

行政院國家科學委員會專題研究計畫總報告

智慧型寬頻光纖通信網路研究－

Study of Intelligent Broadband Optical Fiber

Communication Networks

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一、 中文摘要

光纖具有寬頻高容量的特性，為通信網路最重要的傳輸媒體；由於光纖的頻寬遠大於目前電子系統所能達到的頻寬，所以如何充分利用光纖的頻寬乃是建立高速光纖網路的一大重要課題。另外高密度分波多工在高容量的骨幹網路已成為主要趨勢；目前除了分波多工 (Wavelength-division multiplexing, WDM) 技術之外，光分碼多工處理 (Optical code-division multiple access, OCDMA) 技術亦是漸獲重視的一個領域；然而由於多媒體資訊時代的來臨，各式網路系統必需做有效的整合，本整合型計畫針對智慧型寬頻光纖通信網路作長期而有系統的研究。

本研究計畫研究成果包括有光分碼多工之編碼，以光分碼多工應用在被動光網路，光波長選取、波長轉換和波長監控，高速光交換機，智慧型光纖分波多工網路之保護網路，光纖耦合與光纖光柵之設計等。

關鍵詞：分波多工，光分碼多工，被動光網路，波長轉換，波長監控，高速光交換機，光纖耦合、光纖光柵

Abstract

The optical fiber has inherent large transmission bandwidth. It plays an important role in broadband wireline communication networks. Three techniques: wavelength division multiplex (WDM), code division multiple access (CDMA) and time

division multiple access (TDMA) are the most commonly used transmission schemes. Due to the growing demand of the Internet transmission bandwidth, the networks are required to provide full multimedia services. Dense WDM (DWDM) technology has become the dominant transmission technology of the backbone networks. On the other hand, OCDMA is getting more and more attention recently and can be used in the access networks. Therefore, different kinds of systems have to be integrated to achieve this goal. This project aimed at studying the intelligent broadband optical fiber communication systems thoroughly.

The main research results include coding in optical CDMA (OCDMA) systems, passive optical network using OCDMA, wavelength selection, wavelength conversion, wavelength monitoring, high-speed optical switch, protection of WDM networks, design of long-period fiber gratings and fiber coupling, etc.

Keywords: wavelength-division multiplexing (WDM), optical code-division multiple access (OCDMA), passive optical network, wavelength conversion, wavelength monitoring, high-speed optical switch, fiber coupling, fiber gratings

二、重要研究成果

【第一年】

子計畫一

IP over ATM over SDH/WDM 為目前網路最常見的架構，主要是 ATM 及 SDH 含有營運所需的 OAM 功能。但 IP over ATM over SDH/WDM 的層次太多，且每一層的功能協定重複，在 host 或 router 處理規約太過繁瑣，因此必須改變其結構。我們提出一個新的 IP over WDM 架構，這個架構使用了標籤交換 (label switching) 與 WDM 波長資訊的觀念，同時也將 MPLS 流量控制概念與 IP 傳輸與交換平面做有效整合。

另一方面我們也分析了兩種可以處理非同步變動長度封包的高速光網路路由：部份分享暫存器 (PSB) 與多重路徑路由 (MPR)，模擬結果顯示 PSB WDM 光路由器比較適合使用在高速 IP 流量的核心網路 (core networks)。

子計畫二

本子計畫第一年我們將數個不同特性的光濾波器組合成新型的光濾波器，經比較數種不同濾波方式的調幅速度及可調範圍，選定兩種複合式可調式光濾波器架構，一為以光纖光柵及可調式聲光濾波器組合而成；光纖光柵則設計使用取樣光柵以在光譜上具有多個相同的反射尖峰，此部分已完成分析；另一則是使用 AWG 加上光電開關，可利用較大通道間隔的 AWG 以組合成具有較密間隔的濾波器。此外，我們設計適用於 1.3/1.55 μ m WDM 系統的新型波長多工器，結合 MMI 與光柵的優點，大幅縮小多工器的尺寸，適於光積體化製作，且具有

非常好的波長分離效果。我們也分析模擬波長轉換器的波形整形效果，此可有效抑制雜訊及串音。初步以 OPALS 軟體模擬發現利用 XPM 或 XGM 效應的 SOA 波長轉換器皆可此特性將使用於我們所設計的新型光電開關中。

子計畫三

本子計畫目的針對利用質數碼同時容納量的特性設計出智慧型異質碼多工器，並探討其有關之技術。並且針對光環式共振濾波器的特性，探討其有關之技術及延伸。若用傳統的方法來找出適合共存的碼須耗費許多的資源和時間，以修正式質數碼系統為例，當質數碼為 P 、最大同時容納量為 P_{max} 且 DT 為每次決定所選擇的碼是否能共存所須計算的時間，則整體所須

的計算時間為
$$\sum_{K=P}^{P_{max}} C_K^{P^2} \times DT。$$

由上式可知，當質數碼變大時所須的時間會呈現指數型式上昇，因此須利用更好的搜尋方式以找尋最大同時容納量的碼。近年來，基因演算法 (Genetic algorithm: GA) 已廣泛的被應用於解決最佳化的問題，如 multiprocessor scheduling, topology design and bandwidth allocation 和 channel allocation 等問題，也得到很好的結果，因此我們採用基因演算法來降低搜尋所須要的時間，並找出適當的碼。透過基因演算法，一代代複製、交配及突變的過程，我們能找出最好的解，將我們實際所得的結果和其他論文結果比較，我們可發現我們的結果大於其他論文的結果，且利用此方法我們可以直接的得到碼組合。

