouier Transform
CMOS	Complementary Metal Oxide Semiconductor
CRC	Communication Research Centre
CS	Chip Select
DFT	Discrete Fourier Transform
DIT	Decimation In Time
DSP	Digital Signal Processor
FDMA	Frequency Domain Multiplexing Access
FFT	Fast Fourier Transform
GPDS	General Purpose Digital System
HCMOS	High Speed CMOS (LSI Logic)
IC	Integrated Circuit
LFSR	Linear Feedback Shift Register
MCD	Multi-carrier Demodulator
MCM	Multichip Module
MIPC	Multiple Input Pseudorandom Compacter
MSPS	Million Samples per Second
RAM	Random Access Memory
ROM	Read Only Memory
SPL	Sync Pulse Logic
TDM	Time Domain Multiplex
V_{IL}	Logic low input
V_{IH}	Logic high input
VLSI	Very Large Scale Integrated circuits

Chapter One

Introduction

1.1 Thesis Motivation and Objectives

Since the development of the Cooley-Tukey fast Fourier transform (FFT) algorithms [1] over two decades ago, there has been a great deal of advancement in the area of hardware FFT implementation. However, it is only recently that VLSI technology has enabled the realization and exploitation of this algorithm in such real-time applications as radar detection, image processing or frequency-division multiple access (FDMA) to time-division multiple access (TDMA) signal conversion. Furthermore, current advances in the field of very high speed Analogue to Digital Converters (ADC) enhance data acquisition technique and allow the integration of advanced FFT processors with throughputs of the order of tens of millions of samples per second as part of the final system.

Due to the regular structure of this algorithm and its ability to compute in-place, it is particularly popular for FFT implementation and a number of different approaches have been used. The FFT algorithm can be computed by programming a microprocessor or a general purpose digital signal processor. However, for critical applications where very high throughputs are essential, general purpose digital signal processors and even dedicated FFT processors may not be adequate. One way to increase the throughput of an FFT processor is to partition the computational requirements among a number of processing elements. The pipelined FFT does this by mapping the FFT algorithm onto a one dimensional systolic array. The processor can be implemented by means of two basic building blocks, namely an FFT Switch (Switch) and a Complex Butterfly. The basic function of the FFT Switch is to provide the necessary clock delays and to reorder and rout the data between the Complex Butterflies, which are responsible for all numerical computations.

As the demand for satellite communications increases rapidly, especially in the commercial sectors, the need for low space-segment tariffs and simple and cheap earth-terminals is growing. One promising technique which could meet these objectives would use FDMA as the up-link to the satellite and TDMA for the down-link to ground stations, providing that the necessary processing is done on-board the satellite[2]. Figure 1.1 shows the basic concept of the system. This technique demonstrates that multi-channels communication can be accomplished by employing low cost earth-terminals with the expense of extra processing hardware on-board the satellite. The specific hardware in question is the multi-carrier demodulator (MCD), which permits the direct on-board interfacing of FDMA and TDMA communication links by digital signal processing techniques. One part of the MCD is a demultiplexer which can be

Figure 1.1 A Multi-channel Satellite Communication System

implemented by means of a digital polyphase filter bank cascaded with an FFT processor. Figure 1.2 shows the arrangement of the demultiplexer. The initial impetus for this work has been the requirement to develop a 25 Msample/sec (MSPS) per channel 32-point radix-2 FFT processor for the demultiplexer. Although other techniques [3] can be used to implement the demultiplexer, the FFT design is particularly efficient for the situations where a large number of communication channels are present.

The FFT chip set is designed to support the implementation of a wide range of pipelined radix-2 FFT processors. In order to accomplish this, both the FFT Switch and Complex Butterfly must function as generic building blocks capable of operating at different locations of the pipelined processor. The implementation of the FFT Switch is

Figure 1.2 Polyphase Filters cascade with FFT Processor

the focus of this thesis and a description of the Complex Butterfly can be found in the literature [22].

The objective of this thesis is to develop a generic FFT Switch capable of implementing a very high speed radix-2 FFT. Furthermore, the FFT Switch must be cascadeable in order to provide the necessary number of clock delays for higher transform sizes which a single Switch cannot accommodate due to technological limitations. A portion of the Switch circuitry must be dedicated to facilitate testing. The self-test approach is particularly attractive in the testing of integrated circuits with a large number of pin out.

1.2 Thesis Contributions

By utilizing the pipelined architecture, a chip set is designed for high throughput fast Fourier transform implementation and a FFT Switch architecture is proposed in the thesis. The proposed generic architecture of a FFT Switch demonstrated that by exploring the regular structure of the pipelined FFT algorithm and the in-place combination, it is possible to implement a spectrum of transform sizes by using only two integrated circuits in the chip set. The design of the Switch is cascadeable and it shows that, without the need for further external memory elements, it is possible to provide the necessary number of delay for the pipelined architecture. Moreover, the flexibility of the Switch design enables to support the implementation of the forward and inversed, decimation-in-time and decimation-in-frequency radix-2 FFT algorithms.

1.3 Thesis Format

In Chapter 2, a broad survey on the fast Fourier transform implementations and alternatives is presented. Among the available Fourier transform implementations, a pipelined FFT processor provides the highest performance in terms of data throughput. A more thorough investigation into the pipelined FFT chip set and in particular the FFT Switch is provided in Chapter 3. In Chapter 4, the FFT Switch is partitioned into separate functional blocks and the circuitry in each block will be described in detail. Simulation and test results are presented in Chapter 5 in order to verify the functionality of the Switch. Chapter 6 provides a summary and suggestions for future work.

Chapter Two

Fast Fourier Transform Implementations and Fourier Transform Alternatives

2.1 Introduction

For many years, the Fourier transform has been a universal technique which enables one to examine a particular relationship from an entirely different point of view. It identifies the frequency components which make up a continuous waveform. Unfortunately, as a continuous and infinite process, the Fourier transform is not amenable to the machine discrete time computation adopted by all digital systems. However, a special type of Fourier transform suitable for digital hardware and software implementations has been proposed, and is called the discrete Fourier transform (DFT). The DFT is the sampled representation of the continuous Fourier transform (CFT), which operates over a finite time interval. For a sufficiently small sampling interval and large number of sampled points it asymptotically approximates CFT. A class of efficient

algorithms to compute the DFT is known as the fast Fourier transform (FFT). One such algorithm was developed by Cooley and Tukey [1]. This now commonly used FFT algorithm is one of the most important operations encountered in signal processing applications today.

The FFT, with its substantial savings in computer time, can be used in place of the continuous Fourier transform only to the extent that the DFT could be previously. Many applications, such as digital convolution, spectrum analysis, audio and video signal processing and so on, are able to utilize the FFT as an efficient technique to perform computations. Furthermore, through VLSI technology, many FFT processes are able to operate with the very high clock frequency necessary for real-time applications.

In the next section, the DFT and FFT algorithms are examined. This is followed by an implementation overview of the FFT which includes an outline on the pipelined and parallel pipelined architectures.

2.2 The Discrete Fourier Transform and Fast Fourier Transform Algorithms

The Fourier transform function for continuous signals can be written in the form of Equation (2.1)

$$X(f) = \int_{-\infty}^{\infty} x(t) e^{-j2\pi ft} dt \quad (2.1)$$

for $-\infty < f < \infty$ and $-\infty < t < \infty$. The $X(f)$ represents the frequency-domain function and $x(t)$ is the time-domain function.

Although a more complete examination of fast Fourier transform should begin with a detailed look at the continuous Fourier transform, since it is not suitable for digital computation the emphasis will be placed on the DFT.

As mentioned in the previous section, the DFT can be viewed as the sampled representation of CFT and sampled waveform can be expressed in Equation (2.2).

$$x(n) = x(t) \sum_{k=-\infty}^{\infty} \delta(t - kT) \quad (2.2)$$

Sampling is accomplished by multiplying the continuous time-domain signal $x(t)$ with a time-domain periodic impulse train, $\sum_{k=-\infty}^{\infty} \delta(t - kT)$. A time difference T , between periodic sampling impulses, is called the sampling interval or sampling frequency for its reciprocal. In order to avoid aliasing and loss of information, the continuous waveform must be sampled at a frequency of at least twice the largest frequency component of the usually bandlimited continuous signal $x(t)$. Once the sampled time-domain waveform is obtained, further computation to transform the data to frequency-domain is completed by using a discrete-time process. The analogous discrete Fourier function can be expressed in the form of Equation (2.3)

$$X(k) = \sum_{n=0}^{N-1} x(n) W_N^{kn} \quad (2.3)$$

where $W_N = e^{-j((2\pi)/N)}$ $k = 0, 1, 2, \dots, N-1$ and N is the number of transform points. The $x(n)$ is a complex data sample in the time-domain and $X(k)$ is the complex image of the data with frequency-domain.

Since continuous Fourier transform is an infinite process where an infinite number of samples will be introduced, it is necessary to truncate the sampled function $x(n)$ so that only a finite number of points, say N , are considered. Examination of equation (2.3) reveals that N^2 complex multiplications and $N(N-1)$ complex additions are necessary to perform direct computation [4]. As each complex multiplication represents four real multiplications and two real additions, the algorithm becomes extremely computationally expensive as N increases. In addition to the amount of computation being required, provisions have to be made for storing and accessing the N complex input sequence $x(n)$ and the values of complex coefficients. Fortunately, an efficient class of algorithms has been developed to compute the DFT, collectively known as the fast Fourier transform (FFT). Such algorithms exploit the symmetry property and periodicity properties of W_N^{kn} as shown in Equations (2.4) and (2.5) respectively [5].

$$W_N^{k(N-n)} = W_N^{-kn} = (W_N^{kn})^* \quad (2.4)$$

$$W_N^{kn} = W_N^{k(n+N)} = W_N^{(k+N)n} \quad (2.5)$$

where $()^*$ denotes the complex conjugate of $()$. Further savings can be achieved by using the fundamental principle of decomposing the computation of the discrete Fourier transform of a sequence of length N into successively smaller discrete Fourier transforms. For the special case where N is an integer power of 2 (i.e. $N=2^Y$, where Y is an integer), it is possible to compute the $X(k)$ by separating $x(n)$ into a number of $(N/2)$ -point sequences. For example, a partly decomposed flow graph of a 16-point FFT algorithm is as shown in Figure 2.1. The process continues until the sequence is fully decomposed. Since the number of complex multiplications for an N -point DFT reduces to

Figure 2.1 A partly decomposed 16 point signal flow diagram

$(N/2)^2$ for a $(N/2)$ -point DFT, the total multiplication requirement of the system for one decomposed step will be $N+2(N/2)^2$. As the $(N/2)$ -point DFT decomposes to two $(N/4)$ -point DFT and to four $(N/8)$ -point DFT and so on, it reduces the overall number of complex multiplications from N^2 (no decomposition) to $N\log_2 N$ (fully decomposed). It is clear that even for a moderately sized N , a considerable reduction in computational operations can be achieved. Further details can be found in the literature [5][6].

