Title
High power peripheral coupled waveguide electroabsorption modulator for analog fiber-optic link applications

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Author
Xie, Xiaobo

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High Power Peripheral Coupled Waveguide Electroabsorption Modulator For Analog Fiber-Optic Link Applications

A dissertation submitted in partial satisfaction of the requirements for the degree Doctor of Philosophy in Electrical Engineering (Photonics)

by

Xiaobo Xie

Committee in charge:

Paul K. L. Yu, Chair
William S. C. Chang, Co-chair
John E. Crowell
Yeshaiahu Fainman
Sungho Jin
Ivan Shubin

2007
The dissertation of Xiaobo Xie is approved, and it is acceptable in quality and form for publication on microfilm:

Co-chair

Chair

University of California, San Diego

2007
DEDICATION

To my grandparents and my parents.
EPIGRAPH

学而时习之，不亦说乎？

— 孔子 (公元前551–479年)

(English translation) To study, and when the occasion arises to put what one has learned into practice — is that not deeply satisfying?

— by Confucius (551–479 BC)

Science is a wonderful thing if one does not have to earn one’s living at it.

— by Albert Einstein (1879–1955)
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PREFACE

Semiconductor electroabsorption modulator (EAM), as a key component in optical communications, has long been the main focus of my graduate research group. Even before I started to touch this topic extensively, I had already been immersed in all the discussions and presentations by other fellow students during our regular group meetings. Unfortunately due to the funding availability and my time tied up with other research projects, I could only get familiar with it from time to time until I started devoting to it in late 2004. EAM, as a semiconductor device, can be regarded as a connection point between fundamental material studies and more application oriented system level link implementation. Besides, it is an optoelectronic device which involves both electronic and optical parts. That nature of EAM research intrigued and enabled me, a student with solid physics background but limited exposure to practical engineering works, to deepen my understanding of basic physics in this special case while extend my horizon to a multi-level and multi-facet broad picture.

Being trained as a physics student, curiosity and interest had been the main drives of my research work before I entered engineering world. In terms of how topics studied and products made can be integral parts of a more complex system, can be of any application and commercial values, and can be beneficial to human being as individual and to the society as a whole, it had been totally out of my scope. Consequently, detailed investigation had narrowed my thoughts and sometimes confused me. Engineering training was a refreshing experience for me.
Perfection was no longer the only pursuit and its proximity to application made me feel more real about my work. A professor’s statement of the difference between a physicist and an engineer that “a physicist wants to park his car as perfectly as possible while an engineer only concerns if his car is within the limit of lines” still echoes in my mind. During the course of my education, I have been feeling lucky to have a mixed training in both physics and electrical engineering.

The word tradeoff is such a common term that it invades engineering world at every corner. While in physics, it always seems to me that a little bit deviation from perfection is regarded unacceptable and lack of true beauty. However, the real world, from my mundane point of view, is full of tradeoffs and a balance of extremities. A graduate student life, as of mine, is a mixture of pain, sacrifice, tolerance from one end and growth, development, satisfaction from the other end.

Although engineering provided me with a new view, physics, no matter how far it is from reality, is still the foundation. I have been so amazed to see how much work physicists have done that engineers can just pick up for granted. In some sense, engineers are lazy animals. They only pick half-developed products as starting points, seldom seeking first principle solutions and frequently leaving those strenuous tasks to physicists. So, from the deepest corner of my heart, I appreciate all the work physicists have laid down for good engineering projects.

As I have mentioned, EAM as a subject has provided me with varieties of opportunities to delve into its multiple aspects. The core absorption region in the device, usually multiple quantum well layers, is closely related to semiconduct-
tor material science and quantum physics. The optical waveguide design employs knowledge in electromagnetism and optics. The microwave design and optimization require background in electronic circuits and microwave propagation. The heat issues related to high power operation involves heat generation and conduction. And the link characterization needs a thorough understanding of system performance. Moreover, fabrication process bridges the gap between device design and its final deliverable version. All of these have fueled my graduate life and propelled me through my journey to the next stop of success.

Time is fleeting and seven-year is just like a breeze. There are still so many things in my graduate passage that I do not have time to digest or even to have a taste. Yet being a graduate student is a long safari as well. There are not many seven-years in one’s life time. It is time for me to move on to a new stage and expand my life. When I was sitting by the La Jolla cove, watching and listening to the ebb and flow of the ocean water several days ago, I felt I had been well loaded in this beautiful campus and ready for a new adventure.
Without the financial support and academic guidance from my advisor, Professor Paul K. L. Yu, this dissertation would have been definitely impossible. His constant encouragement and helpful discussions during his hectic years of chairing the department have always been the sources of my improvement in all aspects. I am at the same time deeply indebted to my co-advisor, Professor William S. C. Chang, whose insightful and incisive comments have led me to the right directions of solutions for numerous times. My gratitude also goes to my other dissertation committee members, Professor Yeshaiahu Fainman, Professor Sungho Jin, Professor John E. Crowell, and Dr. Ivan Shubin for all their time and effort to make my graduation happen.

Working in a research group and solving problems collaboratively are always rewarding and memorable experiences for me at UCSD. I have learned so much from all fellow students and researchers and I will be benefitted even beyond my student life. Among all my colleagues, needless to say, Dr. Ivan Shubin is one of the best I have ever met. His keen mind loaded with extensive knowledge in optics as well as in semiconductor physics and his skillful hands have made all the magic fabrication work possible. All the discussions we went through have added values to my dissertation. I also appreciate Ms. Weixi Chen’s kindness of offering me careful training in my first several years that kindled my interest in and led me into the wonderful world of semiconductor device processing.

Furthermore, I need to thank Dr. Yang Wu who lent me a hand in under-
standing analog fiber-optic link, Dr. Yimin Kang and Dr. Tsai-sheng Liao whose knowledge on photodetector was always easily accessible to me, Dr. Yuling Zhuang whose work on peripheral coupled waveguide electroabsorption modulator was the base of my dissertation work, Dr. Guoliang Li for his stimulating discussions on microwave design, Dr. Dongsoo Shin who spent his precious time on helping me with some measurements, Dr. Carmine Sapia for assistance with optical setup, and Ms. Jessica Fischer whose continuous encouragement on my English language learning had been so fruitful. I have also enjoyed all the fun I have with other group members including Mr. Art Clawson, Dr. Phil Mages, Dr. Felix Lu, Dr. Jianxiao Chen, Mr. Fan Chang, Mr. Clint Novotny, Mr. Justin Bickford, Ms. Lauren Friedman, Mr. Sheldon Wu, Mr. Jeff Bloch, Ms. Meredith Draa, Ms. Su-ju Kuo, Ms. Winnie Chen, and Mr. Jian Ren.

My special thanks go to Dr. Susan Lord and Dr. Justin Hodiak from SPAWAR who not only provided me with experimental equipment and components but also infused me with confidence in my research; to Professor S. S. Lau and Professor Charles Tu for granting me access to the equipment in their labs; to Dr. Gary Betts for beneficial discussions on photocurrent effect of electroabsorption modulator; to Mr. Robert Saperstein for collaboration on Analog Optical Signal Processing (AOSP) project. I am very appreciative to all the secretaries in Pod 3600 of EBU1 during my stay at UCSD, including Ms. Cheryle Wills, Ms. Kim Newin, Mr. David Grayson, Ms. Jennifer Young, a lady with first name Dawn who only stayed for a very short period of time, and especially Ms. Michell Parks,
whose secretarial work was nothing to complain and whose kind words never slipped away from my mind.

I have to give credit to all my buddies who I have been hanging out with during my leisure time, including Dr. Yuguang Hong, Dr. Dongjiang Qiao, Dr. Huaxin Lu, Dr. Xiaotian Zhou, Dr. Song Yang, Dr. Bing Shao, Dr. Hongtao Zhang, and Mr. Haijiang Zhang. Without them my graduate life would be only a solo performance lacking a spectrum of color.

Finally I am grateful to all people who helped me during my hard times in my life, to those whose names have already been mentioned and those whose names I have to omit here at this moment.

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VITA

1996 Bachelor of Science, Peking University (a.k.a. Beijing University), P.R. China

1999 Master of Science, Peking University, P.R. China

2000–2001 Teaching Assistant, Department of Physics
University of California, San Diego

2001–2007 Graduate Student Researcher, Department of Electrical and Computer Engineering
University of California, San Diego

2007 Doctor of Philosophy, University of California, San Diego

PUBLICATIONS


**FIELDS OF STUDY**

Major Field: Electrical Engineering  
Studies in Photonics.  
Professor Paul K. L. Yu
ABSTRACT OF THE DISSERTATION

High Power Peripheral Coupled Waveguide Electroabsorption Modulator
For Analog Fiber-Optic Link Applications

by

Xiaobo Xie
Doctor of Philosophy in Electrical Engineering (Photonics)
University of California, San Diego, 2007
Professor Paul K. L. Yu, Chair
Professor William S. C. Chang, Co-chair

Semiconductor electroabsorption modulator (EAM) has found its way in fiber-optic communications thanks to its small size, good modulation efficiency, and promising integration with other semiconductor optoelectronic devices. High power handling capability and low insertion loss are desirable for EAM to achieve high link gain, low noise figure (NF), and high spurious-free dynamic range (SFDR) in analog fiber-optic link applications.
By applying peripheral coupled waveguide (PCW) structure to EAM, the optical mode is buried down below the absorption layer with small confinement factor $\Gamma$, leaving only an evanescent tail for modulation. Consequently, the coupling between the EAM waveguide and fiber along with the waveguide propagation loss is improved, resulting in a reduced fiber-to-fiber insertion loss that is comparable to low loss LiNbO$_3$ Mach-Zehnder modulators (MZMs). A record low loss of 4 dB was demonstrated by a PCW EAM. Besides, small $\Gamma$ results in low absorption and low photocurrent densities, which enables EAM to survive high optical power and delays the onset of saturation. Devices withstanding 590 mW input optical power and generating 222 mA photocurrent were made and measured. Link gain of within -10 dB was commonly achieved with PCW. PCW with tapered waveguide structure was also investigated to further increase EAM power handling capability.

At high optical power, multiple physical mechanisms kick in to play roles in the performance of EAM, among which are carrier screening, band-gap shrinkage due to temperature increase, material index of refraction change due to large photocurrent, and most prominently the junction resistance reduction as a feedback effect on the input microwave signal. An analysis based on EAM equivalent circuit model was presented for the photocurrent feedback effect with experimental support, yielding a gain limit for EAM at high power. The analysis also led to an alternative conjecture of using blue-shift QCSE material with negative differential resistance to overcome the EAM gain limit. In addition, the linearity of the EAM at high power was theoretically investigated, showing improvement with the pres-
ence of photocurrent feedback. An SFDR of 135 dB/Hz$^{2/3}$ was estimated to be obtainable at 700 mW input optical power.
Chapter I

Introduction

Analog fiber-optic links, as one of the forms of optical communications technology, are used in both commercial and military applications to transfer analog electric signals through optical fiber [1, 2]. The low propagation attenuation of fiber at 1.3 $\mu$m and 1.55 $\mu$m wavelength windows enables signals to travel in long distance from a transmitter to a receiver. An optical modulator, as a constituent component of an optical transmitter, performs the function of converting analog electric signal to analog optical signal. The performance of the optical modulator greatly affects the performance of the whole link such as link gain, link bandwidth, and link linearity. Therefore an optimization of the modulator design is desirable to achieve better link performance.

Based on working mechanism, optical modulators can be divided into two categories: electrooptic modulator (EOM) and electroabsorption modulator (EAM) [3]. The former relies on material index of refraction change while the
latter relies on material absorption coefficient change.\textsuperscript{1} EOM utilizes electrooptic effect in materials such as LiNbO\textsubscript{3} [4], semiconductor [5], or polymer [6] to achieve light intensity or phase modulation. Mach-Zehnder interferometer (MZI) and directional coupler [7] are two popular configurations for intensity modulation. EAM is based on Franz-Keldysh effect (FKE) [8, 9] in bulk semiconductor materials and quantum confined Stark effect (QCSE) [10, 11, 12] in quantum well semiconductor materials. The property of electric field controlled absorption coefficient change from both effects enables light transmission to be modulated while it travels through the material.

EAM, as compared with EOM, has advantages of small size, good modulation efficiency, and potential integration with other optoelectronic devices (e.g. lasers and optical amplifiers). However, it also suffered from larger insertion loss and lower operation power, which is directly detrimental to achieving higher link gain. Therefore reducing insertion loss and raising power handling capability become the major design objectives to improve EAM performance. Peripheral coupled waveguide (PCW) [13] structure employed in EAM design can improve both insertion loss and power handling capability, which greatly boosts the link performance.

This chapter, served as an introduction to my dissertation, will start with a background account of a history of optical communications, followed by brief descriptions of analog fiber-optic link as a system and EAM as a component device.

\textsuperscript{1}If index of refraction is treated as a complex number, then the index of refraction here refers to its real component. The absorption coefficient is related to its imaginary component.
An outline of the dissertation will also be listed at the end of the chapter.

### I.1 A History of Optical Communications

Using light as a means for communications has a long history and can be traced back to ancient time. As you can see from the Great Wall in China, the massive ancient fortification initially built about 3rd century BC during Qin dynasty to protect against the invasion from northern nomadic tribes, its regularly spaced watch towers were once used to launch fire and smoke to rally the message that invaders were approaching [14]. Although in its very rudimentary format, this form of information conveyance has the basic constituents of modern-time optical communications: fire and smoke as transmitter, free space as media, human eyes as receiver, and the subsequent launchings of fire and smoke as repeaters or regenerators.

In modern days, the interest of optical communications firstly came from its potential to provide large information bandwidth attributed to its high carrier frequency[15]. Light, as a special form of electromagnetic wave in a specific frequency range, has several orders of magnitude higher frequency than radio frequency (RF) signals. For example, popularly used infrared (IR) light at 1.55 \( \mu \text{m} \) has a frequency of \( 1.94 \times 10^{14} \text{ Hz} \) (194 THz), which is about 100,000 times higher than the frequency of the electromagnetic wave used to transmit cell phone signals (around GHz).
The success of fiber-optic communications today has come to reality in a long way. Without the dramatic improvement of its components, including optical fiber and semiconductor diode lasers, fiber-optic communications would have been just a wonderful dream.

In 1966, pioneering work done by Charles K. Kao [16], which predicted 20 dB/km attenuation for optical fiber with impurities removed, initiated research leading to nowadays seemingly impossible 0.2 dB/km fiber loss in its transmission window (a typical attenuation spectrum of single mode fiber (SMF) is shown in Fig I.1). This low loss silica optical fiber, first demonstrated by Corning Inc. [17], offers numerous benefits as a communication media. The core of the fiber, with a little bit higher index of refraction than its cladding surrounding, provides a waveguide channel for light to propagate. Its extremely low loss makes it superior to its conventional electrical counterparts — twisted pair cable and coaxial cable. Besides, it is of low cost, flexible, light weighted, and immune to electromagnetic interference (EMI) [18, 19].

Along with the development of optical fiber, optoelectronic devices suitable for fiber-optic communications also experienced revolutionary advancement. Shortly after stimulated light emission had been demonstrated in solid state material in 1960 [20] and gaseous media in 1961 [21], lasing action was observed in compound semiconductor material GaAs by independent endeavors from at least three research groups in 1962 [22, 23, 24, 25]. The breakthrough of technology at the end of 1960s by introducing double-heterostructure (DH) brought semicon-
Figure I.1: Typical attenuation spectrum of single mode fiber.

Although optical fiber can provide tremendous amount of bandwidth in the
vicinity of its low loss wavelength windows (\(\sim 25 \text{ THz}\)), the realizable bandwidth in the real system is limited by how fast electrical signals can be impressed onto light carrier and how fast they can be retrieved. This bandwidth restricted by transmitter and receiver (up to 100 GHz at present) only occupies a very small fraction of the total bandwidth optical fiber provides. Wavelength division multiplexing (WDM) technology utilizes a comb of wavelengths to carry different signals so that one optical fiber channel can be shared by an array of sub-channels and intrinsic large bandwidth of fiber can be fully accessed [33].

Thanks to all these technology breakthroughs, fiber-optic communications quickly and successfully underwent an evolution from dream to reality in 70’s and matured in 80’s and 90’s. Now it has deeply penetrated into telecom and datacom industries and profoundly impacted our everyday life from long distance phone calls to high speed Internet services.

I.2 Analog Fiber-Optic Link

An analog fiber-optic link is an optical communications system to convey analog signals through optical fiber from one site to another site. Unlike a digital link, in which original signal is digitized and only discrete states of bits are transmitted, an analog link transmits a replica of the original signal without digital conversion. Compared with a digital fiber-optic link, an analog link has a simpler structure and does not need analog-to-digital conversion (ADC) and
digital-to-analog conversion (DAC). So it is used to lower the system cost or in some applications where ADC and DAC are not available in the band of interest. Without time delay caused by signal conversions, it is also favorable for real time operations especially in military applications.

I.2.1 Link Components

As in all communications systems, an analog fiber-optic link consists of a transmitter, a transmission channel and a receiver as shown in Figure I.2. Optical fiber, with all aforementioned benefits, acts as a transmission channel. A transmitter provides a light source and imprints RF electrical signal onto optical carrier, which is termed as optical modulation. The light carries RF signal and propagates along optical fiber and finally reaches a receiver. The receiver detects RF electrical signal in light and converts it back to its original form. The transmitter here is an electrical-to-optical signal converter while the receiver plays a reverse role to convert optical signal back to electrical signal.

![Figure I.2: Schematic of an analog fiber-optic link.](image)

In some fiber-optic links, the transmitter only consists of a laser diode. The
electrical-to-optical signal conversion is done by direct modulation of laser injection current (see Figure I.3(a)). A change of electrical current through laser junction is converted into a corresponding change of generated optical power, thus achieving intensity modulation. In some other links, a separate external optical modulator is connected after the laser diode (see Figure I.3(b)). Thus the transmitter has two parts. In this case, laser diode only provides the light source. The modulation voltage across the optical modulator generates optical intensity change. Directly modulated semiconductor laser diode usually has a bandwidth limited by its relaxation oscillation frequency and a modulation efficiency restrained by its internal quantum efficiency. It also has unfavorable chirp characteristics for long distance signal transmission [34]. Externally modulated link offers more flexibility for the link performances such as modulation efficiency and bandwidth.

Figure I.3: Schematics of optical transmitter in (a) direct modulation and (b) external modulation.

In analog fiber-optic links, intensity modulation-direct detection (IMDD) is the dominant modulation detection format. The receiver usually consists of a photodetector (photodiode). The photodetection process is through the absorption of photons and it converts optical intensity to electrical power by square-law
detection.

Optical amplifiers such as EDFA and semiconductor optical amplifier (SOA) are sometimes added to analog fiber-optic links to increase the optical power so that larger signal can be received by photodetector. However, the added noise and nonlinear processes degrade other aspects of the link performance. At high optical power, long optical fiber also experiences nonlinear interactions such as stimulated Brillouin scattering (SBS) and four-wave mixing (FWM), which cause reflection and frequency mixing.

To boost the performance of analog fiber-optic links, sometimes electrical components are added such as pre- or post- amplifiers. However, while they can increase the output RF power, they may impair other link performances such as linearity and noise level. To clarify the function of intrinsic analog fiber-optic link, this dissertation only limits the electrical components to impedance matching circuits.