同時我們採用了一種新式的碼區別多重擷取技術—相異質數碼系統

區別多重擷取技術及其設計方法一來提昇整合容量。在光纖波域分工系統中，陣列波導光柵已被頻繁得使用作多工和解多工器，再加上迴路的原理形成陣列波導光柵的加-降多工器(ADM)。參考陣列波導光柵的技術，我們利用光環式共振濾波器來達成加-降多工器，在光纖通訊系統中的功能，而且還具有可調性等，這是新型的架構。光環式共振濾波器的特性是可在選擇端濾下想要的波形，其它剩餘的波形會在傳輸端濾出，利用此來將波長分工，而在波形上的時間軸與頻率軸與波長軸之間必須相互映射的一個重要問題。以元件的觀點，來考慮光環式共振濾波器中的參數:耦合係數，共振腔長度等及可調範圍，並協調參數之間的影響，形成合適的元件。而濾下的波長也有範圍的限制。

子計畫五

本子計畫第一年，我們發展出模擬長週期光纖光柵理論模型，特別是二元光纖光柵的模型；並製作長週期光纖光柵且應用在光通信系統上，例如使頻寬超螢光光纖光源之頻寬增加與增加耦光時之工作距離。我們也開發出損耗可調式之新型鋸齒狀長週期光纖光柵。

子計畫六

在本子計畫第一年，藉由適當佈置光感測器，可提供網路偵錯、容錯等功能。在本年度計畫中，我們提出以光纖布拉格光柵為溫度感測器，並利用模糊-類神經之演算法，以達到高效率、高精度之感測技術。並利用光功率計與計頻器量測光經由光纖布拉格光柵後，反射光頻譜分布情形。經由適當改變外在溫度，我們發現：光頻譜將隨溫度變化而產生飄移情形。

並藉由分析比較，我們發現其監測精度可達到 0.005°C 。我們亦嘗試設計分析新型 SOI 積體光波導感測器，利用 SOI 之低成本、高速率之 CMOS 積體電路製程特色，未來可達到光電元件積體化整合之效果。我們使用 SOI 為基材所設計之波導麥克生干涉儀結構圖。經由分散矩陣法(scattering matrix method)，我們可以得知此波導感測元件之輸入與輸出光場間關係，並利用耦合模態方程式(couple mode equation)及光譜隨溫度飄移之原理達到溫度感測之目的。利用麥克生干涉儀、SOI 波導布拉格光柵與光纖布拉格光柵於外在溫度變化為 10°C 之光譜飄移作比較。由此可知，我們所設計的麥克生干涉儀，對於溫度的靈敏度將 20 倍高於傳統之光纖布拉格光柵感測器。且其 FWHM 亦將遠小於單一 SOI 波導布拉格光柵與光纖布拉格光柵感測器。

由於 SOI 麥克生干涉儀具有高溫靈敏度及低 FWHM 特性，我們嘗試將其應用至分波多工光通訊網路中，並經由分析其於階梯型、星型及樹型等拓樸結構下之訊雜比(SNR)，了解其於網路監測之功能。而星型拓樸結構在分波多工光通訊網路中，可容納較多數目的 SOI 麥克生干涉儀。我們亦針對感測範圍(光纖佈放長度)做分析比較。經由上述分析，我們發現星型拓樸結構在分波多工光通訊網路中，將可容納 82 個高溫度靈敏度之 SOI 麥克生干涉儀，且其感測範圍將廣達 220 公里以上。其優越之網路監測特性，並可作為未來智慧型光通訊網路之參考。

個別子計畫內容請參考附件。

【第二年】

子計畫一

我們已提出一穩定且具有路由交換及傳輸功能的規約。以此規約為基礎，我們研究了叢發交換（burst switching）和封包交換（packet switching）所需具備的條件，進而提出了一多平面之光交換器架構。此交換器之架構簡單，而其所能達到的最高流通率（maximum throughput）為百分之百。同時，其輸入之封包不須在輸入端對齊等待排程，其長度也不需限定為一定長度，所以可簡化光儲存器的設計。理論分析以及模擬結果顯示，此交換器之效能優異。

現代網路的交換器必須提供多點發送交換（multicast switching）的功能，即允許同一進入的封包同時到達多個輸出端。此功能可以支援諸如視訊會議、隨選視訊系統和遠距教學等等之網路應用。我們的交換器可以有效地提供多點發送（multicast）。同時，我們研究了一般可變長度封包之多點發送之效能分析，並提出了最佳化的設計原則。

子計畫二

本子計劃在波長選取方面已設計一新型半導體可調光濾波器，可應用於波長塞取多工器或光交換開關開關中，利用取樣光柵（SGDBR）做為預設波長的光濾波器，SGDBR 的反射率尖峰的間隔（ ΔP ）為 0.8nm，可以發現此光濾波器不但有眾多且平坦的反射率頻譜，而且其相鄰通道的具有足夠的抑制效果，藉由調整溫度或電流以改變波導的折射率，使得通道的位置符合 ITU 的規範。另外，本計畫也已提出了一以四波混合的方式