A fully decomposed 16-point radix-2 decimation-in-time (DIT) FFT signal flow diagram is shown in Figure 2.2. Two interesting features of this algorithm are shown in the diagram. First of all, it is possible to perform the in-place computation using this algorithm. An in-place computation requires the least storage capacity during operation and hence it is considered very suitable for the pipeline implementation. Secondly, in order for the computation to be done in-place, it is necessary to either pre-format the input sequence into bit-reversed order before it is accessed or post-format the bit-reversed output for a normally ordered input sequence. Although it is possible to arrange the signal flow diagram such that both input and output sequences are normally ordered, the in-place property of the algorithm will be lost [6].

2.3 FFT Implementations Overview

2.3.1 Introduction

In the previous section, a brief examination of the FFT algorithm and its computational advantage over the DFT algorithm were presented. Substantial computational saving is the primary reason that has led to the acceleration of development in the field of DFT implementation using FFT algorithms. In this chapter,

Figure 2.2 A fully decomposed flow diagram of a 16-point FFT.

an overview of the FFT implementations will be provided in the next few sections starting with general-purpose digital systems. This is followed by descriptions of digital signal processors and dedicated FFT hardware.

2.3.2 General Purpose Digital Systems

In the early days of FFT use, most implementations were automatically processed on general-purpose digital systems (GPDS) [7][8][9] where code sequences were written in either assembly or high-level languages to achieve a high degree of flexibility. Furthermore, researchers were able to experiment with various forms of FFT algorithms used in many different applications. Apart from the high degree of

flexibility, additional techniques have been introduced to upgrade system performance. Most noticeable are the implementations of FFT application specific microcode [7] and the “number crunching” array processors [10]. For the former technique, the FFT algorithm can be partly or completely integrated into microcode at the expense of additional program memory and an extensive amount of programming effort to accomplish the transformation. Although FFT is an efficient algorithm to compute DFT, nonetheless it remains an expensive computational process. Since array processors are capable of high speed mathematical executions, they will significantly improve the overall system performance. In order to operate an array processor effectively, one must obtain parallelism in a particular algorithm and minimize data organizational overheads. Routing between individual processors can be considered as an operation and incurs additional overhead in term of the arithmetic complexity of the algorithm. Moreover, since the cost of such processors is usually high, unless they are used intensively, this may not be a feasible option to take.

Since the execution speed of most general-purpose digital systems is relatively slow compared to that of the dedicated FFT hardware, this implementation will be limited to lower bandwidth applications. Furthermore, the overall system cost and size are ultimately unsuitable both for wide-spread use and for our application in particular.

2.3.3 General-purpose Digital Signal Processor

Advances in the development of programmable Digital Signal Processors (DSP) since their commercial introduction in the early 1980s provide an attractive alternative for computing FFT algorithms. DSPs can support a wide variety of FFT algorithms and a number of FFT implementations have been reported in the literature [11][12]. Similar

to the general-purpose digital system, DSPs are generally programmed in assembly languages. However, unlike the GPDS, the architecture of the DSP is optimized for signal processing applications where specific numerical operations are required. Researchers have been introducing more and more efficient hardware implementations to fulfil the demands of high FFT execution speed and digital signal processing applications in general. Implementations such as the hardware multiplier, barrel shifter, auxiliary registers and floating point capability enhance the overall performance of the DSPs. The internal random access memory (RAM) structure of the DSPs is able to store a number of twiddle factors necessary to execute FFT algorithms.

In the past, many signal processing algorithms were carefully structured in order to minimize the number of multiply operations at the expense of more addition and memory operations. Such compromise is inevitable for GPDS due to the fact that multiplications require more computational procedures than both addition and memory operations. As for DSPs, multiplications require the same amount of computation as do the other two operations. Thus some optimized algorithms may not be the most favourable routines to implement on a DSP. However, since FFT algorithms minimize all types of operations, they remain popular as a DSP implementation.

Since the cost of DSPs has been steadily decreasing, they have been introduced into many commercial products. Many existing DSP based digital signal processing systems can be arranged in parallel such that it is possible to achieve greater processing power at the expense of higher system cost. However, VLSI will always be faster than a general purpose DSP. For extremely high speed FFT operations, VLSI processors dedicated to compute DFT and FFT are the only option.

2.3.4 Dedicated Processors for DFT and FFT

2.3.4.1 Introduction

When the execution speeds of previously described techniques are found to be either too slow or too costly for the particular application in mind, dedicated VLSI processors optimized for DFT and FFT operations should be used. A number of processor implementations have been reported in the literature. These can be roughly classified into two groups due to their differences in hardware architecture approach, namely single chip processors and multichip processing units. When compared, multichip configuration has the advantage of maximum speed and flexibility over single chip implementation. However, single chip implementation can achieve reasonable performance with lower implementation cost and design effort. Most processors adopt a pipelined architecture approach [13][14] to accomplish the very high performance requirements and maximize the hardware utilization for a given data throughput.

In the next section, the architectures of the pipelined and parallel pipelined FFTs are presented. Selected design features and considerations are highlighted. This is followed by a discussion of single chip and multichip approaches. The multichip approach is adopted in this project.

2.3.4.2 Pipelined and Parallel Pipelined FFT Architecture

Many digital signal processing applications demand an extremely high throughput rate. The pipelined and parallel pipelined FFT processors discussed here may be able to achieve such performance requirements. The one dimensional systolic array (pipelined) architectures generate a great deal of interest since they offer a considerable speed improvement over the traditional Von-Neumann architecture.

Figure 2.3 shows a generic pipelined architecture and the pipeline is partitioned into N number of modules.

Figure 2.3 A Generic Pipelined Architecture.

The criteria for partitioning depend upon various factors such as characteristics of the algorithms and physical constraints of the hardware implementation. One must carefully select the correct algorithm for a particular application and map the algorithm onto the pipelined architecture. In order to achieve a high throughput rate and low system latency, the functionality of each module must be optimized to take advantage of the characteristic of the pipelined architecture. Although the VLSI implementation technologies are advancing rapidly, there remains an upper limit at which the number of transistors can be packed onto a single silicon substrate. For some FFT applications where a large number of transform points is required, one single IC may not be able to contain all the necessary hardware to complete the transformation. This technological constraint may result in mapping the FFT algorithm onto a multiple chip pipelined

architecture in order to achieve a reasonable transform size for various digital signal processing applications.

A single pipelined FFT architecture can be implemented by using only two types of processing elements, namely the FFT Switch (Switch) and Complex Butterfly (Butterfly). A block diagram is shown in Figure 2.4. Whereas the FFT Switches, which act like data commutator elements with delay, rout and reorder the correctly sequenced data between the Complex Butterflies, the Butterflies contain all the computational elements and twiddle factors needed to perform numerical calculations of the FFT algorithm.

Figure 2.4 A Block Diagram of a Single pipelined Fast Fourier Transform.

There are several forms of FFT algorithms. The radix-2 in-place decimation-in-time algorithm was chosen for this project. For a radix-2 FFT implementation, an N -point pipelined FFT will consist of $\log_2 N - 1$ switching stages and $\log_2 N$ Complex Butterflies. A 32-point radix-2 forward and inverse pipelined FFT is shown in Figure 2.5. For the forward and inverse pipelined FFTs, the delays $\{k_1, k_2, k_3, k_4\}$ of the Switches correspond to $\{8, 4, 2, 1\}$ and $\{1, 2, 4, 8\}$ respectively. As for the Complex Butterflies, different numbers of twiddle factors are required for each computational stage in order to achieve a successful transform. Architecturally, the inverse pipelined

FFT corresponds to a transpose of the forward pipelined FFT [15]; hence the same set of integrated circuits can be arranged to perform both forward and inverse transforms.

Figure 2.5 A 32-point radix-2 pipelined fast Fourier transform

The signal flow diagram in Figure 2.2 can be mapped directly onto the single pipelined architecture in Figure 2.5, where the arrows and crosses represent the operations of the FFT Switches and Complex Butterflies respectively.

The diagram in Figure 2.5 corresponds to one Butterfly per stage of the pipelined FFT and thus has one degree of parallelism. However, it is possible to reduce the latency of the system by decreasing the total number of pipelined stages and to perform more of the computations in parallel. The parallel pipelined FFT [16] extends the concept of the pipelined FFT and provides an even greater processing capability. A 32-point radix-2 parallel pipelined FFT is shown Figure 2.6. No Switches are necessary between the first and second Butterfly of each of the two branches as the interconnections are hardwired. It is possible to further increase the processing speed through the utilization of parallel pipelined architectures with a higher degree of parallelism.

Figure 2.6 A 32-point Radix-2 parallel pipelined fast Fourier transform

In order to achieve a 100% hardware efficiency for both pipelined and parallel pipelined implementations, it is necessary to pre-format the input data stream to the parallel inputs of the structure [6][17]. Hardware efficiency is used to measure quantitatively the amount of time (as a percentage) that computational elements are kept busy. Thus for the 32-point single pipelined FFT, the sequential data samples in each 32 sample window must be formatted such that samples 0, 1, 2, ..., 15 are routed to port A and samples 16, 17, 18, ..., 31 are routed to port B of Figure 2.5. A corresponding timing diagram is shown in Figure 2.7. As for the parallel pipelined 32-point FFT, the sequential data samples are formatted such that samples 0-7, 16-23, 8-15 and 24-31 are routed to ports A, B, C and D respectively in Figure 2.6. The timing diagram of the parallel pipelined FFT is shown in Figure 2.8. From the timing diagrams, it can be seen that the parallel pipelined architecture is able to compute a complete set of 32 samples in approximately half the number of clock periods required for the single pipelined architecture.

Figure 2.7 A timing diagram of the single pipelined FFT.

2.3.4.3 Single Chip Approach

In order to achieve high speed operation and minimize the chip count of the entire processing system, one must examine the single chip architecture, whose overall power consumption is generally lower than the multichip approach. A number of single chip FFT processors have been reported in the literature [18][19][20]. Each processor contains all the necessary computational, delay memory, coefficient memory, switching and control elements to compute the transform.

Since technology constraints limit the amount of hardware to be implemented on a single processor, the number of transform points for a processor remains low. Subsequently, some processors are implemented with the discrete Fourier transform

Figure 2.8 A timing diagram of the parallel pipelined FFT.

algorithm rather than the fast Fourier transform algorithm to minimize control overhead. In the case of higher transform sizes, most of the single chip processors are designed such that they are capable of operating in parallel to accomplish, for instance in [19], a 1024-point FFT implementation. However, an excessive amount of external hardware such as multiplier-accumulators and storage memory elements is required in such an implementation. As a result, the eventual size and cost of the system increases a great deal.

2.3.4.4 Multichip Approach

For many FFT applications where extremely high throughput is essential, multichip architecture may be the only feasible option to achieve performance

requirements. In fact, the multichip processor implementations are popular and have been reported in the literature [21]-[26] and more recently some chip sets have become available commercially [33][34]. Pipelined architecture is used to increase throughput. As mentioned in section 2.3.3.2, a pipelined processor can be implemented using only two building blocks, namely an FFT Switch and a Complex Butterfly. It is possible to achieve a spectrum of transform sizes by cascading a number of the two building blocks with minimal external circuitry required.