I.2.2 Link Performance

As a communication system, the output RF signal of an analog fiber-optic link should truly duplicate the input RF signal without adding spurious modifications\(^2\). That is to say, under ideal conditions, the output RF signal \(S_{\text{OUT}}(t)\) should be different from the input RF signal \(S_{\text{IN}}(t)\) only by a scaling factor \(a\) such that

\[
S_{\text{OUT}}(t) = aS_{\text{IN}}(t) \tag{I.1}
\]

\(^2\)The intentional manipulation and alteration of signal fall into the category of signal processing.
without any corruption. Moreover, \( a \) is expected to be large for practical purpose. However, in real world, transmission of signal is always accompanied by degradation.

There are several causes for this signal degradation. Firstly, the analog fiber-optic link is not necessarily a linear system. In a general case, the output RF signal can contain high order power terms of the input RF signal as:

\[
S_{\text{OUT}}(t) = aS_{\text{IN}}(t) + a_2 S_{\text{IN}}^2(t) + a_3 S_{\text{IN}}^3(t) + \cdots
\]  

(I.2)

Consequently, the spectrum of the output signal will have harmonics and intermodulation terms of the input signal frequencies.

Secondly, even if the system is linear, the link has limited bandwidth with uneven frequency response. As a result, the link only transmits signal within the bandwidth and different frequency components have different scaling factors within the bandwidth. Hence, the output signal deviates from the input signal. Let the Fourier transform of the input and output signals be \( S_{\text{IN}}(f) \) and \( S_{\text{OUT}}(f) \), then we can get:

\[
S_{\text{OUT}}(t) = \mathcal{F}^{-1}(S_{\text{OUT}}(f)) = \mathcal{F}^{-1}(H(f)S_{\text{IN}}(f)) = \mathcal{F}^{-1}[H(f)\mathcal{F}(S_{\text{IN}}(t))]  
\]  

(I.3)

where \( H(f) \) is the frequency response function of the link. Only when \( H(f) \) is a constant does the output signal preserve the quality of the input signal. This can only be achieved by a transmission system with infinite bandwidth and with flat frequency response for arbitrary input waveforms.

Finally the transmission process of a signal inevitably introduces extra
noises to the system, which is totally irrelevant to the transmitted signal. These
noises blur the true image of the signal. In this case,

\[ S_{\text{OUT}}(t) = aS_{\text{IN}}(t) + N(t) \]  \hspace{1cm} (I.4)

where \( N(t) \) denotes added noise. Although in most situations the noises are small
compared to signal level, they still impose serious problems to the link if high
quality signal is required.

Several figures of merit can be used to qualify the performance of an analog
fiber-optic link. The commonly used ones are 1) link RF gain; 2) bandwidth;
3) noise figure (NF); and 4) spurious-free dynamic range (SFDR) [1, 2]. The
overall performance of the analog fiber-optic link depends on all these parameters.
Different applications may favor some parameters over the others. In most analog
fiber-optic link applications, the optical fiber used is only several to ten kilometers
long. So I assume fiber dispersion and nonlinearity (SBS and FWM) do not cause
much problem to the link performance in such short distances\(^3\). In this dissertation,
I will only discuss the impacts of optical transmitter and optical receiver on the
link performance and will mainly focus on the performance of optical modulator.

The following is a brief introduction to four figures of merit. As we shall
find out later, they are not independent parameters from both link and device
perspectives. Thus there are trade-offs in link and device design.

\(^3\)This is not true for long haul optical communications systems (mostly digital), where fiber
can be thousands of kilometers long. The dispersion and nonlinear optical processes can be severe
impacts on the transmission quality.
Link RF Gain

There might be a confusion that low loss fiber should guarantee a low loss analog fiber-optic link. This is not true because the ultimate signal conveyed is electrical but not optical. Two opposite processes of electrical-to-optical conversion and optical-to-electrical conversion are involved in the fiber-optic link. Hence, the efficiencies of both processes need to be taken into consideration for the link RF gain. The link gain is determined mainly by modulation efficiency and detection efficiency.

Though in most cases an analog fiber-optic link experiences a power loss instead of a gain, the term link RF gain is still widely adopted in the community. A loss is represented by a negative gain.

In microwave engineering, there are several different definitions for the gain of a two terminal network depending on whether source and/or load are/is impedance matched: Power Gain, Available Gain, and Transducer Power Gain [35]. In this dissertation, we define link RF gain as the ratio of the power $P_{\text{RF OUT}}$ dissipated in the receiver load resistor $R_L$ at the link output terminal to the power available from the RF source $P_{\text{IN}}^{\text{RF}}$. This is similar to the Transducer Power Gain.

$$g = \frac{P_{\text{RF OUT}}^{\text{RF}}}{P_{\text{IN}}^{\text{RF}}} \quad (I.5)$$

Figure I.4 shows a simple circuit of an impedance $Z$ connected to a signal source. The power on the impedance $Z$ reaches a maximum when it equals to the
complex conjugate of the source impedance $Z_S$. This maximum power is the source available power. In most cases, the source impedance $Z_S$ is set to be $R_S = 50\Omega$ and the matched $Z$ is also $50\Omega$. Therefore the available power from the source is

$$P_{\text{IN}}^{\text{RF}} = \frac{v_S^2}{4R_S} \quad (I.6)$$

where $v_S$ is the root mean square (RMS) value of the source voltage amplitude.

Figure I.4: A simple circuit of an impedance $Z$ connected to a signal source.

Figure I.5 shows an analog fiber-optic link with simple circuit elements. The link output RF power, which is the power dissipated in the receiver load resistance $R_L$, is given by

$$P_{\text{OUT}}^{\text{RF}} = i_D^2 R_L \quad (I.7)$$

where $i_D$ is the RMS value of the electrical current flowing through the receiver load resistor $R_L$.

In the case of externally modulated link, the dc characteristic of the optical modulator is determined by its optical transfer curve $T(V)$, which is the optical transmission $T$ versus applied voltage $V$. In practice, $T(V)$ is usually normalized
Figure I.5: A schematic of analog fiber-optic link with simple circuit elements.

to the maximum transmission bias point. In the case of Mach-Zehnder modulator (MZM), the reference point is the peak of the sinusoidal optical transfer curve (see Figure I.6(a)). As for conventional EAM, the reference point is usually conveniently chosen to be zero bias (see Figure I.6(b)). The optical loss at the reference point is usually referred to as link optical insertion loss $t_{INS}$. The optical power received by the receiver $P_D$ is determined by input laser power $P_L$, link insertion loss $t_{INS}$, normalized modulator optical transfer curve $T_{NORM}(V)$, and its associated bias point $V_B$. The receiver photocurrent $I_D$ is related to received optical power by its responsivity $\eta_D$ as expressed by:

$$I_D = \eta_D P_D = \eta_D P_L t_{INS} T_{NORM}(V)$$  \hspace{1cm} (I.8)

If the voltage applied to the optical modulator includes both dc component $V_B$ and a small signal ac component $\sqrt{2}v_S \sin(2\pi ft + \phi)^4$, the modulator optical transfer curve can be expanded by Taylor series around its dc bias point $V_B$ with first two

\footnote{Here $\sqrt{2}$ is added so that $v_S$ is the RMS value.}


\[ T_{\text{NORM}}(V) = T_{\text{NORM}}(V_B) + T'_{\text{NORM}}|_{V_B} \sqrt{2}v_S \sin(2\pi ft + \phi) \quad (I.9) \]

where \( T'_{\text{NORM}}|_{V_B} \) is the normalized slope efficiency at the bias voltage. The ac photocurrent received by the receiver is thus as follows.

\[ \sqrt{2}i_D \sin(2\pi ft + \phi) = \eta_D P_{\text{LINS}}T'_{\text{NORM}}|_{V_B} \sqrt{2}v_S \sin(2\pi ft + \phi) \quad (I.10) \]

Therefore the link RF gain can be expressed as:

\[ g = \frac{P_{\text{OUT}}^{\text{RF}}}{P_{\text{IN}}^{\text{RF}}} = \frac{4i_D^2 R_L R_S}{v_S^2} = 4P_{\text{LINS}}^2 \left( T'_{\text{NORM}}|_{V_B} \right)^2 \eta_D^2 R_S R_L \quad (I.11) \]

(a) \hspace{2cm} (b)

\[ \text{Figure I.6: Typical optical transfer functions of MZM (a) and EAM (b).} \]

As seen from Figure I.6(a), MZM has a sinusoidal optical transfer curve. The largest slope efficiency point is its quadrature point. The slope of that point on the normalized optical transfer curve \( T'_{\text{NORM}}(V)|_{V_B} \) is related to MZM’s half wave voltage \( V_\pi \), which is the voltage difference between adjacent maximum and
minimum transmission points on the optical transfer curve.

\[ T'_{\text{NORM}}|_{V_B} = \frac{\pi}{2V_\pi} \]  

(I.12)

For EAM, the concept of half wave voltage is borrowed from MZM by calculating from its slope at the best bias point on its normalized optical transfer curve. So sometimes it is called equivalent \( V_\pi \) or \( V_{\pi e} \) for EAM.

By using \( V_\pi \), the link gain equation Equation I.11 can be rewritten as:

\[ g = \frac{\pi^2 P_L^2 L_{\text{INS}} \eta^2 D R_S R_L}{V_\pi^2} \]  

(I.13)

The link RF gain is quadratically dependent on incident laser power, link insertion loss (mostly modulator insertion loss), modulator optical transfer curve slope efficiency, and receiver responsivity. To achieve a higher link gain, larger laser power, smaller insertion loss, and larger modulator slope efficiency (or smaller \( V_\pi \)) are desired. A higher link RF gain is favorable not only for boosting output signal power but also for lowering NF and increasing SFDR.

**Bandwidth**

*Bandwidth* of an analog fiber-optic link is defined as the frequency range within the 3-dB gain roll-off frequencies. If the system is narrow-band impedance matched to a certain frequency, the link gain will have a band-pass spectrum. There will be two 3-dB roll-off frequencies, one above the peak gain frequency \( f_H \) and the other below the peak gain frequency \( f_L \). The bandwidth here is the difference of the two, \( B = f_H - f_L \). In most cases, the gain spectrum rolls off from
dc and the bandwidth equals the frequency where gain drops to 3 dB lower than
low frequency gain near dc. In device design, link RF gain and bandwidth can
always be traded off.

**Noise Figure**

*Noise figure* specifies the degree of degradation of signal to noise ratio (SNR)
when an analog signal transmits through a link. It is defined as the ratio of input
SNR to the output SNR.

\[
NF = \frac{\text{SNR}_{\text{IN}}}{\text{SNR}_{\text{OUT}}} = \frac{S_{\text{IN}}/N_{\text{IN}}}{S_{\text{OUT}}/N_{\text{OUT}}} = \frac{S_{\text{IN}}}{S_{\text{OUT}}} \cdot \frac{N_{\text{OUT}}}{N_{\text{IN}}} = \frac{1}{g} \frac{g N_{\text{IN}} + N_{\text{ADD}}}{N_{\text{IN}}} = 1 + \frac{N_{\text{ADD}}}{g N_{\text{IN}}}
\]

(I.14)

where \( g \) is the link RF gain and \( N_{\text{ADD}} \) is the link additive noise. Since the definition
of noise figure depends on the input noise level, it is standardized to be the noise
power resulting from a matched resistor at \( T_0 = 290\text{K} \), which means \( N_{\text{IN}} = k T_0 B \),
where \( k \) is the Boltzmann constant and \( B \) is the noise bandwidth.

In a regular optical link, there are generally three kinds of output noise
sources: thermal noise from all resistive components in the system; shot noise
resulting from the corpuscular nature of electric transport process in the p-i-n
junction of EAM as well as photodetector; relative intensity noise (RIN) from
the laser source. Thermal noise does not depend on the optical power. Shot
noise and RIN noise depend on optical power linearly and quadratically. Besides,
optical amplifiers involved in the link add more noise such as amplified spontaneous
emission (ASE) noise to the link.
Spurious-free Dynamic Range

Spurious-free Dynamic Range is the dynamic range of fundamental signal power from link output noise level to the power level when spurious distortions start to emerge above the output noise. It is related to the link linearity, link RF gain, output noise level, and usually characterized by a standard two-tone test for simple cases. In a two-tone test, two equal power input signals of slightly different frequencies \( f_1 \) and \( f_2 \) are launched to the link. The output signal spectrum usually consists of fundamental frequencies \( f_1 \) and \( f_2 \) along with high-order frequencies such as harmonics \( 2f_1, 2f_2, 3f_1, 3f_2 \) and intermodulation distortions \( f_1 \pm f_2, 2f_1 \pm f_2, 2f_2 \pm f_1 \), etc. as shown in Figure I.7. Due to their proximity to the fundamental signals, third-order intermodulation (IM3) signals \( 2f_1 - f_2 \) and \( 2f_2 - f_1 \) are the most concerned components in sub-octave applications. In most cases they increase by the third-order power of the input signal power level while output fundamental signal is proportional to the input signal. When EAM is biased at its third-order null point \( (T^{(3)}_{\text{NORM}}(V)|v_B = 0) \), the output distortions at \( 2f_1 - f_2 \) and \( 2f_2 - f_1 \) can depend on the fifth-order power of the input signal level. This faster increase IM3 distortions will finally be larger than the noise level and become dominant factor of the dynamic range in sub-octave frequency range. Third-order SFDR is defined as the output SNR when the IM3 distortion power starts to emerge above the noise floor. This is illustrated in Figure I.8.

Different applications require SFDR related to different spurious distortion signals. In sub-octave (narrow band) applications, third-order distortions \( 2f_1 - \)
Figure I.7: An illustration of harmonics and intermodulation distortions in two-tone test.

Figure I.8: An illustration of third-order spurious-free dynamic range.
$f_2$ and $2f_2 - f_1$ are of the primary concern because they are within the system bandwidth. Second-order harmonics $2f_1$, $2f_2$ and intermodulation $f_1 \pm f_2$ fall outside the passband of the sub-octave system and thus are not important. In multi-octave (broad band) applications, both second-order and third-order SFDRs are important because both of them are within the band of interest.

Nonlinear distortion signals at a link output can be from components of the link such as optical modulator and photodetector. In most cases, optical modulator is the dominant nonlinear component in the link. The nonlinearity of the optical modulator originates from its nonlinear optical transfer curve $T_{\text{NORM}}(V)$. If we expand $T_{\text{NORM}}(V)$ by Taylor series as in Equation I.9 to include more terms, high-order distortions will show up and they are related to the high order derivatives of the normalized optical transfer curve.

$$
T_{\text{NORM}}(V) = T_{\text{NORM}}(V_B) + \sum_1^\infty \frac{T_{\text{NORM}}^{(n)}(V)|_{V_B}}{n!} (\sqrt{2}v_S)^n \sin^n(2\pi ft + \phi)
$$  (I.15)

For a detailed derivation of optical modulator nonlinearity from its optical transfer curve, please refer to Appendix B.

Since high-order distortion signals are related to optical transfer curve of optical modulator, they can be minimized by adjusting the modulator bias voltage. For EAM, the best bias point is the third-order null point of its optical transfer curve for sub-octave applications. In multi-octave applications, both second-order and third-order distortions need to be minimized.
I.2.3 Applications

Nowadays probably more than 99% of fiber-optic links are digital links using simple and robust on-off-keying (OOK). The rational for using analog fiber-optic links is basically economic, in which the conversion between analog and digital forms can be avoided [36]. At high frequencies, where ADC and DAC are not readily available, analog fiber-optic link is the only solution. Analog links have applications in both commercial and military applications.

Among those applications there are community antenna television (CATV\(^5\)), phase arrayed antenna, and antenna remoting (as shown in Figure I.9). In CATV signal distribution system, subcarrier multiplexing (SCM) technique is employed to combine different channels of TV programs, which are carried by subcarriers at different RF frequencies [37, 38, 39]. The signal is then added to the main optical carrier and sent from the head-end to multiple remote distribution nodes. Due to different system requirement, carrier-to-noise ratio (CNR) and composite triple beat (CTB) are used instead of NF and SFDR to evaluate the link quality. As the CATV format is being revolutionized and striding into digital age, only signal within each channel is digitized and they ride on different RF subcarriers. The optical carrier still sees the combined digital channels as an analog modulation. Therefore the employed fiber-optic link with subcarrier multiplexed digital channels will still have analog characteristics. Phased array antenna utilizes analog

\(^5\)The acronym CATV has been used interchangeably to represent community antenna television or cable television in different contexts. The actual meanings, however, remain very similar.
fiber-optic links to distribute signals to elements with appropriate phase relationship to accomplish beam forming function in radar application. Antenna remoting is suitable to separate signal feeding and processing node and remote antenna site which is not easily accessible. It can also minimize cost and power consumption at antenna elements by placing most of the processing units at the central node.

Figure I.9: Analog fiber-optic link applications: (a) CATV; (b) Phased arrayed antenna; (c) Antenna remoting.
I.3 Electroabsorption Modulator (EAM)

Most semiconductor EAMs are waveguide type modulators. They have small size (several hundred microns in length versus several centimeters for LiNbO$_3$ MZMs). Since they are fabricated from the same family of semiconductor materials that are also used for laser diode, optical amplifier, and photodetector, they can be potentially integrated with other optoelectronic devices. Electroabsorption modulated laser (EML), which is an EAM integrated with a laser diode on the same chip, has been commercially available for a while.

Two physical mechanisms are available for the operation of EAMs: Franz-Keldysh effect (FKE) [8, 9] and quantum-confined Stark effect (QCSE) [10, 11, 12], which are schematically shown in Figure I.10. In bulk semiconductor materials, when no electric field is applied, photons with the energy close to material band-gap get absorbed and the absorption edge exhibits an exponential tail. In the presence of electric field, this absorption edge is modified, resulting in an electric field dependent absorption tail [40, 41]. In quantum well semiconductor materials, both electrons and holes are populated in discrete energy levels formed by the potential barrier surrounding the quantum well. In addition, electrons and holes can form excitons due to their confinement in the well. Photon energy required for absorption is the combination of electron and hole energy level difference and excitonic binding energy. As electric field is applied, the energy gap between electron and hole energy levels becomes smaller, forming a red-shift absorption
spectrum known as QCSE. Electron and hole wave functions also changes, causing a reduction of overlap integral between each other. Besides, excitonic binding energy changes while exciton linewidth broadens, showing in absorption spectrum as broadening exciton peak. This results in an electric field dependent absorption coefficient change. Therefore, in FKE and QCSE, laser radiation at wavelength just below the band-gap in the absorption tail will have absorption coefficient dependence on the applied electric field. Both FKE and QCSE have been used to make EAMs. The devices based on FKE have less polarization dependence, but QCSE devices offer better modulation efficiency with the assistance of QCSE and exciton.

To access to the telecommunications wavelengths of 1.3 $\mu$m and 1.55 $\mu$m, we need material structure with band-gap just above the laser radiation wavelength. $\text{In}_{1-x}\text{Ga}_x\text{As}_y\text{P}_{1-y}$ family material, with its band-gap wavelength ranging from 0.92 $\mu$m to 1.65 $\mu$m when lattice matched to InP, is widely used to construct EAMs. By tuning the composition of constituent elements x and y, it can be lattice matched to InP substrate to preserve material quality while we still have the freedom to adjust its band-gap to tailor to our application wavelength. It is fortunate that smaller band-gap material tends to have larger index of refraction. Therefore it can also be used as the core and cladding of the optical waveguiding channel by tuning the composition of these layers.