來作波長轉換，藉由加入輔助光提高半導體光放大器的飽和輸出功率，改善整體波長轉換效率。另外運用輔助光技術設計兩段式串接之光放大器結構，可簡化操作方式，提供更高的轉換效率；而且具有可積體化之優點，能實用在真正光通訊網路。

子計畫三

本子計劃中，我們提出的智慧型壓縮決策的光電 ATM 交換器架構，討論如下的兩個基本論點：(1)如何決定壓縮決策的花費時間，(2)當所有細胞/訊框還是處於壓縮狀況中的同時，如何避免輸出連結是可用的情況。為了解決此兩問題，我們提出一簡單且有效的壓縮決定演算法（compression decision algorithm, CDA），是根據細胞的路由資訊，由 CDA 決定是否來執行壓縮決策，由模擬結果得知系統的產量有明顯的改進；因此我們提出的演算法將可達最佳化的而不會浪費連接容量。

子計畫五

本子計劃第二年延續第一年研發研究成果，我們利用自行研發之長週期二元光纖光柵的理論模型，成功建立 quasi-periodic 結構 LPFG 的頻譜模擬與相關應用之光濾波器設計，並將長週光纖光柵進一步應用在光纖通信系統上包括：寬頻超螢光光纖光源進一步改進其架構而產生出多波長的光纖光源，半導體雷射光源與由 LPFG 輔助的光纖透鏡的直接耦合而提高耦合效率，並且利用新開發出之新型鋸齒狀長週期光纖光柵形成損耗及波長可調式的光纖濾波器。

子計畫六

本子計畫第二年著重於智慧型分碼多工網路監測訊號與次載波多工通信訊號同時傳輸技術之研究，將光纖雷射應用於智慧型分碼多工通信訊號及次載波多工訊號傳輸，分析光纖雷射載著智慧型分碼多工通信網路監測訊號並應用於次載波多工系統中之可行性。為了能有效載入檢測訊號，必須製作雙埠環型光纖雷射，利用其雙埠同時調變訊號傳輸特性，以同時載入網路偵測 CDMA 訊號。在此計畫中，我們自行研發完成一寬頻光源並供一種使用光纖布拉格光柵之類神經型光監測網路使用。吾人所設計之寬頻光源可達 100nm 頻寬，而雙埠環型光纖雷射也已開發完成，並申請專利，其特性也作完量測與分析，以提供智慧型寬頻通信網路監測使用。

個別子計畫內容請參考附件。

【第三年】

子計畫一

本子計畫提出一個以光分碼多工應用在被动光網路之實驗網路；此實驗網路可以解決目前被动光網路的問題；可以實現全服務接取網路(FSAN)，研究內容包含同步光分碼多工之編碼研究，非同步光分碼多工之編碼研究，以以太網路訊框為基礎結合完美差異碼之被动光網路。我們在本計畫中設計出最好的完美差異碼及非理想交互關係碼。前者是利用組合數學理論中的完美差異集合建構出一組編碼，具有完美的相關性；經由我們設計的接收機架構可以完全消除其他用戶干擾，是同步光分碼多工系統中效能最好的編碼方式；後者則是用方塊設計(block design)從完美差異集合當中選取出來的非同步編碼方式，在

滿足最大系統(maximal system)理論下可以使用戶數達到最多，經過數值分析結果證明出這兩組編碼方式比傳統的同步碼與非同步碼有更好的性能；將此兩組編碼應用被动光網路可以真正達到寬頻接取的服務。

子計畫二

本計畫自行設計一款分佈式布拉格反射鏡可調式半導體雷射，藉由調動光柵區的注入電流，改變反射鏡的頻譜分佈，進而使雷射發光在特定選擇波長。增益區(gain)提供足夠的能量使雷射發光，最大輸出光功率大於 5mW，相位區則作為波長微調之用。

此外，我們也使用被动 FP etalon 濾波器具有對溫度不敏感的特性，利用溫控將濾波器的透射與反射頻譜的交叉點對準在標準 ITU 規範波長，觀察透射與反射量的差異，可監控波長的細微偏移量。通道監控方面，利用 LED 或 SOA 等半導體光電元件(SOD)的穿透特性，將 SOD 做為光感測器，光打入 SOD，SOD 的接面電壓會隨著注入光的波長不同，有不同的對應電壓，本技術可監控小於 0.008nm 以下的波長漂移現象，滿足 ITU DWDM 100GHz 通道間距規範的品質，並精密且快速監控並回應相關訊息給相關控制單元。

子計畫三

光 CDMA 是利用特殊的光碼將傳輸資料加以編碼後再傳送，如此可容納大量的使用者在同一光纖上同時傳送，故可大量地提升光纖網路的頻寬。然而受限於使用者資料的互相干擾(multi-user interference, MUI)，當同時使用者的數量增加時，錯誤率將隨之增大。本計畫以曼徹斯特編碼的方式，將原本雙極性的 Walsh 碼轉成適合

光纖傳送的單極性碼，我們稱之為曼徹斯特編碼式 Walsh 碼（Manchester Coded Walsh Code, MCWC）。該碼的定干擾量特性讓接收端得以估計出干擾量而加以消除，故可有效地降低錯誤率。此外，針對不同的網路型態，我們提出了三種接收架構：固定多工式、排序式及非排序式接收架構。分析結果顯示，三種接收架構均能大量地提昇網路頻寬，並維持良好的通信品質。尤其是前二種架構，當傳輸量越大時，其通信品質反而越好。

