A DISTRIBUTED AMPLIFIER BASED ACTIVE TRANSVERSAL FILTER MMIC WITH POSITIVE AND NEGATIVE ADJUSTABLE TAP WEIGHTS FOR VERY HIGH SPEED LIGHTWAVE SYSTEMS

BY

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Finally, to my best friend Marcia. Thank you for your love, dedication and support I am truly looking forward to enjoying our lives together.

This thesis is dedicated to my parents,
Brenda and Bill and
my girlfriend, Marcia.
ABSTRACT

A distributed amplifier based 5-tap transversal filter using MESFET implementation of the Gilbert Cell as the tap weights has been designed for use as a fractionally spaced adjustable equalizer in high speed optical communication systems. The effective transconductance of the Gilbert Cell can be adjusted from negative to positive values and therefore the transversal filter designed has adjustable positive and negative tap weights. With these tap weight variation capabilities the frequency response of the filter can be significantly altered making it an attractive candidate for adaptive pulse shaping in optical receivers. This is the first time a transversal filter with adjustable positive and negative has been implemented. In this thesis a review of intensity modulation/direct detection (IM/DD) optical communications is provided with an emphasis placed on the requirement for adaptive linear equalization using the transversal filter. The fundamental theory behind the transversal filter is then presented. Next, the design techniques used in order to achieve a transversal filter with positive and negative adjustable tap weights are discussed. These include a discussion of the transversal filter building blocks: the distributed amplifier, lumped element artificial transmission lines and the Gilbert Cell. Once the building blocks are outlined the entire transversal filter MMIC design is introduced. Finally, simulated and measured frequency response and transient response results are presented for various tap weight configurations. Frequency response results demonstrate the transversal filter is capable of achieving a variety of filtering functions. Transient response results demonstrate the transversal filter’s ability to alter the pulse shape in a 2.5Gb/s system.
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LIST OF SYMBOLS AND ABBREVIATIONS

IM/DD  Intensity modulation / Direct detection
BER    Bit error ratio
ISI     Intersymbol interference
GaAs   Gallium Arsenide
MESFET Metal-semiconductor field-effect transistor
MMIC   Monolithic microwave integrated circuit
$H_f(f)$ Optical fiber transfer function
$n(t)$ Input referred additive white Gaussian noise
$E_{Tx}(t)$ Optical carrier electric field at input of fiber
$E_{Rx}(t)$ Optical carrier electric field incident on receiver photodiode
$i_p$ Electric current generated in the photodiode
$x(t)$ Voltage output from transimpedance amplifier / input to transversal filter
$T_s$ Data signal bit rate
$A_n$ Conversion constant from incident electric field to amplifier output voltage
$A_n'$ Conversion constant from incident electric field to amplifier output voltage
$y(t)$ Transversal filter output
$w_n$ Transversal filter tap weight
$\omega$ Angular frequency
$X(j\omega)$ Frequency spectrum of transversal filter input signal
$Y(j\omega)$ Frequency spectrum of transversal filter output signal
$H(j\omega)$ Frequency response of transversal filter
$h(t)$ Impulse response of transversal filter
$\tau$ Transversal filter tap delay
SSE    Symool spaced equalizer
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>FSE</td>
<td>Fractionally spaced equalizer</td>
</tr>
<tr>
<td>$H_{EQ}(j\omega)$</td>
<td>Desired equalizer response</td>
</tr>
<tr>
<td>$X_d(t)$</td>
<td>Distorted data input pulse</td>
</tr>
<tr>
<td>DA</td>
<td>Distributed amplifier</td>
</tr>
<tr>
<td>$Z_{in}$</td>
<td>Characteristic impedance of the input transmission line</td>
</tr>
<tr>
<td>$Z_{out}$</td>
<td>Characteristic impedance of the output transmission line</td>
</tr>
<tr>
<td>$l_s$</td>
<td>Distance between successive gain element inputs along input transmission line</td>
</tr>
<tr>
<td>$l_d$</td>
<td>Distance between successive gain element inputs along output transmission line</td>
</tr>
<tr>
<td>$g_{mn}$</td>
<td>Transconductance of the nth gain element (Gilbert Cell) in the transversal filter</td>
</tr>
<tr>
<td>$\tau_s$</td>
<td>Time delay for signal between successive gain element inputs along input transmission line</td>
</tr>
<tr>
<td>$\tau_d$</td>
<td>Time delay for signal between successive gain element inputs along output transmission line</td>
</tr>
<tr>
<td>$C$</td>
<td>Total capacitance in a unit cell of a lumped element transmission line unit cell</td>
</tr>
<tr>
<td>$L$</td>
<td>Total inductance in a unit cell of a lumped element transmission line unit cell</td>
</tr>
<tr>
<td>$\gamma$</td>
<td>Effective propagation constant for signal through a unit cell</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>Real portion of effective propagation constant for signal through a unit cell</td>
</tr>
<tr>
<td>$j\beta$</td>
<td>Imaginary portion of effective propagation constant for signal through a unit cell</td>
</tr>
<tr>
<td>$k$</td>
<td>Normalized frequency parameter</td>
</tr>
<tr>
<td>$f_c$</td>
<td>Cutoff frequency</td>
</tr>
<tr>
<td>xjn</td>
<td>Transistor n</td>
</tr>
<tr>
<td>IDP</td>
<td>Input differential pair</td>
</tr>
<tr>
<td>DPL</td>
<td>Differential pair load</td>
</tr>
<tr>
<td>$I_b$</td>
<td>DC bias current drawn by current source of the Gilbert Cell</td>
</tr>
<tr>
<td>$I_{syn}$</td>
<td>DC bias current through transistor n in the Gilbert Cell</td>
</tr>
<tr>
<td>$\beta$</td>
<td>Transistor conductance parameter</td>
</tr>
<tr>
<td>$V_{ref1}$</td>
<td>Bias voltage for input differential pair of the Gilbert Cell</td>
</tr>
<tr>
<td>$V_{ref2}$</td>
<td>Bias voltage for differential pair loads of the Gilbert Cell</td>
</tr>
<tr>
<td>$V_{gain}$</td>
<td>Bias voltage used to control gain of Gilbert Cell</td>
</tr>
</tbody>
</table>
\( v_{in} \)  
Small signal source

\( i_{sg} \)  
Small signal current

\( i_{sn} \)  
Small signal current through transistor n

\( i_{out} \)  
Small signal output current drawn by the Gilbert Cell

\( G(V_{gsn}) \)  
Gilbert Cell small signal gain factor

\( R_I \)  
Gilbert Cell load resistor

\( Z_{load} \)  
Load impedance at output of transversal filter

\( Z_o \)  
Transmission line impedance

\( \epsilon_{eff} \)  
Effective dielectric constant

\( C_t \)  
Capacitance per unit length of transmission line

\( L_t \)  
Inductance per unit length of transmission line

\( c \)  
Speed of light

\( C_p \)  
Total parasitic capacitance

\( L_p \)  
Total parasitic inductance

\( R_s \)  
Sheet resistance

\( C_w \)  
Planar inductor crossover capacitance

\( C_{in} \)  
Gilbert Cell input capacitance

\( C_{out} \)  
Gilbert Cell output capacitance

\( C_g \)  
Lumped element input delay line unit cell capacitance

\( C_d \)  
Lumped element output delay line unit cell capacitance

\( W \)  
Transversal filter weighting vector
CHAPTER 1

INTRODUCTION

1.1 Optical Fiber Communication Systems

As the information age moves forward the demand for communication systems capable of large information carrying capacities over longer distances is increasing. Due to the extremely large low-loss transmission bandwidth available in optical fibers, optical fiber communication systems can satisfy the requirements for such high bit rate-transmission products. These systems provide communication by transmitting laser light, modulated in some fashion through an optical fiber and into an optical receiver. Modulation techniques vary and consequently so do the reception or demodulation techniques. Lightwave communication systems can be classified into two main categories that depend on the receiver configurations they use. Coherent or heterodyne lightwave systems [1]-[3] have receiver configurations that spatially combine the incoming signal from the optical fiber with light from a local oscillator laser. When the combined optical signal is incident on the receiver the photodiode's nonlinear characteristics allow for mixing action to occur resulting in the down conversion of the combined optical signal to an intermediate frequency which can be further processed in the electrical domain. Coherent detection systems allow for high receiver sensitivity, but they are complex and expensive in terms of the component requirements and operational environments. In
contrast to coherent lightwave systems, direct detection systems [1]-[3] are configured so that the incoming optical signal is directly incident on the photodiode. Since the photodiode produces an electrical signal that is proportional to the intensity of the optical signal this reception technique easily lends itself to a baseband lightwave communication technique referred to as intensity modulation/direct detection (IM/DD) [1]-[3].

Intensity modulation means that digital data is represented according to the intensity of the laser light that is transmitted along the optical fiber. In binary IM/DD systems the modulation format is on-off keying. The laser output intensity or power is $P_1$ for a binary '1' and $P_0$ for a binary '0'. In order for the receiver to properly distinguish whether a '1' or a '0' is sent $P_1$ must be distinguishably greater than $P_0$. The laser output power is controlled according to the drive current applied to it. At the receiver of an IM/DD system the intensity modulated optical signal is incident on a photodiode. The photodiode generates a current proportional to the intensity of the optical signal it receives. This base band current is electronically processed in the receiver before a '1' or '0' decision is made by subsequent decision circuitry. The decision is based on the amplitude of the processed electronic signal, which is ideally proportional to the received optical power. Although direct detection receivers are less sensitive than their coherent counterparts, they are less complex and therefore are deployed in virtually all of today's commercial lightwave communication systems [4].

The ultimate performance measurement for a receiver in a lightwave communication system is its sensitivity[1]-[3]. Sensitivity is a measure of the required optical power in order to attain a bit error ratio (BER) of $10^{-9}$. As sensitivity increases the required optical power decreases. Theoretical calculations have demonstrated that the
sensitivity of an IM/DD system can be as low as 20 photons per bit [3] which is easily converted to optical power if the wavelength of light and quantum efficiency of the photodiode are known. In practical IM/DD systems sensitivities are degraded to values of 8000 photons per bit due various impairments that exist within the system [5]. Since it is desirable to consume as little power as possible in these systems it is advantageous for a system to operate with as high a sensitivity as possible. From an overall lightwave communication system point of view the bit rate-transmission distance product is the most important performance measurement since it is a measure of the maximum bit rate achievable over a given distance of fiber while maintaining an acceptable BER. The bit rate-transmission distance product is directly related to the receiver sensitivity since the BER in an IM/DD system is dictated by the receiver sensitivity once an optical transmission power is established.

In order to cater to the increasing speed, span and reduced complexity requirements for lightwave communication systems the bit rate-transmission distance product must be increased. Lightwave systems with high bit rate-transmission distance products exploit the low-loss bandwidth offered by the optical fiber, allowing for higher data transfer rates over longer fiber distances thereby reducing the number of repeater stations required in order to achieve reliable transmission over a given distance. The receiver sensitivity and therefore the bit rate-transmission distance product in IM/DD systems is limited due to the signal distortion which arises as a result of the combination of noise and intersymbol interference (ISI) [1]-[16]. Noise results from the random arrival of photons and thermal noise generated within the receiver electronics. ISI results from the overlap of data pulses that have become distorted due to various transmission
impairments inherent in IM/DD systems. Despite such limitations the simplicity of IM/DD systems still make it an attractive candidate for optical communications. This fact is emphasized not only with the use of IM/DD in current commercial, single wavelength lightwave systems, but also with its use in emerging wavelength division multiplexed systems [4]. Methods have been developed to compensate for transmission impairments in order to reduce signal distortion and improve bit rate-transmission distance products and it can be expected that market demands will promote further work in this area.

1.2 Techniques for Reducing ISI in IM/DD Systems

The primary purpose of an IM/DD optical communication system is to accurately transfer data from one point to another. This means that the original data at the transmitter must be reconstructed to its original form at the receiver. The reconstruction process is performed in the receiver by means of a decision circuit. The decision circuit periodically samples (at the bit rate) the electrical signal, generated in the receiver that corresponds to the incident optical signal. It then compares this sample to a given threshold value, decides whether the sample corresponds to a '1' or a '0' (for the case of a binary IM/DD system) and reconstructs the interpreted data. Ideally, the electrical signal sampled in the receiver is an accurate representation of the original data so that the decision circuit can perform the reconstruction without error. Unfortunately, an exact representation is not possible since the electrical signal sample is distorted from its intended value due to the contributions from noise and ISI inherent in the system. Noise occurs in the receiver and includes signal dependent shot noise resulting from the random arrival of photons at the receiver and thermal noise generated within the receiver.
electronics [1]-[3]. ISI occurs as a result of various transmission impairments within IM/DD systems. These transmission impairments include chromatic dispersion [1]-[3][6][7][9][10][17] and polarization dispersion [6][7][9][18][19], laser nonlinearity [3][6]-[9] and nonideal optoelectronic and electronic receiver component response [6]-[9].