The EAM devices usually adopt p-i-n structure and have reverse bias on it as shown in Figure I.11(b). The middle intrinsic layer contains material that has FKE
Figure I.10: Schematics of (a) Franz-Keldysh effect (FKE) and (b) quantum-confined Stark effect (QCSE).
or QCSE when electric field is applied. With a reverse biased p-i-n junction, the bias voltage drops mostly on the intrinsic layer where photo-absorption takes place. After incoming photons are absorbed, electrons make transitions from valence band to conduction band and electron-hole pairs are generated. They are driven out by the electric field across the junction to form the photocurrent.

The physical configuration of an EAM consists of an optical waveguide structure along with p and n contact metals. Figure I.11(a) shows an EAM waveguide structure alone without p and n contacts. As shown in Figure I.11(b), the cross section of the device has a mesa structure to confine light in lateral direction. The semiconductor layers are also designed to have index of refraction profile to confine light in vertical direction. Therefore light can be well guided along the optical waveguide. Usually p-contact pad is connected to the top p-layer and n-contact is deposited on the n-layer mainly to facilitate biasing and RF signal feeding. When bias voltage is applied to p and n contacts, electric field across the p-i-n junction induces material absorption change in the intrinsic layer. The incoming light is then guided, propagates and attenuates along the waveguide. As shown in Figure I.11(c), the waveguide optical mode not only sees absorptive intrinsic layer but also covers p- and n- layers around. Therefore, not all the optical power within the mode will be absorbed. The percentage of the optical mode within the intrinsic layer is defined as confinement factor $\Gamma$. Due to exponential material absorption dependence on its length, the EAM optical transmission can
be expressed as

\[ T(V) = e^{-\alpha(V)\Gamma L} \]  \hspace{1cm} (I.16)

where \( \alpha(V) \) is the material electric field dependent absorption coefficient and \( L \) is the EAM device length. With different bias voltage, the intrinsic layer has different absorption coefficient and thus the waveguide has varied optical transmission. This transmission versus bias voltage is the modulator optical transfer curve as described in Subsection I.2.2. When an RF voltage is added to the dc bias voltage, the output optical power follows the applied RF voltage around its dc value, thus achieving optical modulation.

I.3.1 Performance and Design Considerations

As an optical modulator for analog fiber-optic link applications, the performance of an EAM is characterized by its modulation efficiency, bandwidth, linearity and additive noise.

As described in Subsection I.2.2, the dc characteristic of an EAM is presented by its optical transfer curve \( T(V) \), which is conventionally divided into two parts: insertion loss \( t_{\text{INS}} \) and normalized optical transfer curve \( T_{\text{NORM}}(V) \). The link RF gain \( g \) is quadratically dependent on input laser power \( P_L \), insertion loss \( t_{\text{INS}} \), and slope efficiency at the bias point on the \( T_{\text{NORM}}(V) \) curve. Since optical modulator can be treated as an electrical-to-optical conversion device, we can group the
Figure I.11: Schematics of (a) EAM waveguide structure, (b) EAM material layer structure and (c) EAM optical waveguide mode (shaded area as intrinsic layer).
terms in Eq. I.11 relating to modulator to define modulation efficiency [42].

\[ r_M^2 = (P_{\text{INS}}T_{\text{NORM}}|_{V_b})^2 = \left( \frac{\pi P_{\text{INS}}}{4V_\pi} \right)^2 \]  

(I.17)

Therefore, high input laser power, low insertion loss, and small \( V_\pi \) are favorable for a better modulation efficiency.

EAM's bandwidth is mainly determined by its \( RC \) time constant with lumped electrode configuration. The \( R \) is related to source resistance, device series resistance, load resistance, and matching circuit. The \( C \) is mainly the p-i-n junction capacitance. With a smaller capacitance \( C \), EAM can be operated at larger bandwidth.

The linearity of EAM is related to its optical transfer curve \( T_{\text{NORM}}(V) \). The relationship between them is detailed in Appendix B. Bias voltage can be used to tune the linearity of EAM. However, unlike MZM with a regular sinusoidal shape transfer curve, EAM’s transfer curve is a complex function of material property as well as device structure. Hence it is very hard to be optimized.

The noise sources from EAM include thermal noise from resistive elements and shot noise from its photocurrent generation in the p-i-n junction. At low optical power, the EAM photocurrent is small and the link RF gain is low, so it is negligible at the link output. However as optical power increases, the EAM shot noise will play an important role in EAM NF performance. This will be further discussed in Chapter IV.

In the design of EAM, different aspects of its performance need to be optimized for a specific application. To improve EAM modulation efficiency, the device
needs to be operated at high optical power, has low insertion loss and small $V_\pi$. The light propagates through EAM waveguide. A well designed waveguide should only support the fundamental mode. A large optical mode size is desired to improve input and output coupling from and to optical fibers. Waveguide scattering loss and material residual absorption need to be minimized as well to reduce insertion loss. The absorption material for the intrinsic layer needs to be selected to have good property of absorption coefficient change under bias voltage. A longer device and thinner intrinsic layer thickness also improve $V_\pi$. However, as a tradeoff, they will increase the device capacitance and thus reduce device bandwidth.

I.3.2 Peripheral Coupled Waveguide (PCW) EAM

In an EAM, photon absorption is used to achieve optical modulation and high power is desired to achieve a better modulation efficiency. However, at high input optical power level, the device modulation efficiency saturates due to carrier screening effect [43] and thermal effect. With large amount of electron-hole pairs generated in the device junction, they can not be swept out by the applied electric field quickly and will pile up around the junction. As a result, the applied electric field will be screened by these carriers and the effective electric field applied to the junction will be smaller. The modulation efficiency is thus reduced. High power also generates large photocurrent, which will be converted into Joule heat. This causes temperature increase, which is also detrimental to device operation. At sufficiently high optical power, the device will be damaged permanently. So to
raise device power handling capability and reduce its saturation, a reduction of photocurrent density is needed.

According to Equation I.16, a smaller $\Gamma$ apparently generates smaller amount of photocurrent per unit length with a certain input optical power. Hence it can bring the device to operate at higher optical power and only generates same amount of photocurrent per unit length as a larger $\Gamma$ device at lower optical power does.

The peripheral coupled waveguide (PCW) structure has the most of its optical mode in a non-absorptive waveguiding layer down below the intrinsic absorption layer. Only the evanescent tail of the mode reaches the absorption layer. Figure I.12 shows optical modes of a conventional EAM and a PCW EAM. The intrinsic layer of the PCW EAM only occupies a very small portion of the whole optical mode. Its confinement factor $\Gamma$ thus can be dramatically reduced to be less than 10%. As a result, device saturation at high power can be greatly improved. PCW also has the advantage to separate optical waveguide design and microwave waveguide design.

I.3.3 EAM at High Power

The main objective for high power EAM is to increase its modulation efficiency and the link gain since they depend on the square of the optical power. EAM operating at high power levels has some characteristics different from low power operation. The most prominent effect is the gain saturation caused by the photocurrent feedback effect which effectively reduces the modulation voltage across
the active absorption layer. This results in a gain limit for EAM at high power. This effect will be analyzed in detail in Chapter IV. EAM shot noise also becomes more important as the input optical power and link gain increase. It will finally overcome link gain to determine the link noise figure. The link NF therefore exhibits a minimum at certain optical power level [44]. The photocurrent effect also affects EAM linearity with its negative feedback. As will be discussed in Chapter V, the EAM link SFDR will improve with the increase of optical power [45].

To maintain normal operation of EAM at high power, cares need to be taken to ensure current density and temperature increase are within the sustainable limit. PCW structure, tapered waveguide along with temperature control can bring device to operate at higher power levels.
I.3.4 Lump Element EAM vs. Traveling Wave EAM

At low frequencies, waveguide EAM can be considered as a lumped element, where microwave wavelength is much longer than the device dimension and the RF phase difference between different parts of the device is not significant. In this case, the bandwidth of the device is limited by the $RC$ time constant. As we increase the modulation frequency, the wavelength of the RF signal becomes smaller and smaller and finally is comparable to the dimension of the device. The RF phase difference at different parts of the device is significant and consequently the device can not be viewed as a lumped element but a transmission line. The device resistance and capacitance now become distributed along the waveguide. To maintain the modulation efficiency at high frequencies, traveling wave (TW) electrode structure can be constructed to overcome the $RC$ time limitation. In the TW configuration, microwave signal is fed from front end of the waveguide and terminated at the other end as shown in Figure I.13. Optical wave and electrical wave propagate along the waveguides together so that RF signal can modulate the optical carrier efficiently. To keep electrical signal in phase with optical carrier during propagation, the velocity mismatch between the microwave and the optical wave needs to be minimized. To avoid microwave reflection from the back end, the RF signal needs to be terminated properly. In addition, low microwave attenuation along the waveguide is desirable because as the loss increases with frequency it will limit the device bandwidth. Therefore for TW EAM, its bandwidth is no longer limited by $RC$ time constant but by velocity mismatch, impedance mismatch and
microwave loss [42].

For the PCW EAM, the confinement factor is significantly smaller than the convention EAM and it needs significantly longer device to achieve the same modulation efficiency. The resulting capacitance and $RC$ time constant is large and lumped element structure has very limited bandwidth. Hence TW structure is desired to push the device to high frequencies.

\section*{I.4 Dissertation Outline}

This dissertation focuses on peripheral coupled waveguide electroabsorption modulator with lumped electrode, especially when it is operated at high optical power. Although TW EAM will be necessary to provide large bandwidth, several characteristics such as saturation, thermal influence, and index of refraction change at high power can be investigated at low frequency without the TW electrode.
Studies of these factors at low frequency are the objectives of this dissertation.

During my dissertation research work, device design and optimization were carried out by simulation. Several different designs were fabricated into devices. The measurement of intrinsic link with the devices showed improved performance compared to conventional waveguide EAM, especially of insertion loss, power handling capability, and link gain. A best fiber-to-fiber insertion loss of 4 dB was achieved. Devices were able to operate at 590 mW input optical power to produce a link gain within -10 dB. The link gain saturation at high optical power was observed and then analyzed based on EAM equivalent circuit model. The theoretical prediction agrees with experimental data very well. With the same circuit model that was interpreted as a negative feedback system, link linearity performance at high power was theorized to improve over non-saturation case. Taper structure was added to PCW to further improve the device performance. The devices were also measured at different temperatures. The result showed that higher temperature degrades device link gain performance.

The dissertation is arranged in six chapters. After this introduction chapter, Chapter II describes in detail PCW EAM structure and its benefits. Material and device designs are presented followed by device fabrication procedure. In Chapter III, PCW EAM performances are discussed and analyzed such as low insertion loss and high power operation. We presented taper structure to further increase device power handling capability and also measured device performance at different temperatures. Some of the measurement results are shown. High power induced
index of refraction change is also discussed. Chapter IV is dedicated to link gain saturation due to photocurrent effect. The link gain and noise figure expressions are derived based on EAM equivalent circuit model. The analysis is compared with experimental link gain result to show a close match. Chapter V deals with link linearity performance under high saturation. By separating intrinsic and extrinsic optical transfer curves, the circuit negative feedback effect proves to improve link linearity over non-saturation case. Finally, Chapter VI summarizes the dissertation and gives suggestions for future work.
References


Chapter II

Peripheral Coupled Waveguide

Electroabsorption Modulator

To improve the performance of an EAM, we have implemented peripheral coupled waveguide (PCW) structure in the device design[1]. With the help of PCW, we have greatly improved the power handling capability of our EAMs and thus have improved the link gain in analog fiber-optic links involving such EAMs. This chapter is dedicated to describing PCW structure and corresponding device design and fabrication.

II.1 Basic Concept of PCW

Electroabsorption modulator relies on its electric field controlled absorptive property change to obtain intensity modulation of the light going through it. As briefly described in Chapter I, it employs p-i-n structure and the intrinsic
The layer between p-type and n-type layers is the absorption layer. When the device is reversely biased, most of the voltage applied to the device is dropped on the intrinsic undoped layer. The voltage change causes intrinsic layer material absorption coefficient to change accordingly. Hence light modulation occurs.

EAM modulation performance is related to how large the absorption of optical power in each section of the waveguide will happen due to a certain voltage change. It also related to the fraction of optical power that sees the absorption layer, which is the confinement factor $\Gamma$ because only that portion of the optical radiation is absorbed. In the earlier EAM waveguide design, a large portion of optical mode is in the intrinsic layer with a large $\Gamma$ such as 30%. As input optical power increases, a large $\Gamma$ causes early device saturation because the large photo-generated current causes shielding of the electric field. In the later large optical cavity (LOC) EAM waveguide design [2], the n-type layer below the intrinsic layer is made thicker with relatively higher index of refraction than the earlier design. This design enables optical mode to submerge and locate mostly in the n-type layer. Therefore the portion of optical power in the intrinsic layer drops and $\Gamma$ decreases. The second purpose of LOC design is to form a large optical mode in EAM waveguide so that it is compatible with the fiber mode that couples light in and out. In this way, the mode mismatch between EAM and fiber can be greatly reduced and coupling coefficient can be improved. PCW waveguide structure goes to more extreme. It employs two-step mesa structure. The lower mesa is used for

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1 This is actually only the case at low optical power. At extremely high optical power levels, it will change. That case will be discussed in Chapter IV.
lateral confinement of optical mode and the upper mesa is mainly for microwave electrode. The n-type layer is made even thicker and sometimes a layer of relative lower index of refraction is inserted between intrinsic layer and the major n-type layer. The resultant optical mode mainly resides in the n-type layer with only an evanescent tail penetrates to the intrinsic layer. The confinement factor $\Gamma$ in PCW configuration can be small such as below 10%.

![Figure II.1: A schematic of PCW EAM waveguide cross section.](image)

The advantages of PCW waveguide structure for EAM are as follows [3].

- It lowers the device confinement factor $\Gamma$, which reduces absorption per unit volume for the same input optical power. Therefore it improves device power handling capability.

- By adjusting the lower mesa width and n-type layer thickness, the optical mode size can be made large enough to be comparable to a fiber mode size, thus reducing device coupling loss.

- The optical mode is now submerged down in the n-type layer. It sees less
waveguide side walls which causes light scattering. Hence the device propagation loss can be reduced.

- The optical design is mainly dictated by the lower mesa and the upper mesa controls device microwave properties such as capacitance per unit length of the electrode transmission line. Now designs of the microwave electrode and the optical waveguide mode can be partially decoupled.

II.2 Device Design

The main part of the PCW EAM device design studied in this dissertation is the optical waveguide design, which optimizes material index of refraction and device geometrical structure to ensure PCW configuration. However, for a complete device design, material selection and microwave structure design are indispensable. They are all responsible for the overall performance of the final device.

II.2.1 Optical Waveguide Design

EAM optical waveguide configuration is determined by the geometry of the structure as well as by indices of refraction of different material layers involved in the device. We used commercial simulation software FIMMWave from Photon Design [4] to simulate optical modes in the device, calculate the coupling coefficient and confinement factor.

Table II.1 shows the layer structure of a typical PCW EAM. The QW layer
(layer 5–7) is the active absorption layer. The thick InGaAsP (PL 1.111 µm) layer (about 1.68 µm thick) below is the main waveguiding layer. It has a relative high index of refraction of 3.30. The InP layer between the IQW layer and the waveguiding layer separates further the main optical mode and the evanescent tail in the high index (3.50) QW layer. The top highly doped InGaAs layer is used for p ohmic contact. The InP layer below pushes the optical mode from reaching highly absorptive InGaAs layer.

Table II.1: A typical PCW EAM material layer structure with QW as active layer.

<table>
<thead>
<tr>
<th>Layer</th>
<th>Description</th>
<th>Thickness (µm)</th>
<th>Refractive Index</th>
</tr>
</thead>
<tbody>
<tr>
<td>11</td>
<td>p⁺-InGaAs</td>
<td>0.05</td>
<td>3.592</td>
</tr>
<tr>
<td>10</td>
<td>p-InP</td>
<td>0.75</td>
<td>3.17</td>
</tr>
<tr>
<td>9</td>
<td>p-1.208µm -InGaAsP</td>
<td>0.07</td>
<td>3.36</td>
</tr>
<tr>
<td>8</td>
<td>u-InP</td>
<td>0.03</td>
<td>3.17</td>
</tr>
<tr>
<td>7</td>
<td>u-1.148±0.01µm -InGaAsP</td>
<td>0.007</td>
<td></td>
</tr>
<tr>
<td>6-2</td>
<td>5×u-1.542µm -InGaAsP</td>
<td>0.01</td>
<td>3.50</td>
</tr>
<tr>
<td>6-1</td>
<td>5×u-1.148µm -InGaAsP</td>
<td>0.007</td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>u-1.148±0.01µm -InGaAsP</td>
<td>0.014</td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>n-InP</td>
<td>0.05</td>
<td>3.17</td>
</tr>
<tr>
<td>3</td>
<td>n-1.111µm -InGaAsP</td>
<td>0.58</td>
<td>3.30</td>
</tr>
<tr>
<td>2</td>
<td>u-1.111µm -InGaAsP</td>
<td>1.1</td>
<td>3.30</td>
</tr>
<tr>
<td>1</td>
<td>u-InP Buffer Layer</td>
<td>0.5~1</td>
<td>3.17</td>
</tr>
<tr>
<td>0</td>
<td>S.I. InP Substrate</td>
<td>350 ± 25</td>
<td>3.17</td>
</tr>
</tbody>
</table>

The simulated optical mode of a PCW EAM is shown in Figure II.2. In this figure, the top mesa and bottom mesa widths were set to be 2 and 4 µm. The effective index of refraction and confinement factor were simulated to be 3.2653 and 5.49%. The coupling coefficient from a 2.5 µm waist size Gaussian mode is about 95%. The intrinsic layer is located at the bottom of the top mesa. Most of
the optical mode is within the lower thick InGaAsP (PL 1.111 µm) waveguiding layer. The active layer only sees an evanescent tail of the mode. The simulated $\Gamma$ is very small, enabling low photocurrent density when in operation.

![Simulated optical mode of PCW EAM waveguide.](image)

**Figure II.2: Simulated optical mode of PCW EAM waveguide.**

**II.2.2 Microwave Design for Lumped EAM**

During my research, we mainly focused our EAM design on simple lumped element structure to better understand PCW performance. Although a TW electrode design can improve PCW EAM bandwidth performance, I will limit PCW EAM device to lumped element electrode structure in this dissertation.

The electrode design for lumped EAM is quite straightforward. It only involves a p-contact and an n-contact. Either ground-signal-ground (GSG) or ground-signal (GS) contact pads are made to facilitate microwave probe touching for measurement. Signal contact (p-contact) usually sits on a dielectric pad to be insulated from n-type layer. The area of the p-contact along with the intrinsic
layer thickness determines device capacitance.

II.2.3 Material Design

In order to obtain large slope efficiency for the modulation, the material needs to be designed to have large absorption coefficient change at certain bias voltage. During our device design, both multiple quantum well (MQW) and intra-step quantum well [5, 6] (IQW) were used. The advantages of IQW lie in two folds. Firstly an extra step in the quantum well enables holes to escape from the well more easily. Secondly the structure causes a delay of the increase of absorption coefficient with voltage, which moves the best bias voltage to be higher. The current drifts out of the intrinsic layer faster at higher bias electric field. Both of the advantages improve device’s operation at high optical power and reduces saturation.