子計畫五

我們利用光束傳播法來分析去年所執行的光纖耦合架構。此架構乃是利用在光纖透鏡之後埋入長週期光纖光柵，以達到光源與光纖之間光耦合的目的。我們利用光束傳播法來分析光纖的核蕊對光纖透鏡的影響以及分析兩種典型的光纖，即色散偏移光纖與單模步階式光纖對此架構的影響。利用光束傳播法，我們可以決定出在不同的光源光腰下有最佳耦合效率的透鏡半徑與工作距離。另外，我們也比較兩種分析方法，即 ABCD 法與光束傳播法，可發現光束傳播法與實驗結果比較吻合。此外，我們提出了在二元長週期光纖光柵中非線性開關的理論模型。與過去利用非線性模場耦合理論分析在正弦長週期光纖光柵的非線性開關相較起來，此一新的模型清楚說明了非線性相位調制如何改變光柵的相位匹配條件，而且能顯示出週期個數對穿透率的影響。利用這個模型，我們探討入射脈衝波形對非線性開關的影響，證實了在某特定強度的理想方波入射時會有完全的非線性

耦合。同時我們進一步發現在長週期光纖光柵中完全耦合的條件，而這些條件廣義地適用在線性與非線性耦合的情形。再將此模型應用在類週期性長週期光纖光柵中，我們在理論上預測了非線性開關存在於此種光柵，並且在某特定強度的理想方波入射時亦有完全的非線性耦合。

子計畫六

本子計畫提出一有效的路由自動切換方式，在此我們設計出一具波長選擇性的元件，並研究智慧型光纖分波多工網路之保護網路，以提高智慧型光纖分波多工網路的可靠度。此波長選擇性的元件是一積體化的可調式光分叉復用器，利用結合布拉格光柵及多模干涉區所組成之光分插復用器元件，我們在布拉格光柵兩側摻雜硼和磷離子，再利用電壓調變布拉格光柵中雜質的濃度變化，利用雜質的變化來改變折射率，藉此從波導中取出特定信號，以達到可控制特定波長信號的路由路徑，接著，我們將此可調式光分插復用器應用於高密度分波多工的保護網路系統中，如此才可充分利用有限的波長資源。

個別子計劃內容請參考附件。

三、計畫成果自評與討論

本計畫在光纖通訊研究上有相當多的成果，在分波多工電路元件之設計，分碼多工系統之研究均有許多論文發表或投至國際期刊與國際會議，成果豐富。因期刊及國際會議論文篇數甚多且已分列在各分項計畫報告中，故不再重列。

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子計劃一：高密度波長多工通信網路研究

計畫編號：NSC89-2215-E-002-012

NSC89-2215-E-002-061

NSC90-2215-E-002-012

主持人：吳靜雄教授 台灣大學電機系

附件二

子計畫二：應用於寬頻光通訊網路之波長轉換與波長選取技術研究

計畫編號：NSC89-2215-E-011-006

NSC89-2215-E-011-014

NSC91-2215-E-011-001

主持人：李三良教授 台灣科技大學電子系

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子計劃三：智慧型光 CDMA 網路

計畫編號：NSC89-2215-E-032-001

NSC89-2215-E-032-002

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主 持 人：李揚漢教授 淡江大學電機系

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子計畫五：光柵模組在光通信系統之應用

計畫編號：NSC89-2215-E-002-013

NCS89-2215-E-002-062

NSC90-2215-E-002-007

主持人：王倫教授 台灣大學光電光程學研究所

附件五

子計畫六：智慧型寬頻光通信網路監測技術之研究

計畫編號：NSC89-2215-E-155-002

NSC89-2215-E-155-011

NSC90-2215-E-003-001

主持人：曹士林教授 國立師範大學 光電研究所

附件一

行政院國家科學委員會專題研究計畫成果報告

智慧型寬頻光纖通信網路研究－

子計劃一：高密度波長多工通信網路研究

計畫編號：NSC89-2215-E-002-012

NSC89-2215-E-002-061

NSC90-2215-E-002-012

執行期間：88 年 8 月 1 日至 91 年 7 月 31 日

主 持 人：吳靜雄 教授 台灣大學電機系

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黃柏鈞、楊銘樑、熊致融、張智豪、簡義文、康欣祐、王迺鈞

一、 中文摘要

【第一年】

網際網路使用者日益增加，使得通訊量快速成長；爲了提供使用者更大的服務頻寬，光纖高密度波長多工技術逐漸成爲主要技術，但由於網路營運的需要，如 ATM 及 SDH 仍爲網路之基本架構。故 IP over ATM over SDH/WDM 爲目前網路最常見的架構，主要是 ATM 及 SDH 含有營運所需的 OAM 功能。但 IP over ATM over SDH/WDM 的層次太多，且每一層的功能協定重複，在 host 或 router 處理規約太過繁瑣，因此必須改變其結構。我們提出一個新的 IP over WDM 架構，這個架構使用了標籤交換(label switching)與 WDM 波長資訊的觀念，同時也將 MPLS 流量控制概念與 IP 傳輸與交換平面做有效整合。另一方面我們也分析了兩種可以處理非同步變動長度封包的高速光網路路由器：部份分享暫存器(PSB)與多重路徑路由(MPR)，模擬結果顯示 PSB WDM 光路由器比較適合使用在高速 IP 流量的