More recently, due to the advances in packaging technology, researchers have been able to further improve system performance by introducing techniques such as the Wafer-Scale integration (WSI) [25][26] and the Multichip Module (MCM) [31][32]. These techniques allow the reduction of the propagation delay (due to smaller interconnection parasitics) between each integrated circuit. Traditionally, a solution which employs a printed circuit board with off-the-shelf components and custom-built VLSI circuits is used. However as the clock frequency and data throughput increase, data transfer between ICs on the printed circuit board becomes a problem. Although the very expensive ECL and GaAs technologies (at the time of writing) can provide the necessary density and speed required for many digital signal processing applications, the requirement of high power consumption will be marked as a severe constraint for some systems. Hence, both WSI and MCM become very attractive alternatives for digital signal processing hardware implementations. Nevertheless, several architectural problems exist including testability and power consumption, which require more thorough examination before either can be fully realized.

2.4 Summary

The multi-carrier demodulator for digital signal processing on-board of a satellite consists of a demultiplexer; one of the techniques to construct a demultiplexer is to use a polyphase filter bank cascaded with a FFT processor. A number of architectural alternatives can be used to implement the processor. However, in order to achieve the 25 MSPS per channel performance specification within a reasonable size and cost, the multichip pipelined architecture approach is adopted in this project. The initial requirement of a 32-point radix-2 processor can be realized by using a chip set consisting of a FFT Switch and a Complex Butterfly. In the next chapter, the architectural considerations of the FFT processor for our application are discussed.

Chapter Three

Architectural developments for the Pipelined Fast Fourier Transform Chip Set

3.1 Introduction

The design of the pipelined FFT chip set is based upon a number of architectural considerations which reflect both technology constraints and certain performance and system level requirements. These considerations will be discussed for the FFT Switch and briefly for the Complex Butterfly.

The chip set implements the decimation-in-time (DIT) radix-2 algorithm for high performance fast Fourier transform applications. The operations of the FFT algorithm are partitioned such that it is possible to obtain maximum flexibility. Both Switch and Butterfly are implemented in CMOS Gate Array technology.

3.2 Problems Definition

As mentioned in chapter one, the 25 MSPS per channel radix-2 32-point FFT processor is required as part of a demultiplexer in a multi-carrier demodulator for on-board digital signal processing. In order to achieve the system throughput requirement, a multichip pipeline approach is adopted. The chip set consists of two generic building blocks namely, an FFT Switch and a Complex Butterfly operating in a pipelined architecture as shown in Figure 2.4. Although the initial requirement for the demultiplexer is a 32-point FFT processor, it would be desirable to implement processors of higher transform sizes for other applications using the same chip set. In order to achieve this objective, the FFT Switch must be capable of providing the necessary number of delays and the correct reordering and routing operations between Complex Butterflies. Since a limited number of delay elements can be implemented on a single Switch, it is necessary for the Switch to be operated in a chain (cascadeable). In addition, the Complex Butterflies must be able to access the complex coefficients (twiddle factors) efficiently for the numerical computations.

In order to ease the increasingly more difficult task of testing an integrated circuit (IC), both chips should include some testability features. As the number of transform points grows, the number of ICs in a pipeline is expected to increase. Hence, an effective technique must be implemented such that the overall system control overhead is minimized, and as a result, the complexity of the system control circuitry is reduced. These issues which concern the design of the chip set are addressed in the next few sections. Although the architectural problems of the chip set have been identified, the solutions will be limited by technological constraints, such as total gate count,

component library, device speed, power consumption and so on, of the available gate array family.

3.3 Architectural Considerations for the FFT Switch

3.3.1 Introduction

The basic function of the FFT Switch is to provide the necessary clock delays and to reorder and rout the data between the Complex Butterflies. Routing is different from reordering in the sense that it represents a fixed data path in the commutator. A block diagram of the Switch is shown in Figure 3.1. Detailed circuit descriptions of all the functional blocks will be included in the next chapter. However, architectural issues such as cascadeability, control and testability measures which influence the design of the Switch will be discussed in the next few sections.

Figure 3.1 A block diagram of the FFT Switch

3.3.2 Cascadeability

The amount of storage required by a Switch in an N-point pipelined FFT depends upon the value of N and the position of the Switch within the pipeline. As the value of N increases, the number of storage elements required to implement the clock delays will increase. Furthermore, the exact position of the Switch within a pipeline is vital in determining the necessary number of clock delays before and after the reordering of data. Both technologically imposed limitations on the Switch storage capacity and the requirement for a single generic Switch capable of supporting large FFT sizes dictate that the Switch must be cascadeable. A generic Switch implies that the module must be capable of implementing all the switching functions in a pipelined FFT architecture. In order to accomplish the cascadeability requirement, a three-commutator architecture is adopted for the Switch (see Figure 3.1). The commutators are used to rout or to reorder the data through the Switch. All three commutators are identical and their select inputs are controlled by signals which originate from the "Control Unit". The select signals determine the operation of the commutators and subsequently a number of modes can be implemented by the Switch. There are all together five separate modes of operation as shown in Figure 3.2. In order to keep the generic nature of the Switch, every Switch must be capable of operating in all five modes.

Only one Switch is responsible for the reordering of data when a number of Switches are cascaded. This particular Switch is named, appropriately, the Principle Switch. The rest of the Switches on both sides of the Principle Switch provide the necessary delay for the pipeline. The Principle Switch is represented by the Middle and End switch modes depicted in Figure 3.2. As their names imply, the Middle and End

Figure 3.2 Five modes of operations

switch modes correspond to the cases where the second (i.e. middle) and third (i.e. end) of the three commutators of the Principle Switch reorder the data. In cases where a number of Switches are cascaded, those to the left and right of the Principle Switch operate in the Left and Right switch modes respectively. Operation in the two switch modes differs only in the manner in which the synchronization pulse is processed. In all instances where two or more Switches are cascaded, the last Switch must operate as the Output switch mode to ensure data is routed correctly before it proceeds to the next Butterfly stage.

A block diagram of a 128 point radix-2 pipelined FFT processor is shown in Figure 3.3, and Figure 3.4 shows the first set of the cascaded Switches in a 128-point radix-2 pipelined FFT where all the commutators in the Switches are included. Since a

Figure 3.3 A block diagram of a 128 point pipelined FFT processor

maximum of eight clock cycle delays per channel is possible for each Switch and a total delay of 32 clock cycles is required for this particular stage (stage #6), four Switches are necessary for the cascaded structure. The thick lines represent the delay path of the cascaded structure. All data will pass through the same number of delay elements and hence data at output A' and B' is once again aligned before the next processing stage. In this instance, the Principle Switch is operated at the second module of the structure.

Figure 3.4 Four cascaded Switches

3.3.3 Switch Control

Architecturally, both the Switch and the Butterfly are aimed at minimizing control overheads on the system level for the pipelined FFT. The proposed processor architecture should adopt a distributed control methodology where almost all control functions are generated on-board the Switches and Butterflies. The necessary information will be loaded into the modules only once and the control signals will be generated accordingly. Information must include the location of the Switch relative to the whole pipeline and the switch mode of each module.

In order to identify and to ensure the synchronization of valid data in the pipeline, the synchronization (Sync) pulse technique is proposed. The Sync pulse will be generated by the system control of the pipelined processor and propagated along with the data. A negative going edge indicates the start of the valid data and initiates the reordering process. The Sync pulse will remain active low as long as the input data is valid and terminates the reordering process when the Sync pulse returns to active high. Figure 3.5 illustrates how valid data is defined by the Sync pulse. Same number of delay clock cycles will be imposed on both data and Sync pulse in every stage of the processor.

Figure 3.5 A diagram to illustrate the operation of the Sync pulse.

As the number of cascaded modules increases in a pipelined FFT processor, power consumption for the overall system becomes a critical issue. Consequently, it is most desirable to disable those portions of the Switch which are not used. Since some of the delay elements contained in the Switches located at the last few stages, preceding

the pipeline output are not required at all, they are automatically disabled to reduce power consumption. More registers will be disabled in a parallel pipelined architecture where there are more unused registers in the structure.

3.3.4 Testability Issues

Although functional testing traditionally constitutes the basic test strategy for integrated circuits, the cost of functional testing is becoming excessively expensive due to the rapid increase in circuit density and pin count. Subsequently, the automatic test equipment (ATE) used to complete functional testing is considered either too costly or inadequate to accomplish the task. Test pattern generation also requires (particularly for complex circuits) long computation. It appears that the most attractive solution to reduce testing expenses is to include circuitry on each chip to facilitate testing. The test approach adopted for both the FFT Switch and Complex Butterfly is the built-in self-test (BIST) technique based on the generation of pseudorandom test vectors to provide the stimulus of the circuit and analyses the results at the output by using response compression[27][28]. The response compression technique is used to reduce the volume of output data by first allowing the response pattern to pass through a circuit called a compacter which reduces the number of bits after a data compression is performed. A compacted output is compared with a predetermined test response known as the signature. Generally, a single bit pass/fail output will define the status of the circuit. Figure 3.6 shows a block diagram of the BIST arrangement. All BIST procedures are performed separately from normal operations, hence the "off-line" test method. Since testing is accomplished "off-line", there are no constraints imposed on the operating frequency to complete the testing procedures. As a result, testing can be done at design speed or higher.

Figure 3.6 A built-in self-test arrangement.

The disadvantage of BIST technique is that extra circuit area is required to accommodate testing and often BIST related performance degradation of the circuit also occurs. Although there is no strict guideline on the percentage of circuitry in an IC dedicated to testing, normally every effort is made to minimize it.

In essence, the main motivation in pursuing the BIST approach has been to reduce the cost of testing, and this technique has proved to be very popular. Even with the disadvantages mentioned above, the benefits of the BIST make it a worthwhile technique.

3.4 Architectural Considerations for the Complex Butterfly

In this section, a brief account of some design considerations for the Complex Butterfly will be presented. A more thorough discussion of the Butterfly implementation is included in the literature [22]. The Butterfly was designed in advance of the FFT Switch. A block diagram of the Butterfly is shown in Figure 3.7.

Figure 3.7 A Block Diagram of a Complex Butterfly

The decimation-in-time FFT algorithm implemented by the Butterfly can be expressed by Equation 3.1-3.4

$$a' = a + (c \times \cos\theta - d \times \sin\theta) \quad (3.1)$$

$$c' = a - (c \times \cos\theta - d \times \sin\theta) \quad (3.2)$$

$$b' = b + (c \times \sin\theta + d \times \cos\theta) \quad (3.3)$$

$$d' = b - (c \times \sin\theta + d \times \cos\theta) \quad (3.4)$$

where $A = a + jb$ and $B = c + jd$ represent the complex inputs and $A' = a' + jb'$ and $B' = c' + jd'$ represent the complex outputs of Figure 3.7. In order to achieve a throughput of 25MSPS per channel system requirement, multiple-levels of pipelining have to be incorporated into the architecture of the multiplier accumulator of the Complex Butterfly. Techniques such as the modified Booth Algorithm, Wallace Adder Tree and Redundant Binary Arithmetic are implemented in a synchronous pipelined architecture. Since there is always a limit for the number of transistors to be implemented on a single die, it is possible that only a relatively small number of twiddle factors can be stored on-board the Complex Butterfly. The solution is such that the Butterfly can either read the necessary twiddle factors stored in on-board memory or be loaded with the twiddle factors from an external RAM or ROM in a sequential manner. Since not all butterfly stages require a large number of different twiddle factors, they are loaded into the Butterfly and can be accessed in a cyclic fashion. Similarly, the Butterfly circuit also accommodates the distributed control and test methodology to minimize control overheads on the system level.