Quaternary In$_{1-x}$Ga$_x$As$_y$P$_{1-y}$ was used for designing QWs. Even with lattice matching to the InP substrate, the quaternary material system still have one degree of freedom to change its composition to obtain the desired band-gap energy and to tune the index of refraction [7, 8]. We conducted quantum mechanical electron and hole wavefunction, energy level, and exciton binding energy calculations to obtain the desired exciton peak wavelength for the 1.55 $\mu$m laser radiation. Several parameters such as quantum well width, barrier band-gap, and well band-gap can be used to optimize the design. The operating wavelength is set to be 70-80 nm longer than the exciton peak wavelength to avoid residual absorption.
Figure II.3 shows photocurrent spectrum measurement of an IQW material sample at several bias voltages. The material exciton peak was designed to be 1483 nm. The operating wavelength 1550 nm is 67 nm longer than the exciton peak. At 1550 nm, measured photocurrent increases as the reverse bias voltage is increased, showing QCSE effect.

Figure II.3: Experimental measured photocurrent at different bias voltages versus wavelength for an EAM material (Sample GLDPQ0504222-B).
II.3 Device Fabrication

We fabricated PCW EAMs on designed wafers, which were grown by an outside vendor. First, wafer was cleaved into small pieces and chemically cleaned and prepared. Photoresist was spun on the sample to define p-metal contact. After metal deposition and lift-off, p-metal was formed on the sample. Photoresist was again used to define microwave waveguide, which aligned with p-metal pad. Chemical etching is used to form ridge waveguide. The same procedure was used to form the wide optical waveguide. Afterwards, BCB was applied to passivate waveguide sidewall. Finally the n-contact was deposited. After sample was processed, it was lapped down, cleaved, and mounted for measurement.

Figures II.4 and II.5 show a schematic picture of lumped EAM (n-contact not shown) and its cross section, respectively. Figure II.6 shows scanning electron microscopic (SEM) images of the fabricated device, its facet, and p-contact pad. Two-step ridge waveguide can be clearly seen from the facet.
Figure II.4: A typical lumped element EAM waveguide structure.

Figure II.5: EAM device cross section.
Figure II.6: SEM pictures of (a) lumped EAM, (b) its facet, and (c) its contact pad.
References


Chapter III

PCW EAM Performance

Compared with LiNbO₃ Mach Zehnder modulator (MZM), one of the disadvantage of semiconductor electroabsorption modulator (EAM) is its power handling capability. While LiNbO₃ can handle up to 500 mW input optical power with its commercial devices above which it may suffer photorefractive effect and degrade its performance [1], EAM usually can not survive at 100 mW power and it can cause permanent damage to the device. The apparent difference comes from different operation mechanism. As for LiNbO₃ modulator, it is based on material’s electrooptic effect and is driven purely by voltage. Therefore only negligible amount of light is absorbed by the material and there is almost no photo-generated current flowing through the electrodes. This eliminates the causes of current crowding and temperature increase. On the contrary, EAM relies on its voltage controlled absorptive property to modulate the light going through it. It is inevitable to have photocurrent generation and thermal elevation within the device. The large pho-
tocurrent will also cause a feedback effect to the modulation voltage and the device will have a saturated link gain at high power (See Chapter IV for details). Therefore at high power illumination, the device performance will degrade due to large current density in the absorption layer and material failure at high operational temperature.

From system point of view, high power handling capability of the components in an analog fiber optic link is desirable mainly because of the quadratic dependence of RF link gain on received laser power. More power the components can provide and handle, higher RF link gain the link can realize. This is the reason behind all efforts to increase power handling capabilities of optical modulators and photodetectors.

There are several considerations in EAM design to strengthen its robustness under high power. First of all, reducing current density in the device at the same power is a basic strategy. As is described in chapter II, peripheral coupled waveguide (PCW) design decreases the device confinement factor which also decreases the absorption per unit length in the device. Hence the density of photocurrent generation is proportionally reduced in the front section, enabling device to operate at higher current throughput while maintaining the same maximum current density handling capability. When the optical power is high, as the optical power is reduced in propagation by absorption, the confinement factor $\Gamma$ can be increased to increase gain without causing excessive current density. Secondly, with the exponentially decay absorption profile along the device waveguide, the current density
and resulting temperature increase throughout the device are non-uniform. The maximum current density and temperature at the front end of the device are much higher than the average values across the device. Therefore the situation at certain points in the device will cause failure of the whole device. Waveguide taper structure changes the confinement factor along the waveguide so as to compensate the exponentially absorption profile in straight waveguide. When properly designed, both the current density and temperature across the device can be made more uniform. Thirdly, using more heat conductive material in EAM fabrication or using temperature controller during device operation can greatly mitigate the thermal problem.

In the following sections, we describe PCW EAM and its related taper waveguide structure to increase device power handling capability, device performance under high power, and a high power phenomena: high power induced refractive index change in EAM. Other high power effect of EAM such as gain limit and linearity improvement will be described separately in Chapter IV and V.

III.1 Reducing Insertion Loss

Compared with LiNbO$_3$ modulator, EAM has too much insertion loss. A commercial LiNbO$_3$ MZM has a typical fiber-to-fiber loss of 3 dB. Conventionally designed EAM, however, has an average of fiber-to-fiber loss of about 10 dB.

There are three sources of loss involved in the optical modulator: coupling
loss, propagation loss, and facet reflection loss. Facet reflection loss results from
the mismatch of index of refraction between air and the waveguide material. For
quasi-normal incidence of light, the transmission reduction due to reflection can be
estimated by Fresnel equation to be \(4n/(n+1)^2\), where \(n\) is the index of refraction
of waveguide material. For InGaAsP family material, the reflection loss per facet
is about 1.5 dB. Facet reflection loss can usually be eliminated by anti-reflection
coating of the facet for certain wavelength. Therefore, it is not an important
issue in device design. Coupling loss comes from the mode mismatch between the
coupling fiber and the modulator waveguide. The fiber has a circular mode. The
waveguide mode is determined by its indices of refraction of different layers and
the lateral geometry of the structure, which is usually asymmetric. The overlap
integral of the two modes determines the coupling loss. Propagation loss is caused
by material residual absorption and scattering loss by semiconductor-air interface.

LiNbO\(_3\) modulator with a waveguide fabricated by Ti diffusion has a di-
mension of several microns, which is close to fiber mode. On the contrary, EAM
made of InGaAsP semiconductor materials confines the optical mode by epitaxially
grown layers with different index of refraction in vertical direction and by etched
mesa in lateral direction. The mode of EAM is not circular symmetric and quite
small compared with a fiber mode. Therefore, the coupling loss of EAM is larger
than LiNbO\(_3\) modulator. As light propagates along the waveguide, it does not ex-
perience much attenuation in LiNbO\(_3\) modulator since the material is transparent
and the scattering loss of a diffused index profile is very small. However, EAM
relies on its absorption property change and operates pretty close to the band-gap wavelength, which introduces large residual loss even when voltage is not applied. The waveguide mode could experience significant scattering loss from roughness in the etched mesa depending on the mode profile. Reducing coupling loss and propagation loss for EAM is beneficial for improving EAM link gain.

III.1.1 Reducing Coupling Loss

Making the waveguide mode large, symmetric, and similar to the input and output fiber modes reduces the coupling loss. Conventional EAM with absorption InGaAsP layer sandwiched between p-type and n-type InP layers has very small dimension in the vertical direction. The optical waveguide and electrical structure to create the EA effect shares the same intrinsic MQW layer. To reduce device capacitance for bandwidth consideration, the width of the ridge waveguide can not be made too large either. Such a structure suffers large coupling loss. Peripheral coupled waveguide structure has a thick quaternary layer below the absorption layer as the major waveguiding layer (1.5 - 2 \( \mu \)m), which greatly increases the vertical size of the mode. In the lateral direction, device capacitance is mainly affected by the top microwave waveguide while the second step etched optical waveguide width can be made large to confine the optical mode. Thus the mode in PCW structure is enlarged in both horizontal and vertical directions compared with the conventional design.

The optical modes of conventional EAM waveguide structure and PCW
structure were simulated by FIMMWave software as shown in Figure III.1. PCW optical mode is apparently much larger and more circularly symmetric. The coupling losses between waveguides and 3 μm diameter lensed fiber (emulated by a Gaussian mode) are about 3.5 dB and 0.5 dB for conventional EAM waveguide and PCW waveguide, respectively.

![Waveguide modes](image)

**Figure III.1**: Waveguide modes of (a) conventional EAM and (b) PCW structure EAM (on the same scale).

### III.1.2 Reducing Propagation Loss

As discussed earlier, EAM waveguide propagation loss is mainly caused by residual absorption and scattering. To reduce the residual absorption, PCW structure lowers the confinement factor $\Gamma$ so that less absorption per unit length occurs. Besides, the submerged optical mode is away from waveguide ridge side walls, edges and corners, highly absorptive top InGaAs layer grown for ohmic p-contact, p-type InP layer with free carrier absorption. All these factors reduces EAM propagation loss in PCW structure. The best propagation loss was obtained
from a intra-step quantum well (IQW) PCW EAM device to be 0.8 dB/mm, which
is a substantial improvement over about 10 dB/mm for conventional waveguide
design.

III.1.3 Loss Measurement Using Febry-Perot Effect

Since EAM has pretty large propagation loss in the conventional design, the
characterization of its loss is usually carried out by cut-back method [2]. Different
lengths of devices with the same structure are prepared and fiber-to-fiber loss is
measured. By plotting the fiber-to-fiber loss versus device length, a straight line
fitting can be obtained as illustrated in Figure III.2. Propagation loss per unit
length can be extracted from the line slope. The fitted line can be extrapolated to
zero length to estimate coupling loss.

However, cut-back method suffers several drawbacks. Firstly, it assumes
all the devices have the same material layer structure and waveguide structure.
Growth and fabrication non-uniformity can cause variance of individual device.
Secondly, it assumes coupling loss for testing of different devices is a constant. This
can hardly be achieved due to the alignment difficulties. Thirdly, several lengths of
devices need to be prepared, adding inconvenience to the measurement. Fourthly,
lack of sufficient large number of data points also introduces error. Finally, if the
propagation loss is small and the device is not anti-reflection (AR) coated, there
will be Fabry-Perot effect (FPE), which adds another degree of uncertainty to the
measurement. Thus the cut-back method is only valid for large propagation loss
Figure III.2: An illustration of cut-back method measurement plot.

case in which FPE is negligible.

FP method is an easy method that can be employed to measure propagation and absorption in EAM device if the propagation loss is low and device facets are not AR coated. In this case, light experiences multi-reflection between two facets as shown in Figure III.3. By tuning the incoming light wavelength, the light transmission, reflection and absorption exhibit periodic patterns, which give us information of both coupling loss and propagation loss.

Following analysis of multi-reflection in an FP cavity with coupling fibers [3], optical power transmission \( T \) through an optical waveguide can be derived an-
alytically as follows.

\[ T = \frac{(1 - R_1)(1 - R_2)C_1C_2e^{-\alpha_{\text{eff}}L}}{(1 - R_{12}e^{-\alpha_{\text{eff}}L})^2 + 4R_{12}e^{-\alpha_{\text{eff}}L}\sin^2\left(\frac{2\pi n L}{\lambda}\right)} \quad (\text{III.1}) \]

where \( R_1 \) and \( R : 2 \) are reflectivities of two waveguide facets, \( C_1 \) and \( C_2 \) are coupling coefficients between waveguide and fibers at two facets, \( \alpha_{\text{eff}} \) is the effective attenuation coefficient (including confinement factor \( \Gamma \)) due to absorption and scattering, \( L \) is the waveguide length, \( R_{12} \) is the geometric average of \( R_1 \) and \( R_2 \), which is \( \sqrt{R_1 R_2} \), \( \lambda \) is the light wavelength, and \( n \) is the effective index of refraction of the waveguide. The period of FP pattern is a fraction of nm and absorption coefficient change in one period is negligible. Hence, we treat \( \alpha_{\text{eff}} \) as a parameter independent of wavelength.

When the propagation loss \( e^{-\alpha_{\text{eff}}L} \) is high, Equation III.1 reduces to:

\[ T = (1 - R_1)(1 - R_2)C_1C_2e^{-\alpha_{\text{eff}}L} \quad (\text{III.2}) \]

Equation III.2 shows that three types of loss do not depend on light wavelength as propagation loss is high.

Figure III.3: A schematic of multi-reflection in EAM waveguide.
From transmission equation III.1, waveguide effective index of refraction $n$ and effective attenuation coefficient $\alpha_{\text{eff}}$ can be obtained.

$$n = \frac{\lambda^2}{2L\Delta \lambda} \quad (\text{III.3})$$

$$\alpha_{\text{eff}} = \frac{1}{L} \ln \left( \sqrt{T_{\text{max}}} + \sqrt{T_{\text{min}}} \right) + \frac{1}{L} \ln(R_{12}) \quad (\text{III.4})$$

where the $T_{\text{max}}$ and $T_{\text{min}}$ are maximum and minimum transmission while light wavelength is tuned, and $\Delta \lambda$ is the FP pattern period.

The reflectivity $R_1$ and $R_2$ can be calculated by the Fresnel equations.

$$R_1 = R_2 = R_{12} = \left( \frac{n - 1}{n + 1} \right)^2 \quad (\text{III.5})$$

The product of coupling coefficients $C_1C_2$ can also be obtained from Equation III.1.

$$C_1C_2 = \frac{T_{\text{max}}(1 - R_{12}e^{-\alpha_{\text{eff}}L})^2}{(1 - R_1)(1 - R_2)e^{-\alpha_{\text{eff}}L}} \quad (\text{III.6})$$

or

$$C_1C_2 = \frac{T_{\text{min}}(1 + R_{12}e^{-\alpha_{\text{eff}}L})^2}{(1 - R_1)(1 - R_2)e^{-\alpha_{\text{eff}}L}} \quad (\text{III.7})$$

When same type fibers are used for both front and back facets, similar coupling coefficients of $C_1$ and $C_2$ are expected.

Based on the above discussion, EAM propagation loss (equivalent to $e^{-\alpha_{\text{eff}}L}$) and coupling loss ($C_1$ and $C_2$) can be extracted from FP measurement with just one device. Figure III.4 shows a measurement result done on an EAM device for transmission, reflection, and photocurrent by tuning the incoming light wavelength. Clear periodic patterns appear for all three curves and the period is about 0.44
nm for this 600 µm device. The contrast ratio of transmission $T_{\text{max}}/T_{\text{min}}$ variation in the measurement wavelength range is negligible. This validates the assumption that $\alpha_{\text{eff}}$ in such a small wavelength range can be treated as a constant if only one or two periods are all our focus. The propagation loss $e^{-\alpha_{\text{eff}}L}$ can be easily calculated to be about 1 dB. The total coupling loss $C_1C_2$ is about 3.3 dB. If it is evenly distributed, each facet contributes 1.65 dB.

![Graph](image)

Figure III.4: Transmission (solid dot), reflection (hollow dot), and photocurrent (solid triangle) patterns due to Fabry-Perot effect in EAM loss measurement.

### III.2 Tapered Waveguide

For PCW with uniform waveguide cross section throughout, the optical confinement factor $\Gamma$ is a constant along the waveguide. Due to the exponential
decay profile of the absorption in the waveguide, maximum absorption occurs at the front facet of the device. The high current density caused by this absorption drives the device to saturate at that spot first. The accompanying temperature elevation and thermal reflection by the facet interface exacerbates the situation, causing device to fail. By incorporating a waveguide taper in the PCW, the $\Gamma$ at the front facet can be lowered to reduce the electroabsorption saturation at that facet. The photocurrent density along the waveguide can be made more evenly distributed and failure at hot spots can be avoided. Moreover, lower $\Gamma$ is usually achieved by allowing the optical mode to be more submerged and enlarged, which helps to improve the coupling coefficient. Compared with uniform PCW EAM, tapered PCW EAM also increases $\Gamma$ in the middle and back section of the optical waveguide where current density and heat generation are not as serious as the front section. Therefore, large slope efficiency can be obtained with a shorter device length $L$ [4].

III.2.1 Simulation Results of Confinement Factor Tuning

As described in Chapter II, PCW structure consists of two mesa steps. Both mesas’ width and depth can be adjusted to tune the confinement factor. Figure III.5 shows an example of tuning the width and depth of the lower optical waveguide. The simulation results of confinement change due to geometrical parameters variation are presented in Figure III.6. It can be seen that $\Gamma$ can be reduced by increasing optical waveguide width, reducing optical waveguide depth,
or reducing microwave waveguide width. In practice, only waveguide width can be continually tapered along the waveguide. Waveguide depth change needs to be done discretely by section.

![Figure III.5: EAM waveguide with tapered optical waveguide width and depth.](image)

### III.2.2 Experimental Results

To verify the confinement tuning capability by adjusting geometrical parameters, we fabricated non-tapered PCW IQW EAM with different optical waveguide (lower mesa) width $W_1$ and microwave waveguide (upper mesa) width $W_2$ values, with $W_1 = W_2 + 4 \mu m$. The waveguides studied are 1.2 mm long. Figure III.7 shows optical transfer curves of three cases with $W_2 = 1.5$, $2$, and $3 \mu m$ respectively. It is clear that with the increase of $W_2$, the device propagation loss increases and $V_\pi$ decreases, indicating an increase of confinement factor.

The detailed simulation and measurement results are shown in Table III.1. The ratio of simulated confinement factors are close to that of the propagation losses (in dB/mm) for the corresponding devices.
Figure III.6: Confinement factor change by tuning of (a) optical waveguide width, (b) optical waveguide depth, and (c) microwave waveguide width.

Table III.1: Simulated and measured results from PCW IQW EAM with various $W_1$ and $W_2$ values ($W_1 = W_2 + 4 \mu m$).

<table>
<thead>
<tr>
<th>$W_2$ ($\mu m$)</th>
<th>$\Gamma$ (simulated)</th>
<th>$V_\pi$ (measured)</th>
<th>Propagation Loss (measured) $dB/mm$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.5</td>
<td>2.64</td>
<td>2.95</td>
<td>0.81</td>
</tr>
<tr>
<td>2</td>
<td>4.26</td>
<td>2.37</td>
<td>1.7</td>
</tr>
<tr>
<td>3</td>
<td>7.11</td>
<td>1.89</td>
<td>2.2</td>
</tr>
</tbody>
</table>
Figure III.7: Measured normalized optical transfer curves and EAM photocurrent versus bias voltage for EAMs with different $W_2$ values.
PCW EAM device with tapered waveguide structure was also fabricated with a structure shown in Figure III.8. The optical and microwave waveguides widths along with optical waveguide depth were varied along the waveguide to tune the $\Gamma$. The $\Gamma$s at both input and output facets were made small to not only reduce the absorption and thermal problem but also to improve the coupling from and to the fibers. The maximum $\Gamma$ occurs in the middle of the device, which is the work force of modulation.