核心網路(core networks)。

關鍵詞：高密度波長多工，協定，標籤交換，路由器

【第二年】

多協定標籤交換 (Multi Protocol Label Switching, MPLS) 與光交換 (Optical Switching) 的整合已經成爲光通訊網路的趨勢。我們已提出一穩定且具有路由交換及傳輸功能的規約。以此規約爲基礎，我們研究了叢發交換 (burst switching) 和封包交換 (packet switching) 所需具備的條件，進而提出了一多平面之光交換器架構。此交換器之架構簡單，而其所能達到的最高流通率 (maximum throughput) 爲百分之百。同時，其輸入之封包不須在輸入端對齊等待排程，其長度也不需限定爲一定長度，所以可簡化光儲存器的設計。理論分析以及模擬結果顯示，此交換器之效能優異。

現代網路的交換器必須提供多點發送交換 (multicast switching) 的功能，即允許同一進入的封包同時到達

多個輸出端。此功能可以支援諸如視訊會議、隨選視訊系統和遠距教學等等之網路應用。我們的交換器可以有效地提供多點發送 (multicast)。同時，我們研究了一般可變長度封包之多點發送之效能分析，並提出了最佳化的設計原則。

關鍵詞：交換器，多協定標籤交換，光交換，叢發交換，多點發送，可變長度封包，非同步。

【第三年】

光纖通信技術為固定網路的主流，目前被動光網路被視為可以真正實現全服務接取網路(FSAN)的方法。現在提出的被動光接取網路以時域多工(TDMA)與非同步傳輸模式(ATM)為主，然而對於網際網路或乙太網路封包而言，時域多工並不適用；為了實現真正寬頻接取服務，必需解決寬頻接取網路的問題；本計劃提出一個以光分碼多工應用在被動光網路之實驗網路；研究內容包含同步光分碼多工之編碼研究，非同步光分碼多工之編碼研究，以乙太網路訊框為基礎結合完美差異碼之被動光網路。我們設計

出最好的完美差異碼及非理想交互關係碼。前者可以完全消除其他用戶干擾，是同步光分碼多工系統中效能最好的編碼方式；後者則是在非同步系統中有最多用戶數的編碼方式，這可以有效提高網路容量。我們提出的光分碼多工應用在被動光網路上可以真正達到寬頻接取的服務。

關鍵詞：被動光網路，時域多工，非同步傳輸模式，光分碼多工，完美差異碼，非理想交互關係碼

Abstract

【1st year】

Because the Internet traffic increases rapidly, the Internet Protocol (IP) has become very popular. Large transmission bandwidth is required to provide more services. The Dense Wavelength Division Multiplexing (DWDM) technology is growing in optical fiber transmission and can provide a lot of transmission capacity. Currently many companies have the network architectures as IP over ATM

over SDH/WDM. Because these architectures have too many layers, they need a lot of processing power in the host and the router. Thus the architectures must be simplified to make efficient use of resources. A new IP over WDM protocol with routing and transport capabilities is proposed. This protocol utilizes the concept of label switching and wavelength information of Wavelength Division Multiplexing (WDM). The transport adaptation and encapsulation scheme of this protocol is simple and transparent to protocol data unit. It also integrates the Multiprotocol Label Switching (MPLS) traffic engineering control concept as the IP transport and switching control plane. On the other hand, we have fulfilled the performance analysis for two types of asynchronous Wavelength Division Multiplexing (WDM) optical routers that can handle asynchronous and variable length packet switching for high-speed Internet core networks. We compare the network using Partially

Shared Buffer (PSB) and using Multiple Path Routing (MPR) by simulation results. We conclude that a PSB type asynchronous WDM optical router using the same number of input/output ports as the re-circulation ports without using void filling algorithm is more suitable for handling IP traffic in the high-speed optical core networks.

Keywords: dense wavelength division multiplexing (DWDM), multiprotocol label switching (MPLS), router, partially shared buffer (PSB), multiple path routing (MPR)

【2nd year 】

Multiple Protocol Label Switch (MPLS) and Optical Switching integration is a trend for future all optical networks. We had proposed a very stable protocol with transport and switching capabilities. Based on this protocol, we study the feasibility of using this protocol on burst switching and packet optical wavelength switching architecture. The architecture may fulfill

the requirements of both burst switching and packet switching networks. The architecture can switch variable length packets asynchronously, i.e., the input packets are not necessary to enter the switch at the inputs ports simultaneously. This property will simplify the buffer design. The result shows that the throughput of this switch is very high.

A modern switch must provide multicasting to allow multiple copies of an incoming packet to be sent to their destinations simultaneously. Such feature is necessary to support video conferencing, video-on-demand, distant learning, etc. An enhanced version of the variable length self-routing packet switch is proposed to perform multicast switching. The optimal design rule and analysis of multicast switching of variable length packets are discussed in this report.