3.5 System Control

Since most control signals are generated within the FFT Switches and Complex Butterflies, external interference to the flow of data during normal operation will be kept at a minimum. Nevertheless, the system control must be capable of distinguishing the validity of the data and producing the correct Sync pulse signal accordingly.

Furthermore, the system control will supervise the loading of information from an external source to all the modules in the pipeline, apart from the more general distributed signals such as clock pulse and clear pulse.

3.6 Functional Simulation of the Pipelined FFT

3.6.1 Introduction

Prior to the development of a detailed Switch design at the schematic level, a functional simulation of a 32-point radix-2 single pipelined FFT is performed. Functional simulation at a high level of abstraction enables one to gain a better understanding of the overall system. The modified Splice is used as the simulator. Splice is a circuit and logic simulator which can handle multiple levels of abstraction from transient analysis of MOS and bipolar integrated circuits to functional block models. The block models are implemented as a sequence of code based on their arithmetic or logical functions. The models will be assumed as basic primitives for simulations. All functional models are written in "Ratfor" which is a FORTRAN preprocessor language. The models' code, will be compiled and linked with the simulator as subroutines, and can be used like any other Splice source deck. Both FFT Switch and Complex Butterfly models use the electrical analysis mode. The electrical analysis is chosen to obtain detailed analog voltage waveforms for the circuit blocks under analysis. Using the analog representation simplifies the interconnects between models by replacing 4 sets of 16 bit digital connections with 4 analog connections and consequently reduces the evaluation time.

3.6.2 Models for the FFT Switch and the Complex Butterfly

A 32-point radix-2 pipelined FFT model can be implemented by using four FFT Switch models and five Complex Butterfly models. Figure 3.8 and Figure 3.9 show block diagrams of the Switch and the Butterfly models as they are implemented in Splice. The four analog connections represent the real and imaginary parts of the two complex data channels

Figure 3.8 A model for FFT Switch.

The Switch is modelled as three commutators with delays in between. The delay is represented by Z^k blocks in the Figure 3.8. The commutators are controlled by separate select flags. The operation of the commutator is shown in Figure 3.10.

The input to each commutator either passes straight across or crisscrosses, depending on the select flag. For a 32-point pipelined FFT implementation, only the centre commutator of all the Switches is required for data reordering. Figure 3.11 and

Figure 3.9 A model for the Complex Butterfly.

select				
1	$a \rightarrow a'$	$b \rightarrow b'$	$c \rightarrow c'$	$d \rightarrow d'$
0	$a \rightarrow c'$	$b \rightarrow d'$	$c \rightarrow a'$	$d \rightarrow b'$

Figure 3.10 Operations of the commutator

Figure 3.12 shows the Ratfor code for the delay element and commutator respectively in Ratfor.

The arithmetic unit in the butterfly implements the DIT algorithm with equations shown in Equations 3.1-3.4. Since the Splice source desk is written in Ratfor, the Sine and Cosine functions in the equations can be determined with reasonable precision. Twiddle factors for the equations are generated in a different block called the Angle block, separately from the Complex Butterfly model. Only a small number of twiddle factors in each stage is required to compute the 32-point pipelined FFT and these are

```

# model for Delay Element
.define elmod3
.numout 1      # define the number of input
.numinp 1      # define the number of output
.*begin

                evop1=evip1    # one time step delay

.end

```

Figure 3.11 Model code for the Delay Element

accessed in a cyclic manner. The Ratfor code for the Complex Butterfly and Angle models is shown in Figure 3.13 and Figure 3.14 respectively.

3.6.3 Simulation Results

It is possible to examine the functionality of a pipelined forward FFT structure by using a Sine wave as the time-domain input. Since the mapping of a correctly sampled Sine wave in the time-domain to frequency-domain is a periodic impulse train located at the fundamental frequency and its harmonic, it will be a simple task to verify the functionality of the system. Figure 3.15 and Figure 3.16 show some of the simulation results which verify the system functionally.

3.7 Summary

An FFT processor can be implemented by using a chip set which consists of an FFT Switch and a Complex Butterfly. The two building blocks must be cascadeable in order to realize a pipelined structure. Furthermore, due to technological constraints, the Switch is designed such that it can be cascaded with another Switch. Cascadeable

```

# model for Commutator
.define elmod1
.numout 2
.numinp 3
.nummem 2
.*begin
# to determine whether the data
# is straight across or crisscross
# evin1: data
# evin2: data
# evin3: switch flag
#define store1 VAR1          # define variables
#define store2 VAR2

    real flg1
    genflg=.true.

    if (lastim == -1)        # initialization
    {
        flg1=0
    }

# evin3=1 -> straight across
# evin3=0 -> crisscross
# test input flag

    if (simtim != lastim)    # one clock pulse after
    {
        if (evin3==1)
        {
            if(flgl==0)
            {
                flg1=1
            }
            store1=evin1
            store2=evin2
        }

        if (evin3==0)
        {
            if(flgl==1)
            {
                flg1=0
            }
            store2=evin1
            store1=evin2
        }
        # crisscross if flag 0 is false

        evop1=store1
        evop2=store2
        # output results
    }

#undef store1
#undef store2
.end

```

Figure 3.12 Model code for the Commutator


```

# model for Complex Butterfly
.define elmod1
.numout 4
.numinp 6
.nummem 4
.*begin
#define a      VAR1
#define b      VAR2
#define c      VAR3
#define d      VAR4
#define ang    VAR6
#define mut1   VAR7
#define mut2   VAR8
# evin1 is for input a
# evin2 is for input b
# evin3 is for input c
# evin4 is for input d
# evin5 is for angle input
# evin6 is for run flag

                                genflg=.true.
                                if ( lastim == -1 )
                                {
                                    rflg=1
                                }
                                if ( simtim != lastim )      # initialization
                                {
                                    if ( evin6==0 )
                                    {
                                        if ( rflg==1 )
                                        {
                                            rflg=0
                                        }
                                        # when run flag is low, the butterfly is active
                                        # to calculate the four equations

                                        ang=evin5
                                        mut1=evin3*cos(ang)-evin4*sin(ang)
                                        a=evin1+mut1
                                        b=evin1-mut1
                                        mut2=evin3*sin(ang)+evin4*cos(ang)
                                        c=evin2+mut2
                                        d=evin2-mut2
                                    }
                                    if ( evin6==1 & rflg==0 )
                                    {
                                        rflg=1
                                    }
                                    evop1=a
                                    evop2=b
                                    evop3=c
                                    evop4=d
                                }
                                # output results

#undef a
#undef b
#undef c
#undef d
#undef rflg
#undef ang
#undef mut1
#undef mut2
.end

```

Figure 3.13 Model code for the Complex Butterfly

```

# model for Angle Block
.define elmod2
.numout 1
.numinp 1
.nummem 2
.*begin
#define angle VAR1
#define point IVAR2

integer x,z,i,y,ang(32)

genfig=.true.

data pi/3.1415926/
# stored twiddle factors and read in a cyclic manner
data (ang(i),i=1,8)/0,0,0,8,0,4,0,12/
data (ang(i),i=9,16)/0,2,0,10,0,6,0,14/
data (ang(i),i=17,24)/0,1,0,9,0,5,0,13/
data (ang(i),i=25,32)/0,3,0,11,0,7,0,15/

if (lasttim == -1)      # initialization
{
  angle=0
}
point=par1
if (sintim != lasttim)
{
  x=mod(sintim,32)      # read data in cyclic manner
  z=mod(sintim,point)
  if (x==0 & z==0)
  {
    y=0
  }
  if (z == 0)
  {
    y=y+1
  }

  # calculate the angle
  angle=ang(y)*2*pi/32  # evaluation
  evop1=angle           # output
}

#undef angle
#undef point

```

Figure 3.14 Model code for the Angle Block

Switches are necessary to increase the number of delays per channel so that it is possible to implement other pipelines of larger transform sizes. Both the Switch and the Butterfly adopt a distributed control and test methodology in order to minimize the system control overheads. The most attractive strategy to reduce the expense of IC testing is to

Figure 3.15 The simulated Splice output with a 2 KHz input source

Figure 3.16 The simulated Splice output with a 10 KHz input source

include circuitry on the chip to facilitate testing. As described earlier, one of the self-test techniques is based on the generation of pseudorandom test vectors as the stimuli of the circuit and it analyses the result at the output by using response compaction. This built-in self-test technique is implemented in both the Switch and the Butterfly. A Splice simulation of the pipelined FFT has been performed and the functionality of the model is verified. In the next chapter, the detailed circuit description of the FFT Switch is presented.

Chapter Four

Circuit Description of the FFT Switch

4.1 Introduction

The architectural considerations of the pipelined FFT described in chapter three clearly indicate the directions for the FFT Switch implementation. Before the detailed circuit description of FFT Switch is presented, a brief account of Computer Aided Design (CAD) tools, which are used in the design of the Switch, will be given. The design is implemented by using the CMOS Gate Array technology to achieve shorter design cycles and lower fabrication cost over a full custom design, which occurs at the expense of higher gate count and hence higher power consumption. Circuit description of the Switch will be partitioned into sections which include the delay element and commutator, control unit, input/output blocks and circuit for testability. Simulations of the FFT Switch will be presented in the following chapter.

4.2 Computer Aided Design Tools

4.2.1 Introduction

Since the traditional “breadboard” prototype techniques are impractical for complex I.C. designs, alternative design methodologies are required. The fact IC designers are facing the need for increasingly more complex designs and shorter design cycles further spells the importance of Computer Aided Design (CAD) tools. The introduction of the use of advanced CAD tools has vastly reduced design cycle time and improved the quality of products. By using such tools, designers are able to put together circuits and systems in a much quicker, more accurate and more convenient manner than ever before. CAD tools allow designers to capture the schematic of the circuit and simulate the design thoroughly. Extensive simulation helps shorten the total development time by reducing the number of prototype iterations. Designs remain in the iteration loop, which consists of schematic capture refinements and accurate circuit simulations, until they have fully met performance specifications. Finally, the design of the integrated circuit is expanded¹ and the netlist extracted. The layout of the circuit must be completed before the design is sent to the foundry. All CAD tools used in the FFT Switch design are supplied by Mentor Graphics and run on version 10 of the Aegis² Operating System.

4.2.2 Schematic Capture Tools

In order to perform hierarchical IC designs, a circuit editor (NETED³) and a symbol editor (SYMED²) is used. These tools allow a designer to create a schematic

1. A process which maps a hierarchical design of a circuit into a “one-level” design in order to extract the netlist of the circuit.