![Figure III.8: Schematic layout of the tapered PCW EAM.](image)

In order to compare the optical power handling capability, the link RF gain was measured at 100 MHz as a function of the optical power for both non-tapered and tapered PCW MQW EAMs with the same length of 1.6 mm. Each device had an optical insertion loss ($\sim 7$ dB) and an equivalent ($V_\pi \sim 1.4V$). Figure III.9 shows the measurement results for the non-tapered and tapered devices respectively. Both results show gain saturation at high optical power levels, which is due to the reduction of effective modulation voltage as a result of photocurrent feedback effect. This gain saturation will be discussed in detail in Chapter IV.
Theoretical curves match the measured data closely, especially for the tapered case. At high optical power above 20 dBm, the link RF gain for tapered device still follows the trend of the prediction. The non-tapered device, however, shows a trend of lowering link gain. This is attributed to the high current density and thermal effect at the front facet.

### III.3 Thermal Effects

Thermal effects can be very serious for EAM operating at high optical power levels. The temperature increase in EAM is mainly caused by the Joule heating in the device junction when large photocurrent flows through. High temperature can cause material properties change which is detrimental to the device operation. The non-uniform distribution of the temperature due to non-uniform distribution of the current density also impacts device’s mechanical robustness.

#### III.3.1 Thermal Runaway

Thermal runaway is a positive feedback loop caused by increased photocurrent in EAM. As input optical power increases, heat generation and temperature increase accordingly. The semiconductor material has reduced band-gap at elevated temperature, which induces more absorption. This positive feedback loop goes along and will cause temperature to rise rapidly at a high optical power level and finally destroys the device under test. This phenomenon is called catastrophic
Figure III.9: Measured link RF gain for (a) non-tapered and (b) tapered PCW EAMs. Dots represent the measured RF gain and the lines represent the predictions by the theory detailed in Chapter IV.
optical damage (COD).

Figure III.10 shows the SEM pictures of EAM device before and after COD. It can be seen that the device front facet was mechanically damaged due either to high current density breakdown or thermal melting. It might also be caused by large temperature gradient. Material debris can be spotted around the damaged facet. The COD happened at the front facet is due to the maximum current density and poor thermal circulation.

![SEM pictures showing EAM device facet (a) before and (b) after catastrophic optical damage.](a) (b)

Figure III.10: SEM pictures showing EAM device facet (a) before and (b) after catastrophic optical damage.

### III.3.2 Link Gain Reduction

To see how the temperature can change the device operation at high power, EAM was tested on a heat sink with constant temperature as schematically shown in Figure III.11. The real temperature of the device tested under high power is still much higher than the heat sink temperature because of the device substrate thickness and low heat conduction. However, by adjusting the heat sink temper-
ature, the device temperature is expected to change accordingly. Figure III.12 shows EAM link gain under heat sink temperature of 20 °C, 30 °C, and without temperature control. The curve with no temperature control has the worst saturation at high power while the curve of 20 °C was best. The result clearly shows the link gain deterioration with raised temperature.

Figure III.11: Schematic of EAM device test on a heat sink.

### III.3.3 Thermal Considerations

Since high temperature is harmful for EAM device operation, thermal considerations need to be taken care of during the device design. PCW structure can reduce current density per unit length, thus reduce heat generation per volume. Tapered structure can decrease the excessive absorption and lower the heat load at the front facet. It also evens out the temperature throughout the device. Besides, high thermal conductive material such as AlN can be used to deposit on the device surface to help heat conduction. The device can also be flip-chip bonded to high
Figure III.12: Measured link gain of EAM on a heat sink with 20 °C (diamond), 30 °C (square), and no temperature control (triangle).
thermal conductive materials such as diamond. During the application, heat sink and active temperature controller are beneficial for high power operation.

**III.4 EAM Operating Under High Power**

With the help of PCW structure, we were able to operate our EAM devices under high power. Both dc and RF characteristics were measured and the results showed that the devices could survive at high power without damage [5]. The performance, however, degraded as the optical power was increased.

Since EAM is a light absorbing device, it can be treated as a photodetector if the input light is a modulated light. The responsivity is low though because EAM operates at the absorption edge of the material. However, if bias voltage is increased or light wavelength is decreased, EAM can act very reasonably as a photodetector.

Figure III.13 shows dc photocurrent generated at different powers and normalized transmission of a MQW PCW EAM as a function of the bias voltage. The measurement was done with light wavelength of 1560 nm. A fiber-to-fiber insertion loss was measured to be 7.4 dB.

The link gain at 50 MHz and at different input optical power levels is displayed in Figure III.14. A link gain of about -8.5 dB was obtained at 20 dBm optical power, above which the gain decreases due to current screening and thermal effects.
Figure III.13: Normalized EAM transmission and device photocurrent at different input optical power levels. The optical powers are 2.2 mW (hollow triangle), 45 mW (solid triangle), 125 mW (hollow circle), 290 mW (hollow square), and 590 mW (solid square).
Figure III.14: Measured EAM link gain at 50 MHz versus input optical power.
The device was also measured as a photodetector at a bias voltage of -2 V without heat sink and active cooling as shown in Figure III.15. The maximum optical power launched into the device reached 27.7 dBm (590 mW), which was limited by the source available in the experiment. Photocurrent was measured to be 222 mA at that optical power level. The device responsivity drops at high power and a 1 dB small-signal compression current of 43 mA was measured at 300 MHz.

Figure III.15: Device output dc current ($V_B=-2V$) and corresponding responsivities as a function of optical input power.

These measurement results showed that PCW EAM devices we made can successfully operate under high optical power and high photocurrent. The link gain involving these devices improved from around -20 dB of conventional design
to higher than -10 dB.

III.5 High Power Induced Index Change

At high optical power, the photocurrent flowing through the device can be very substantial. The index of refractive change induced by high inject current level in forward biased devices can also happen in reversed biased EAM if the photogenerated current is high enough [6]. In dc optical transfer curve measurement at high power, step-like modulator photocurrent and multiple-peak light transmission (equivalent to the received photodetector current) were observed in un-coated EAM devices. Figure III.16 shows modulator photocurrent and photodetector current at 10, 15, and 20 dBm input optical power levels. The device measured had a propagation loss of around 1 dB. At 10 dBm, the optical transfer curve along with the modulator current curve are still quite smooth. As optical power increased to 15 and 20 dBm, modulator current curves show several steps and photodetector current curves show several peaks with increased bias voltage. This is basically caused by the change of index of refraction with large current flow. The curves shapes are the manifestation of the Febry-Perot effect (such as shown in Figure III.4) overlapped with the electroabsorption effect (low power curves). At some bias points, it is confusing that both modulator and photodetector currents increase and they are not complementary. The puzzle can be solved by FPE.

The high power induced material index of refraction can also be verified by
Figure III.16: Febry-Perot effect caused by high power induced material index of refraction change in EAM. Both the modulator current and received photodetector current are shown at 10 dBm (solid), 15 dBm (dash), and 20 dBm (dot) input optical power, respectively. The laser wavelength was at 1548 nm.
changing the input optical wavelength and keeping optical power to be 20 dBm such as shown in Figure III.17. As the wavelength is tuned, both modulator and photocurrent detector curves shift. If the wavelength is changed to one FP period, the curve shapes return to their original form, closing a cycle.

![Figure III.17: Febry-Perot effect caused by high power induced material index of refraction change in EAM at different wavelengths. The input optical power was 20 dBm.](image)

With a sharp slope presented at some bias voltages on the optical transfer curve with FP effect at high power, a very good link gain should be expected. However, the real link gain measured is about 10 dB lower than predicted by transfer curve slope efficiency. This shows the complication of the combination of
voltage controlled electroabsorption effect and current induced index of refraction change. Further studies need to be carried out for a better understanding.

III.6 Acknowledgement

Chapter III, in part, uses the material as it appears in two published paper:

References


Chapter IV

Gain Limit and Noise Figure of High Power EAM

The electroabsorption modulator operated at high power has very different behaviors as at low power. The most prominent one is the gain saturation. As the input optical power increases, the EAM link gain as a function of input power does not follow the conventional quadratic curve any more and drops under the predicted curve. The reason for the saturation can be from many effects such as screen effect due to photocurrent crowding [1] and material property change due to temperature increase. The most important reason, however, is the circuit feedback effect caused by photocurrent generation from the photon absorption. The simple explanation is that the increased photocurrent results in more voltage dropped on other resistive elements than the p-i-n junction. As the effective modulation voltage decreases, the modulation efficiency drops.

The photocurrent generated in EAM also added shot noise to the link. As
the link gain increases with the increased optical power, the portion of the EAM shot noise seen from the output end becomes more and more important. This results in a minimum NF at certain optical power level. Below that level, the increased link gain is beneficial for NF. Above that level, the EAM shot noise outperforms link gain and brings NF larger again.

IV.1 EAM Equivalent Circuit Model

Gain saturation due to photocurrent feedback in circuits can be most easily described in terms of a lumped EAM at low frequencies. Lumped EAM can be presented in a circuit environment with a small signal model as described in G. L. Li’s Ph.D. dissertation [2]. The model is duplicated here in Figure IV.1. The EAM is modeled as a current path in parallel with a capacitor in his analysis. As input optical power increases, however, the photocurrent flowing out of the junction also flows through other circuit elements, and in turn affects the current path itself. Therefore, EAM is more accurately modeled as a power dependent resistance in parallel with a capacitor at high optical power.

IV.1.1 Circuit Model

Figure IV.2 shows the small signal ac equivalent circuit model of lumped EAM driven by an RF source. The RF source is modeled as a current source $i_S$ and a source resistor $R_S$. $R_L$ is the EAM termination load resistance. $R_M$
is the modulator series resistance coming from doped semiconductor layers and Ohmic contacts between semiconductor and electrode metals. $i_P$ represents the photocurrent going through the junction. $C_M$ is the junction capacitance, which is ignored in the following analysis for very low frequency effect. The photocurrent generation process, as we will see later, can be modeled as a linear voltage-current relationship, thus a resistor $R_P$.

The fiber-optic link is shown in Figure IV.3. The input and output coupling

![Figure IV.1: EAM circuit model constructed by G. L. Li.](image1)

![Figure IV.2: Small-signal ac equivalent circuit of electroabsorption modulator. The resistor $R_P$ represents the voltage-dependent modulator photocurrent source.](image2)
losses are represented as $t_I$ and $t_O$. As the light propagates through the waveguide, it experiences a propagation loss of $t_P$. The device is assumed to be anti-reflection (AR) coated and there is no reflection loss. $P_{IN}$ and $P_{OUT}$ are optical powers at two ends of the waveguide after input coupling and before output coupling.

Figure IV.3: Diagram of the link. The link output is the power delivered to $R_D$.

$P_{IN}$ and $P_{OUT}$ are related by device propagation loss and normalized optical transfer curve $T_{NORM}(V)$.

$$P_{OUT} = P_{IN} t_P T_{NORM}(V)$$  \hspace{1cm} (IV.1)

where $t_P$ is the waveguide propagation loss. When the device is fed by a dc bias voltage $V_B$ and an ac modulation voltage $v_M$, Equation IV.1 can be approximated by a Taylor series expression. If we only include the first term of the Taylor’s series in a small signal analysis, we obtain:

$$P_{OUT} = P_{IN} t_P T_{NORM}(V_B + v_M) = P_{IN} t_P \left(t_B - \frac{\pi}{2V_{pe}} v_M \right)$$  \hspace{1cm} (IV.2)

Here $t_B$ is the loss caused by dc bias voltage. The $V_{pe}$ is the EAM equivalent half wave voltage defined as $V_{pe} = \pi/(2T_{NORM}'|v_b|)$. When FP effect can be neglected
with reasonably large absorption coefficient or facet AR coating, the photocurrent generation \( i_P \) is given by \((P_{IN} - P_{OUT})\eta_M\), where \( \eta_M \) is the modulator responsivity at the bias point, so

\[
i_P = P_L t_I \eta_M (1 - t_P t_B) + P_L h_P t_P \eta_M \frac{\pi}{2V_{pe}} v_M \quad (IV.3)
\]

Here \( P_L \) is the laser power and it is related to \( P_{IN} \) by input coupling loss as \( P_{IN} = P_L t_1 \). The current caused by the reverse junction itself is neglected here for simplicity. The coefficient of \( v_M \) has the units of inverse resistance, so we can define the effective small-signal ac photocurrent resistance \( R_P \) as

\[
R_P = \frac{2V_{pe}}{P_L h_I t_P \eta_M \pi} \quad (IV.4)
\]

This is a resistor that depends on the input laser power \( P_L \). As the input optical power increases, \( R_P \) decreases.

The equivalent circuit can be solved to give the modulator voltage \( v_M \) in terms of the source current \( i_S \) as

\[
v_M = i_S \frac{R_L R_S}{R_L + R_S} \frac{1}{1 + \frac{P_L h_I t_P \eta_M \pi}{2V_{pe}} \left( \frac{R_M}{R_L + R_S} + \frac{R_L R_S}{R_L + R_S} \right)} \quad (IV.5)
\]

This equation shows that the modulator ac voltage is inversely proportional to the optical power at high optical power. As power goes up, less and less effective voltage is utilized for modulation.

### IV.1.2 Junction as an Equivalent Resistor

The EAM equivalent circuit presented above shows its junction can be modeled by a differential resistance that is inversely proportional to the input optical
power. The current flowing through the junction includes both photocurrent and
current conduction due to junction resistor. They both can be represented by $R_P$.

The $R_P$ lowering at high power can also be seen from the scattering matrix
parameter $S_{11}$ measurement as shown in Figure IV.4. The $S_{11}$ for a one port de-
vice such as EAM is the microwave reflection coefficient caused by the impedance
mismatch between source impedance and device impedance. It can be expressed as
\[
\frac{Z_M - Z_S}{Z_M + Z_S} \quad [3],
\]
where $Z_S$ is the source impedance and $Z_M$ is the EAM device impedance.
The source impedance $Z_S$ is commonly 50Ω for microwave circuit. In the case of
no EAM termination resistance, the EAM impedance is $R_M + R_P || C_M$. It simpli-
fies to $R_M + R_P$ at low frequency. Since $R_M$ does not change much at different
optical power levels (about 5Ω for our EAM devices), the information of $R_P$ can
be extracted from the EAM $S_{11}$ data. The Smith chart is one form to present the
scattering matrix [3]. At low frequency, the lines in the Smith chart approach the
center horizontal real impedance axis, which indicates 0 at the left end and $\infty$ at
the right end. The measured Smith chart of our EAM device at 15 dBm input
optical power clearly shows lower resistance than that at no optical input.

**IV.2 Link Gain Calculation**

Based on the EAM equivalent circuit model, a link gain calculation can be
carried out. As described in Chapter I, the link gain of an analog fiber-optic link is
defined as the ratio of the link output power to the input available power from the
source. When $R_L$ is adjusted to obtain maximum power transfer from the source, the input available power in this circuit model is $\frac{i^2}{2}R_S/4$ and the link output power is the power delivered to the photodetector load resistance $R_D$.

The link gain can be derived as shown in Equation IV.6, where $\eta_D$ is the detector responsivity.

$$g = \left[ \frac{P_L t_i t_p t_O \eta_D \pi}{2 V_{\pi e}} \right]^2 R_D R_L \left[ \frac{4 R_L R_S}{(R_L + R_S)^2} \right] \left[ 1 + \frac{P_L t_i t_p \eta_M \pi}{2 V_{\pi e}} \left( \frac{R_M + R_L R_S}{R_L + R_S} \right) \right]^2$$

(IV.6)

The gain is the product of three terms: the link gain for an external modulation link with impedance-matched input, the effect of an impedance mismatch between the source and termination, and a third term with the dependence on the input optical power. In the limit of small $P_L$, the third term approaches unity and the link behaves as expected for an external modulation link. In the limit of large $P_L$, the third term becomes inversely proportional to $P_L$. In this limit, the gain becomes independent of either $P_L$ or $V_{\pi e}$.

Figure IV.4: EAM $S_{11}$ measurement as shown in Smith chart. The left and right charts are at no optical input and 15 dBm input optical power, respectively.
IV.2.1 Gain Limit at High Power

From Equation IV.6, the EAM link gain approaches a limit as the input laser power $P_L$ becomes high enough.

$$g_{\text{Limit}} = \left( \frac{t_O \eta_D}{\eta_M} \right)^2 \frac{4 \frac{R_D}{R_S}}{\left( 1 + \frac{R_M}{R_S} + \frac{R_M}{R_L} \right)^2} \quad (\text{IV.7})$$

The effect of this gain limit is shown in Figure IV.5. The case of $\eta_M = 0$ is the standard external modulation result with no photocurrent effect. The case with $\eta_M = 1 \text{A/W}$ approximates performance expected from a high-power electroabsorption modulator. For a high-performance modulator, the limiting value is near 0 dB. The limit can be increased if the modulator responsivity $\eta_M$ is reduced, but even at a low responsivity such as 0.1 A/W, the photocurrent effect has an impact.

The input impedance $Z_M$ of the modulator varies with optical power. The relation is given by

$$Z_M = \frac{1}{\frac{1}{R_L} + \frac{1}{\frac{1}{R_M + \frac{2t_{\text{ne}}}{R_L t_O \eta_M \eta_D^2}}}} \quad (\text{IV.8})$$

The input impedance is plotted in Figure IV.6. While the impedance mismatch does affect the gain, it is not the primary factor in the gain limit.

IV.2.2 Performance Affected by Other Parameters

The influence of modulator responsivity $\eta_M$ on the link gain limit is presented in Figure IV.5. Besides $\eta_M$, the EAM link gain limit is also affected by several circuit parameters including $R_S$, $R_M$, $R_L$, $R_D$, $t_O$, and $\eta_D$. The photode-
Figure IV.5: Link electrical gain as a function of laser power, for various values of the modulator responsivity $\eta_M$ (units of $\eta_M$ are A/W). The dc component of the modulator photocurrent is also plotted. The parameter values are: $V_{\pi e}=1$ V, $R_S = R_L = R_D = 50\Omega$, $R_M = 5\Omega$, $\eta_D = 0.8$ A/W, $t_I = -2$ dB, $t_P = 1$, $t_O = -2$ dB, and $t_B = 0.5$.

Figure IV.6: Modulator input impedance as a function of laser power, for various values of the modulator responsivity. The modulator capacitance is zero in this calculation. $R_M$ is the modulator series resistance. Parameters are as in Figure IV.5.
tector responsivity $\eta_D$ is constrained by its quantum efficiency. The larger $\eta_D$, however, the higher the gain limit. The modulator input coupling loss $t_1$ and propagation loss $t_P$ do not affect the link gain limit because they can be compensated by raising the input laser power. Modulator series resistance $R_M$ is unavoidable. But with a lower $R_M$, less voltage is wasted when $R_P$ goes down at high optical power. $R_L$ as an impedance matching termination resistance is usually selected to be matched to $R_S$ at low power. However, this value can also be chosen to purposely match the circuit impedance at high power to improve link gain limit. $R_S$ and $R_D$ are usually limited by system and set to be 50 $\Omega$. With the help of impedance transform, a decrease of $R_S$ and an increase of $R_D$ will be beneficial for raising the gain limit.

**IV.3 Experimental Verification**

We have verified the gain limit by measuring the gain of a link using an electroabsorption modulator at high optical power levels. The modulator structure is similar to that described in [4]. The $V_{ze}$ was 0.85 V and the input and output losses were approximately $t_1 = t_O = 0.5$. The bias point was about $t_B = 0.5$, which occurred at 1.5-V reverse bias. The ac input voltage was 0.063-V peak-to-peak. The modulator’s apparent dc responsivity varied from 0.7 to 1.5 A/W, indicating some mechanism creating additional photocurrent beyond simple absorption. An RF responsivity $\eta_M = 0.8$ A/W was used to fit the calculation to the measured
data. The measurement frequency was 50 MHz, well below the $RC$ bandwidth, which makes the assumption of neglecting junction capacitance valid.