Keywords: switch, MPLS, burst switching, packet switching, optical switching, variable length packet,

asynchronously

【3rd year】

Optical fiber is an important medium of the wireline communication systems. Passive optical network (PON) is considered to be a solution to realize full-service access networks (FSAN). It is a point-to-multipoint optical network with no active devices through the signal transmission. The network is easy to maintain and to be upgraded. Combining Ethernet frame with PON (EPON) is a promising candidate for next-generation broadband access networks. PON uses time division multiple access (TDMA) and asynchronous transfer mode (ATM) in upstream and down stream direction. The two schemes have fixed time slot and are not suitable to variable length packets, such as IP packets and Ethernet frames. We proposed an EPON configuration using optical code division multiple access (OCDMA) in its upstream direction. The downstream direction is still broadcasting in nature,

which is the same as the ordinary EPON. We proposed a synchronous OCDMA code: Perfect difference code (PDC). PDC has better performance than the other codes when the multiple user interference (MUI) canceller is used. We also proposed an asynchronous OCDMA code: optical orthogonal codes (OOCs) with nonideal cross correlation. Given the same code weight and code length, the size of the proposed new codes is ten times more than that of ideal optical orthogonal codes. It can thus accommodate more simultaneous users and increases the network utilization. The proposed system can achieve the goal of broadband access networks.

Keywords: Passive optical network (PON), Perfect difference code (PDC), multiple user interference (MUI), optical orthogonal codes (OOCs), cross correlation

二、 緣由與目的

【第一年】

The explosive growth of Internet traffic demands more bandwidth on the core networks. While IP backbone systems are being realized using Synchronous Optical Network / Synchronous Digital Hierarchy (SONET/SDH) and Asynchronous Transfer Mode (ATM) technologies, existing transport networks will not yield the bandwidth-effective transporting and routing schemes for IP traffic. The integration of IP and Dense WDM (DWDM) has been the most popular research and development area for next generation Internet. These new emerging technologies, i.e., Generalized Multiprotocol Label Switching (GMPLS) and the related signaling protocols play an important role of the control plane that could integrate IP and DWDM capability. They aim at providing an efficient and flexible end-to-end

lightpath for IP centric and non-IP centric services. MPLS is rapidly emerging as an Internet Engineering Task Force (IETF) standard. MPLS intends to enhance the speed, scalability, and service provisioning capabilities in the Internet. MPLS uses the technique of packet forwarding based on labels, to enable the implementation of a simpler high-performance packet-forwarding engine. MPLS is best viewed as a 2.5th layer protocol, integrating layer 2 and layer 3 functions. With this point of view, optical label switching has been deemed as the most probable technique for the future all optical networks.

However, at the present time, due to the gap to success of optical elements, optical labeling is not so realizable. Other techniques using optical sub-carriers to carry the labels may suffer the optical transparent problem in all optical networks. For this sake, we need to reconsider 1) the properties of optical transmission, 2) the properties of IP traffic, and 3) future requirements for

IP backbone traffic. What we need are as follows: 1) an IP over WDM transport mechanism that can handle physical framing, data link delineation and part of layer 3 routing functions at the same time. This mechanism also needs to speed up the processing, while keeping all possible optical networks transparent 2) As for IP traffic properties, we need additional traffic engineering models as the control plane for IP networks. In other words, we need a high-speed asynchronous transport mechanism integrating with label switching techniques and the advantages of control plane functions. This mechanism must be able to migrate to future all optical networks smoothly.

Although GMPLS could support packet, Time Division Multiplexing (TDM), wavelength, waveband and fiber switching, it still treats DWDM as a big bandwidth pipeline and acts more like cross-connect provisioning mechanism. This may not be the most efficient and flexible solution that could be achieved

directly by combining the advantage of DWDM with all optical packet switching capabilities to yield WDM optical router for IP centric services. However, at the present time, due to the availability of optical elements, all-optical packet switching and processing are not so realizable. Almost all-optical packet switching using optical-to-electrical-to-optical (O-E-O) techniques for the header processing is more practical. At the same time, the optical buffer can only be emulated by using fiber delay lines with fixed delay unit. On the other hand, considering the present Internet network scenario and IP properties, the transfer mode based on variable length packets and asynchronous operation is more effective.

【第二年】

Recently, the traffic of backbone and access networks increase dramatically. The transmission capacity of optical networks doubles every nine

months. However, the switching and routing capacity increases slowly compared with the transmission capacity. Many switching or access schemes are proposed [1]-[4]. Among them, integration of MPLS and wavelength division multiplexing technologies is a promising choice.

Combination of electronic and optimal routing for the hybrid Internet Protocol (IP) networks is inevitable. A WDM optical packet switch consists of input interface unit, the switch fabric, the output interface unit and the control unit. In a conventional switch, the input interface unit processes the packet header information extraction, packet delineation and alignment. The switching fabric is the most important element. The output interface unit regenerates the optical signals. Based on the information in the packet header, the control unit controls the switch. Because of the synchronization requirements, the packet size must be fixed. Currently, the

packet header can be out-of-band signaling including sub-carrier multiplexing and dual wavelength, or in-band-signaling.

In this report, we propose a self-routing variable-length packet switch, which can provide unicast and multicast switching. The performance of the switch operating in unicast or multicast switching is verified both by simulation and queueing analysis. The first part of this report focuses on the unicast switching, and the second part discusses the multicast switching.