2. Trademark of HP/Apollo.

3. Trademarks of the Mentor Graphics.

sheet at a lower hierarchical level by using NETED, which can be represented by a symbol block, generated in SYMED, for a schematic sheet at a higher hierarchical level. Consequently, the complexity of the schematic at high hierarchical levels will be reduced and visual verification by the designer can be simplified. The component library used throughout the design of the FFT Switch is based on the LSI Logic LMA 9K library. It is a 1.5 μ m HCMOS Gate Array technology and the available device family has a maximum gate utilization of 15,000.

4.2.3 Simulation Tools

To analyse and verify an electronic design created in the schematic capture tools, a logic simulator (QUICKSIM¹) and a fault simulator (QUICKFAULT¹) are used. QUICKSIM is an interactive logic simulator that allows designers to verify the functionality of designs produced with NETED and SYMED. With the logic simulator, designers are allowed to supply stimulus to the designs, run the simulation, analyse the results, and modify the design based on those results. Designers can then re-apply stimulus to the designs and start the cycle again for further simulations. QUICKFAULT is a fault simulator which models manufacturing defects by using the stuck-at fault models. The design is first of all simulated under fault-free condition. Then faults are inserted one at a time before the design is simulated again using the same set of test vectors. Results of the fault and fault-free simulations are compared and subsequently the simulator predicts the fault coverage of the test vectors set.

1. Trademarks of Mentor Graphics

4.3 Design Implementation

4.3.1 Design of the Delay Circuit and Commutator blocks

The two delay blocks (Z^{-k} blocks) and the three commutators in Figure 4.1 collectively occupy the majority of circuit area on the FFT Switch. The delay elements are required in the pipelined FFT for the data realignment. Realignment is necessary for

Figure 4.1 A block diagram of a FFT Switch

the data reordering process and is performed before and after the reordering. All delay elements in the Z^{-k} blocks are implemented by means of shift registers. The delay path in each Z^{-k} block is eight delay elements long. For a 32 bit wide data path, an array of 32×8 shift registers is needed. A Random Access Memory (RAM) structure, which can be implemented as the delay element by simply displacing the read and write addresses with a constant value, probably consumes less power than the shift register array.

Figure 4.2 A two bits delay line structure

data on channel A does not pass through any delay element at all, data on channel B propagates through both Z^{-k} blocks, hence a total of sixteen clock cycles delay will be imposed.

In Figure 4.2, the delay line structure is controlled by four separate clock cycles, namely CP4a, CP4b, CP4c and CP4d. All clock cycles originate from the “Control Unit”. The four clock cycles approach is employed so that it is possible to distribute the load onto several clock cycle generators and to disable parts of the shift register array by disabling the clock inputs.

Data routing is accomplished by using three separate commutators in each FFT Switch. Their locations are shown in the block diagram of Figure 4.1. The select inputs of all three commutators are also controlled by signals originated from the Control Unit. Each commutator is made up of four sets of multiplexers each set consisting of sixteen two-to-one multiplexers. In order to rout two complex data paths through the commutator, the four sets of multiplexers are arranged as shown in Figure 4.4. Two sets of multiplexers are responsible for routing the real parts of the complex data and the other two sets are associated with the imaginary parts.

Data reordering is done by using either the second or third commutator of the Principle Switch. If a particular commutator is responsible for data reordering, the select inputs for the commutator will be controlled by a periodic square wave called the Reordering Frequency, also generated in the Control Unit.

Figure 4.4 A Commutator with four sets of multiplexers

4.3.2 Control Unit Design

4.3.2.1 Introduction

The Control Unit generates all the necessary control signals to ensure the correct operations of the Switch. External information required by the Switch is loaded onto the module only once prior to the start of the operation and the internal control signals are generated accordingly. Almost all control signals required by the Switch are generated within the Switch in order to minimize control overheads at the system level.

The Control Unit is divided into four parts which include: the clock cycles generator, synchronization pulse logic, cascadeability control and information latch. A description of each part of the Control Unit follows.

4.3.2.2 The Clock Cycles Generator

The function of the generator is to scale down the external master clock and provide the necessary clock cycles to all parts of the Switch. The initial specification of the master clock is 50MHz with a 50% duty cycle. This specification is used throughout the design and simulation processes of the Switch. Altogether, there are four clock prescalers within the generator that are responsible for the four internal clock cycles namely CP4a, CP4b, CP4c and CP4d. Each prescaler, shown in Figure 4.5, divides the master clock by four to provide a 12.5MHz clock cycle. As mentioned in section 4.3.1, each bit of delay line in Z^{-k} block is partitioned into four sets of delay elements and these are clocked by the four separate CP4 cycles. Except for CP4a, the other three clock cycles are solely for clocking the inputs of the delay elements in the Z^{-k} blocks. Cp4a is responsible for the first set of registers, CP4b is responsible for the second set and so on. Each of the four clock cycles could be disabled to reduce the overall power

Figure 4.5 A Clock Prescaler

consumption of the FFT processor. For example, Figure 4.6 shows a block diagram of a 32 point radix-2 pipelined FFT processor. The figure indicates that less than half of the delay elements are operating simultaneously and the rest are disabled. This example emphasizes that it is essential to be able to disable the unused delay elements in order to reduce the power consumption of the processor.

Figure 4.6 Delay elements of a 32 point pipelined FFT processor

4.3.2.3 The Sync Pulse Logic (SPL)

As mentioned in chapter three, the synchronization between data and reordering operations is ensured by using a Sync pulse which is generated at the system control of the FFT processor. The Sync pulse propagates through the Switch along with the data

stream; however both “travel” in two independent channels. The reordering process is activated if the incoming data is valid and the valid data is identified by using a logic low signal on the Sync pulse. The negative going edge of the Sync pulse initiates the reordering process and the process continues indefinitely until the pulse returns to a logic high, thus terminating the process.

The function of the SPL is to provide exactly the same delays for the Sync pulse as for the data stream when it passes through the Switch. The delays in the SPL are also implemented by means of a shift register. Each switch mode processes the Sync pulse in a different fashion except the Output switch mode which handles the pulse similarly to the Right switch mode. Once the Principle Switch is identified in a set of cascaded Switches, additional delays are added onto the Sync pulse to compensate for the delays imposed by the Complex Butterfly to the data stream. This is necessary since the Sync pulse does not propagate through the Complex Butterfly. Sync pulse and data stream will be aligned such that both the start of the valid data and the negative edge of the Sync pulse will appear at the output of their respective channels of the Switch stage within the same clock cycle. The output of the Sync pulse (Syout) is directly fed into the Sync pulse input (Syin) of the following Switch even if the Butterfly is located between the Switches.

4.3.2.4 The Cascadeability Control

In order for the Switch to be cascadeable, it must be capable of operating in all five modes as explained in section 3.2.2. The FFT Switch responsible for the reordering of data is called the Principle Switch. It can be implemented as either the Middle or End Switch mode (as shown in Figure 3.2). Data reordering is accomplished by using a

commutator with its select input controlled by the Reordering Frequency¹. Since a Principle Switch is designed such that it can operate at any position in the pipeline, it must be capable of producing a spectrum of frequencies required for reordering. Figure 4.7 shows the reordering of the data through the Switches in a 32-point radix-2 pipelined FFT. It shows that the Reordering Frequency of the Principle Switch in stage 4 is half the frequency of the Switch in stage 3, which in turn is half the frequency of the Switch in stage 2, and so on. A binary up-counter with synchronized clear is chosen to generate the Reordering Frequencies. A twelve bit counter, with a tap at every stage, is used to generate all of the Reordering Frequencies required to support a single pipelined FFT processor of up to 8192 points. The twelve tapped frequencies are fed to a twelve-to-one multiplexer which is controlled by four Position signals. The Position signals are used to indicate the stage of the Switch within the pipeline. Subsequently, these signals will determine the correct Reordering Frequency for this particular stage.

The five Switch modes are selected by three single bit Control signals. Once a Switch is assigned to a particular switch mode, it remains in the same mode until another set of Control signals is loaded.

4.3.2.5 Information Latch

The information latch block in the Control Unit is used to store and decode the Position and Control signals. Both Position and Control signals are loaded through the information bus. For a 32-point pipelined FFT processor, all Switches and Butterflies on the same structure share the same bus. The Butterflies utilize the bus for the loading

1. The Reordering Frequency is essentially a square wave which determines that the data either passes straight through or crisscrosses the commutator as shown in Figure 4.7.

Figure 4.7 Operation of the 32 point pipelined FFT

of twiddle factors. Information is sent through the bus to a Switch only once before the start. In order for the Switch to retain the loaded information, a series of latches, controlled by the Chip Select (CS) and Load Enable (SLOAD) inputs, is used. Information is loaded onto the Switch only when both CS and SLOAD inputs are at a logic low. Under this arrangement, SLOAD can be implemented as a broadcast signal for the FFT processor provided that a separate CS is implemented for each Switch and Butterfly. Once the information is latched, it is decoded and transferred to other parts of the Switch.

4.3.3 Input and Output

The input and output blocks are implemented such that data is clocked in and out of the Switch sequentially. Furthermore, a Linear Feedback Shift Register (LFSR) circuit is implemented in the input block such that it generates the pseudorandom test vectors necessary for self-test operation. The output block, it includes an LFSR based Compacter circuit and predetermined signatures which are required to verify the results of the self-test.

4.3.4 Design for Testability

4.3.4.1 Built-in Self-test

With the rapidly increasing size of the VLSI designs, traditional methods of circuit testing become either impractical or very expensive. The built-in self-test (BIST) techniques are more popular than ever. They provide an efficient way of testing integrated circuits with large gate count and high pin out. Although testing is not required in both off-line and at-speed circumstances, the FFT Switch employs both

techniques for implementing the BIST while remaining compatible with the Complex Butterfly on the system level.

The most popular techniques for built-in self-test involve providing a sequence of pseudorandom numbers on board as the stimulus and comparing the output of the circuit with the predetermined result (called a signature) at a fixed interval. One of the techniques involves using a read only memory (ROM) structure large enough to store a set of predetermined vectors and using them as input stimulus for testing. However, such an approach is very time consuming since the effort necessary to determine the required set of test vectors to achieve the best fault coverage is substantial. Furthermore, a circuit with a large number of inputs will require a large ROM structure to hold the necessary test set. By far the most popular method of generating test vectors in-circuit is to use an in-circuit pseudorandom sequence generator. The name pseudorandom reflects the fact that the output sequence is cyclic and not truly random. One such generator is called the Linear Feedback Shift Register (LFSR). An LFSR is a finite state machine consisting of interconnections of essentially two elements, namely memory element and modulo-2 summer. The memory element and modulo-2 summer can be implemented as a shift register and exclusive-or gate respectively. The circuit overhead to produce test vector for a circuit with a large number of inputs is considerably less for LFSR than when using an ROM structure. The maximum period for an n -stage LFSR is $2^n - 1$ clock cycles (no all zero vector) and the generating function corresponding to a maximum period LFSR is called a primitive polynomial.