A number of signal processing techniques are used to compensate for the transmission impairments inherent in IM/DD systems, thereby reducing ISI and improving system performance. The various techniques include equalization by microstrip lines [20], equalization by microstrip waveguides [21], precompensation techniques at the transmitter, nonlinear cancellation [6][7][22], maximum likelihood detection [6] and linear equalization with a transversal filter [6]-[9][18][21][23]-[25][27]. It is advantageous for the compensation technique that is used to be adaptive so that time varying transmission impairments such as polarization dispersion can be continuously compensated for. Furthermore, adaptive compensation can also accommodate for the different nonideal receiver responses that will exist due to the variation of transmitter/receiver pairs used in IM/DD systems. Therefore, some of these mitigation techniques are more attractive than the others.

Equalization using microstrip lines and microwave waveguides makes use of the dispersive nature of these components in order to compensate for the chromatic dispersion in the fiber. This type of equalization is not adaptive and can only be used in coherent lightwave systems. By using dispersion shifted fiber, dispersion compensating fibers or gratings chromatic dispersion can be further compensated for, however the compensation technique is not adaptive and does not compensate for polarization
dispersion. Precompensation techniques involve tailoring the transmitter output so that the received signal is not distorted as it travels through the lightwave system. In order for precompensation to be adaptive, feedback between the transmitter and receiver is required. Nonlinear cancellation and maximum likelihood detection can be made adaptive and therefore are effective means of compensating for transmission impairments in IM/DD systems thereby reducing ISI and improving system performance. Linear equalization through means of a transversal filter is also an ideal candidate for adaptive equalization provided that the tap weight values of the filter can be adjusted [9]. Although the concept of equalization in IM/DD systems with transversal filtering has been known to be effective for some time, the actual use of transversal filters in IM/DD lightwave systems is still relatively new.

1.3 Adaptive Linear Equalization Using a Transversal Filter

Linear equalization with a transversal filter that is placed between the preamplifier and the decision circuitry in the receiver has been shown as an effective means of reducing the effects of ISI thereby improving system performance[6]-[9][18][21][23]-[25][27]. A transversal filter consists of a series of delay elements referred to as the tap delays and gain elements referred to as tap weights. The circuit topology is arranged in such a fashion so that the output of the filter consists of a superposition of linearly weighted past and present signal responses. The characteristics of the filter depend on the number of taps or length of the filter, the time delay imposed by the tap delays and the values of the tap weights. Hence, if the tap weight values are made adjustable the filter response will be adjustable. As discussed previously, for the
purposes of adaptive equalization in IM/DD systems an equalizer that has adjustable characteristics is a necessity.

References [9][24]-[27] present monolithic microwave integrated circuit (MMIC) implementations of transversal filters, which are attractive since they illustrate the potential for a fully integrated optical receiver system. Reference [26] presents a transversal filter with a fixed bandpass response and is not applicable to equalization in IM/DD lightwave systems. Reference [24] presents a distributed amplifier based transversal filter for the purposes of equalization in lightwave systems. However, the filter designed has fixed tap weight values and is therefore not capable of adaptability. References [9] and [27] presents a novel adjustable transversal filter for the purposes of adaptive equalization in IM/DD lightwave systems. The design is based on the distributed amplifier with the tap delay implementation performed using lumped elements and the tap weight implementation using cascode GaAs MESFETs. The gain of each cascode stage can be individually adjusted and controlled according to the voltage that is applied to the gate of the top transistor and therefore adjustable tap weights are achieved. However the cascode gain cell is only capable of adjusting its gain within a range of positive tap weight values. This tap weight value constraint limits the versatility of the filter in that the number of achievable filter responses is restricted.

In response to the positive tap weight limitation in the transversal filters presented by [9] and [27] this thesis will demonstrate that a transversal filter with positive and negative adjustable tap weights can be achieved in MMIC form. Specifically, a 5-tap transversal filter with 50ps tap delays and adjustable positive and negative tap weights is designed, fabricated in the form of a MMIC and characterized. Like the design of the
transversal filters limited to adjustable positive tap weights [9][27] the design of this transversal filter is based on the distributed amplifier and the delay segments are implemented using lumped elements. However, in order to achieve tap weights that are adjustable within a range of positive and negative values the Gilbert Cell is used to implement the adjustable tap weight in this filter. As in the case of the cascode gain cell used in [9] and [27] the gain provided by each Gilbert Cell can be electronically controlled thereby providing the potential for adaptive linear equalization. In this thesis simulated and measured results demonstrate the adjustable nature of the transversal filter designed.

1.4 Organization of Thesis

The purpose of this thesis is to demonstrate that a transversal filter in the form of a MMIC with positive and negative adjustable tap weights can be achieved. For that reason the bulk of the thesis will focus on the design and implementation of the MMIC. However, since the driving force behind the transversal filter is the need for reducing the effects of ISI and noise in IM/DD systems, Chapter 2 will outline the operation of IM/DD systems in more detail. It will also discuss the use of the transversal filter as an adaptive equalizer in these systems. Chapter 3 will outline the design techniques used in order to create the 5-tap transversal filter presented here and Chapter 4 will discuss the implementation of the filter in MMIC form and also present simulated results. Chapter 5 will present measured results that demonstrate the filter's adjustability as well as other functions that this type of filter is capable of performing. Finally, Chapter 6 will summarize the findings of the work performed and provide recommendations for future work.
CHAPTER 2

BACKGROUND

2.1 IM/DD Optical Communication System Considerations

Figure 2.1.1: IM/DD lighwave communication system block diagram [9]

Figure 2.1.1 illustrates the block diagram of an IM/DD communication system. It consists of a laser transmitter, optical fiber communication channel with transfer function $H_f(f)$ and a receiver comprised of a photodiode, preamplifier and a decision circuit. $n(t)$ represents an input referred additive white Guassian noise source attributable to the circuit noise in the receiver. The laser output, which depends on the bit level and is controlled by the laser drive current is coupled into the optical fiber. The value of the electric field of the optical carrier signal at the beginning of the fiber is represented as $E_{Tx}(t)$. $E_{inc}(t)$ represents the value of the electric field of the optical carrier that is incident on the photodiode after travelling through the fiber. When the optical carrier is incident
on the photodiode it generates an electric current, $i_p$ that is proportional to the square of the magnitude of the electric field of the carrier signal. The current is injected into a transimpedance preamplifier which outputs a voltage $x(t)$ that is ideally linearly proportional to the input current. The data stream $x(t)$ is sampled by the decision circuitry at the bit rate, $T$. The decision circuit compares each sample value to a threshold value in order to determine whether a given bit that has been received represents a ‘1’ or a ‘0’, thereby reconstructing the original data stream. As outlined in Chapter 1 the pulses which make up $x(t)$ ideally would have an optimum shape so that there would be no possibility of an error being made by the decision circuit in interpreting the received data. Unfortunately, this is not possible due to the fact that the signal becomes distorted as it travels along an optical fiber and through the receiver towards the decision circuitry. The primary causes of distortion are noise at the receiver and intersymbol interference (ISI)[1]-[16].

Noise in the system [1]-[3] occurs as a result of shot noise and thermal noise. Shot noise occurs due to the random arrival times of photons in the optical signal. The arrival times are governed by Poisson statistics and the amount of shot noise present is dependent on the signal level. Thermal noise is generated in the receiver by the preamplifier. It can be modeled as independent additive white Guassian noise with zero mean [8]. The actual amount of thermal noise present in a receiver depends on the preamplifier design.

ISI results from transmission impairments in the system including fiber dispersion [1]-[3][6][7][9][10][17]-[19], laser nonlinearities [3][6]-[9] and nonideal receiver response [6]-[9]. The two types of fiber dispersion that cause signal distortion
and therefore ISI to occur are chromatic dispersion and polarization dispersion. Chromatic dispersion [1]-[3],[6],[7],[9],[10],[17] occurs as a result of material and waveguide dispersion. Material dispersion arises from the dependence of the index of refraction of the fiber on frequency and waveguide dispersion is a result of the dependence of group velocity on frequency. Chromatic dispersion causes signals of different frequencies to travel along the optical fiber with different delays. Since an optical pulse consists of a spectrum of frequency components the frequency dependent delay imposed by the fiber will cause the pulse to become spread out in time at the fiber’s output ultimately resulting in signal distortion. Chromatic dispersion effects are constant for a fixed length of fiber, but its contribution to signal distortion increases as the bandwidth of the data pulses (or the data rate) increases. Most of today’s IM/DD communication systems operate at a carrier wavelength of 1.55μm, which corresponds to the wavelength at which the effects of chromatic dispersion are minimal in single mode dispersion compensated optical fiber [9], however at higher bit rates chromatic dispersion is still significant. Polarization dispersion [6],[7],[9],[18],[19] is a result of birefringence which occurs in a fiber when the circular symmetry of the fiber is broken as a result of stresses that already exist or are induced in the fiber. Polarization dispersion and can be characterized in terms of first and second order (in frequency) effects. The first order polarization dispersion effect is a delay in the signal in one polarization relative to the delay in the signal in the other polarization. Since orthogonal polarizations add power-wise at the receiver, first order polarization dispersion causes data pulse spreading and even separation. The delay difference between the different polarizations increases with optical fiber length and varies with time as a result of varying temperature. First order
Polarization dispersion effects become more prominent as the system bit rate increases since higher data rates have smaller bit periods and therefore the delay difference between the orthogonal components of the signal become significant. Second order polarization effects are linear delay distortion that differs in sign in the two polarizations and a cross coupling of power between the polarizations at the receiver. In higher data rate lightwave systems where the effects of chromatic dispersion will be eliminated by necessity it is expected that the ISI caused by polarisation dispersion will be the limiting factor in terms of system performance.

Laser nonlinearities [3][6]-[9] that result in ISI include chirp, linewidth enhancement and relaxation oscillations. Chirp is the variation of the optical carrier frequency of a laser resulting from changes in laser drive current. Linewidth enhancement refers to the increased width of the spectrum of the laser output due to phase noise. Both chirp and linewidth enhancement cause an increase in the bandwidth of the optical signal transmitted which in turn causes chromatic dispersion effects to increase, ultimately resulting in increased ISI. Oscillation relaxation is a nonideal laser response that effects the output intensity of the laser. When the amplitude of the laser drive current is altered from one level to another the amplitude of the laser output intensity changes to a level corresponding to the change in the current drive level. However, oscillations in the amplitude of the laser output intensity occur around the new output level as a result of the sudden transition in the drive current. Oscillation relaxation can distort a data pulse in such a fashion so that it has a nonzero value outside its assigned bit period that in turn can result in ISI.
Finally, ISI can result from nonideal receiver response [6]-[9]. The photodiode and the preamplifier response combine to impose an overall receiver bandwidth limitation that in turn can result in data pulse distortion and therefore ISI. The degree of ISI that can result due to the receiver response is dependent on the receiver input signal shape and the response characteristics. It should also be noted that a tradeoff exists between thermal noise level in the receiver and the ISI caused due to finite receiver bandwidth. A receiver that is broadband will reduce ISI and increase the noise while a receiver that is narrow band will reduce the amount of thermal noise in the system at the expense of increased ISI [9].

To obtain some insight into the effects of noise and ISI consider an IM/DD system with a preamplifier in the receiver that provides constant gain over a large enough bandwidth so that the effects of a nonideal receiver response can be neglected. Assume that a train of data pulses incident on the receiver's photodiode can be represented as,

\[
e_{\text{inc}}(t) = \sum_{n} E_n(t - nT_s)
\]

where \(e_{\text{inc}}(t)\) represents the time dependent magnitude of the electric field of the optical carrier for the entire pulse train and \(E_n(t-nT_s)\) represents the electric field amplitude associated with the \(n\)th pulse in the data pulse train with bit period \(T_s\). The shape of each pulse depends on the assigned bit value (0 or 1 for binary systems) and also carries with it the effects of fiber dispersion and laser nonlinearities. \(e_{\text{inc}}(t)\) is converted to a current according to the square law characteristics of the receiver photodiode and injected into the preamplifier. The resulting voltage output from the preamplifier that is to be sampled by the sampling circuitry can be represented as,
where $A_n$ and $A_{n'}$ accounts for the optical to electrical conversion performed by the photodiode and the current to voltage conversion of the preamplifier. As can be seen in Equation 2.1.2, $x(t)$ is comprised of three terms. The first two terms corresponds to the square of the magnitude of the electric field of each pulse, and the product of the electric field of a given pulse with the conjugate of another in the pulse train. These two terms arise from the fact that the photodiode generates a current that is proportional to the intensity of the optical signal incident on it. The third term, $n(t)$ represents noise in the system.