【第三年】

Recently, fiber-optic code-division multiple-access (CDMA) systems have attracted much attention. Because synchronous optical CDMA (SOCDMA) systems are more efficient to utilize the bandwidth than any asynchronous optical CDMA systems, many studies of SOCDMA systems have been reported in the literature. There are several

possible ways to achieve slot synchronization. One way is that we add a synchronizer node located at the star coupler. This node broadcasts a pulse called the sync pulse periodically. Using the periodic pulses, the synchronization may be achieved among all the transmitters. One of the major concerns of designing a direct-detection SOCDMA system is the multiple-user interference (MUI) because the performance of this system is usually interference limited. That is, there is an asymptotic error floor no matter how much power the receiver received. Unlike in the electric domain, the signature sequence in a direct-detection SOCDMA consists of unipolar (0, 1), sequences. As a result, there are no strict orthogonal codes under the constraint that the total number of codes is equal to the code length. We study the “perfect difference sets” and propose a new SOCDMA system using these sets as the signature code sequences instead of modified prime sequences. The perfect

difference codes have the following interesting characteristics:

1) Any two distinct codes are cyclic-shifted with each other.

2) The cross correlation between any two distinct codes is exactly one.

However, the modified prime sequence codes have the cross correlation either one or zero. Although these codes are not strictly orthogonal, we may say that they are quasi-orthogonal. Using the first property, we propose a simple transmitter structure, in which the code sequence can be cyclic-shifted to get another transmitter. Using the second property, we also can easily design the receiver to remove the effect of the MUI. However, from the viewpoint of asynchronous systems, the perfect difference codes are basically identical because they are cyclically shifted with one another. In other words, we have only one code for asynchronous systems. Therefore, we must modify the perfect difference codes to get a larger code size. To do so, we consider the case that the

correlation is relaxed to two. When the two codes are not aligned—that is, they are cyclically shifted with each other—the cross correlation between the two codes is exactly one. In such a situation, it does not violate the constraint of correlation. However, when there is no cyclic shifting between them, the value of the cross correlation is equal to the code weight, which is far from the constraint. To overcome this, we can drop some marks appropriately from marks of each code, such that the modified code weight is reduced and the code length is still equal to original size. Moreover, the cross-correlation property should satisfy the constraint.

We further discuss the implementation of OCDMA system in high-speed access networks. EPON is promising a solution of broadband full service access networks and is illustrated in Fig. 1.

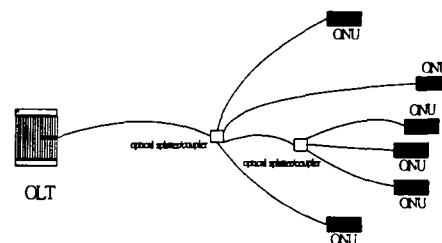


Fig. 1. An EPON configuration

The downstream direction is a point-to-multipoint transmission and is broadcasting in nature. Downstream packets are sent from the optical line terminal (OLT) and received by optical network units (ONU) according to the specific address. In the upstream direction, each ONU transmits packets to the OLT and may collide with other packets if more than one packet arrive at the OLT simultaneously. To solve this problem, three multiple access schemes can be used: time division multiple access (TDMA), wavelength division multiple access (WDMA), and code division multiple access (CDMA). In TDMA, each ONU is allocated a time slot telling when to send data and for how long. This allocation can be static or dynamic. In WDMA, each ONU transmits data at different wavelengths. The WDM components such as tunable laser and tunable optical receiver are much more expensive than that of

TDMA and CDMA. The operation and maintenance are also more complicated. Thus makes WDMA not suitable for FSAN.

We propose an EPON configuration using synchronous OCDMA (SOCDMA) in its upstream direction. The downstream direction is still broadcasting in nature, which is the same as the ordinary EPON. Bidirectional transmission is performed over a single fiber by wavelength multiplexing of $1.5\ \mu\text{m}$ wavelength in the downstream direction and $1.3\ \mu\text{m}$ wavelength in the upstream direction. In the downstream direction, the packets are broadcasted to all ONUs from the OLT. Only the frames with the desired ONU's Ethernet address are extracted by that ONU and the others are discarded. In the upstream direction, Ethernet frames from the ONU are spread by a spreading circuit, which uses PDC as the code signature. The OLT decodes the incoming OCDMA signals by using the despreading receivers. Upstream packets

of different addresses can be received simultaneously without collision. The EPON configuration with maximum range of at least 20km can split at most 32 subscribers. The synchronization strategy is worked out as the following. In the proposed network, the OLT has a very high precision clock. When there is an Ethernet packet to be sent, the OLT just transmits the packet. Otherwise the OLT transmits the preamble signal. Thus the OLT plays the role of master clock and transmits clock information to the downstream ONU. Initially, when an ONU is powered, the local clock of the ONU locks the down stream clock signal transmitted from the OLT. In order to synchronize all the ONUs, the OLT polls the inactive ONUs sequentially and periodically. The polled ONU responses immediately and the OLT will estimate the round trip delay and send synchronization information back to the ONU. The ONU synchronizes the local clock and sends training packet to the OLT. The

procedure will be repeated until satisfactory synchronization is obtained.

三、 研究結果與討論

【第一年】

A. An IP over WDM Protocol with Routing and Transport Capabilities

The concept of the proposed IP over WDM protocol is not merely an IP over DWDM adaptation and encapsulation scheme, but also the concept of utilizing MPLS Label Distribution Protocol (LDP/CR-LDP) as the DWDM network signaling protocol. It also integrates MPLS traffic engineering control for optical network elements. However, each transport protocol does have its own framing and encapsulation method. In this section, we will first introduce the transmission format of the newly proposed IP over WDM protocol.