For the FFT Switch, a sixteen bit LFSR, shown in Figure 4.8, is employed to generate two primitive polynomial pseudorandom sequences. It is located at the input

block of the Switch. The two polynomials shown in Equations (4.1) and (4.2) are chosen to generate the pseudorandom sequences.

$$p = x^{16} + x^{14} + x^{13} + x^{11} + 1 \quad (4.1)$$

$$p' = x^{16} + x^5 + x^3 + x^2 + 1 \quad (4.2)$$

These two polynomials are published in the literature [28][29] along with a number of other polynomials for different bit sizes. Although both polynomials are primitive, the pseudorandom sequence of the Equation (4.1) is in reverse order of the Equation (4.2).

A sixteen bit Multiple Input Pseudorandom Compacter (MIPC) along with two signatures is located at the output block of the Switch. The MIPC is based on an LFSR architecture with an additional exclusive-or gate at every stage. The MIPC structure is shown in Figure 4.9. The compacted vector is clocked sequentially into the signature analyser and compared with the predetermined signatures. A single output pin called GO_NOGO indicates the status of the circuit. The GO_NOGO output is at the logic high level at the beginning of the self-test procedure and it will toggle to a logic low for a CP4 clock cycle (80ns) when a compacted vector matches the signature. Since there are two pseudorandom sequences, two signatures are chosen in each test cycle with one located at the middle and the other at the end of each cycle. When the compacted vector from the first sequence matches the first signature located at the middle of the test cycle, all registers in the Switch are reset and ready for the second sequence generated from the same LFSR structure. Once the second signature is matched, all registers will be reset again and the self-test procedure will repeat itself by using the first sequence. Hence, if the Switch is held in the self-test mode, a periodic waveform will be present at the GO_NOGO output provided that the module passes the self-test routine. The

Figure 4.8 A Linear Feedback Shift Register

Figure 4.9 A Multiple Input Pseudorandom Compactor

operation of a two-signature scheme reduces the frequency of aliasing from $1/2^n$ for a single signature to $1/2^{2n}$ for two signatures [30], thus increasing fault coverage. Aliasing occurs when the circuit under test produces the same compacted vectors even when it contains faults. The additional hardware cost to implement such a scheme is small, and it has the additional benefit of doubling the test length for pseudo-exhaustive testing. However, due to the memory constraint¹ of the available IC tester, it has not been possible to apply the full length of the pseudorandom sequence to the FFT Switch. Instead, the LFSR will produce a sequence for each polynomial of just over 2000 clock cycles.

4.3.4.2 Interconnectivity Test

Apart from the implementation of LFSR and MIPC as a self-test technique to cover the core of the Switch, additional circuitry is added for both the Switch and the Butterfly to test the I/O buffers and more importantly, the interconnectivity of cascaded modules. The method is based on the usage of a parametric NAND tree structure present at the input buffer of every ASIC. The NAND tree structure is required by the foundry for the input threshold voltage measurements. Figure 4.10 shows the input buffer incorporated with a NAND gate for the parametric NAND tree². All input pins are held at logic high at the beginning of the test. Then the inputs are toggled to logic low sequentially. The output of the NAND tree is sent to the TESTOUT pin and it is expected to toggle continuously if there are no disconnections between the modules. It is possible to test the connectivity of a series of ASICs by connecting the NAND trees in series and testing each ASIC in sequence. Additional multiplexers are implemented

1. The I.C. tester is able to store a set of simulated results of up to 4000 clock cycles.

2. A more detailed explanation of the parametric NAND tree can be found in Appendix C.

PI : Parametric tree input from the NAND gate output of the previous stage
 PO : Parametric tree output to the following stage

Figure 4.10 An input buffer with a NAND gate for a parametric NAND tree.

at the output of each module so that any TESTOUT signal is able to bypass a number of chip modules providing the required controllability and observability. The test commences with the first module on the pipelined structure, then the second module, and so on down the pipeline.

4.4 Summary

In this chapter, a circuit description of the FFT Switch, divided into separate sections based upon the functionality of each block has been presented. The design of each functional block was carried out so that the technological, performance and system constraints of the FFT Switch were reflected. Simulation results to verify the implementation of the FFT Switch will be given in the next chapter.

Chapter Five

Simulation and Test Results

5.1 Introduction

In this chapter, simulations and test results of the FFT Switch are presented to verify the functionality of the module. The Switch is simulated either as a stand-alone module or a two-module cascaded structure. The stand-alone module represents just one Switch in the simulation and the two-module cascaded structure represents three different combinations, namely “Switch-Switch”, “Switch-Butterfly” and “Butterfly-Switch” where all combinations include at least one FFT Switch. All simulations are performed under the worst case, the best case and the nominal case commercial conditions. The worst case commercial condition simulations are done under conditions where the power supply is assumed to be 4.5V and the ambient temperature 70°C; the best case commercial condition assumes a 5.25V power supply and 0°C ambient temperature; and the nominal case corresponds to 5.0V and 20°C.

5.2 Simulations of the FFT Switch

5.2.1 Stand-alone module

5.2.1.1 Introduction

The simulations of the FFT Switch are divided into two major sections, namely functional and non-functional verifications. For functional simulations, emphasis will be placed on the various switch modes and different number of delays by changing the combinations of Positional and Control signals. Further simulations are also done in the self-test mode. As for non-functional simulations, these include fault simulation, toggle coverage simulation and post layout verifications. The post layout verifications are required by the foundry and will be explained in a later section.

5.2.1.2 Functional Test

As mentioned in Chapter 3 and 4, in order for the Switch to operate as a generic building block for a pipelined FFT processor, architecturally it must be capable of implementing five different switch modes, namely the Middle End, Right, Left and Output switch modes. Besides the Middle switch mode, Switches which operate in the other four modes do not function as a single module in an FFT processor and they are required to cascade with other Switches during normal operations in order to provide the necessary number of delays between Complex Butterflies. Nevertheless, to verify the functionality of the different switch modes, all five modes are individually simulated and the data stream and Sync pulse outputs are compared with the predicted results.

Example 1: The following two simulations correspond to Switch A and Switch D of a 32 point FFT processor (shown in Figure 5.1) where Switches A and D represent the last and the first Switch of the pipeline respectively. All four switches operate in the Middle switch mode.

Figure 5.1 A 32 point pipelined FFT processor

Figure 5.2 shows the simulation results of Switch A. The input stimulus is selected so that results at the output can be easily interpreted. Besides the two clock cycles delay due to the input and output blocks, the simulation shows that the Z^{-k} blocks impose only one clock cycle delay on each data stream, hence the location of the Switch can be verified as the last FFT Switch in the pipelined FFT processor. The Reordering Frequency, which controls the select input of the middle commutator in the Switch, is identified as the signal called “/A91/Z6/swi2” in Figure 5.2. The Reordering Frequency shown in the plot represents the highest frequency that can be provided by the FFT Switch. Depending upon the location of the FFT Switch relative to the pipeline, the Reordering Frequency may alter. This is required to ensure that the data reorders correctly throughout the FFT processor. As the simulation results have indicated, apart from the clock pulse CP4a, the other three internal clock pulses (CP4b, CP4c and CP4d)

Figure 5.2 Simulation result of Switch A in Example 1

are shown to be disabled. Since the simulation represents the last FFT Switch, only the first set of shift registers in the Z^{-k} blocks are utilized. The remaining registers are disabled in order to reduce the power consumption of the Switch. Initially, the valid data at the input is three CP4 clock cycles behind the Sync pulse. These three clock cycles are imposed on the data by the Complex Butterfly which is located just before the Switch A. As the simulation shows, extra delays are placed on the Sync pulse such that the data stream and the pulse are once again aligned at the output.

Figure 5.3 shows the simulation of Switch D which also operates in the Middle switch mode. The Positional signals of the Switch for this simulation have been altered in order to change the delay in the Z^{-k} blocks and the Reorder Frequency for a different location on the pipeline. The number of delayed clock cycles imposed on the data stream shown in Figure 5.3 is increased to eight clock cycles. This is the maximum number of delayed clock cycles provided by a single Z^{-k} blocks hence there is no clock pulse being disabled since both Z^{-k} blocks are fully utilized. The Reordering Frequency of Figure 5.3 is reduced eight-fold when compared with the same signal in Figure 5.2. Results for Switches B and C can be found in Appendix A in order to complete all simulations of FFT Switches in a 32 point pipelined FFT processor.

Example 2: Apart from the Middle switch mode, the module operating in the End switch mode is also considered a Principle Switch which is responsible for data reordering. An example of a Switch which operates in the End switch mode is shown in Figure 5.4. The two-Switch cascaded structure includes an End switch mode and an Output switch mode. In Figure 5.5¹, the simulation result of the End switch mode shows

1. The simulation diagram is drawn according to the actual simulated result.

Figure 5.3 Simulation of $\hat{\theta}_n$

Result of Switch D in Example 1

Figure 5.4 A two Switches cascaded structure

the input data on channel B which is delayed and has crisscrossed to output A'. No delay is applied on input data on channel A, however, part of the data will be crisscrossed to output B'. In order to direct the overall output of the cascaded structure so that there is no data misalignment and misplacement at the outputs A'' and B'', a second FFT Switch is used. The function of the second Switch is to apply the correct amount of delay on the data from output B' so that both outputs A'' and B'' will be aligned once again. Figure 5.6 shows the simulation of the second Switch and the Sync pulse defines the valid data at the output of the cascaded structure.

Example 3: Another example of a Switch which operates in the End switch mode is shown in Figure 3.4 and again in Figure 5.7. As the data passes through the inputs of the cascaded Switches, the data on channel B will be delayed by sixteen clock cycles at the output of the first Switch which operates in Left switch mode. An additional sixteen clock cycles delay is imposed on channel B by the second Switch. Figure 5.8 shows the simulation results at the output of the second Switch which is operated in the End Switch mode. The negative going edge of the Sync pulse is aligned

* 0001-0016 denotes the samples 0001, 0002, 0003, ..., 0016 and each sample represents by one clock cycle.

Figure 5.5 Simulation result of End switch mode in Example 2

0000	0017-0032	0081-0096	AI(15:0)
0000	0049-0064	0113-0128	BI(15:0)
0001-0016	0065-0080	0000	CI(15:0)
0033-0048	0097-0112	0000	DI(15:0)
			SYIN
			/A91/Z6/swi2
			/A91/Z6/swi3
0000	0017-0032	0081-0096	AO(15:0)
0000	0049-0064	0113-0128	BO(15:0)
0000	0001-0016	0065-0080	CO(15:0)
0000	0033-0048	0097-0112	DO(15:0)
			SYOUT

Figure 5.6 Simulation result of Output switch mode in Example 2

Figure 5.7 A four Switches cascaded structure

with the start of valid data of channel A. The simulation results show no clock pulse in the Switch which is disabled since both Z^k blocks are fully utilized. The simulations for Left, Right and Output switch modes for the same cascaded structure can be found in Appendix B.

Combining the simulations of the previous examples, including the four Middle switch mode modules, the two-Switch cascaded structure and the four-Switch cascaded structure, operations for Switches in all six stages of a 128 point pipelined FFT processor are demonstrated.