The results are shown in Figure IV.7. The gain follows the theoretical prediction very closely. The gain deviates from the prediction of this model only at the highest powers used (> 250 mW) due to heating.

![Figure IV.7: Experimental measurement of a link using an electroabsorption modulator at 1550 nm, compared with the theoretical gain calculation.](image)

IV.4 Noise Figure

Due to the EAM photocurrent effect, the link noise figure also needs to be modified from its conventional form as detailed in [2]. At low power, EAM acts as a perfect external modulator with voltage controlled characteristics and dc photocurrent can be ignored. At high power, however, dc EAM photocurrent
increases substantially and generated shot noise in the device junction. Moreover, the high power raises the link gain and makes the EAM shot noise more visible at the link output. It finally plays a critical role in the noise performance of the link.

IV.4.1 Link Noise Performance

With the same low frequency circuit model in Figure IV.2 and IV.3, the link noise figure can be derived as

$$f = \frac{N_{\text{OUT}}}{gkT_0} = 1 + \frac{f_R}{g} + \frac{2eP_Lt_Bt_O\eta_D R_D}{gkT_0} + \frac{eP_Lt_B\eta_M(1 - t_P t_B)R_S}{2kT_0} \left[1 + \frac{R_S(R_L + R_o)}{R_S R_L}\right]^2$$  \hspace{1cm} (IV.9)

where \(N_{\text{OUT}}\) is the total output noise, \(f_R\) is the receiver noise figure (\(f_R = 1\) is assumed here), \(k\) is Boltzmann’s constant, \(T_0\) is 290 K, and \(e\) is the elementary charge. We have assumed there is no relative intensity noise. The thermal noise from \(R_M\) and \(R_L\) are neglected in comparison with the shot noise.

The first three terms are the familiar input, receiver, and detector shot noise terms. The fourth term is due to shot noise from the dc component of the modulator photocurrent. The noise figure is plotted in Figure IV.8.

The NF plot shows a minimum NF at certain optical power level if the modulator responsivity \(\eta_M\) is not zero. This minimum value is a balance between the EAM shot noise and link gain. At low power, link gain increase from raised optical power is dominant and results in a NF reduction. At high power, EAM shot noise takes over and NF increases again.
IV.4.2 Influence of $V_{\pi e}$

Even though the gain limit is independent of $V_{\pi e}$, there is some advantage to a low $V_{\pi e}$. The gain reaches its limit faster and at a lower optical power with a low $V_{\pi e}$. The minimum noise figure is also lower and it occurs at a lower optical power. Figure IV.9 shows these effects.

IV.5 EAM based on Blue-shift QCSE

The preceding analysis on EAM link gain limit is based on the fact that optical transfer curve has a negative slope, which results in the negative sign before $V_{\pi e}$ in Equation IV.1 and a positive junction resistance $R_P$. This is true for devices based on material absorption behavior of FKE in bulk semiconductor and QCSE in simple quantum well structures. These effects generally have a red-shift optical
absorption spectrum when bias voltage is applied as illustrated in Figure IV.10(a). EAM typically operates at the absorption spectrum tail (e.g. 1.57 µm) to avoid residual attenuation at zero bias. More absorption occurs when bias voltage is increased. Consequently this causes a positive junction resistance in the modulator circuit, which directly limits link gain and link noise figure at high optical power. A different behavior would occur, as shown in Figure IV.10(b), if the quantum wells can be specially designed to have a blue-shift optical absorption spectrum. In such a structure, transition energy from heavy hole to electron is smaller at lower bias voltage. While the bias voltage is increased, transition energy increases as well, which results in less absorption at higher bias voltage. With a blue-shift material incorporated into EAM absorption region, the equivalent differential junction resistance becomes negative. This negative resistance can cancel collective positive resistances of other circuit elements of EAM at a certain input laser power level.
Consequently, the link gain can approach higher values than red-shift material based link theoretically. The link noise figure will be benefitted as well. This is expected to extend the gain limit of EAM based analog fiber-optic links. The above mentioned effect can be regarded as a resonant effect between the EAM circuit and photogenerated current. At the frequencies where EAM junction capacitance cannot be neglected, the negative resistance can only cancel the real part of the circuit impedance. Hence the link gain resonant peak will be lowered and linewidth will be broadened.

### IV.5.1 Circuit Analysis

The blue-shift EAM circuit analysis can be carried out with the same small signal equivalent circuit model as presented by Figure IV.2. The only different assumption here is that the optical transfer curve now has a positive slope at bias point and it is normalized to sufficiently large bias voltage where absorption is minimum. As a result, Equation IV.2 has a sign change.

\[
P_{\text{OUT}} = P_{\text{INT}} t_p \left( t_B + \frac{\pi}{2V_{\pi e}} v_m \right) \quad (IV.10)
\]

Following the same derivation, we can obtain a negative resistance as

\[
R_P = -\frac{2V_{\pi e}}{P_L t_1 t_p \eta_M \eta_l} \quad (IV.11)
\]

In actual case, this is only the equivalent resistance caused by the photocurrent. It should be paralleled by an intrinsic junction resistance when no photocurrent is generated in the circuit. Our analysis does not include this junction resistance.
Figure IV.10: Schematic absorption spectra for red-shift (a) and blue-shift (b) materials at various bias voltages. Arrows denote direction of peak absorption wavelength shift with bias. The plots are based on simple calculation with assumptions of 5 meV transition linewidth and 7.5 meV exciton linewidth.
In conventional red-shift EAM, ac photocurrent flowing through the modulator junction resistance $R_P$ is limited by the ac current provided by the source $i_S$, which also limits the maximum achievable link gain. A negative $R_P$, on the contrary, extends the photocurrent to be beyond the limit set by the source current $i_S$. The relationship between the ac photocurrent $i_P$ and source current $i_S$ is given by

$$i_P = i_S \frac{R_L R_S}{R_L + R_S} \frac{1}{R_P + R_M + \frac{R_L R_S}{R_L + R_S}}$$  \hspace{1cm} (IV.12)

It is very clear to see maximum $i_P$ is available only when $R_P$ approaches zero for a positive $R_P$. With a negative $R_P$, however, the bound limitation is lifted. Especially when the condition is reached for the denominator to be zero, $i_P$ amplitude can be arbitrarily large. This condition can be expressed as

$$P_L = \frac{2V_{\pi e}}{t_1 t_P \eta_M \pi \left( R_M + \frac{R_L R_S}{R_L + R_S} \right)}$$  \hspace{1cm} (IV.13)

Similar relationship between modulation voltage $v_M$ and the source current $i_S$ can be derived, which results in an arbitrarily large $v_M$ when equation is satisfied. This is important because the modulator is a voltage controlled device. With larger modulation voltage, higher link gain can be obtained.  \(^1\)

The link gain is given by

$$g = \left[ \frac{4R_L^2 R_S R_D t_0^2 \eta_D^2}{(R_L + R_S)^2 \eta_M^2} \right] \left[ \frac{2V_{\pi e}}{R_L t_1 t_P \eta_M \pi} - \left( R_M + \frac{R_L R_S}{R_L + R_S} \right) \right]^2$$  \hspace{1cm} (IV.14)

\(^1\)In the case of high frequency where the inclusion of EAM junction capacitance is necessary, the denominator of Equation IV.12 can not reach absolute zero but it has a minimum value when the real part becomes zero. Thus $i_P$ and $v_M$ amplitudes will have a maximum at certain optical power.
The major difference between this link gain expression and a conventional red-shift EAM link is the denominator of the second term. In the limit of low power $P_L$, the term in the parentheses can be ignored and the link gain behaves as an unsaturated external modulation link. In the limit of high optical power $P_L$, the term in the parentheses dominates and the link gain approaches the limit of a conventional EAM link. When an appropriate optical level is reached so that Equation IV.13 is met, the link gain of negative differential resistance EAM demonstrates a peak.\textsuperscript{2}

This optical power level is determined by $V_{\pi e}$, $t_I$, $t_P$, $\eta_M$, and circuit parameters as $R_M$, $R_S$, and $R_L$. With a small $V_{\pi e}$, a large input coupling coefficient $t_I$, low residual propagation loss $t_P$, and a large modulator responsivity $\eta_M$, the optimal optical power is expected to be within reasonable range.

As a result of the link gain enhancement, the link noise figure is also benefited. The link noise figure is given by

\begin{equation}
 f = 1 + \frac{f_R}{g} + \frac{R_S R_M}{R_L} + \frac{2eP_L t_I t_B \eta_D R_D}{gkT_0} + \frac{\eta_M (1 - t_B) R_S}{2kT_0} \left[ 1 + \frac{R_M R_S (R_L + R_M)}{R_S R_L} \right]^2
\end{equation}

(IV.15)

Which includes thermal noise from $R_M$ and $R_L$. The relative intensity noise (RIN) is still excluded in the expression.

The first two terms are the input and receiver noise. The third and fourth terms are from the thermal noise of $R_L$ and $R_M$. The last two terms are shot noise terms from photodetector and modulator. At high optical power and high link gain, modulator shot noise due to the dc photocurrent becomes dominant.

\textsuperscript{2}When EAM junction capacitance is included, the link gain will still have a peak but will not reach infinity.
IV.5.2 Device Performance

Based on the circuit analysis, the predicted performance of the blue-shift EAM can be plotted. Figure IV.11 shows the link gain peaking behavior of negative resistance EAM along with the link gain of conventional positive resistance EAM. The parameters used in the plot for both kinds of EAMs are the same. For $V_{\pi}=0.5V$, the link gain of negative resistance shows at about 17 mW laser power with more than 30 dB, which is a dramatic improvement compared with conventional EAM. If the EAM propagation loss increases, the optimal optical power will increase proportionally.

Figure IV.12 shows the link noise figure at varied laser power levels. Again, there is a minimum noise figure for negative resistance EAM similar to the conventional EAM. However, this minimum is lower than that of conventional EAM if all the parameters are the same except for the sign of the junction resistance. For smaller $V_{\pi e}$, the optical power where the link gain peak and the minimum link noise figure occur is reduced. The minimum link noise figure is also lowered.

IV.5.3 Material Design Consideration

The implementation of negative differential resistance EAM requires semiconductor material with low $V_{\pi e}$, low residual propagation loss, and most importantly the behavior of reduced absorption with increased bias voltage. A quantum well semiconductor material with a pre-biased shape and based on type-II band lineup was demonstrated to have blue-shift QCSE effect[5]. It is shown schemati-
Figure IV.11: Link gain as a function of laser power for various values of $V_{\pi e}$. Solid lines and dashed lines are for negative resistance EAM and conventional positive resistance EAM, respectively. Other parameters used in the plot are $R_S = R_L = R_D = 50\Omega$, $R_M = 5\Omega$, $\eta_M = \eta_D = 1$ A/W, $t_I = t_O = -2$ dB, $t_D = 1$, and $t_B = -3$ dB.
Figure IV.12: Link noise figure as a function of laser power for various values of $V_\pi$. Solid lines and dashed lines are for negative resistance EAM and positive resistance EAM, respectively. Other parameters are the same as in Figure IV.11.
cally in Figure IV.13. Without any bias, it has a step-like profile which mimics a pre-biased condition. The well floor becomes flatter when a bias is applied.

![Figure IV.13: Schematic of pre-biased quantum well structure for blue-shift QCSE application.](image)

By taking the structure from Ref. [5], a simulation was done for EAM application. The structure used is InP/InAs_{0.4}P_{0.6}/In_{0.53}Ga_{0.47}As/InP type II asymmetric quantum well. The thicknesses of InAs_{0.4}P_{0.6} and In_{0.53}Ga_{0.47}As layers are 5.5 nm and 3.5 nm respectively. The band-gap offset ratios (conduction band offset to valence band offset) are 0.6/0.4 for InAsP and 0.4/0.6 for InGaAs. With 600 µm long waveguide and 10% confinement factor, we have calculated the residual waveguide propagation loss to be about 3 dB and $V_{\pi e}$ to be 2.24 V. Assuming 3dB coupling coefficient at both input and output facets, a peak of link gain occurs at about 100 mW input laser power. The minimum noise figure is about 16 dB at 70 mW laser power. For actual blue-shift materials, the absorption coefficient at
bias voltage and the $V_{pe}$ may be much larger than the simulation results. In that case, optical power will need to be proportionally raised to achieve the peak link gain and minimum noise figure.

It should be noted that the simulation done on this blue-shift QCSE material is quite crude. We calculated electron and hole wave functions in the quantum well under different bias voltages and did overlap integrals. We also assumed Lorentzian electron to hole energy level transition with full width at half maximum (FWHM) linewidth of 5 meV and Gaussian exciton FWHM linewidth of 7.5 meV. The absorption coefficient we obtained by this simple calculation does not reveal all the details of material property so it is quite inaccuracy. To better calculate the blue-shift QCSE effect, a more complete model is needed to include strain, exciton linewidth broadening, etc.

### IV.6 Acknowledgement

References


Chapter V

Linearity Performance of High Power EAM

V.1 Fiber-Optic Link Linearity

The linearity along with related third-order intercept point (IP3) and SFDR are important performance parameters of an analog fiber-optic link. They greatly affect the quality of the signal transmitted. In CATV systems, for example, the image quality will degrade if the channel being watched also contains intermodulation distortions from other channels. The link linearity is determined by both the optical transmitter and receiver. At the transmitter end, the nonlinear shape of modulator optical transfer curve introduces nonlinearity to the link. At the receiver end, the nonlinear response of the generated photodetector current to the input optical power also alters the quality of the signal received. However, the photodetection process, which originates from an electron-hole pair generation from
an incoming photon, is a much more linear process. Only the deviation caused by either current crowding effect or thermal effect will change its response. On the contrary, optical transfer curves of different types of modulators are intrinsically nonlinear. Therefore, the linearity of optical modulator is of primary importance among the components of a fiber-optic link. Many studies of its properties and improvement methods have been carried out.

The dc optical transfer curves of optical modulators can be used to estimate their linearities and corresponding parameters. When a small single sinusoidal signal is applied to the device at a bias point, both the fundamental signal and its harmonics will be present at the link output due to the high order nonlinearities of the optical transfer curve. LiNbO$_3$ MZMs and other EOMs usually have well defined and stable optical transfer curves for analytical analysis. As an example, MZMs have sinusoidal shape optical transfer curve. Its Taylor expansion coefficients are Bessel functions. This characteristics eases the modeling and demonstration of combination of multiple MZMs for canceling high order harmonics and thus improving the link linearity. In contrast, the optical transfer curve of an EAM is dependent on both material properties and the device structure. It varies from device to device and can not be easily described analytically. Therefore, it becomes very complicated to employ similar schemes to linearize an EAM link.

As described in Chapter IV, EAM experiences link gain saturation at high input optical power level. The mechanism behind this saturation is due to the negative feedback effect generated by the EAM photocurrent, which causes effective
voltage reduction across the modulation layer. The resultant link gain will deviate from the quadratic dependence on the input optical power and finally approach a gain limit. This same mechanism also affects the EAM optical transfer curve and changes its linearity performance. However, this is not the sole mechanism responsible for the optical transfer curve change at high power. Other effects such as material property change and thermal effect also play roles in it due to the complicated nature of EAM. To get a full understanding of the undergoing picture, all the effects need to be studied to determine the EAM linearity. If we assume other effects only have minor impacts on modulator linearity change from low power to high power, we can analyze the impact of photocurrent feedback effect on the device linearity analytically with the help of the equivalent circuit model.

V.1.1 Modulator Optical Transfer Curve

The dc transfer curve of an optical modulator is usually a continuous and well behaved curve. It can be expanded in Taylor series at its bias point as:

\[
T(V) = T(V_B) + T'(V)|V_B(V - V_B) + \sum_{n=2}^{\infty} \frac{T^{(n)}(V)|V_B(V - V_B)^n}{n!}
\]  

(V.1)

The first and second terms correspond to the output dc and fundamental signal respectively. All other terms are responsible for high order distortions at the output of the link. The link IP3 and SFDR can be calculated based on Equation (V.1). For detailed derivation, please refer to Appendix B of this dissertation.
V.1.2 Spurious-free Dynamic Range

As briefly described in Chapter I, spurious-free dynamic range (SFDR) is an important figure of merit of an analog fiber-optic link. Two-tone test is one commonly used means to measure SFDR. In two-tone tests, two RF tones $f_1$ and $f_2$ at equal amplitude with a small frequency offset modulate the optical carrier. Different orders of distortion signals are measured at the link output. The third-order intermodulations (IM3) are of the most critical distortions since their frequencies fall around the vicinity of the desired fundamental signal. In two-tone tests, SFDR is a parameter describing the link output SNR as the IM3 is equal to the noise level. This is the maximum output SNR a link can obtain without any significant IM3 spurious signal contamination. The link SFDR depends not only on link linearity, but also link gain and link output noise floor. The higher link gain, lower output noise, and more linear transfer curve yield a larger SFDR.

V.2 Intrinsic and Extrinsic Transfer Curves of EAM

As introduced in Chapter I, an optical transfer curve of a modulator is its transmission versus applied voltage curve. However, applied voltage of an optical modulator is an ambiguous term when the voltage from the outer source and the effective voltage across the device intrinsic layer are different. Concepts of intrinsic and extrinsic transfer curves are used to clarify the difference and to better serve
the analysis of the device linearity.

We define intrinsic optical transfer curve as a function of junction voltage $T(V_M)$. It depends on device properties and does not include any circuit effects. We also define extrinsic optical transfer curve as a function of source voltage $T_e(V_{IN}) = T_e(V_S/2)$. The scaling factor 1/2 comes from the conventional definition for analog fiber-optic link gain, where the input RF power is taken with a modulator load matched to that of the RF source [1]. The extrinsic transfer curve includes circuit effect that rises from photocurrent, resistivity and capacitance changes and alters the effective modulation voltage across device junction to deviate from the source voltage. For an optical modulator without any photocurrent generation under operation such as LiNbO$_3$ MZM, these two definitions of transfer curve are almost identical if the device is impedance matched to the source. For an EAM, however, photocurrent generation in the p-i-n junction lowers its equivalent resistance and there is a substantial difference between two kinds of transfer curves. At low input optical power level, the EAM photocurrent is small and the junction resistance lowering is still negligible, which makes intrinsic transfer curve still valid for device analysis. As optical power increases, device circuit effect must be included for an accurate calculation.

Figure V.1 shows the small circuit model of a lumped EAM at low frequency as used in an analog fiber-optic link, in which the optical signal transmits from

---

1Although the original definition of link gain was for matched circuit, it is also used for un-matched cases in practice. The RF synthesizers usually have an isolator built in so that the reflection from device under test does not affect its internal operation. So the definition is still valid for un-matched case in measurement. Here the linearity is defined the same way for un-matched circuit.
the optical modulator to the photodetector. To simplify the following equation
derivation, the modulator termination resistance is not included here. At very low
frequency analysis which is compatible with linearity estimation from dc optical
transfer curve, the device junction capacitance \( C_J \) is also omitted in the subsequent
analysis.