The $kth$ sample of $x(t)$, taken by the decision circuitry, will be at time $kT_s$ and will therefore have the form,

$$x(kT_s) = |A_k E_k(0)|^2 + \sum_{n \neq k} |A_n E_n(kT_s - nT_s)|^2 + \sum_{n, n'} \sum_{n \neq n'} (A_n E_n(kT_s - nT_s))(A_{n'} E_{n'}(kT_s - n'T_s)) + n(kT_s)$$

The first term in Equation 2.1.3 represents the desired value to be detected by the decision circuitry, the second and third terms represent ISI and the fourth is noise. It can be seen that if the pulses other than the $kth$ one had a zero value outside their own assigned bit periods, or at least a zero value at the sampling time $kT_s$ and the noise at $kT_s$ was zero that the intended data value, $|A_k E_k(0)|^2$ would be interpreted correctly. However,
as discussed previously transmission impairments in the system cause the data pulses to become distorted so that they contain nonzero values outside their own assigned bit period which ultimately leads to ISI. Furthermore, noise is always inherent in the receiver. Hence, it can be seen from Equation 2.1.3 that there exists a nonzero probability that the decision circuit could incorrectly interpret the actual value of the desired data sample $|A_kE_k(0)|^2$ due to the distortion of the signal resulting from contributions of ISI and noise.

The amount of ISI and noise in a lightwave system can be investigated through the means of an eye diagram. An eye diagram is made up of a superposition of the voltage waveforms at the output of the receiver resulting from all possible bit transitions in one bit period. Eye diagrams can be used in order to determine the optimal decision time and threshold level for a specific lightwave system. Typically, the optimum decision time and threshold level for a specific lightwave system.

Figure 2.1.2: IM/DD system eye diagram [2]
time occurs on the horizontal axis (time) where the eye has the largest opening in the vertical direction (amplitude) and the optimal threshold can be chosen according to the width of the eye at the optimum decision time. The optimum decision time and threshold are chosen so that the probability of the decision circuit incorrectly interpreting the data is minimized. Once the optimal decision time and threshold are determined the decision circuit can interpret the incoming signal values by comparing the signal level at the decision time to the decision threshold level, thereby reconstructing the original signal. Figure 2.1.2 illustrates an eye diagram and has the optimum decision time and threshold labeled on it. ISI and noise cause the eye to become smaller in both the vertical and horizontal directions thereby reducing the tolerable margin of error for detection in time and amplitude. An eye with a small opening increases the probability of a bit being interpreted incorrectly by the decision circuit thereby increasing the BER for the system. In order to improve the BER the area of the eye must be increased by increasing the signal power and thereby decreasing the system sensitivity. The increase in the required signal power is referred to as the power penalty. As discussed earlier since power is a premium it is advantageous to open the eye through other means such as linear equalization with a transversal filter rather than increasing the system power.

2.2 Transversal Filter Theory

Figure 2.2.1 illustrates the general structure of a transversal filter. It is comprised of constant delay line segments or tap delays, tap weights and an output that combines the delayed and weighted samples of the input signal. It should be noted that although the tap delays in this particular transversal filter implementation are equal this is not a constraint
on the transversal filter system, however, in order to simplify the transversal filter analysis equal tap delays will be used here. This analysis is desired since the transversal filter designed in this thesis utilizes equal tap delay lengths. The transversal filter operates

![Block Diagram of transversal filter with constant tap delays](image)

**Figure 2.2.1:** Block Diagram of transversal filter with constant tap delays [9]

as follows. A signal \( x(t) \) enters the input of the transversal filter and travels through \( L \) tap delays of equal length \( \tau \). At the output of each delay line segment the signal, which is delayed in time but unchanged in shape is continuously sampled and linearly weighted by a corresponding tap weight \( w_n \). Finally, at the transversal filter output the linearly weighted present and delayed signal samples are summed together to form \( y(t) \). The time representation of the transversal filter output can be written as,

\[
y(t) = \sum_{n=0}^{L} w_n x(t - n\tau) \quad 2.2.1
\]

and the corresponding frequency spectrum of the signal output is,
\[ Y(j\omega) = \sum_{n=0}^{L} w_n X(j\omega) e^{-j\omega n\tau} \quad 2.2.2 \]

where the Fourier Transform relation \( x(t-\tau) \Rightarrow X(j\omega)e^{j\omega\tau} \) is used and \( y(t) \), \( x(t) \) and \( Y(\tau) \) and \( X(\tau) \) are Fourier transform pairs respectively. The filter frequency response can therefore be determined as,

\[ H(j\omega) = \frac{Y(j\omega)}{X(j\omega)} = \sum_{n=0}^{L} w_n e^{-j\omega n\tau} \quad 2.2.3 \]

and the corresponding impulse response of the transversal filter is,

\[ h(t) = \sum_{n=0}^{L} w_n \delta(t-n\tau) \quad 2.2.4 \]

since the inverse Fourier Transform of \( e^{-j\omega n\tau} \) is \( \delta(t-n\tau) \). It can be seen that the response of the transversal filter of length \( L+1 \) in Figure 2.2.1 is solely a function of the filter tap weight values, \( w_n \). Hence, if the tap weights are adjustable then the filter response will be adjustable and therefore can be used to perform adaptive equalization.

The length of the tap delays, \( \tau \) also plays an important role in determining the filter’s operational characteristics, especially when being used as an equalizer in a digital system. As can be seen in Equation 2.2.3 the frequency response of a transversal filter with constant tap delays of length \( \tau \) is periodic in frequency with period \( 1/\tau \). Because of this relationship between the filter’s tap delay length and its frequency response
transversal filters are divided into two groups; symbol spaced equalizers (SSEs) and fractionally spaced equalizers (FSEs).

SSEs have a frequency response that is periodic in frequency with period $1/\tau$, where $\tau$ is both the period of the input signal and also the length of the delay segments of the transversal filter equalizer. Generally, a digital signal with bit period $\tau$ has a frequency spectrum that extends beyond $1/\tau$. Hence, the periodicity of the SSE frequency response combined with the bandwidth of the input signal can lead to aliasing which in turn results in a reduction in the overall effect that the equalization process will have. This effect is referred to as band-edge distortion. Despite the inherent band edge distortion in synchronously spaced equalizers it has been found that they can still improve system performance [8], [24].

FSEs have tap delay lengths that are shorter than the signal bit period and therefore the period of the frequency response is spaced far enough apart so that band edge distortion can be avoided. In most lightwave systems a FSE with a tap delay of $\tau/2$ is sufficient [9] to avoid the any significant effects of band edge distortion, where $\tau$ is the bit period of the digital signal. Fractionally spaced equalizers have demonstrated improvements over synchronously spaced equalizers [8][23].
2.3 Linear Equalization with Transversal Filters in IM/DD Lightwave Communication Systems

As discussed in Section 1.3 linear equalization by means of a transversal filter can be used in order to reduce the effects of signal distortion. A transversal filter equalizer, placed in the receiver as shown in Figure 2.3.1 can perform pulse shaping in order to compensate for the effects of ISI and noise, thereby opening the eye, improving the system BER and increasing system sensitivity. The desired equalizer response can be written as,

$$H_{EQ}(j\omega) = \frac{Y(j\omega)}{X(j\omega)}$$                              \hspace{1cm} 2.3.1

where $Y(j\omega)$ is the spectrum of the equalizer output pulse (before the decision circuitry) required to minimize the system’s BER, $X(j\omega)$ is the spectrum of the equalizer input pulse and $H_{EQ}(j\omega)$ is the desired equalizer transfer function. Of course, the required equalizer response is system dependent since the input signal, $X(j\omega)$ is dependent on the system / channel characteristics.
To gain insight on how the transversal filter can be effective from a time domain point of view, consider a distorted data pulse, \( X_p(t) \) that arrives at the input of a transversal filter with \( L+1 \) stages (assume \( L \) is even), the corresponding filter output \( y(t) \) will be,

\[
y(t) = \sum_{n=0}^{L} w_n X_p (t - n\tau) \tag{2.3.1}
\]

where \( w_n \) is the tap weight value of the \( n \)th tap weight and \( \tau \) is the tap delay length of the filter. When the transversal filter output is sampled by the decision circuitry the sample value will be,

\[
y(kT_s) = \sum_{n=0}^{L} w_n X_p (kT_s - n\tau) \tag{2.3.2}
\]

where \( T_s \) is the sampling period of the decision circuitry. Nyquist’s first method for zero ISI [28] states that the output pulse from the transversal filter will not interfere with any other pulses at their designated time of sampling provided,

\[
y(kT_s) = C \quad k = 0 \tag{2.3.3}
\]

and

\[
y(kT_s) = 0 \quad k = 1, 2, \ldots, L \tag{2.3.4}
\]

where \( C \) is the desired value for the data pulse sample.
The tap weights required to force the zeros of the transversal output can be found by solving,

$$[W] = [X_p]^{-1}[Y]$$  \text{2.3.5}

where $[W]$ is a column matrix with L+1 elements representing the tap weight values, $[X_p]$ is a square matrix with side length equal to $L+1$. Each row of $[X_p]$ corresponds to a specific sampling time between $k = 0$ and $k = L$ and contains the values of each term in the transversal filter output for that sampling time. $[Y]$ is a column matrix representing the desired transversal filter output at sampling times between $k = 0$ and $k = L$. $[Y]$ contains $L+1$ elements all of which are zero except for the first element which is $C$. Hence, the tap weight values depend on the shape of the distorted pulse, $X_p(t)$ that enters the transversal filter. This method ensures that the pulse has a value of zero at sampling times previous to and after its own sampling time. As can be seen from Equation 2.3.5 the transversal filter can only force its output pulse to have values of zero at $L$ sampling times. Generally, a transversal filter is designed with enough stages so that the effect of ISI from the currently sampled pulse will not be significant in pulses which are sampled at times outside $k = 0$. It should be noted that the above demonstration of ISI reduction though transversal filter tap weight adjustment, provides an intuitive understanding of how the transversal filter can make use of its delay elements and tap weight values in order force outputs which have values of zero at specific times. The demonstration did not take into account other practical system considerations, such as noise that can cause further signal distortion resulting in an alteration of the required optimum tap weight settings.
2.4 Adaptive Linear Equalization with Transversal Filters in IM/DD Lightwave Communication Systems

It is of practical interest for the transversal filter equalizer to be adjustable. As mentioned in Section 2.1 there exists a tradeoff between receiver response (in this case including the equalizer) and the amount of thermal noise in the system. An adjustable equalizer has the advantage in that it can be optimized according to the signal and noise characteristics of a specific lightwave system. Furthermore, an equalizer that is electronically adjustable lends itself to adaptability. Adaptive equalization allows for time varying optimization which can be important when compensating for time varying system phenomena such as polarization dispersion and electronic component drift/degradation, or advantageous when installing the equalizer in lightwave systems with different component characteristics so that a system specific calibration can be done automatically.

As demonstrated in Section 2.2 the transversal filter can have a response that is adjustable. Since a transversal filter's frequency response is defined solely as a function of its tap weight values adjustability can be achieved by implementing tap weights that have variable values. The range of adjustability that is available from a practical transversal filter is limited by the range in which the tap weights can be adjusted and the number of stages in the filter. Once a filter is capable of adjustability adaptive equalization can be performed.