Example 4: Whereas both the Middle and the End switch modes are handled as Principle Switches, Switches which operate in the other three modes are used solely for routing and to provide extra clock delays for data before and after the reordering process. The difference between the operations of the Left and Right switch mode can be demonstrated by using the same data input and Sync pulse in both simulations.

AI(15:0)	0001-0016				0017-0032	0033-0048	0049-0064	0000	
BI(15:0)	0065-0080				0081-0096	0097-0112	0113-0128	0000	
CI(15:0)	0000				0129-0144	0145-0160	0161-0176	0177-0192	0000
DI(15:0)	0000				0193-0208	0209-0224	0225-0240	0241-0256	0000
SYIN									
/A91/Z6/swi2									
/A91/Z6/swi3									
AO(15:0)	0000				0033-0048	0049-0064	0161-0176	0177-0192	
BO(15:0)	0000				0097-0112	0113-0128	0225-0240	0241-0256	
CO(15:0)	0001-0016				0017-0032	0129-0144	0145-0160	0000	
DO(15:0)	0065-0080				0081-0096	0193-0208	0209-0224	0000	
SYOUT									

Figure 5.8 Simulation result of End switch mode in Example 3

Figure 5.9 and Figure 5.10 show the simulations of Left and Right switch modes respectively. The only major difference between the operations of the two switch modes is the response of the Sync pulse. It is shown in Figure 5.10 that the Sync pulse is delayed by sixteen clock cycles. This is necessary to ensure the Sync pulse is aligned with the start of valid data in the pipeline. The same input data as the Left and Right Switch modes is used to simulate a Switch which operates in the Output mode in order to distinguish the differences between these modes. Simulation results of the Output switch mode in Figure 5.11 shows that in comparison to the previous two switch modes, the data outputs of the two channels are crisscrossed. However, the response of the Sync pulse is the same as that of the Right switch mode.

Example 5: Whereas the five switch modes mentioned above are supposed to operate under on-line processes, the self-test mode is an off-line operation. Figure 5.12 shows the simulated response of the GO_NOGO output when the Switch is operating in the self-test mode. The response of the GO_NOGO output is a periodic waveform and along with the 80ns logic low pulses, it indicates that the self-test is operating successfully.

5.2.1.3 Non-functional Simulations

Non-functional simulations of the FFT Switch include a fault simulation, a toggle coverage test and foundry-required post layout verification tests. The fault simulation is done by using a fault simulator called QUICKFAULT. A stuck-at fault coverage of 84% is recorded using the self-test procedure. As for the toggle coverage, 99% coverage is achieved by using the self-test procedures and a functional test routine. Additional simulations are required by the foundry for three other verifications. These

Figure 5.3 Simulation of the

Figure 5.10 Simulation

2 of/de 2

PM-1 2H/24" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO Z39.50-1968

PRECISIONTM RESOLUTION TARGETS

0 Simulation result of Right switch mode in Example 4

															CP
															/A91/26/16.4
															/A91/26/16.4B
															/A91/26/16.4C
															/A91/26/16.4D
															AI (15 0)
															BI (15 0)
															CI (15 0)
															DI (15 0)
															SYIN
															/A91/26/16.4E
															AO (15 0)
															BO (15 0)
															CO (15 0)
															DO (15 0)
															SYOUT
2400.0	2600.0	2800.0	3000.0	3200.0	3400.0	3600.0	3800.0	4000.0	4200.0	4400.0	4600.0	4800.0	5000.0	5200.0	

ation result of Right switch mode in Example 4

Figure 5.11 Simulation result of Output switch mode in Example 4

put switch mode in Example 4

+			GO_NOGO
000.0	140000.0	160000.0

/A91/Z6/cp4a
+				GO_NOGO
41100.0	41200.0	41300.0	41400.0	

Output.
 magnified GO_NOGO pulse

are necessary in order to test the tri-state buffers, and to measure the circuit for threshold voltages and current from the power supply. The test pattern for these three simulations are organized depending on the number of the input buffers and their nature. A more detailed explanation of these simulations can be found in Appendix C.

5.2.2 Cascaded Structure Simulations

After all switch modes are simulated as stand-alone modules, they are cascaded with either a Switch or a Butterfly for further simulations. The main purpose of these simulations is to verify that the Switch design is capable of operating as part of a cascaded structure. Three different combinations of cascaded structures can be included in an FFT Switch, namely “Switch-Switch”, “Switch-Butterfly” and “Butterfly-Switch”. These are shown in Figure 5.13.

The “Switch-Switch” combination corresponds to two cascaded Switches. Table 5.1 shows all the possibilities of the “Switch-Switch” combination for various switch modes. The input stimulus of the cascaded structure is loaded onto the first Switch and the result of the complete simulation is recorded at the output of the second Switch. Although Table 5.1 shows all the possibilities of the “Switch-Switch” combination, the majority of them are not applicable to the pipelined FFT processor.

Example 6: Figure 5.14 shows the simulation of two cascaded switches where both operate in the Left switch mode. As explained in the previous chapters, this is used primarily to provide the necessary delay before the reordering process. As the simulation shows, since each Switch imposes a sixteen clock cycles delay, data on channel A is thirty-two clock cycles in advance of data on channel B. No delay is

Figure 5.13 Three two modules combinations include a Switch

channel A is thirty-two clock cycles in advance of data on channel B. No delay is imposed on the Sync pulse, hence it is still three clock cycles in advance of the valid data on channel A.

Example 7: Figure 5.15 shows another "Switch-Switch" simulation result where one Switch operates in End switch mode and the other in Output mode. This is the same combination as in Example 2 of the section 5.2.1.2. When the two simulations are compared, both yield the same output result. The Figure shows that the data stream in both channels and the Sync pulse at the output of the second module are aligned. This is important since the Output mode must represent the last Switch when a number of Switches are cascaded. In order to pass data onto the next Butterfly in the pipeline, the

First Second	Right	Left	Output	End	Middle
Right	√	x	x	√	x
Left	x	√	x	x	x
Output	√	x	x	√	x
End	x	√	x	x	x
Middle	x	x	x	x	x

☒ Applicable

☐ Not Applicable

Table 5.1 A table of all possibilities for “Switch-Switch” combination

data stream and Sync pulse must be synchronized and routed correctly at the output of the cascaded structure.

For the “Switch-Butterfly” and “Butterfly-Switch” combinations, shown in Figure 5.13, the FFT Switch is cascaded either in front of or after the Complex Butterfly. Table 5.2 shows all the possibilities allowed for these two combinations. There are five different combinations that are applicable for meaningful simulations out of the ten possible combinations..

AI(15:0)	0001-0016	0017-0032	0033-0048	0049-0064	0000
BI(15:0)	0065-0080	0081-0096	0097-0112	0113-0128	0000
CI(15:0)	0129-0144	0145-0160	0161-0176	0177-0192	0000
DI(15:0)	0193-0208	0209-0224	0225-0240	0241-0256	0000
SYIN					
AO(15:0)	0001-0016	0017-0032	0033-0048	0049-0064	0000
BO(15:0)	0065-0080	0081-0096	0097-0112	0113-0128	0000
CO(15:0)	0000	0129-0144	0145-0160	0161-0176	0177-0192
DO(15:0)	0000	0193-0208	0209-0224	0225-0240	0241-0256
SYOUT					

Figure 5.14 Simulation result of Example 6

AI(15:0)	0001-0016	0017-0032	0033-0048	0049-0064	0000
BI(15:0)	0065-0080	0081-0096	0097-0112	0113-0128	0000
CI(15:0)	0129-0144	0145-0160	0161-0176	0177-0192	0000
DI(15:0)	0193-0208	0209-0224	0225-0240	0241-0256	0000
SYIN					
AO(15:0)	0000	0001-0016	0129-0144	0033-0048	0161-0176
BO(15:0)	0000	0065-0080	0193-0208	0097-0112	0225-0240
CO(15:0)	0000	0017-0032	0145-0160	0049-0064	0177-0192
DO(15:0)	0000	0081-0096	0209-0224	0113-0128	0241-0256
SYOUT					

Figure 5.15 Simulation result of Example 7

	Butterfly	
	First	Second
Switch Modes	Left	√
	Right	x
	Middle	√
	End	√
	Output	x

√

 Applicable

x

 Not Applicable

Table 5.2 "Switch-Butterfly" and "Butterfly-Switch" combinations

5.3 Prototype for the FFT Switch

The FFT Switch prototypes were implemented by using 1.5 μ m HCMOS gate array technology from LSI Logic. The total gate count of the Switch is 9262 gates distributed over 6134 nets. The 9.98mm² die is housed in a 180-pin ceramic package. Post-layout simulations were completed with the collaboration of the Alberta Microelectronics Centre. A die photo of the FFT Switch is shown in Figure 5.16 and the pin layout and bonding diagram are located in Appendices D and E respectively.

5.4 Prototype Tests

Ten prototype FFT Switch ASICs were fabricated and functionally tested at the foundry. All ten chips passed the functional test carried out at 10MHz. These prototypes were further tested at CRC in the self-test mode. A test board was constructed so that it

Figure 5.16 A die photo of the FFT Switch

was possible to examine just the self-test mode for both the FFT Switch and Complex Butterfly. A signal generator was equipped to produce the master clock frequencies and a frequency counter was used to observe the signal at the GO_NOGO output. All ten prototypes operated satisfactorily at the master clock frequency of 50MHz which met the system requirement. The frequency of the signal generator was then increased until the logic low pulse disappeared from the GO_NOGO output in order to find out the maximum operating frequency. The frequency just before the pulse vanished from the

output was recorded along with the power consumption of the Switch at that frequency. Table 5.3 shows the test results of the ten prototypes which were operated with a power supply of 5V. Further testing was completed at 3.3V to determine the response of the prototype at a lower voltage. The primary reason to reduce the voltage supply was to decrease the power consumption of the Switch and the FFT Processor on the whole. Table 5.4 shows the test results of the Switch prototypes at 3.3V.

Prototypes	Maximum Frequency Master Clock (MHz)	Current (mA)
1	150.8	350
2	150.2	340
3	149.8	340
4	151.9	350
5	150.4	340
6	149.4	340
7	150.5	340
8	150.2	340
9	152.3	350
10	150.1	340

Table 5.3 Prototype results at self-test mode for 5.0V power supply

The Switch prototypes were found to operate in the ranges of 149 to 152MHz at 5V and 97 to 101MHz at 3.3V. The power dissipation at 150MHz and 5V was 1.70W and at 100MHz and 3.3V was 0.43W.

Prototypes	Maximum Frequency Master Clock (MHz)	Current (mA)
1	97.4	130
2	100.1	130
3	100.6	130
4	100.1	130
5	100.1	130
6	100.6	130
7	99.6	130
8	100.2	130
9	101.0	130
10	96.7	130

Table 5.4 Prototype results at self-test mode for 3.3V power supply

At the time of writing, a 32 point pipelined FFT processor board is being built. Both the FFT Switch and the Complex Butterfly will be tested together on this processor board.