![Equivalent circuit model of EAM](image)

**Figure V.1:** An equivalent circuit model of EAM used in an analog fiber-optic link

From the circuit model in Figure V.1, it is not difficult to obtain:

\[
V_S = V_M + I_P(R_S + R_M) \quad (V.2)
\]

and

\[
I_P = P_L t_I t_P \eta_M [1 - T(V_M)] \quad (V.3)
\]

Where \( P_L \) is the input laser power, \( t_I \) is the input coupling coefficient, \( t_P \) is the de-
vice propagation loss, and \( \eta_M \) is the EAM responsivity. Therefore the incremental
change of \( V_{IN} = V_S/2 \) and \( V_M \) can be related:

\[
\frac{dV_{IN}}{dV_M} = \frac{1}{2} \left( 1 - P_L t_I t_P \eta_M \frac{dT}{dV_M} \right) \quad (V.4)
\]
Since the difference between intrinsic and extrinsic transfer curves is the argument of the function, $V_M$ for intrinsic transfer curve and $V_{IN}$ for extrinsic transfer curve, which means $T_e(V_{IN}) = T(V_M)$. Therefore different orders of derivatives of both intrinsic and extrinsic optical transfer curves with respect to their arguments can be evaluated and related according to Equation V.4:

\[
\frac{dT_e}{dV_{IN}} = \frac{2 \frac{dT}{dV_M}}{1 - P_L t_P \eta_M (R_S + R_M) \frac{dT}{dV_M}} \quad (V.5)
\]

\[
\frac{d^2T_e}{dV_{IN}^2} = \frac{4 \frac{d^2T}{dV_M^2}}{(1 - P_L t_P \eta_M (R_S + R_M) \frac{dT}{dV_M})^3} \quad (V.6)
\]

\[
\frac{d^3T_e}{dV_{IN}^3} = \frac{8 \frac{d^3T}{dV_M^3} \left(1 - P_L t_P \eta_M (R_S + R_M) \frac{dT}{dV_M}\right) + 24 P_L t_P \eta_M (R_S + R_M) \left(\frac{d^2T}{dV_M^2}\right)^2}{(1 - P_L t_P \eta_M (R_S + R_M) \frac{dT}{dV_M})^5} \quad (V.7)
\]

Equation V.5 is the relationship between first order derivatives of the transfer curves. It accounts for the link gain saturation as detailed in Chapter IV. The term $dT/dV_M$ is considered negative due to the fact that larger voltage causes less optical transmission in conventional quantum well designs. The denominator on the right side of Equation V.5 becomes much larger than unity when the input optical power is high enough, which reduces the link gain. It can be lumped as an EAM saturation factor $k$.

\[
k = 1 - P_L t_P \eta_M (R_S + R_M) \frac{dT}{dV_M} = 1 + \frac{P_L t_P \eta_M (R_S + R_M) \pi}{2 \pi_e} \quad (V.8)
\]

where $V_{\pi e}$ is the EAM equivalent half wave voltage defined as $(\pi/2)(dT/dV_M)^{-1}$. Equations V.6 and V.7 represent higher order derivatives of optical transfer curves. The derivatives of extrinsic and intrinsic optical transfer curves are related by
a factor of \( k^3 \) for the second order, and \( k^4 \) for the third order when EAM is biased at its largest slope efficiency voltage point where the second order derivative nulls out. The second order null point is also the bias point for multi-octave operation (see Appendix B). It is clear that the derivatives of extrinsic optical transfer curves become much smaller than that of intrinsic optical transfer curves when the saturation factor \( k \gg 1 \). Therefore the extrinsic optical curve becomes more linear than the intrinsic one.

As will be discussed in the next section, EAM at high power is actually a negative feedback system. The physical meaning of \( k \) is actually the ratio of the system response without feedback to the system response with feedback. The part of \( k - 1 \) is related to the system feedback coefficient. As negative feedback is strong, the link gain suppression becomes large. However, the system becomes more linear.

### V.3 Linearity Improvement by Feedback Effect

As described in Chapter IV, an EAM experiences photocurrent effect at high input optical power levels. When the input optical power is increased, both dc and ac photocurrent generation increase. This increases the portion of the voltage drop on the source resistance \( R_S \) and the EAM serial resistance \( R_M \) relative to the total source voltage, leaving a smaller portion on the modulator layer. From a feedback point of view, what happens at the EAM circuit parallels a negative feedback sys-
tem. The incoming voltage $V_S$ modulates the active layer and leads to the intensity modulation of the optical carrier, expressed as $P_L t_t P p t_O [T(V_B) - T(V_B + v_M)]$, where $t_O$, $V_B$, and $v_M$ are EAM output coupling coefficient, the dc bias voltage, and the effective ac voltage across the modulation layer, respectively. At the same time the modulated light generates ac photocurrent which effectively reduces the voltage across the junction. At steady state, it becomes a negative feedback system, the output of which is coupled into the photodetector and generates output voltage $v_L$ across load resistance $R_D$. This description can be schematically shown in Figure V.2, where $\eta_M$ and $\eta_D$ denote the responsivity of EAM and photodetector, respectively.

![Optical Fiber](image)

Figure V.2: Negative feedback system formed by the effect of photo-generated current on EAM circuit in an analog fiber-optic link configuration

With a low input optical power to the EAM, the photocurrent feedback can be ignored and the linearity of the electro-to-optical conversion is mainly determined by that of the intrinsic optical transfer function $T(V)$. When the optical power increases, the effective voltage across the EAM junction is no longer $v_S$, but $v_M$ which is modified by the photocurrent feedback. The extrinsic optical transfer
V.3.1 Background from Electronic Amplifier Design

In electronic amplifier design, negative feedback system has been used for a long time to trade gain for linearity [2]. This can be shown with a simple negative feedback system with open loop response of $g$ and feedback coefficient $f$. The closed loop response of the system is:

$$A = \frac{g}{1 + gf} \quad (V.9)$$

When the $gf$ product is much larger than unity, the closed loop response can be approximated by

$$A \approx \frac{1}{f} \quad (V.10)$$

The feedback coefficient $f$ in electronic amplifier design can be implemented with perfectly linear elements, such as resistive voltage dividers, so that the overall closed-loop behavior of the system is linear even though the amplifier open loop response $g$ is not. As a trade-off, the closed loop gain is smaller than the open loop gain. This effect is frequency limited only when the loop does not provide a negative feedback at certain frequencies.

Taking the preceding description as an analogue, we can take a look at the EAM under saturation from a new prospective. We can define a voltage gain function for EAM link $v_{OUT} = g(v_{IN})$ at no feedback, which is equivalent to the open loop response of electronic amplifier. The relationship between $v_{IN}$ and $v_{OUT}$...
can be constructed based on Figure V.2.

\[ g(v_{IN} - f v_{OUT}) = v_{OUT} \]  \hspace{1cm} (V.11)

where \( g \) is a function that includes the nonlinear harmonics caused by the optical transfer curve \( T(V) \) and \( f \) is the negative feedback coefficient as described by:

\[ f = \frac{(R_S + R_M)\eta_M}{2\eta_D R_D t_O} \]  \hspace{1cm} (V.12)

Here \( v_{IN} = v_S/2 \) and the output voltage \( v_{OUT} \) is equivalent to \( v_L \), the ac voltage across the load resistance of the photodetector. This situation is equivalent to the electronic amplifier case. As the open loop response is large enough, the system closed loop response is mainly determined by \( f \). Since \( f \) does not consist of any elements of the nonlinear optical transfer curve, the linearity of EAM under negative feedback is expected to be more linear. However, the EAM and photodetector responsivities involved in \( f \) can still affect the system linearity.

By taking a close look and comparison, the \( 1 + gf \) in electronic amplifier design is equivalent to the saturation factor \( k \) in EAM. Hence, when the EAM is under high saturation, the above discussion will be valid.

Compared with electronic amplifier design, the difference here is that negative feedback is intentionally added to the electronic amplifier to improve linearity while it naturally exists in EAM. Although link gain suffers from saturation, the link linearity actually benefits from it.
V.3.2 EAM Linearity under High Power

Optical modulator linearity can be calculated from its optical transfer curve as detailed in Appendix B. The input second- and third-order intercept points (IIP2 and IIP3) can be derived to be related to derivatives of optical transfer curve as:

\[
IIP2 \propto \left( \frac{dT}{dV} \right)^2 \frac{d}{dV^2}
\]

\[
IIP3 \propto \frac{dT}{dV} \frac{d}{dV^3}
\]

With the help of Equations V.5–V.7, IIP2 and IIP3 of highly saturated EAM link can be improved by a factor of \(k^4\) and \(k^3\) compared with the no-feedback situation. The output second- and third-order intercept points (OIP2 and OIP3) increases by the same factor when the gain saturates. On the other hand, the link output noise only increases linearly with optical power even when EAM shot noise dominates in the saturation case, which is approximately proportional to \(k\). Hence, in the absence of laser intensity noise (RIN), the link SFDR improves by \(k^{4/3}\).

The discussion carried out so far compares linearity performance of the EAM link with and without photocurrent feedback effect, both under high optical power when link gain saturates. Although the observed high power operation of EAM has already the photocurrent feedback built-in, a view from the no-feedback perspective gives us more insight in understanding the involved mechanism. The evaluation of the EAM link linearity change from low optical power state to high optical power state would require knowledge not only of the difference between
intrinsic and extrinsic optical transfer curves at high optical power but also of the intrinsic optical transfer curve change as the optical power increases, which involves a variety of effects that can modify material absorption properties. If we assume the change of the intrinsic transfer curve imposes only minor effect on the link linearity compared with the feedback effect and also exclude the nonlinearities of the EAM and photodetector responsivities, we can calculate the link performance based on the IIP3 value of the EAM at low power. Figure V.3 shows the dependence of the link IIP3 and output noise floor on the input optical power in the absence of the laser RIN. A low power IIP3 of 20 dBm is assumed[3]. The calculation clearly shows faster increase of IIP3 than link output noise at high optical power levels. Figure V.4 shows the dependence of RF link gain, multi-octave OIP3 and SFDR on input optical power. Before the EAM experiences gain saturation, the OIP3 increase is mainly due to the gain increase. At high power, the EAM link gain does not increase substantially but IIP3 starts to rise up rapidly. The OIP3 thus keeps increasing. As mentioned before, the link output noise does not increase faster than OIP3, therefore the link SFDR keeps improving as the input optical power is raised. At optical power of more than 700 mW, SFDR can reach 135 dB/Hz^{2/3}.

The derivation of Equations V.5–V.7 assumes \( \eta_M \) and \( \eta_D \) to be constants with respect to the input voltage, which is the ideal case. In practice, a complete calculation should include nonlinearities of \( \eta_M \) and \( \eta_D \) as well. More studies are needed for a better understanding of the dependencies of \( \eta_M \) and \( \eta_D \) upon input voltage at high optical power before their effects on link linearity can be evaluated.
Figure V.3: Calculated link output noise floor and IIP3 as a function of input optical power. Laser RIN noise is not included. Low power EAM IIP3 of 20 dBm is assumed. Additional optical loss of 3 dB is caused by dc bias. Other parameters used in the calculation are: $t_I = t_O = -3$ dB, $t_P = -1$ dB, $R_S = R_D = 50 \ \Omega$, $R_M = 5 \ \Omega$, $\eta_M = \eta_D = 1$ A/W, $V_{\pi e} = 1.5$ V.
Figure V.4: Calculated RF link gain, multi-octave link OIP3 and SFDR dependence on input optical power. Conditions and parameter values in the calculation are the same as Figure V.3.
V.4 Comparison between EOM and EAM

Compared with EOM links, SFDR improvement of EAM links at increased input optical power is due to a different mechanism. In EOM links, IIP3 stays constant as optical power is raised. The SFDR improvement comes from faster increase of link gain over link noise floor. In the absence of RIN, SFDR of EOM links improves only by a power of two-thirds of the increased optical power. This is less than a power of four-thirds as for the case of EAM links at high power with contributions only from circuit negative feedback effect. With RIN involved, SFDR of EOM links will finally reach a limit as optical power is raised. In contrast, when feedback saturation is the dominant nonlinear mechanism, SFDR of EAM links will keep increasing.

SFDR of single MZM link is mainly limited by its nonlinear transfer curve. Experimental result showed a multi-octave SFDR of only 119.5 dB/Hz^{2/3} at 240 mW laser power [4]. Linearized EOM links can improve sub-octave SFDR to be beyond 130 dB/Hz^{4/5}. However, the multi-octave SFDR remained low. A dual-wavelength EOM scheme with optical powers of 200 mW at 1320 nm and 30 mW at 1550 nm yielded an SFDR of 121 dB/Hz^{2/3} [5]. These numbers are lower than what we estimated for a high power EAM link due to negative feedback effect.
V.5 Acknowledgement

References


Chapter VI

Conclusions and Future Work

This dissertation studies electroabsorption modulator with peripheral coupled waveguide structure. The design allows device to operate at higher optical power than earlier designs. Its application in analog fiber-optic links has yielded better link RF gain, an important link parameter.

The major contributions of the dissertation are as follows. 1) PCW EAMs were designed and fabricated with confinement factor as low as several percent. It raised the device power handling capability to be more than 500 mW, which is comparable to commercial LiNbO$_3$ MZM [1, 2, 3]; 2) The mechanism behind link gain saturation of EAM link at high power was investigated and analyzed using an equivalent circuit model. It was proved to be a photocurrent circuit effect which effectively reduces modulation voltage across device junction. The measurement data and calculation were compared to have a close match [4]; 3) The EAM linearity due to circuit properties was analyzed by comparing it with an
electronic negative feedback system. The linearity was calculated to improve with large photocurrent at saturation stage [5].

Besides, the following achievements are also beneficial for understanding the EAM device operation and further investigations. 1) By optimizing the optical mode design, the coupling between fiber and EAM has been greatly improved. A coupling coefficient of about 2 dB was achieved; 2) The waveguide propagation loss has been improved as well by pulling the optical mode to be submerged in the lower waveguiding layer to avoid scattering loss. The minimum propagation loss was measured to be 0.8 dB/mm; 3) Febry-Perot method was implemented to measure EAM device insertion losses. It improved the measurement accuracy for low loss devices over cut-back method. The propagation loss measured by this method does not depend on coupling condition. The propagation loss can be separated from coupling loss based on the experimental data; 4) The EAM noise figure at high power was analyzed based on the same circuit model used to explain gain saturation effect. It involves shot noise generated by EAM photocurrent. The link noise figure was shown to have a minimum at certain optical power level [4]; 5) The taper waveguide structure with lumped electrode only was designed and implemented with PCW to further reduce device saturation at high power [6]; 6) Preliminary study on temperature dependence of EAM operation was conducted to show high temperature degrades device link gain performance; 7) Index of refraction change induced by high photocurrent in EAM was observed and identified.
VI.1 Conclusions

Electroabsorption modulator is favorable in the application of analog fiber-optic links due to its small size, good modulation efficiency, and potential integration with other semiconductor optoelectronic devices. However, its larger insertion loss and lower power handling capability compared with LiNbO$_3$ MZM makes it inferior to achieve a high link gain in an analog link. PCW structure with a low confinement factor and a submerged optical mode improves propagation loss and coupling loss as well as device saturation at high optical power. The taper structure added to the PCW evens out exponential decay profile of light absorption, photocurrent generation, and heat dissipation along the waveguide. It pushes the device power handling capability further.

The PCW EAM devices were experimentally measured and their performances were analyzed with varied device parameters. By using an equivalent circuit model for the EAM, the device link gain saturation at high power was identified to be attributed to the photocurrent generated in the EAM junction, which lowers the effective modulation voltage across the junction due to its negative feedback effect. The model explained the phenomenon very well and fitted the experimental data closely. The device’s other figures of merit including noise figure and linearity were also explored by employing the circuit model and concept adapted from the electronic amplifier design. Their characteristics serve as a guidance for the optimization of the device performance.
VI.2 Suggestions for Future Work

Due to the project funding limit and time constraint, this dissertation only investigates the major properties of PCW EAM device and some related subjects. There are still some interesting aspects left for future work, which will either further improve the device performance or better understand how the device operates. The following is a list of my suggestions for future work.

1. The major focus of this dissertation is to use PCW structure to boost EAM power handling capability, thus improving link RF gain. As the confinement factor is decreased to reduce generated photocurrent density, the waveguide length needs to be increased to achieve the same modulation efficiency. However, it increases the device capacitance and degrades device 3-dB bandwidth. To get the device to operate at high frequencies, traveling wave structure is a necessity to adopt. TW-EAM has been studied extensively in both LiNbO₃ electrooptic and semiconductor electroabsorption modulators. However, the optimization of the structure in PCW, which has a length of semiconductor waveguide on the order of millimeter, has not been well touched. It is definitely worthwhile to explore the TW structure to extend the device bandwidth and maintain the benefits obtained by PCW Lumped EAM.

2. Using the blueshift QCSE material to overcome the link gain limit imposed by the EAM photocurrent feedback effect is briefly investigated in this dissertation. There are still many aspects to investigate further on this subject.
Among them are material modeling accuracy and sensibility, feasibility and optimization of material design, insertion loss consideration at the bias point, stability analysis of the positive feedback system, and extending the negative resistance concept to TW structure.

3. The shot noise contribution from the EAM photocurrent was calculated in the dissertation. However, an experimental verification is still necessary to validate the calculation and at the same time to support the circuit model adopted in the derivation.

4. The current equivalent circuit model for EAM is well suited for explanation of gain saturation and gain limit. However it is based on lumped element approximation. An equivalent circuit model for TW EAM at high power under the dominance of photocurrent will serve as a guidance for the design of TW EAM as well as provide insight of internal operational mechanisms.

5. The linearity analysis for high power EAM was carried out in this dissertation. There are several assumptions made for the final result. The experimental exploration and validation will clear the way for implementing the theory to applications. A direct measurement of the high power EAM linearity will be a supplement for the theory.
References


Appendix A

Index of Refraction of InGaAsP Material

Index of refraction of compound semiconductor material is a key important parameter in the design of photonic devices such as semiconductor laser, optical modulator, and photodetector. It affects all ranges of optical phenomena such as light confinement and optical mode formation in optical waveguides, light reflection at an interface, and light absorption. As one of the most important and widely used semiconductor optoelectronic material system, InGaAsP/InP, the modeling and calculation of its index of refraction especially in the telecommunication infrared wavelength range becomes a necessity for material and device design. Moreover, its accuracy sometimes is very critical and can greatly affect the performance of the device fabricated.

There are several methods in the literature to calculate the values of index of refraction of InGaAsP/InP material system, either semi-empirical or based on theoretical models. However, almost all of them are only valid at wavelengths below
the material band-gap, which means that they are in the transparent region. As the wavelength starts to approach the material band-gap, there are more interactions between photons and the material and the index of refraction will not only have a real component but also an imaginary component related to material absorption. Besides, the values of index of refraction calculated from different methods can be quite different even in the transparent region. The designers need to be cautious to select values based on their experience to reflect the real situation of the device under design.

Since some methods use material band-gap energy to calculate index of refraction, a conversion from element composition in InGaAsP to its band-gap energy needs first to be carried out. InGaAsP is usually described as $\text{In}_{1-x}\text{Ga}_x\text{As}_y\text{P}_{1-y}$ with two independent compositional parameters. If InGaAsP is lattice matched to its substrate InP, one free parameter is eliminated. Based on Vegard’s law, the lattice constant of $\text{In}_{1-x}\text{Ga}_x\text{As}_y\text{P}_{1-y}$ can be expressed as follows. [1]

$$a(x, y) = 0.1894y - 0.4184x + 0.0130xy + 5.8696 (\text{Å}) \quad (A.1)$$

By equating the above equation to InP lattice constant 5.8696 Å, a relationship between $x$ and $y$ for lattice matched InGaAsP can be obtained.

$$x = \frac{0.1894y}{0.4184 - 0.0130y} \quad (A.2)$$

Besides, other similar equations have also used in literature to relate $x$ and $y$.