Adaptive equalization with a transversal filter in IM/DD systems requires an algorithmic process to be carried out that analyzes the BER of the system and then
adjusts the equalizer tap weights in order to achieve optimum performance for the system under consideration. In this case the tap weight values and hence system performance are optimized when a minimum BER is achieved. Hence, the ultimate goal of an adaptive transversal filter is to provide and maintain an optimal response that will minimize the BER. Reference [9] outlines in detail adaptive filtering algorithms for optimization of transversal filters (via tap weight adjustment) so that a minimum BER can be achieved in IM/DD systems.
CHAPTER 3

TRANSVERSAL FILTER DESIGN

The transversal filter designed in this thesis is based on the architecture of the distributed amplifier which is also comprised of delay and gain elements. In its simplest form the delay elements are implemented with lengths of transmission line and the gain elements are single stage common source transistors. However, in the case of the transversal filter designed here the delay elements are implemented using lumped elements and the gain element is implemented using the Gilbert Cell. As a premise to Chapter 4 which discusses the transversal filter implementation, the following section will outline the theory behind the design of a transversal filter based on the distributed amplifier, discuss how an artificial transmission line can be implemented using lumped elements in order to provide the required tap delay and also describe the operation of the Gilbert Cell and how it can be used to achieve positive and negative adjustable tap weights.

3.1 A Distributed Amplifier Based Transversal Filter Design

The distributed amplifier [29] (DA) provides gain over a large bandwidth and is therefore attractive for high speed electronic processing in lightwave communication applications. Furthermore, the DA is comprised of delay and gain elements that are
Figure 3.1.1: Configuration of an N-stage distributed amplifier [29], the distributed amplifier output (DA Output) and the transversal filter output (TF Output) are at opposite ends of the output transmission line.

connected in a manner that allows for a transversal filter to be implemented using the same architecture. Figure 3.1.1 illustrates the architecture of the DA, where it can be seen that gain elements are periodically placed along input and output transmission lines. The result of this is that the input and output capacitance of the each gain element is absorbed, effectively becoming a part of the input and output transmission line respectively. What is left behind is an ideal gain element capable of broadband amplification. In operation the input voltage, $x(t)$ enters the DA structure and propagates along a match terminated input transmission line of characteristic impedance $Z_g$. The gain elements in the DA ($G_1$-$G_N$) which are spaced at equivalent distances $l_g$ along the input transmission line, tap off a portion of the energy of the input signal and amplify it. The output of each active device is connected at equal intervals $l_d$ along a match terminated, output transmission line of characteristic impedance of $Z_d$. The amplified output signals from each active element combine to form forward travelling and backward travelling output signals that propagate
to either end of the match terminated output transmission line. The propagation constants and length of each of the periodically loaded input and output transmission lines are chosen so that the delay between each active element is the same and therefore constructive interference occurs at the distributed amplifier output (DA Output) and gain is achieved over a large frequency range. By taking the output at the opposite end of the transmission line, labeled TF Output in Figure 3.1.1 the desired transversal filter action can be realized. As can be seen by inspection of the DA in Figure 3.1.1 the output signal, \(y(t)\) at the TF Output is equal to,

\[
y(t) = -\frac{1}{2} Z_d \sum_{n=0}^{N-1} g_{mn+1} x\left(t - n(\tau_g + \tau_d)\right)
\]

where \(x(t)\) is the input signal, \(Z_d\) is the termination impedance of the output transmission line, \(g_{mn+1}\) is the transconductance of the \((n+1)th\) gain element, \(G_{n+1}\) and \(\tau_g\) and \(\tau_d\) are the time delays invoked as a result of signal propagation through transmission line length \(L_g\) and \(L_d\) respectively. It should be noted that for reasons of simplicity it was assumed that the gain elements have infinite input and output impedance and have negligible delay in this analysis. It can be seen that Equation 3.1.1 is similar in form to Equation 2.2.1, where \(1/2Z_d g_{mn+1}\) is equivalent to the transversal filter tap weight values, \(w_n\) and \(\tau_g + \tau_d\) combine to provide the tap delay, \(\tau\). Hence, by redefining the output of the distributed amplifier to be located at the end of the output transmission line opposite the DA's defined output a transversal filter with bandwidth capabilities large enough for lightwave signal processing can be created.
3.2 Tap Delay Implementation using Lumped Element Transmission Line Approximation

The distributed amplifier based transversal filter described in Section 3.1 utilized transmission line segments in order to introduce the required delay between the gain stages. If the desired delay is long the required length of the transmission line will also be very long, and as a result the overall size of the physical circuit will be large. Lumped elements can be arranged in such a fashion so that a desired electrical length and impedance of transmission line can be simulated and very often can take up less circuit space. Consider the periodic lumped element structure illustrated in Figure 3.2.1a). It consists of a connection of unit cells that are comprised of two shunt capacitors, $C/2$ on either side of a series inductor, $L$. The cell can be analyzed by considering it as a two-port network and by applying the transmission or ABCD matrix [29] method of analysis to it. For this unit cell $V_n$, $I_n$, $V_{n+1}$ and $I_{n+1}$ are related according to,

$$
\begin{bmatrix}
V_n \\
I_n
\end{bmatrix} =
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix}
\begin{bmatrix}
V_{n+1} \\
I_{n+1}
\end{bmatrix} =
\begin{bmatrix}
1 - \frac{\omega^2 LC}{2} & \frac{j\omega L}{2} \\
\frac{j\omega C}{4} & 1 - \frac{\omega^2 LC}{2}
\end{bmatrix}
\begin{bmatrix}
V_{n+1} \\
I_{n+1}
\end{bmatrix}
$$

3.2.1

Since it is intended for the unit cell to create an artificial transmission line propagating solutions to Equation 3.2.1 are desired such that,

$$
V_{n+1} = V_n e^{\gamma d} \quad 3.2.2a)
$$

$$
I_{n+1} = I_n e^{\gamma d} \quad 3.2.2b)
$$
where $\gamma$ is the effective propagation constant and $d$ is the effective length of the transmission line being approximated with the lumped element unit cell. Substitution of Equations 3.2.2a) and b) into Equation 3.2.1 yields,

$$\begin{bmatrix} A-e^{\gamma d} & B \\ C & D-e^{\gamma d} \end{bmatrix} \begin{bmatrix} V_{n+1} \\ I_{n+1} \end{bmatrix} = 0 \quad 3.2.3$$

The nontrivial solution for Equation 3.2.3 can be found by taking the determinant of the first matrix on the left side and setting it equal to zero. This yields,

$$1 + e^{2\gamma d} - (A + D)e^{\gamma d} = 0 \quad 3.2.4$$

where the fact that $AD-BC=1$ was applied since the unit cell is reciprocal. Equation 3.2.4 can then be simplified to,

$$\cosh(\gamma d) = \frac{A + D}{2} \quad 3.2.5$$

By breaking the effective propagation constant down into its real and imaginary components such that $\gamma=\alpha+j\beta$, and using the values for $A$ and $D$ given in Equation 3.2.1 Equation 3.2.5 can be transformed to,

$$\cosh(\alpha d)\cos(\beta d) + j \sinh(\alpha d)\sin(\beta d) = \frac{A + D}{2} = 1 - \frac{\omega^2 LC}{2} \quad 3.2.6$$
Figure 3.2.1: a) Artificial transmission line implementation using lumped element unit cells and b) the corresponding effective characteristic impedance, $Z_t$ (normalized to $(L/C)^{1/2}$) and c) the unit cell time delay (normalized to $(LC)^{1/2}$) vs. normalized frequency parameter $k = \omega(LC)^{1/2}$. The transmission line approximation is reasonably accurate provided the operational frequency is less than $f_c/2$. 
Since the right hand side of Equation 3.2.6 is entirely real there are two cases which can be considered in order to satisfy the relationship; \( \alpha = 0 \), \( \beta \neq 0 \) and \( \alpha \neq 0 \) and \( \beta = 0 \). Since the solution which propagates is of interest the case that will be considered here is when \( \alpha = 0 \) and \( \beta \neq 0 \). Therefore Equation 3.2.6 becomes,

\[
\cos(\beta d) = 1 - \frac{\omega^2 LC}{2} \quad 3.2.7
\]

which can be solved for \( \beta \) provided the magnitude of the right hand side is less than 1.

The cutoff frequency, \( f_c \) beyond which real solutions for \( \beta \) can no longer be found and therefore signal propagation no longer occurs through the unit cell. The cutoff frequency occurs at,

\[
f_c = \frac{1}{\pi \sqrt{LC}} \quad 3.2.8
\]

where \( L \) and \( C \) are the values of the lumped elements in the unit cell in Figure 3.2.1a).

The impedance of the unit cell can be found according to the ratio of \( V_n \) to \( I_n \) and is calculated as,

\[
Z_L = \sqrt{\frac{L}{C}} \left( \sqrt{1 - \frac{\omega^2 LC}{4}} \right)^{-1} = \sqrt{\frac{L}{C}} \left( \frac{k^2}{4} \right)^{-1} \quad 3.2.9
\]

where \( Z_L \) is effective characteristic impedance and \( k \) is the normalized frequency parameter \( \omega^2 LC \). The normalized effective impedance (normalized by \( \sqrt{L/C} \)) of the unit
cell is plotted in Figure 3.2.1b) as a function of normalized frequency, $k$. Provided the signal frequency is below one half of the cutoff frequency ($f_c/2$ or $k=1$) the effective characteristic impedance of the unit cell deviates from its nominal zero frequency value of $\sqrt{L/C}$ by no greater than 15%.

The phase delay associated with the unit cell is,

$$\phi = \beta d = \arccos \left( 1 - \frac{k^2}{2} \right) \quad 3.2.10$$

and the corresponding delay is,

$$\tau = \frac{\partial \phi}{\partial \omega} = \sqrt{LC} \left( \sqrt{1 - \frac{k^2}{4}} \right)^{-1} \quad 3.2.11$$

The unit cell time delay is plotted in Figure 3.2.1c) as a function of normalized frequency, $k$. As in the case for the effective impedance the time delay of the unit cell deviates by no more than 15% from its nominal zero frequency value of $\sqrt{LC}$ provided the signal frequency is below one half of the cutoff frequency ($f_c/2$ or $k=1$).

Hence, provided the signal frequency is maintained to below $f_c/2$ and the appropriate lumped element values for $L$ and $C$ are used an artificial transmission line with a desired effective characteristic impedance and unit cell time delay can be created using a series of lumped element unit cells as illustrated in Figure 3.2.1a). The transversal filter designed in this thesis utilizes this lumped element transmission line.
approximation technique in order to implement the required tap delays. The reason for this is that the area of the circuit can be greatly reduced. For example, for a tap delay of 50ps, used for the transversal filter designed here the required length of coplanar transmission line on GaAs is approximately 6mm while the lumped element structure achieves the same delay in a length of less then 200μm.

3.3 Positive and Negative Adjustable Tap Weight Implementation using the Gilbert Cell

In order to increase the versatility of the transversal filter which has only positive adjustable tap weight values [9] while maintaining the same number of taps, positive and negative tap weight capability is required. By implementing each tap weight in the transversal filter with a Gilbert Cell this requirement is achieved. A MESFET implementation of the Gilbert Cell is illustrated in Figure 3.3.1a. Generally, the Gilbert Cell is used to perform mixing operations [30], however through proper implementation it can provide an effective transconductance which can not only change its amplitude but also its sign.

The Gilbert Cell is comprised of an input differential pair (IDP), consisting of two matched transistors (xj1-xj2) and two matched differential pair loads (DPLs) (xj3-xj4 and xj5-xj6). In the DC analysis of the circuit in Figure 3.3.1a) the input differential pair formed by xj1 and xj2 each draw a current of I/2 through their respective differential pair loads (DPL). In this case the DPL connected to the drain of xj1 is formed with xj3 and xj4 and the DPL connected to the drain of xj2 is formed with xj5 and xj6.
Figure 3.3.1: a) The Gilbert Cell constructed with MESFETs and b) the DC current relationship between the transistors in each differential pair load vs. $V_{\text{gain}}$. 