5.5 Summary

The simulations of the FFT Switch were performed under the best, the nominal and the worst case commercial conditions. All functions of the Switch operated satisfactorily under all of these conditions and met the system throughput requirement of 25 Msample/sec per channel. Ten prototypes were fabricated by LSI Logic and the self-test procedure was performed on all ten prototypes initially with the system

requirement of 50MHz as the frequency of the master clock. Further testing was completed to investigate the maximum operational master clock frequency of each prototype in self-test mode.

Chapter Six

Conclusions

6.1 Summary

The increasing demand for low-cost and simple earth stations for satellite communications has signalled the need for more efficient techniques to tackle the issue. One such technique involves the use of frequency-division multiple access (FDMA) as the up-link to satellites and time-division multiple access (TDMA) as the down-link to ground stations. The required transformation of FDMA to TDMA is executed on-board the satellite. The multi-carrier demodulator (MCD) is responsible for the digital signal processing and part of the MCD is a demultiplexer. The demultiplexer can be implemented by using a digital polyphase filter bank in cascade with an FFT processor. A pipelined FFT processor, which can be implemented by using two generic building blocks, namely an FFT Switch and a Complex Butterfly, has been designed such that it is capable of supporting throughput of tens of millions of samples per second. The

implementation of such an FFT Switch for the radix-2 algorithm is described in this thesis.

The in-place computation and regular structure of the FFT algorithm permits the implementation of either a forward or an inverse pipelined FFT by means of the same two generic building blocks. By mapping the FFT algorithm onto a one dimensional systolic array, the pipelined processor is able to offer speed improvement over the traditional Von-Neumann architecture. A trade-off between processing speed, flexibility and external control/hardware support is demonstrated by some processor architectures. Though a number of proposed FFT processors can operate at a very high rate of throughput, they lack the flexibility to implement the wide range of transform sizes and the high data throughputs offered by this chip set in a parallel pipelined architecture. Furthermore, the chip set employs a distributed control methodology in order to minimize the system control overheads, thus reducing external circuitry at the system level. Although IC testing is essential for circuit verification, there has been no mention in the literature on pipelined FFT architectures which address the issue of testing by means of built-in self-test technique. The incorporation of BIST enables the FFT Switch to operate off-line testing with minimum external test equipment.

The FFT Switch is fabricated in 1.5 μ m HCMOS gate array technology. Simulations have shown that the FFT Switch is able to support the data throughput of 25 Msamples/sec per channel, which represents the minimum system requirement. It is possible to increase the processing power of a single pipeline through the use of a parallel pipelined FFT configuration, which reduces the processing time at the expense of extra hardware.

6.2 Future Work

As the density of integrated circuits grows and the geometry of transistors decreases, the propagation delay due to packaging and interconnects between modules is becoming a significant portion of total signal delay. As mentioned briefly in Chapter 2, the multichip module (MCM) offers a considerable improvement in system level performance due to advances in packaging techniques. By placing several tested dies on a substrate and utilizing the substrate as interconnects between these dies, it is possible to reduce the overall propagation delay of the system.

During the design of the FFT Switch, the possibility of implementing an FFT processor MCM was always a consideration on our part. Attempts have been made to address architecturally a number of issues such as testability, power consumption, number of pin out and so on. Although the idea of MCM is not new, due to technological constraints, it is only recently that a great deal of interests and activity has been directed into this area. Moreover, no reliable and efficient CAD tools have been available on the market thus far and the technology has yet to mature.

Through implementation as a multichip module (a 32 point processor consists of four FFT Switches and five Complex Butterflies, a total of over 100,000 gates), the FFT processor may assist in identifying some areas where further investigations and refinements are required.

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Appendix A :

Simulations for Switch B and Switch C in Example 1

Simulation result of Switch E in Example 1

Simulation result of Switch C in Example 1

Appendix B :

Simulations for Left, Right and Output Switch modes in Example 3

0001-0016	0017-0032	0033-0048	0049-0064	0000	AI(15:0)
0065-0080	0081-0096	0097-0112	0113-0128	0000	BI(15:0)
0129-0144	0145-0160	0161-0176	0177-0192	0000	CI(15:0)
0193-0208	0209-0224	0225-0240	0241-0256	0000	DI(15:0)
					SYIN
					/A91/Z6/swi2
					/A91/Z6/swi3
0001-0016	0017-0032	0033-0048	0049-0064	0000	AO(15:0)
0065-0080	0081-0096	0097-0112	0113-0128	0000	BO(15:0)
0000	0129-0144	0145-0160	0161-0176	0177-0192	CO(15:0)
0000	0193-0208	0209-0224	0225-0240	0241-0256	DO(15:0)
					SYOUT

Figure B.1 The simulation result of the Left Switch mode.

0000	0033-0048	0049-0064	0161-0176	0177-0192	AI(15:0)
0000	0097-0112	0113-0128	0225-0240	0241-0256	BI(15:0)
0001-0016	0129-0144	0145-0160	0000		CI(15:0)
0065-0080	0081-0096	0193-0208	0209-0224	0000	DI(15:0)
					SYN
					/A91/Z6/swi2
					/A91/Z6/swi3
0000	0033-0048	0049-0064	0161-0176	0177-0192	AO(15:0)
0000	0097-0112	0113-0128	0225-0240	0241-0256	BO(15:0)
0000	0001-0016	0017-0032	0129-0144	0145-0160	CO(15:0)
0000	0065-0080	0081-0096	0193-0208	0209-0224	DO(15:0)
					SYOUT

Figure B.2 The simulation result of theRight Switch mode.

0000	0033-0048	0049-0064	0161-0176	0177-0192	AI(15:0)
0000	0097-0112	0113-0128	0225-0240	0241-0256	BI(15:0)
0001-0016	0017-0032	0129-0144	0145-0160	0000	CI(15:0)
0065-0080	0081-0096	0193-0208	0209-0224	0000	DI(15:0)
					SYIN
					/A91/Z6/swi2
					/A91/Z6/swi3
0000	0033-0048	0049-0064	0161-0176	0177-0192	AO(15:0)
0000	0097-0112	0113-0128	0225-0240	0241-0256	BO(15:0)
0000	0001-0016	0017-0032	0129-0144	0145-0160	CO(15:0)
0000	0065-0080	0081-0096	0193-0208	0209-0224	DO(15:0)
					SYOUT

Figure B.1 The simulation result of the Output Switch mode.

Appendix C :

Additional notes on the functional simulations necessary for post-fabrication test¹

C.1 Introduction

The functional simulations for post-fabrication is divided into three separate tests:

- to measure the input threshold voltages (V_{IL}/V_{IH})
- three-states simulation
- to measure power supply current (I_{dd})

A maximum of 12000 test patterns can be used and it is divided into three blocks where each block consists of 4000 test patterns.

C.2 V_{IL}/V_{IH} Simulation

The simulation pattern allows the tester to measure the circuit for input threshold voltages. The simulation is performed under the following conditions: $V_{dd}=5V$, Temperature= $25^{\circ}C$, process=worst case commercial condition and test period=2000ns.

To Check input voltage levels accurately, the design must contain V_{IL}/V_{IH} test circuitry called a "parametric NAND tree". It is possible to build the parametric NAND tree by interconnecting the 2-input NAND gates already attached to the inputs and dedicate one pin as a TESTOUT pin. If a pin cannot be dedicated for use in the

1. From the LSI Logic "Databook and design manua, HCMOS Macrocells and Macrofunctions, October, 1986

parametric NAND tree, then multiplex the TESTOUT pin with a normal output pin. The tester monitors only the TESTOUT pin when the V_{IL}/V_{IH} test is executed. Figure C.1 shows a simplified parametric Nand tree. The first V_{IL}/V_{IH} test pattern must consist of all ones on the inputs. Initially, the output at TESTOUT is either one or zero depending on the number of inputs (for even number inputs, one is the output at TESTOUT and via versa). As zeros are applied sequentially to the inputs along the parametric NAND tree, the TESTOUT toggles between one and zero.

C.3 Three-state Simulation

The three-state simulation typically consists of only a few patterns. The patterns must bring all three-state outputs to high impedance at least once during the simulation. When the tester executes the three-state patterns, a test program stipulates that only pins in the high impedance state be monitored. All output pins in the zero state are masked. Execute the three-state simulation under the following conditions: $V_{dd}=5V$, Temperature= 25° C, process=worst case commercial condition and test period=2000ns.

C.4 I_{dd} Simulation

I_{dd} denotes the power supply current leading from the V_{dd} (power supply) pin into the circuit. The I_{dd} simulation patterns set up the circuit for I_{dd} measurement. The purpose of I_{dd} measurement is to determine how much direct current leakage remains in the circuit after all activity in the circuit has ceased. Normally the I_{dd} simulation very few test patterns and execute under the following conditions: $V_{dd}=5V$, Temperature= 25° C, process=worst case commercial condition and test period=2000ns.

M_1	M_2	M_3	M_i	$i=\text{even}$	$i=\text{odd}$
1	1	1...	1	1	0
1	1	1...	0	0	1
1	1	0...	0	1	1
1	0	0...	0	0	0
0	0	0...	0	1	1

Figure C.1 Operation of a parametric NAND tree.

Appendix D :

The pin layout of the FFT Switch

	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15		
A	VDD	SV2M	AI15	AI14	AI13	AI12	AI11	AI10	AC10	AC11	AC12	AC13	AC14	AC15	VDD	A	
B	VSS		S_TEST_S	AI9	AI8	AI7	AI6	AI5	AC5	AC6	AC7	AC8	AC9	AI1	FC17	B	
C	BI15	SCN060	AI4	AI3	AI2	AI1			AC1	AC2	AC3	AC4	AI2	BC9	FC14	C	
D	BI14	BI9		AI0	VDD	PI1	PI2	PI3	AC0	PI4	VDD	BC0	BC4	BC8	FC13	D	
E	BI13	BI8	BI4	BI0				VSS				BC3	BC2	BI7	FC12	E	
F	BI12	BI7	BI3	S/OUT								BI132	BC2	BC0	BC11	F	
G	BI11	BI6	BI2	CP								VDD	BC1	BC5	BC10	G	
H	BI10	BI5	BI1	VDD	VSS						VSS		CC1	CC5	CC16	H	
J	CI10	CI5	CI1	TSTIN									CC2	CC6	CC11	J	
K	CI11	CI6	CI2	CLR									CC3	CC7	CC12	K	
L	CI12	CI7	CI3					VSS					CC0	CC4	CC8	CC13	L
M	CI13	CI8	CI4	CI0	VSS		DI9	DC0	BYPASS			VDD		CC9	CL14	M	
N	CI14	CI9	TSTOUT	DI4	DI3	DI2	DI1	DI0	DC1	DC2	DC3	DC4	CC	TESTCON	TSTOUT2	CC15	N
P	CI15		DI9	DI8	DI7	DI6	DI5	DC5	DC6	DC7	DC8	DC9	S_TEST_S	SLDAD	VDD	P	
Q	VDD	DI15	DI14	DI13	DI12	DI11	DI10	DC10	DC11	DC12	DC13	DC14	DC15	CS_P	VSS	Q	
	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15		

TOP VIEW OF SWITCH PINOUT

END

2/3.12.92

FIN