\footnote{Sometimes in literature, $x$ and $y$ are interchanged. In other cases, $x$ and $1-x$ or $y$ and $1-y$ are interchanged. A caution needs to be taken to use all formulae based on compositional parameters $x$ and $y$.}
y. The following are some of them.

\begin{align}
    x &= 0.46y \\
    x &= 0.47y \\
    y &= 2.197x \\
    x &= \frac{0.1896y}{0.4176 - 0.0125y}
\end{align}

An empirical equation describing the band-gap energy of In\(_{1-x}\)Ga\(_x\)As\(_y\)P\(_{1-y}\) that lattice matched to InP was provided by Nahory \textit{et al.} [1] as

\[ E_g(y) = 1.35 - 0.72y + 0.12y^2 \text{ eV} \]  \hfill (A.7)

Here \(x\) is eliminated by the relationship between \(x\) and \(y\). This equation matched experimental result better than that was derived from ternary compounds band-gap energy parameters, which is given by:

\[ E_g(x, y) = 1.35 + 0.668x - 1.17y + 0.758x^2 + 0.18y^2 - 0.069xy - 0.322x^2y + 0.03xy^2 \text{ eV} \]  \hfill (A.8)

Among several methods being used to calculate the index of refraction for compound InGaAsP material are modified single-effective-oscillator method [2, 3], Adachi’s method [4], Burkhard’s method [5], Jensen and Torabi’s method [6], and Sellmeier formula [7]. Most of the methods only deal with intrinsic semiconductor materials. However, the index of refraction also changes when the material doping and carrier injection condition change. These effects are not the main focus of this appendix.

The following sections details the respective methods.
A.1 Modified Single-Effective-Oscillator Method

Modified single-effective-oscillator (MSEO) method was presented by Afro- mowitz [2] to improve the original less accurate model by Wemple and DiDomenico [3]. The results of this method deteriorates at energies approaching the material band edge as all other method do.

According to this method, the index of refraction of compound semiconductor material can be obtained by the following equations.

\[ n^2(E) = 1 + \frac{E_d}{E_0} + \frac{E_d}{E_0^3}E^2 + \frac{\eta}{\pi}E^4 \ln \left( \frac{2E_0^2 - E_g^2 - E^2}{E_g^2 - E^2} \right) \]  
(A.9)

\[ \eta = \frac{\pi E_d}{2E_0^3(E_0^2 - E_g^2)} \]  
(A.10)

Here \( E_g \) is the material band-gap energy and \( E \) is the incident photon energy which is related to the photon wavelength by \( E = \frac{hc}{\lambda} \). The single oscillator energies \( E_0 \) and \( E_d \) for InGaAsP can be described by:

\[ E_0 = 3.391 + 0.524x - 1.891y + 1.626xy + 0.595x^2(1 - y) \]  
(A.11)

\[ E_d = 28.91 + 7.54x + (-12.71 + 12.36x)y \]  
(A.12)

All the energy parameters are in the unit of eV. Following the above equations, the index of refraction of InGaAsP of a certain composition for a certain photon energy (or wavelength) can be calculated.
A.2 Adachi’s Method

Adachi [4] derived the index of refraction equation for compound semiconductors by considering free electron-hole pair transition (band-to-band transition) and the Wannier-exciton transition and then applying Kramers-Kröning relations to the material dielectric constant.

According to Adachi’s model, the real part of the material dielectric constant $\epsilon(\omega)$ in the zinc-blende material below the band edge can be written as:

$$\epsilon_1(\omega) = A \left[ f(\chi) + \frac{1}{2} \left( \frac{E_0}{E_0 + \Delta_0} \right)^{3/2} f(\chi_{so}) \right] + B \quad (A.13)$$

with

$$f(\chi) = \chi^{-2} \left[ 2 - (1 + \chi)^{1/2} - (1 - \chi)^{1/2} \right] \quad (A.14)$$

$$\chi = \frac{\hbar \omega}{E_0} \quad (A.15)$$

and

$$\chi_{so} = \frac{\hbar \omega}{E_0 + \Delta_0} \quad (A.16)$$

where $E_0$ and $\Delta_0$ are material band-gap energy and spin-orbit splitting energy, respectively. The value of $E_0 + \Delta_0$ gap energy in the unit of eV can be taken from Ref. [8], which is given by:

$$E_0 + \Delta_0 = 1.038 + 0.299(1 - y) + 0.129(1 - y)^2 = 1.466 - 0.557y + 0.129y^2 \quad (A.17)$$

$A$ and $B$ are parameters which were obtained by fitting the theoretical curve with experimental results.

$$A(y) = 8.40 - 3.40y \quad (A.18)$$
\[ B(y) = 6.60 + 3.40y \]  \hspace{1cm} (A.19)

In the transparent region below material band gap, \( \epsilon_2(\omega) \) can be taken as zero and index of refraction can be obtained as

\[ n(\omega) \approx \epsilon_1(\omega)^{1/2} \]  \hspace{1cm} (A.20)

### A.3 Burkhard’s Method

Burkhard \textit{et al.} [5] took the average of quantity \( (\epsilon - 1)/(\epsilon + 2) \) of binary constituents of InGaAsP according to Vegard’s rule, where \( \epsilon \) is the dielectric constant of the material, to get the index of refraction of quaternary InGaAsP. It was based on Clausius-Mosotti’s relation being the result of a summation of atomic polarizabilities of densely packed atoms. The resulting index of refraction of \( \text{In}_{1-x}\text{Ga}_x\text{As}_y\text{P}_{1-y} \) can be calculated in terms of \( y \) and \( \Delta E = E - E_0 \) as follows.

\[
\begin{align*}
n(\Delta E, y) &= 3.425 + 0.940\Delta E + 0.952(\Delta E)^2 + (0.255 - 0.257\Delta E)y \\
&\quad - (0.103 - 0.0952\Delta E)y^2
\end{align*}
\]  \hspace{1cm} (A.21)

where \( E_0 \) and \( E \) are material band-gap energy and photon energy in the unit of eV, respectively. Since the binary data used were taken in the wavelength range from 365 to 1100 nm, the accuracy of the formula for 1.3 and 1.55 \( \mu \)m wavelengths is questionable.
A.4 Jensen and Torabi’s Method

Jensen and Torabi [6] used a quantum mechanical calculation to obtain the dielectric constant of a compound semiconductor. The index of refraction was given by:

\[ n^2 = 1 + 2c_0 \left\{ (y_B - y_F) - z \left[ \tan^{-1} \left( \frac{y_B}{z} \right) - \tan^{-1} \left( \frac{y_F}{z} \right) \right] \right\} \tag{A.22} \]

where

\[ z = (1 - (\hbar \omega / G))^{1/2} \tag{A.23} \]

\[ y_B = m(a_0 - a) \tag{A.24} \]

\[ c_0 = \frac{3\eta \omega_v^2}{2\omega_g^2} \tag{A.25} \]

\[ \omega_g = \frac{G}{\hbar} \tag{A.26} \]

\[ \omega_v^2 = \frac{4\pi e^2 N_v^*}{m_n} \tag{A.27} \]

\[ N_v^* = N_v \left( \frac{m_r}{m_n} \right)^{3/2} \tag{A.28} \]

\[ \frac{1}{m_r} = \frac{1}{m_n} + \frac{1}{m_p} \tag{A.29} \]

\[ N_v = \frac{4}{3} \frac{2}{\pi^2 \lambda_c^3} \tag{A.30} \]

\[ \lambda_c = \frac{\hbar}{m_n \alpha_0 c} \tag{A.31} \]

\[ \alpha_0 = \left( \frac{G}{2m_n c^2} \right)^{1/2} \tag{A.32} \]

Here \( G \) is the material band-gap energy; \( \hbar \omega \) is the photon energy; \( c \) is the speed of light; \( e \) is electron charge; \( \hbar \) is the reduced Planck constant; \( a \) is the materia
lattice constant. The values of $m$ and $a_0$ are taken as:

$$m = (3.04 \pm 0.08) \text{Å}^{-1} \quad (A.33)$$

$$a_0 = (7.49 \pm 0.04) \text{Å} \quad (A.34)$$

The value of $y_F$ depends on material doping level. For intrinsic material, it can be taken as zero. For extrinsic material, $y_F$ is given by:

$$y_F = 2 \left( \frac{n_e}{N_v} \right)^{1/3}, \quad \text{degenerate n-type materials} \quad (A.35)$$

$$y_F = \lambda_r k_h, \quad \text{degenerate p-type materials} \quad (A.36)$$

$$\lambda_r = \left( \frac{m_n}{m_r} \right)^{1/2} \frac{\lambda_c}{2} \quad (A.37)$$

$$k_h = (3\pi^2 p)^{1/3} \quad (A.38)$$

where $n_e$ and $p$ are free electron and hole concentration for n-type and p-type materials, respectively.

The value of $\eta$ is $2/3$ for small spin-orbit splitting where $\Delta/G \ll 1$. The values of $m_n$ and $m_p$ can be obtained by Vegard’s rule in the unit of electron mass $m_e$.

$$m_n = 0.07xy + 0.13x(1 - y) + 0.028y(1 - x) + 0.07(1 - x)(1 - y) \quad (A.39)$$

$$m_p = 0.5xy + 0.8x(1 - y) + 0.33y(1 - x) + 0.4(1 - x)(1 - y) \quad (A.40)$$

All the above equations can be used to calculate index of refraction of InGaAsP for different composition and doping levels. However, cgs units are used in the original paper instead of SI units to formulate the equations. A conversion of unit needs to be carried out if this method is used.²

²For electric charge unit, 1 esu in cgs unit is equivalent to $3.3356 \times 10^{-10}$ coulomb.
A.5  Sellmeier Formula

Sellmeier formula has been widely used to calculate the wavelength dependent index of refraction of a transparent optical material with empirical fitting parameters. It is in the form as:

\[ n^2(\lambda) = A + \frac{B\lambda^2}{\lambda^2 - C} \]  \hspace{1cm} (A.41)

The values of \( A \), \( B \), and \( C \) for InGaAsP are interpolated from binary materials [7].

\[ A = 7.255 + 1.15y + 0.489y^2 \]  \hspace{1cm} (A.42)

\[ B = 2.316 + 0.604y - 0.493y^2 \]  \hspace{1cm} (A.43)

\[ C = 0.3922 + 0.396y + 0.158y^2 \]  \hspace{1cm} (A.44)

Here \( \lambda \) is in the unit of \( \mu m \).

A.6  Multiple Quantum Well Structures

Multiple quantum well (MQW) structures are widely used in optoelectronic device design. The index of refraction of MQW layer is usually treated as a homogeneous region with an average index of refraction for approximation [9]. For TE and TM modes, the averaging has different optimal forms.

\[ (n_{TE})^2 = \frac{\sum n_j^2 d_j}{\sum d_j} \]  \hspace{1cm} (A.45)
\[
\frac{1}{(n_{TM})^2} = \frac{\sum_j \frac{1}{n_j} d_j}{\sum_j d_j}
\]  

(A.46)

where \(d_j\) is the thickness of the \(j\)th layer in the MQW and \(n_j\) is the refractive index of the \(j\)th layer. The thickness of each layer is assumed to be much smaller than the optical wavelength.

Only when the number of quantum wells becomes large (> 20) the averaging becomes accurate. For small quantum well numbers, the calculated average value is larger than real value.
References


Appendix B

Linearity Calculation from Optical Transfer Curve

For optical modulators, optical transfer curve (optical transmission versus applied voltage) greatly influences the linearity performance of the device. Commonly a two-tone test is used to test the linearity of a fiber-optic link, which involves optical modulator as part of the transmitter. In this case, two radio frequency (RF) tones $f_1$ and $f_2$ with a small frequency offset are used to modulate the optical carrier. At the same time, different orders of distortion signals including $2f_1$, $2f_2$, $f_1 \pm f_2$, $3f_1$, $3f_2$, $2f_1 \pm f_2$, $2f_2 \pm f_1$, and so on, are detected by the link output. The linearity performance of the link can thus be assessed by comparing these distortion signals with the fundamental signals and output noise floor. Among all these distortion signals, the most important ones are third order intermodulation (IM3) $2f_1 - f_2$ and $2f_2 - f_1$ due to their proximity to the fundamental signals. Different figures of merit are used to evaluate the linearity performance of a link,
including spurious-free dynamic range (SFDR), input third-order intercept point (IIP3), output third-order intercept point (OIP3), and 1 dB compression point for input signal. The following are their definitions [1].

**SFDR (for third order distortion)** the output signal to noise ratio (SNR) as the third-order intermodulation signal starts to emerge above the noise floor.

**IIP3** the input RF signal amplitude corresponding to identical signal amplitudes of output fundamental and third-order distortion.

**OIP3** the output RF signal amplitude corresponding to identical signal amplitudes of output fundamental and third-order distortion. It is related to IIP3 by a factor of link gain.

**1 dB compression point** the input RF signal when the output fundamental signal deviates from linear increase by 1 dB.

Although all these figures of merit can be measured experimentally with RF equipment, it is sometimes necessary to do the evaluation from just the dc measured optical transfer curve. Since distortion can be evaluated from the nonlinear optical transfer curve, these figures of merit actually can be calculated from it, at least at low frequencies where the actual transfer curve of the device is still represented accurately by the dc transfer curve [2].
With a known optical transfer curve, we can fit it by polynomial functions, which is equivalent to doing Taylor expansion around a certain bias point \( V_B \).

\[
T(V) = T(V_B) + \sum_{n}^{\infty} \frac{T^{(n)}(V)|_{V_B}}{n!} (V - V_B)^n
\]  

(B.1)

In the case of two RF tone input, \( \Delta V = V - V_B \) is replaced by sinusoidal signals at two frequencies \( f_1 \) and \( f_2 \), and Eq. B.1 becomes

\[
T(V) = T(V_B + v_m) = T(V_B) + \sum_{n}^{\infty} \frac{T^{(n)}(V)|_{V_B}}{n!} v_s^n [\sin(2\pi f_1 t + \phi_1) + \sin(2\pi f_2 t + \phi_2)]^n
\]  

(B.2)

where \( v_m \) is the input modulation signal

\[
v_m = v_s [\sin(2\pi f_1 t + \phi_1) + \sin(2\pi f_2 t + \phi_2)]
\]  

(B.3)

If we ignore the non-linear conversion in the photodetector, its optical-to-electrical conversion can purely modeled by its responsivity \( \eta_d \). In this case, the photocurrent generated in the photodetector is

\[
I_d = P_L t T(V) \eta_d = P_L t \eta_d \left\{ T(V_B) + \sum_{n}^{\infty} \frac{T^{(n)}(V)|_{V_B}}{n!} v_s^n [\sin(2\pi f_1 t + \phi_1) + \sin(2\pi f_2 t + \phi_2)]^n \right\}
\]  

(B.4)

Where \( P_L \) is the laser power, \( t \) is the total optical insertion loss from laser to photodetector. It clearly shows that different orders of distortion current present at the photodetector can be calculated from the nonlinear optical transfer curve.
of the optical modulator. By separating the orders that we are interested in, we can obtain dc, fundamental signal, and all high order distortion signals.

The complexity here for separation is the mixture of polynomial expansion and trigonometrical conversion. Different order signal terms do not come just from the equivalent order of polynomial term. For instance, the second order term \( \sin^2(2\pi f_1 t + \phi_1) \) will also generate dc signal due to the trigonometrical identity \( \sin^2 x = [1 - \cos(2x)]/2 \). This happens for all other terms as well. Only when small signal approximation is applicable, the inclusion of other polynomial orders can be ignored.

Fortunately, the definitions of IIP3, OIP3, and SFDR are based on the extrapolation from small signal approximation. Therefore the calculation is straightforward. Only the 1 dB compression point involves inclusion of high order terms in the fundamental signal. In most cases, output signal levels decrease very quickly when the distortion order goes up. Hence only the inclusion of lowest several order terms is necessary.

Let’s first take a look at the third-order intermodulation signals. Due to symmetry, only one frequency, \( 2f_1 - f_2 \) for example, needs to be examined. The lowest polynomial order involving this frequency is the third order. By following polynomial rules and trigonometrical identities,

\[
\begin{aligned}
(sin x + sin y)^3 &= sin^3 x + sin^3 y + 3 sin^2 x sin y + 3 sin^2 y sin x \\
\sin^2 x \sin y &= \frac{1 - \cos(2x)}{2} \sin y = \frac{\sin y}{2} - \frac{\sin(2x + y) - \sin(2x - y)}{4}
\end{aligned}
\] (B.5) (B.6)
We can obtain the $2f_1 - f_2$ term in the optical transfer curve to be

$$\frac{T^{(3)}(V)|_{V_B}}{24} v_s^3 \sin[2\pi(2f_1 - f_2)t + (2\phi_1 - \phi_2)]$$  \hspace{1cm} (B.7)

It is also easy to get fundamental part in the optical transfer curve by following the same procedure.

$$T'(V)|_{V_B} v_s \sin(2\pi f_1 + \phi_1)$$  \hspace{1cm} (B.8)

According to the definition of IIP3, it happens when the output fundamental and third order intermodulation signal levels become identical. This is equivalent to equating the coefficients of fundamental and third order intermodulation parts in the transfer curve expansion.

$$\frac{T^{(3)}(V)|_{V_B} v_s^3}{24} = T'(V)|_{V_B} v_s$$  \hspace{1cm} (B.9)

which yields

$$v_s^2 = \frac{24T'(V)|_{V_B}}{T^{(3)}(V)|_{V_B}}$$  \hspace{1cm} (B.10)

Then IIP3 can be easily derived to be

$$\text{IIP3} = \frac{1}{2} \frac{v_s^2}{R_s} = \frac{12T'(V)|_{V_B} R_s}{T^{(3)}(V)|_{V_B}}$$  \hspace{1cm} (B.11)

OIP3 signal is just the IIP3 multiplied by the link gain $G$. By definition, SFDR due to third order nonlinearity can also be obtained by taking the two-thirds power of the value of OIP3 divided by the output noise level. In dB scale, it can also be formulated as

$$\text{SFDR} = \frac{2}{3}(\text{OIP3} - \text{Noise})$$  \hspace{1cm} (B.12)
The derivation of 1 dB compression point is more tedious because it needs to include all polynomial terms with fundamental signal in it. However it does not need to have two tone input signals. If we take the approximation to include up to the fifth term, we can get

$$T'(V)|_{V_B v_s} + \frac{3T^{(3)}(V)|_{V_B v_s^3}}{24} + \frac{T^{(5)}(V)|_{V_B v_s^3}}{192} + \ldots \quad (B.13)$$

The 1 dB compression point can be obtained from the $v_s$ value at which the output fundamental signal falls 1 dB below that extrapolated from the first term in the above expression. Since square law detection happens in the photodetector, calculation needs to take the square of the terms expanded from the optical transfer curve.

All the above discussion is intended for a general shape optical transfer curve. In the case of Mach-Zehnder type modulator with sinusoidal shape optical transfer curve, different orders of distortion are represented by Bessel functions $J_n$, which makes the calculation much easier.
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