$(V_{\text{ref2}}=7.5V, \beta=0.02A/V^2 \text{ and } I_s=20mA)$
Provided the transistors are operating in saturation mode it can be shown that the current in \( xj3 \) and \( xj4 \) will be [31],

\[
I_{xj3} = \frac{I_b}{4} + \sqrt{\beta_{xj3}I_b}\left(\frac{V_{\text{gain}} - V_{\text{ref}2}}{2}\right)\sqrt{1 - \frac{\beta_{xj3}(V_{\text{gain}} - V_{\text{ref}2})^2}{I_b}} \tag{3.3.1 a)}
\]

\[
I_{xj4} = \frac{I_b}{4} - \sqrt{\beta_{xj4}I_b}\left(\frac{V_{\text{gain}} - V_{\text{ref}2}}{2}\right)\sqrt{1 - \frac{\beta_{xj4}(V_{\text{gain}} - V_{\text{ref}2})^2}{I_b}} \tag{3.3.1 b)}
\]

where \( \beta_{xjn} \) is the conductance parameter for the MESFET \( xjn \). The total current through \( xj3 \) and \( xj4 \) is always \( I_b/2 \), however as can be inferred from the Equations 3.3.1a) and b), depending on the relationship between \( V_{\text{gain}} \) and \( V_{\text{ref}2} \) the amount of current in \( xj3 \) and \( xj4 \) is not necessarily the same. In fact, if \( V_{\text{gain}} \) is greater than \( V_{\text{ref}2} \) by at least \( (I_b/2\beta)^{1/2} \) then all of the current, \( I_b/2 \) will flow through \( xj3 \) and \( xj4 \) will be turned off. In the same respect if \( V_{\text{gain}} \) is less than \( V_{\text{ref}2} \) by at least \( (I_b/2\beta)^{1/2} \) then all of the current, \( I_b/2 \) will flow through \( xj4 \) and \( xj3 \) will be turned off. Hence, the amount of current that flows through each transistor is controlled according the relationship between \( V_{\text{gain}} \) and \( V_{\text{ref}2} \). By adjusting the value of \( V_{\text{gain}} \) relative to \( V_{\text{ref}2} \) the current through \( xj3 \) (\( xj4 \)) can be altered between 0 and \( I_b/2 \) (\( I_b/2 \) and 0). As can be seen from the symmetry in the Gilbert Cell identical current relationships can be derived for transistors \( xj5 \) and \( xj6 \) where the expression for \( xj5 \) is the same as \( xj4 \) and the expression for \( xj6 \) is the same as \( xj3 \). Figure 3.3.1b) displays the DC current relationship between \( I_{xj3} \) (\( I_{xj6} \)) and \( I_{xj4} \) (\( I_{xj5} \)) for the Gilbert Cell as a function of \( V_{\text{gain}} \) for \( V_{\text{ref}2}=7.5V, \beta=0.02A/V^2 \) and \( I_b=20mA \). Control over the amount of current that
flows through each transistor in each of the DPLs becomes important in the small signal analysis of the Gilbert Cell.

In order to demonstrate the AC operation of the Gilbert Cell a small signal analysis must be performed. Figure 3.3.2a) illustrates the equivalent small signal diagram for the Gilbert Cell. In this diagram the small signal hybrid-\( \pi \) model [31] has been used for transistors \( xj1 \) and \( xj2 \) while the \( T \) model [31] has been used for each of the transistors in the DPLs. Furthermore, low frequency versions of these models are used in order to simplify the analysis. As can be seen in Figure 3.3.2a), the signal source drives the input differential pair (IDP) from one side and causes a small signal current, \( i_{sig} \) to flow with a magnitude equivalent to,

\[
i_{sig} = g_{m_{xj1}} \frac{v_{in}}{2} \tag{3.3.2}
\]

where \( g_{m_{xj1}} \) is the transconductance of the matched transistors \( xj1 \) and \( xj2 \). The IDP drives \( i_{sig} \) through each of the DPLs with a phase difference of \( 180^0 \). \( i_{sig} \) is split between the two transistors in each DPL according to the relationship between their transconductances. The transconductance of each transistor in each DPL depends on the DC current which flows through each of the transistors which as stated earlier depends on the relationship between \( V_{gain} \) and \( V_{ref2} \). By using the current divider rule the fraction of the total signal current, \( i_{sig} \) through each DPL transistor is,

\[
i_{xj3} = i_{sig} \left( \frac{g_{m_{xj3}}}{g_{m_{xj3}} + g_{m_{xj4}}} \right) \tag{3.3.3a)}
\]
Figure 3.3.2: a) Equivalent small signal circuit for the Gilbert Cell and b) the small signal transconductance parameter $g_m$ of the Gilbert Cell vs. $V_{gain}$

$(V_{ref2}=7.5V, \beta=0.02A/V^2$ and $I_s=20mA)$
where $g_{m_{nj}}$ is the transconductance of $nth$ transistor. Hence, the output current, $i_{out}$ that is drawn through the load resistor, $R_l$ is

$$i_{out} = i_{y3} + i_{y5} = i_{sig} \left( \frac{g_{m_{y3}}}{g_{m_{y3}} + g_{m_{y4}}} - \frac{g_{m_{y5}}}{g_{m_{y5}} + g_{m_{y6}}} \right)$$

Due to the symmetry of the Gilbert Cell and the fact all the transistors in the DPLs are matched the transconductance of $xj3$ and $xj6$ are equivalent as well as the transconductance of $xj4$ and $xj5$. Hence, the small signal output current, $i_{out}$ can be simplified to,

$$i_{out} = \frac{v_{in}}{2} g_{m_{y1}} \left( \frac{g_{m_{y3}} - g_{m_{y5}}}{g_{m_{y3}} + g_{m_{y4}}} \right)$$

where Equation 3.3.2 has been substituted for $i_{sig}$.

The small signal transconductance for a saturated MESFET transistor is [23],

$$g_{m_{jn}} = 2 \sqrt{I_{s}} \beta_{jn}$$
where $I_{xjn}$ is the DC current flowing through the transistor $xjn$ and $\beta_{xjn}$ is the MESFET conductance parameter of the transistor $xjn$. Therefore, the small signal output current of the Gilbert Cell can be written as,

$$i_{out} = -\sqrt{(I_b/2)\beta_{xjl}}\left(\frac{\sqrt{I_{xj3}\beta_{xj3}} - \sqrt{I_{xj5}\beta_{xj5}}}{\sqrt{I_{xj3}\beta_{xj3}} + \sqrt{I_{xj4}\beta_{xj4}}}\right)v_in \quad 3.3.7$$

where $I_b/2$ is the DC current flowing through $xjl$.

This expression can be simplified since the conductance parameters for each of the MESFETs in the DPLs are assumed equivalent. Furthermore, since $xj4$ and $xj5$ have equivalent conductance parameters and DC bias conditions due to symmetry it can be seen that $I_{xj4}$ and $I_{xj5}$ must be equivalent. Hence Equation 3.3.7 can be rewritten as,

$$i_{out} = -\sqrt{(I_b/2)\beta_{xjl}}\left(\frac{\sqrt{I_{xj3}} - \sqrt{I_{xj4}}}{\sqrt{I_{xj3}} + \sqrt{I_{xj4}}}\right)v_in \quad 3.3.8$$

and can be further simplified to,

$$i_{out} = -G(V_{gain})\sqrt{(I_b/2)\beta_{xjl}}v_in \quad 3.3.9$$

where $G(V_{gain})$ is the small signal gain factor and is expressed as,

$$G(V_{gain}) = \frac{\sqrt{I_{xj3}} - \sqrt{I_{xj4}}}{\sqrt{I_{xj3}} + \sqrt{I_{xj4}}} \quad 3.3.10$$
Finally, the small signal transconductance, $g_m$ for the entire Gilbert Cell can be determined as,

$$g_m = -\frac{i_{out}}{v_{in}} = G(V_{gain})\sqrt{(I_b/2)\beta_{sj}}$$ \hspace{1cm} 3.3.11

In Equations 3.3.9, 3.3.10 and 3.3.11 $G(V_{gain})$ is the small signal gain factor which, as is evident from Equation 3.3.10 can be adjusted according to the relationship between the DC bias current which flows through each transistor of the DPLs. From Equation 3.3.1 a) and b) it can be seen that the DC current through $xj3(xj6)$ and $xj4(xj5)$ is controlled according to the relationship between the DC bias voltages $V_{gain}$ and $V_{ref2}$. As outlined in the DC circuit analysis, depending on this relationship the current through $xj3$ can vary between 0 and $I_b/2$ and the current through $xj4$ will correspondingly vary between $I_b/2$ and 0 in order to maintain a total current of $I_b/2$ through the DPL. This results in $G(V_{gain})$ taking on values between 1 and -1. Therefore, with $V_{ref2}$ fixed, $G(V_{gain})$ and therefore $g_m$ can be controlled according to the value of $V_{gain}$.

Figure 3.3.2b) displays the small signal transconductance, $g_m$ for the Gilbert Cell as a function of $V_{gain}$ for a fixed $V_{ref2}$ ($V_{ref2}=7.5V$, $\beta=0.02A/V^2$ and $I_b=20mA$). It can be seen that the transconductance can be varied between approximately -14mmho and 14mmho as $V_{gain}$ is adjusted. Hence, by implementing the Gilbert Cell in this fashion so that its small signal gain transconductance can be electronically altered within a range that covers both positive and negative values, positive and negative adjustable tap weights are achieved in the transversal filter designed.
CHAPTER 4

MMIC IMPLEMENTATION OF THE TRANSVERSAL FILTER

In this chapter the design techniques that were used to implement the 5-tap transversal filter with positive and negative adjustable tap weights are discussed. The transversal filter layout is presented and component modeling techniques that were used are described. Finally, simulated results are presented demonstrating the feasibility of the design as well as the transversal filter MMIC's positive and negative adjustable tap weight capability.

4.1 The Distributed Amplifier Based 5-tap Transversal Filter with Positive and Negative Adjustable Tap Weights

Figure 4.1.1a) presents the schematic diagram of the 5-tap transversal filter MMIC with positive and negative tap weight capability that was designed. As outlined in Sections 3.1 – 3.3 the transversal filter design is based on the distributed amplifier architecture and uses lumped elements for the tap delays and the Gilbert Cell for the adjustable tap weights. The lumped element input and output delay lines and the Gilbert Cell gain elements are all labeled in Figure 4.1.1a). Notice from Figure 4.1.1b) that the load resistor, $R_l$ in the Gilbert Cell outlined in Section 3.3 is removed and the output is connected directly to the output delay line. The Gilbert Cell input is connected directly to the input delay line. An analysis of this transversal filter yields an output of,
where it is assumed that the delay introduced by each Gilbert Cell is negligible, \( Z_{\text{load}} \) is real and matched to the output delay line and the operational frequency is less than half of the cutoff frequency for the lumped element delay lines. In Equation 4.4.1 \( \tau_g \) and \( \tau_d \) are the respective delays introduced by the input and output delay line segments between each Gilbert Cell tap weight. \( \tau_{go} \) and \( \tau_{do} \) are the respective delays introduced by the input and output delay line segments that exist before the first Gilbert Cell tap weight. \( g_{mn+1} \) is the transconductance of the \( n+1 \)th Gilbert Cell tap weight. Apart from the constant delay term, \( \tau_{go}+\tau_{do} \) the form of Equation 4.1.1 is the same as the expression for the transversal filter output presented in Equation 3.1.1 and Equation 2.2.1 and therefore the design architecture is validated. The constant delay term, \( \tau_{go}+\tau_{do} \) in Equation 4.1.1 has no effect on the filter characteristics and simply introduces a constant delay in its response.

The total effective tap delay, between each of the tap weights consists of the delay between successive tap weights along the input delay line, \( \tau_g \) and the delay between successive tap weights along the output delay line, \( \tau_d \). The transversal filter designed here intended for each delay segment to introduce 25ps of delay so that the overall effective tap delay would be 50ps. That is,

\[
v_{out}(t) = -\frac{1}{4} Z_{\text{load}} \sum_{n=0}^{4} g_{mn+1} v_{in}(t - (\tau_{go} + \tau_{do} + n(\tau_g + \tau_d))) \quad 4.1.1
\]

\[
\tau_g = \tau_d = 25 \text{ps} \quad 4.1.2
\]
\[ \tau = \tau_g + \tau_d = 50 \text{ps} \quad 4.1.3 \]

Therefore, the transversal filter designed is a fractionally spaced equalizer (FSE) when considering its use as an equalizer in any system operating at a bit rate below 20\text{Gb/s}. Furthermore, in addition to the intended tap delay of 50ps, the intended effective characteristic impedance of the input and output lumped element delay lines was 50\Omega so that a good input and output match could be achieved for the filter.

4.2 Circuit Component Modeling

Every circuit element has parasitic elements associated with it that become especially important when considering the operation of the element at high frequencies. For this reason models were generated for all of the circuit elements used in the transversal filter so that the operation of the transversal filter could be accurately predicted through HSPICE simulations. The models used for these elements as well as how the appropriate model values were generated will be outlined in this section.

4.2.1 Metal Interconnect, Resistor and Capacitor Models

The parasitics associated with interconnects, resistors and capacitors were approximated for by using transmission line theory. The type of transmission line best suited for modeling the element under consideration was first chosen according to the geometry of the element and its surroundings in the circuit. Once the type of transmission line model to use to approximate the element was decided upon (in most microstrip or coplanar waveguide transmission lines were used) the cross sectional dimensions
Figure 4.1.1: a) Schematic diagram of the distributed amplifier based 5-tap transversal filter with positive and negative tap weight capability b) Gilbert Cell tap weight connection.
were measured and the substrate parameters were determined. Using Eesof's LineCalc routine the characteristic impedance, $Z_o$ and effective dielectric constant, $\varepsilon_{\text{eff}}$ of the transmission line were determined. Once these parameters were determined an appropriate lumped element transmission line approximation model could be formed and the associated parasitic values determined. Using $Z_o$ and $\varepsilon_{\text{eff}}$ for the transmission line geometry chosen, the capacitance per unit length, $C_i$ and inductance per unit length, $L_i$ were determined according to [29],

$$C_i = \frac{\sqrt{\varepsilon_{\text{eff}}}}{cZ_o} \quad 4.2.1a)$$

$$L_i = \frac{Z_o \sqrt{\varepsilon_{\text{eff}}}}{c} \quad 4.2.1b)$$

where $c$ is the speed of light in free space. The total parasitic capacitance and inductance associated with the element that is being modeled can then be determined as,

$$C_p = C_i l \quad 4.2.2a)$$

$$L_p = L_i l \quad 4.2.2b)$$

where $l$ is the length of the element being modeled. Equations 4.2.1-4.2.2 are derived from the transmission line equations (which are derived using circuit theory) and come from the fact that a length of transmission line can be modeled as a configuration of
lumped elements provided the length of the transmission line is a lot less than the wavelength of operation [29].

Figure 4.2.1 illustrates the lumped element model used for interconnects and resistors. In this case the total parasitic capacitance, $C_p$ is split in half and placed at either end of the series connection of the parasitic inductance and resistance. This allowed for a better approximation of the distributed nature of the element. The resistance value is determined according to,

$$R = R_s \frac{l}{W} \quad 4.2.3$$

where $l$ and $W$ are the respective length and width of the element being modeled and $R_s$ is the sheet resistance of the metal interconnect or the resistor measured in $\Omega$ square.

![Figure 4.2.1 a) Metal interconnect and resistor lumped element model b) capacitor lumped element model. Both models are derived using transmission line approximations.](image)

The lumped element model used for capacitors can be seen in Figure 4.2.1b). The model is similar to that used for interconnects and resistors, but it has a capacitor in series with the parasitic resistance and inductance associated with the component. This
capacitor, $C$ is assigned the actual capacitance value of the component being modeled. The series resistance, $R$ associated with the capacitor model can be determined using Equation 4.2.3 and by using the dimensions of one of the metal plates of the actual capacitor.

4.2.2 Planar Spiral Inductor Model

The model used for the planar spiral inductors in the delay line is illustrated in Figure 4.2.2b). The values for the components in the model were determined using a two step procedure [9]. Initially the inductor was considered as a solid piece of rectangular metal that could be approximated as a coplanar transmission line. This allowed for the parasitic capacitance, $C_p$ associated with the distributed nature of the inductor to be calculated in the same manner that it was determined for the metal interconnects, resistors and capacitors in Section 4.2.1. Next the crossover capacitance, $C_w$ which represents the capacitance between the inductor lead that extends from the center of the inductor and the windings that lie underneath it was calculated according to [32],

$$C_w = 8.854 \times 10^{-18} \left\{ \frac{Wl}{h} + 1.393(l + W) + \frac{2}{3} \left[ l \ln \left( \frac{W}{h} + 1.444 \right) + W \ln \left( \frac{l}{h} + 1.444 \right) \right] \right\} \quad 4.2.3$$

where $h$ is the distance between the metal forming the air bridge (inductor lead) and the underlying metal of the inductor windings, $W$ is the width of the inductor lead and $l$ is the length of the overlap. All of these dimensions are in microns. The parasitic resistance, $R$ was determined using Equation 4.2.2. The value used for $L$ in the lumped element model was the desired value required for the lumped element delay line design. Once the
lumped element model values were estimated the geometry of the rectangular spiral inductor and the associated lumped element model were entered into Academy. The Academy routine then optimized the inductor geometry and the associated lumped element model in order to achieve an accurate match between the two while maintaining the desired inductance, \( L \). Once an accurate match was achieved the geometry of the inductor was implemented in the circuit and its corresponding lumped element model could be used in HSPICE simulations.

![Diagram](image)

Figure 4.2.2: a) Rectangular spiral inductor and b) its associated lumped element model. The actual inductor geometry and the model were optimized to have equivalent responses.

### 4.2.3 Bondwires

Bondwires are required to connect the MMIC to the real world. They can be modeled using the transmission line model illustrated in *Figure 4.2.1a*). In this case the bondwire can be modeled as a microstrip transmission of width \( W \) (that is equivalent to the diameter of the wire, \( d \) and length \( l \) that sits in air \( (\varepsilon_r=1) \) an average distance of \( h \)
above a ground plane. The capacitance values for the model in Figure 4.2.1a) were determined in the same manner that the capacitance values for an interconnect, resistor or capacitor were determined in Section 4.2.1. The series inductance, $L$ was determined according to [33],

$$L = 2 \times 10^{-4} \left[ \ln \frac{4h}{d} + \ln \left( \frac{l + \sqrt{l^2 + d^2 / 4}}{l + \sqrt{l^2 + 4h^2}} \right) + \sqrt{1 + \frac{4h^2}{l^2}} - \sqrt{1 + \frac{d^2}{4l^2}} - 2 \frac{h}{l} + \frac{d}{2l} \right]$$  \hspace{1cm} 4.2.4

and the series resistance, $R$ was calculated as,

$$R = \frac{4\rho l}{md^2}$$  \hspace{1cm} 4.2.5

where $\rho$ is the resistivity of the bondwire.

4.2.4 Transistor Modeling and Gilbert Cell Input and Output Capacitance

The transversal filter was implemented using Nortel's 0.8-$\mu$m Self Aligned Gate MESFET process. The model used for the MESFET transistor was based on the Raytheon MESFET model and appropriate model parameter values were extracted from measured s-parameter data. The schematic of the model deviates only slightly from the common high frequency small signal transistor model [31]. This model was used in the HSPICE simulations and therefore an accurate representation of the transistor operation could be predicted during simulations.
The primary concern with respect to the design of the lumped element delay lines was the behaviour of the input and output capacitance of the Gilbert Cell with changes in transconductance control voltage, $V_{\text{gain}}$. This was important since the delay lines were designed to incorporate the input and output capacitance of the Gilbert Cell gain elements. If the input and output capacitance were to change drastically with biasing, the properties of the input and output delay line would lead to a variation in tap delay values and the characteristic impedance of the line. Such variations would ultimately effect the desired performance of the transversal filter. Fortunately, due to the inherent structure of the Gilbert Cell the input and output capacitance changes very little (relative to that of a common source gain element) when the transconductance of the circuit is adjusted. As illustrated in Section 3.3 the transconductance of the Gilbert Cell is adjusted by adjusting $V_{\text{gain}}$. Therefore it was important to see how the input and output capacitance of the Gilbert Cell varied over the useful range of $V_{\text{gain}}$. For the Gilbert Cell biasing used in this design the useful range for $V_{\text{gain}}$ is between 6V and 8V. Figure 4.2.4 illustrates the input and output capacitance of the Gilbert Cell versus $V_{\text{gain}}$.

![Figure 4.2.4: Gilbert Cell input and output capacitance versus $V_{\text{gain}}$.](image)

50
It can be seen that the input and output capacitance of the Gilbert Cell used in this design varies by less than 5% and 10% respectively over the useful range for $V_{gain}$. Such stability in input and output capacitance made the implementation of the lumped element input and output delay lines far less complicated.

4.3 Input and Output Lumped Element Delay Line Implementation

As can be seen in Figure 4.1.1 the input and output lumped element delay lines are implemented with inductors and capacitors. The inductors used in the delay are the rectangular spiral inductors introduced in Section 4.2.2. The capacitors in the delay line are comprised of parallel plate capacitors, the input and output capacitance of the Gilbert Cell tap weight and the parasitic capacitance associated with the rectangular spiral inductor. The actual design of the lumped element input and output delay lines was performed by Jamani [9][27] who designed a 5-tap transversal filter using cascode MESFETs for the tap weights was designed. Measured results for the transversal filter designed by Jamani indicated that the delay lines introduced the desired tap delay of 50ps and also provided good input and output matching over a broad frequency range and therefore, the author did not see any point in revising the delay lines. In order to conform to the requirements imposed by the lumped delay line designed by Jamani the input and output capacitance of the Gilbert Cell gain element was matched to the input and output capacitance of the cascode MESFET configuration used by Jamani. The input and output capacitance was adjusted according to the width selection and the biasing of the transistors which make up the Gilbert Cell.

The input and output lumped element delay lines in the transversal filter
presented in Figure 4.1.1a) are broken down into their unit cells in Figure 4.3.1a) and b). The parasitic resistance and the inductor crossover parasitic capacitance, $C_w$ present in the inductor model of Figure 4.2.2b) and the inductor and resistor parasitic elements in the capacitor model of Figure 4.2.1a) have been neglected for the purpose of simplifying the illustration of the delay line.

Figure 4.3.1a) illustrates the lumped element input delay line. $C_g'$ represents the excess capacitance introduced by the parallel plate capacitor $C_{go}$ from Figure 4.1.1a). The value of $C_{go}$ was determined to be incorrect in Jamani's [9] and was carried over into this design. This error was determined after the transversal filter MMIC design was submitted for fabrication. Fortunately, since Jamani's [9] measured results were not significantly affected by this error it was expected that the measured results for this design would also not be affected. $C_g$ represents the combined capacitance of a parallel plate capacitor and the parasitic capacitance associated with each of its nearest neighbor inductors whose value was determined through the inductor modeling discussed in Section 4.2.2. $C_{in}$ represents the combined capacitance that includes the input capacitance, $C_{inGC}$ of the Gilbert Cell and the parasitic capacitance associated with its nearest neighbor spiral inductors.

In Figure 4.3.1b) $C_d'$ represents the excess capacitance introduced by the parallel plate capacitor $C_{do}$ from Figure 4.1.1b). As in the case of $C_{go}$, the size of $C_{do}$ is also incorrect, but was not expected to have a significant impact on results. $C_d$ represents a combined capacitance comprised of a parallel plate capacitor and the parasitic capacitance of the nearest neighbor inductors whose value was determined through the inductor modeling discussed in Section 4.2.2. $C_{out}$ represents the combination of the
Gilbert Cell output capacitance, $C_{outGC}$, an additional parallel plate capacitance, $C_{adj}$ required to compensate for the lower output capacitance (relative to the input capacitance) of the Gilbert Cell and also the parasitic capacitance of the nearest neighbor spiral inductors.

Since the desired delay between successive tap weights along each delay line was 25$ps$ it was necessary for each unit cell to provide a delay of 12.5$ps$. Furthermore, a characteristic impedance of 50$\Omega$ was desired in order to provide good transversal filter input and output matching. Jamani used a unit cell inductance, $L$ and total unit cell capacitance, $C$ of 600$\mu$H and 260$\mu$F respectively. Using the lumped element delay line theory presented in Section 3.2 these values translate to a unit cell delay of 12.5$ps$, an impedance of 48$\Omega$ and a cutoff frequency of 25.5GHz.

By comparing the generic unit cell illustrated in Figure 3.2.1a) with lumped element values assigned as $L=600\mu$H and $C=260\mu$F the necessary values for the reactive elements which comprise the unit cells in the input and output delay lines were determined. Table 4.3.1 summarizes the correct and actual values for the inductors and capacitors that make up the lumped element input and output delay lines for the transversal filter designed. In order to provide the necessary effective capacitance in each unit cell the input capacitance of the Gilbert Cell gain element should be set to 200$\mu$F and the output capacitance of the gain element should be 60$\mu$F. Jamani’s discussion on his implementation of the lumped element delay lines gave the perception that the necessary input capacitance of the gain cell, $C_{inGC}$ was 160$\mu$F and the necessary output capacitance of the gain cell, $C_{outGC}$ was 60$\mu$F. Hence, the widths and biasing of the transistors in the Gilbert Cell were appropriately adjusted in order to conform to these requirements.
Unfortunately, the necessary input capacitance of 200fF was determined after the design had been submitted for fabrication during a more comprehensive review of Jamani’s thesis [9]. Fortunately, as is evident in the simulated results presented in Section 4.5 and the Jamani’s measured results the error did not have a large impact on the performance of the input and output lumped element delay lines.

<table>
<thead>
<tr>
<th>Input Delay Line Parameters</th>
<th>Correct Value</th>
<th>Actual Value</th>
<th>Output Delay Line Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>( L )</td>
<td>600pH</td>
<td>600pH</td>
<td>( L )</td>
</tr>
<tr>
<td>( C_g ) ((\text{includes parasitics}))</td>
<td>260fF</td>
<td>260fF</td>
<td>( C_d ) ((\text{includes parasitics}))</td>
</tr>
<tr>
<td>( C_{in} ) ((\text{includes } C_{aGC} \text{ and parasitics}))</td>
<td>260fF</td>
<td>260fF</td>
<td>( C_{out} ) ((\text{includes } C_{aGC}, C_{dadj} \text{ and parasitics}))</td>
</tr>
<tr>
<td>Gilbert Cell Input Capacitance ((C_{inGC}))</td>
<td>200fF</td>
<td>160fF</td>
<td>Gilbert Cell Output Capacitance ((C_{outGC}))</td>
</tr>
<tr>
<td>( C_g' ) ((\cdot C_{RD}))</td>
<td>0fF ((130fF))</td>
<td>100fF ((230fF))</td>
<td>( C_d' ) ((\cdot C_{RD}))</td>
</tr>
<tr>
<td>( C_{dadj} )</td>
<td>150fF</td>
<td>150fF</td>
<td>( C_{dadj} )</td>
</tr>
</tbody>
</table>

Table 4.3.1: Reactance values used in the lumped element delay line implementations

All parameter values are illustrated in Figure 4.1.1a or Figure 4.3.1a and b)
Figure 4.3.1: Lumped element delay lines broken down into their unit cells. a) input delay line b) output delay line. Note that $C_{q'} = C_{q2} - C_q/2$ and $C_{d'} = C_{d2} - C_d/2$. $C_{q2}$ and $C_{d2}$ are illustrated in Figure 4.1.1a)
4.4 Transversal Filter MMIC Layout

The layout of the 5-tap transversal filter designed is presented in Figure 4.4.1. The layout was performed using Cadence Analog Artist. The filter was designed in a coplanar format in order to avoid the ground vias required with microstrip layouts. Vias are expensive and also have a parasitic inductance and resistance associated with them that can hinder the overall circuit performance. In Figure 4.4.1 it can be seen that the signal enters at the input of the filter and travels along a 50Ω coplanar waveguide (CPWG) before it is transferred to the 50Ω lumped element input delay line. As the signal travels through the delay line towards the input delay line matched termination each Gilbert Cell tap weight taps off some of the energy from the signal and amplifies it. The gain (or transconductance) of each tap weight is controlled electronically according to the voltage applied to its corresponding tap weight value control pad. In Figure 4.4.1 the 2nd Gilbert Cell gain element and its corresponding tap weight control are labeled. The amplified signal is injected into the 50Ω lumped element output delay line and it travels in both directions; towards the 50Ω output CPWG and towards the matched output delay line termination. Both the input and output delay line matched terminations insure that there are no reflections from the end of either delay line which would otherwise interfere with the intended operational characteristics of the filter.
4.5 Simulated Results

The simulations for the transversal filter designed were performed using HSPICE and included all the parasitic elements associated with each circuit component determined using the techniques described in Section 4.2. As can be seen in Figure 3.3.2b) the transconductance of each Gilbert Cell gain element in the transversal filter can be varied according to the value of $V_{gain}$ that is applied to it. With $V_{ref}$ set at 7.5V the transconductance of the Gilbert Cell has a value of approximately -14mmho when $V_{gain}$ is...
7V and a value of approximately 14 mmho when \( V_{\text{gain}} \) is 8V. Therefore, the tap weight convention was set so that a tap weight setting of -1 corresponded to an applied \( V_{\text{gain}} \) of 7V and a transconductance of -14 mmho and a tap weight setting of 1 corresponded to an applied \( V_{\text{gain}} \) of 8V and a transconductance of 14 mmho. A tap weight value between -1 and 1 would indicate that \( V_{\text{gain}} \) is between 7V and 8V and the transconductance is between -14 mmho and 14 mmho. A tap weight setting of 0 corresponds to an applied \( V_{\text{gain}} \) of 7.5V so that the transconductance of that tap weight has a value of zero. In order to describe the entire state of the transversal filter in terms of all of its tap weight settings the weighting vector can be defined as \( W = [w_1, w_2, w_3, w_4, w_5] \), where \( w_n \) corresponds to the tap weight setting of the \( n \)th tap weight. For example when the transversal filter has all of its tap weights set at the maximum positive position each of the five taps has a \( V_{\text{gain}} \) of 8V applied to it and \( W = [1,1,1,1,1] \). Conversely, when each of the five taps has a \( V_{\text{gain}} \) of 7V applied to it \( W = [-1,-1,-1,-1,-1] \).

Simulated results for S11 and S22 of the transversal filter for tap weight settings of \( W = [1,1,1,1,1] \) and \( W = [1,0,0,0,1] \) are illustrated in Figure 4.5.1. It can be seen that S11 is less than 15dB out to 7.4GHz and S22 is less than 15dB beyond 15GHz indicating that the design of the input and output lumped element delay lines provides good matching despite the design error discussed in Section 4.4. The fact that the curve for S11 and S22 remain virtually unchanged for the two different tap weight configurations further confirms that there is little variation in the input and output capacitance of each Gilbert Cell gain element when the tap weights are changed. This is a desirable feature since the input and output matching of the entire transversal filter can be maintained for any tap weight configuration imposed on it.
The simulated responses of the transversal filter for tap weight settings of 
W=[0,0,1,0,0], [1,1,1,1,1] and [-1,-1,-1,-1,-1] are illustrated in Figure 4.5.2. It is evident 
that the transversal filter response is indeed adjusted according to the tap weight settings. 
It can be seen that the transversal filter is capable of performing both low pass and band 
pass filtering on signals with a frequency spectrum less than 13 GHz. The transversal 
filters designed by Jamani [9][27] which utilizes only positive adjustable tap weights is 
not capable of performing band pass filtering operations and therefore the increased 
capabilities of the transversal filter designed here is illustrated. These simulated results 
helped to reassure that the circuit designed would yield the desired outcome of a 
transversal filter MMIC with positive and negative adjustable tap weight capability. 

Group delay simulations for the transversal filter designed were also performed 
for tap weight configurations of W=[1,0,0,0,0], [0,1,0,0,0], [0,0,1,0,0], [0,0,0,1,0], and 
[0,0,0,0,1]. The simulated results are illustrated in Figure 4.5.3. It can be seen that the 
group delay increases by approximately 50ps (especially for frequencies below 5GHz) as 
the position of the tap weight that is turned on is successively increased. This indicates 
that the delay between successive gain elements in the transversal filter is close to the
Figure 4.5.1: Simulated a) S11 and b) S22 characteristics of the transversal filter for $W=[0,0,1,0,0]$ and $W=[1,0,0,0,1]$. 
Figure 4.5.2: Normalized S21 characteristics of the transversal filter for $W=[0,0,1,0,0]$, $[1,1,1,1,1]$ and $[-1,-1,-1,-1,-1]$. Simulated results only.

Figure 4.5.3: Simulated group delay characteristics for transversal filter for $W=[1,0,0,0,0]$, $W=[0,1,0,0,0]$, $W=[0,0,1,0,0]$, $W=[0,0,0,1,0]$ and $W=[0,0,0,0,1]$
intended value of 50ps and therefore, the design error discussed in Section 4.4 has very little impact on the time delay introduced by the lumped element input and output delay line sections. This result provided further emphasis that the filter designed would perform as expected when fabricated in MMIC form. Furthermore, the fact that the group delay for the transversal filter can be controlled according to which tap weight is turned on demonstrates that the filter can be used in applications where an adjustable delay element is required.

Finally, time domain simulations in a 2.5Gb/s receiver system were performed in order to demonstrate the adjustable nature of the transversal filter. Figure 4.5.4 illustrates simulated 2.5Gb/s eye diagrams for various tap weight settings. It can be seen that the profile of the eye diagram can be significantly altered according to the tap weight adjustments, reiterating once more that the transversal filter design would yield the desired performance.

Figure 4.5.4: Simulated eye diagrams at 2.5Gb/s for various tap weight settings. a) Reference eye, b) $W=[0,0,1,0,0]$, c) $W=[1,1,1,1,1]$, and d) $W=[-1,-1,1,1,-1]$
CHAPTER 5

TRANSVERSAL FILTER
CHARACTERIZATION

In this chapter the techniques used in order to characterize the transversal filter MMIC that was designed and fabricated are discussed. Furthermore, measured results along with their corresponding simulated results are presented for frequency response and transient response measurements.

5.1 Circuit Fabrication and Placement in the Test Fixture

The transversal filter MMIC was fabricated using Nortel's 0.8μm GaAs SAGRF MESFET technology. The MMIC is illustrated in Figure 5.1.1. After fabrication the circuit was mounted into a test fixture and then wire bonded in an appropriate manner so that the transversal filter characterization could be performed. Figures 5.1.2 a) and b) present pictures of the transversal filter MMIC mounted in the test fixture.

5.2 Frequency Response Measurements

The frequency response of the transversal filter MMIC was measured using the HP 8703A 130MHz-20GHz Lightwave Component Analyzer. The setup used for the measurements is illustrated in Figure 5.2.1. Bias Ts were required in order to block DC from the ports of the analyzer. A biasing board, constructed using potentiometers was
used in order to set the tap weight voltages. These voltage levels could then be easily adjusted by adjusting the potentiometers. The tap weight convention followed for the measured results was the same as that for the simulated results outlined in Section 4.5.

The measurements performed included S-parameters and group delay measurements.

Figure 5.1.1: Fabricated transversal filter MMIC
Figure 5.1.2: a) and b) Magnified views of the transversal filter MMIC in the test fixture
5.2.1 S-parameter Measurements

The S-parameter measurements that were made consisted of S11 and S22 for tap weight settings of W=[0,0,1,0,0], and W=[1,0,0,0,1] and S21 for tap weight setting of W=[0,0,0,0,0], W=[1,1,1,1,1], W=[0,0,1,0,0], W=[-1,1,-1,1,-1], W=[-1,0,0,0,1] W=[1,0.75,0.5,0.25,0] and W=[1,0,0,0,1].

Figure 5.2.2 illustrates measured S11 and S22 results for W=[0,0,1,0,0], and W=[1,0,0,0,1]. S11 and S22 in this case are different from the values determined through
simulation since they include the entire circuit mounted in the test fixture. It can be seen however that S11 and S22 do not change due to the value of the tap weight settings which was predicted in the simulated results of Figure 4.5.2. S11 and S22 of the transversal filter in the test fixture is less than -15dB out to a frequency of 1.85 GHz and less than -10dB out to a frequency of 2.2 GHz. This indicates the transversal filter in the test fixture has a good input and output low frequency match to 50Ω. Simulated results for S11 and S22 from Figure 4.5.2 indicate that the transversal filter itself should have had S11 and S22 of less than -15dB beyond 15GHz. This discrepancy can be attributed to the lack of modeling performed in the simulations of the transversal filter MMIC in the test fixture. For example, simulations did not account the long wire bonds used for the input and output signal or the insufficient number of wire bonds used to ground the MMIC.

![Graphs showing S11 and S22](image)

**Figure 5.2.2:** Measured a) S11 and b) S22 of transversal filter MMIC in test fixture for tap weight settings of W=[0,0,1,0] and W=[1,0,0,0,1]
S21 measurements were performed for tap weight setting of \( W=[0,0,0,0,0] \), \( W=[1,1,1,1,1] \), \( W=[-1,1,-1,1,-1] \), \( W=[-1,0,0,0,1] \) \( W=[1,0.75,0.5,0.25,0] \) and \( W=[1,0,0,0,1] \). Measured and simulated results are presented in Figures 5.2.3 to 5.2.8. As can be seen by comparing the measured and simulated S21 results for \( W=[0,0,0,0,0] \) there exists an unexpected resonance at 10GHz in the measured response of the transversal filter. This is attributable to the packaging of the transversal filter MMIC; probably due to a combination of an insufficient number of wire bonds used to ground the MMIC as well as the long length of the existing wire bonds themselves. This combination causes the transversal filter MMIC ground plane to sit at a different potential than the package ground and therefore the measurement setup ground. The effect of the different ground potentials on S21 becomes extremely noticeable at a frequency when the effective inductance from the bond wires and the effective capacitance between the transversal filter MMIC ground plane and the package ground resonate.

As is evident in Figures 5.2.3 to 5.2.8 the resonance due to the transversal filter MMIC being mounted into the test fixture significantly distorts any correlation between simulated and measured results in the area of 10GHz. However, at frequencies below 10GHz there is definite correlation between measured results and their corresponding simulated results. Correlation is especially evident when comparing the measured and simulated results curve shapes as well as what frequencies nulls in the response occur. This indicates that the modeling used in order to design and simulate the transversal filter MMIC was accurate.

It can be seen in Figure 5.2.5 that the simulated results when the tap weights are set to \( W=[-1,1,-1,1,-1] \) predict a band pass filter response centered very close to 10GHz.
Unfortunately, the measured response does not demonstrate the filter's bandpass capabilities. This is due to the fact that the bandpass filter response is masked by the resonance in the response at 10GHz. In order to attempt to demonstrate the transversal filter MMIC's bandpass capabilities the tap weights were set to \( W = [-1,0,0,0,1] \) during the measurements. The response of the transversal filter with this tap weight configuration is illustrated in Figure 5.2.6 and it can be seen that the measured results indicate a bandpass filter response centered close to 2GHz. The simulated response for \( W = [-1,0,0,0,1] \) in Figure 5.2.6 indicates that the expected response should resemble that of a response of a number bandpass filters spaced periodically in frequency with a period close to 5GHz. Measured and simulated results in this situation show some promise of correlating with one another at lower frequencies however beyond 3GHz they do not correlate well due to the effects of the test packaging resonance.

![Graph](image.png)

**Figure 5.2.3:** Measured S21 of transversal filter MMIC in test fixture and simulated S21 for transversal filter for \( W = [0,0,0,0,0] \)
Figure 5.2.4: Measured $S_{21}$ of transversal filter MMIC in test fixture and simulated $S_{21}$ for transversal filter for $W=[1,1,1,1,1]$

Figure 5.2.5: Measured $S_{21}$ of transversal filter MMIC in test fixture and simulated $S_{21}$ for transversal filter for $W=[-1,1,-1,1,-1]$
Figure 5.2.6: Measured $S_{21}$ of transversal filter MMIC in test fixture and simulated $S_{21}$ for transversal filter for $W=\{-1,0,0,1\}$

Figure 5.2.7: Measured $S_{21}$ of transversal filter MMIC in test fixture and simulated $S_{21}$ for transversal filter for $W=\{1,0.75,0.5,0.25,0\}$
5.2.2 Transversal Filter Tap Delay Estimation

An estimation of the effective tap delay of the transversal filter can be performed by setting the first and fifth tap weights to 1 and all of the other tap weights to 0 and then investigating $S_{21}$ for the transversal filter MMIC. $S_{21}$ for $W=[1,0,0,0,1]$ for the transversal filter is illustrated in Figure 5.2.8. It can be seen that the first null of the measured $S_{21}$ curve for the filter is at 3.18GHz. This fact can be used in order to estimate the effective tap delay for the filter. From the frequency response representation for a transversal filter with $L+1$ taps and tap delay, $\tau$ that is the presented in Equation 2.2.3 the response for the filter designed with $W=[1,0,0,0,1]$ can be written as,

$$H(j\omega) = 1 + e^{-j\omega \tau} \quad 4.5.1$$
where $\omega$ is the signal frequency. In this case the delay before the first tap weight in the practical design is neglected since it has no impact on the desired result. The response in Equation 4.5.1 can be simplified to,

$$H(j\omega) = 2e^{-j\omega t} \cos(2\omega \tau) \quad 4.5.2$$

and the magnitude response of this filter is therefore,

$$|H(j\omega)| = 2|\cos(2\omega \tau)| \quad 4.5.3$$

From Equation 4.5.3 it can be seen that the magnitude response is periodic and that the first zero (or minimum in practical situations) occurs at a frequency, $\omega_{\text{min}}$ when $2\omega_{\text{min}} \tau = \pi/2$. Hence, by examining the response of the transversal filter with this tap weight configuration and identifying the frequency, $f_{\text{min}}$ at which the first minimum of the response occurs the effective tap delay can be estimated as,

$$\tau = \frac{\pi}{4\omega_{\text{min}}} = \frac{1}{8f_{\text{min}}} \quad 4.5.5$$

Therefore, since the first minimum in the response occurs at 3.18GHz the effective tap delay for the transversal filter designed here is estimated as 39.3ps which is slightly less than the 50ps that the transversal filter design intended. The difference between measured
and simulated tap delay can be attributed to variations during fabrication in the dimensions of the lumped elements used to create the input and output delay lines.

5.2.3 Transversal Filter Group Delay Measurements

Group delay measurements for tap weight settings of W=[1,0,0,0,0], [0,1,0,0,0], [0,0,1,0,0], [0,0,0,1,0] and [0,0,0,0,1] were also included in the frequency response measurements. Figure 5.2.9 illustrates the measured group delay while simulated results appear in Figure 4.5.3. Measured and simulated results can only be compared on a relative basis since the group delay of the measured results contains contributions from the packaging and the test fixture. It can be seen that in the measured results of Figure 5.2.9a) that the group delay does increase as the transversal filter tap which is turned on (gain element which has a tap weight of 1) increases from position one through five. This delay is expected since the time for the signal to propagate through the input delay line to the gain element which is turned and through the output delay line increases as the tap weight position increases. This result indicates that the transversal filter is also effective as an adjustable delay element. The measured group delay results indicate that the variation in the difference in the group delay between successive group delay curves as a function of frequency is between 20ps to 40ps in the frequency range of 1-3GHz. The group delay measured data is very sporadic and it makes it very difficult to extract any other useful information from it. The sporadic nature of the results can be attributed to the accuracy of the measurement device combined with the measurement technique. However, for the purpose of this thesis these measured group delay results are sufficient to demonstrate the operation of the transversal filter MMIC.
Figure 5.2.9: a) and b) Measured group delay of transversal filter MMIC in test for \( W=\{0,0,1,0,0\}, \{0,1,0,0,0\}, \{0,0,1,0,0\}, \{0,0,0,1,0\} \) and \( \{0,0,0,0,1\} \)
5.3 Transient Measurements

Transient response measurements of the transversal filter MMIC were made using the HP 70841B 0.1-3Gb/s Pattern Genrator, HP 70311A Clock Source HP 54120B Digitizing Oscilloscope Mainframe and HP 54121A DC-20GHz Digitizing Oscilloscope. Due to the inherent attenuation in the transversal filter designed a 20dB pad and a 34dB broadband amplifier (VMA10GB-142, 10Gb/s 34 dB Broadband Amplifier) were also required in order to maintain a signal level which could be distinguished from the system noise. The setup for the measurements is illustrated in Figure 5.3.1. Bias Ts were required in order to block DC from the ports of the pattern generator and amplifier. The same biasing board that was used for the frequency response measurements was used in order to set the tap weight voltages. The tap weight convention followed for the measured results was the same as that for the simulated result and is outlined in Section 4.5.

5.3.1 Eye Diagram Measurements

Eye diagram measurements were made at 2.5Gb/s using a superimposed $2^{23} - 1$ NRZ pseudo-random bit sequence for tap weight settings of $W=[1,1,1,1,1]$, $W=[0,0,1,0,0]$, $W=[-1,0,0,0,1]$ and $W=[-0.8,0,0,0,1]$. A through measurement was also made with a through adapter replacing the transversal filter in the test fixture in Figure 5.3.1. Figures 5.3.2-5.3.6 illustrate measured and simulated 2.5 Gb/s eye diagrams for the through measurement as well as for the above transversal filter tap weights.

Analysis of the measured eye diagrams in Figures 5.3.2-5.3.6 demonstrates that the transversal filter MMIC is capable of modifying its response when its tap weight voltages are adjusted. Furthermore, by comparing the curve shapes of the measured and
simulated eye diagrams for the various tap weight settings it becomes apparent the modeling used for simulations was accurate up to bit rates of 2.5Gb/s.

Figure 5.3.1: Setup for transversal filter MMIC transient measurement
Figure 5.3.2: Measured a) and simulated b) reference (through) eye diagrams
\textbf{Figure 5.3.3:} Measured a) and simulated b) W=[1,1,1,1] eye diagrams
Figure 5.3.4: Measured a) and simulated b) W=[0,0,1,0,0] eye diagrams
Figure 5.3.5: Measured a) and simulated b) \( W = [-1,0,0,0,1] \) eye diagrams
Figure 5.3.6: Measured a) and simulated b) W=[-0.8,0,0,0,1] eye diagrams
CHAPTER 6

CONCLUSIONS, OTHER WORK AND RECOMMENDATIONS

6.1 Conclusions on the 5-tap Transversal Filter

High data rate IM/DD optical communication systems suffer from a variety of transmission impairments due to fiber dispersion, laser nonlinearities and nonideal receiver component response. These impairments cause signal distortion that can result in intersymbol interference (ISI) occurring at the receiver decision circuitry. The ISI combined with receiver noise causes the receiver sensitivity to decrease, which in turn limits the bit rate-transmission distance product. In order to increase the bit rate-transmission distance product, transmission impairment compensation is required.

Linear equalization with a transversal filter has been demonstrated to be an effective method of improving receiver sensitivity in IM/DD systems [6]-[9][18][21][23]-[25][27]. Adaptive linear equalization is attractive since it can increase a receiver’s versatility by allowing for time varying transmission impairments to be compensated for accordingly. In order to perform adaptive equalization the equalizer must be adjustable. A transversal filter can implement an adjustable linear equalizer provided its tap weights are made adjustable.
A 5-tap distributed transversal filter with 40ps tap delays has been designed in the form of a coplanar MMIC for the purpose of adjustable linear equalization. The design was based on the transversal filters designed by Jamani [9][27]. The delay segments in the filter are created using lumped element artificial transmission lines and the transversal elements are based on a MESFET implementation of the Gilbert Cell. Due to the nature of the Gilbert Cell the transversal filter tap weights can be adjusted electronically and can achieve both positive and negative values. Measured and simulated results verify the adjustability of the transversal filter and illustrate the impact that the tap weight values can have on the filter frequency response and pulse shaping characteristics. Furthermore, measured and simulated group delay curves demonstrated that the transversal filter also has the potential to be used as an adjustable delay element.

6.2 Recommendations for Future Work

The purpose of this thesis was to further the work that was performed by Jamani [9][27] in demonstrating that a transversal filter MMIC is capable of performing adaptive equalization. Furthermore, by using a Gilbert Cell for the tap weights in the transversal filter, adjustable positive and negative tap weights were implemented thereby increasing the versatility of the filter in terms of the frequency responses it was able to achieve. This thesis did indeed fulfill its purpose; however there is a great deal of work to be conducted before an adaptive equalizer of this type can be implemented in a real optical receiver system. The following points indicate a few of the things that could be done to further the realization of an adaptive transversal filter based equalizer in an optical receiver:
• It is recommended that the 5-tap transversal filter MMIC designed be re-mounted in the test fixture with an increased number of shorter wire bonds so that a true circuit ground can be better approximated. This would probably get rid of the resonance in the frequency response at 10 GHz and allow for a better evaluation of the device.

• Further experimentation can be performed with other transversal filter type topologies in an attempt to further increase the filter’s versatility as an adjustable equalizer. The transversal filters designed by Jamani [9][27], as well as the one designed in this thesis are all finite impulse response filters. The author has submitted an infinite impulse response MMIC filter for fabrication, but its performance remains to be evaluated. This filter can provide both zeros and poles in its response and will therefore be even more versatile than the filter designed in this thesis.

• The adjustable equalizer filters need to put into a real system so that system improvements using this technique can be verified. This would require both the development of an adjustable equalizer element with appropriate gain and gain control as well as the development of an appropriate adaptive algorithm. In terms of the required gain, it is evident in the frequency response measurements for the transversal filter designed here that the signal gain is less than one in all cases and also dependent on the tap weight configuration of the filter. It would therefore be necessary for the transversal filter to have a gain stage before and/or after it (depending on the filter) that could be controlled according to the tap weight configuration of the filter. This way the received signal could be kept out of the noise
floor of the receiver ensuring that the adjustable equalizer itself is not degrading system performance. Once these system considerations are taken into account it will be necessary to develop an adaptive algorithm that will control the gain stages associated with the transversal filter and that will ensure optimization of the tap weight settings. Appropriately implementing the transversal filter as an adaptive equalizer in an optical receiver system would be more than a thesis in itself.
REFERENCES