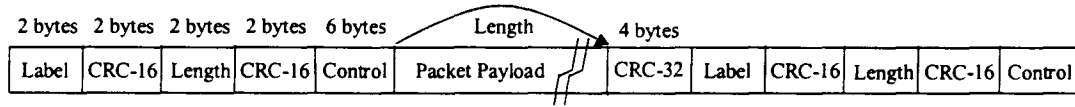


Fig. 2 Physical and Data Link Layer Frame Delineation of the Proposed IP over WDM Protocol

The transmission format of the proposed protocol is shown in Fig. 2. The frame encapsulation is done by using <Label> <CRC-16> <Length> <CRC-16> <Control> and followed by the protocol data unit of variable length and a <CRC-32> field as the trailer for payload CRC checking. As shown in Fig. 2, the “Label” field has two bytes and is used as the label switching function. The CRC-16 circuit with generating polynomial: $g(x) = x^{16} + x^{12} + x^5 + 1$, is used to check if the “Label” field is correct or not during searching process of re-framing. These bytes with “Label” field together form the framing indication of physical layer. The “Length” field together with the second CRC-16 bytes is used to indicate the length of the protocol data unit, and to be the data link layer frame delineation. Once the framer arrives at the “Sync”

state, the CRC-16 check circuit is triggered as the single-bit error correction circuit (so does at the “Post Sync” state). So, the framer just need one CRC-16 and one logic control circuit which records the status of the framer and controls the operation of CRC-16 circuit to be multiple errors detection or single-bit error correction. This simplifies the hardware of the framer a lot. “Control” field is used to indicate the type of protocol data unit, scrambler synchronization status and maintenance information. In order to avoid the inefficiency of using preamble techniques as the bit synchronization method, there must have the inter-packet fills to maintain the channel continuity. Those idle inter-packet fills contain only the necessary <Label>, <CRC-16>, <Length> and <CRC-16> fields. Maintenance packets like those to

deliver OAM messages will have additional <Control> and <CRC-32> fields. These OAM packets will periodically be inserted into optical channel with higher priority than the normal data packets.

We propose 16 bits for labels in the proposed IP over WDM protocols. To accommodate all the addressing range, i.e., 2^{20} , and incorporate the technologies of wavelength routing, we adopt 16 different wavelength, $\lambda_0, \lambda_1, \dots, \lambda_{15}$, as a group of MPLS path addresses. In that way, we can fully utilize the wavelength information while keeping electrical labels as small as possible. In this case, it is possible that we could integrate the MPLS technologies with wavelength routing. These similarities between wavelength and label have been used for combining MPLS traffic engineering control with optical cross-connects (OXC). Additional need for a General Switch Management Protocol (GSMP) or GSMP-like protocol to interface the optical router gear may also be

considered.

I. Analysis of Framing and Encapsulation Performance

To find the frame in “Sync” state at the first time that the system is powered up or at the time of re-framing after loss of frame detected, the framer is in “Hunt” state. The framer uses byte-by-byte search to find the first valid CRC-16 sequence of label field. The framing and delineation state transition diagram of the proposed IP over WDM protocol is shown in Fig. 3. The framing state transition diagram consists of four distinct states: “Hunt”, “Presync”, “Sync” and “Post Sync” states.

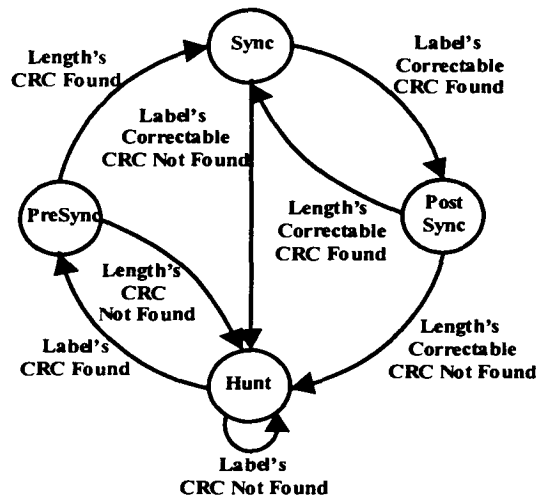


Fig. 3. Framing and Delineation State Transition

Diagram

The procedure of framing and delineation is as follows: When the framer is first powered up or in the state of re-framing, the framer tries to find out the first valid Label field in a byte-by-byte shifting and checking manner. If no valid header is found, the framer continues the shifting and check processes, i.e., staying in the “Hunt” state. Once upon finding such a valid CRC-16 sequence, the framer moves to the “Presync” state, and records the status of framing. The control logic of framer, at this state, continues to perform the second CRC-16 checking for the subsequent specific 4 bytes (Length field + second CRC-16 field) to find out the second correct CRC-16 pattern of Length field. If the framer fails in finding the second valid CRC-16 pattern after subsequent 4 bytes, it then returns back to the “Hunt” state. If it finds the second valid CRC-16 sequence, the framer goes into “Sync” state. At that point, the framer is considered in frame. The framer determines from the

Length field to find out how many bytes there are until the beginning of next Label field of next frame.

Once the framer gets into the “Sync” state, the control logic circuit of framer enables the CRC-16 error checking circuit into a single-bit error correction circuit. The framer now detects the CRC-16 sequence with no error bit and with one error bit pattern for the Label and Length field of next frame. If it does not find the next single-error correctable CRC sequence, it goes back to the “Hunt” state to perform the searching process again. If the framer detects the first correctable CRC sequence of Label field of next frame, it goes into “Post Sync” state (with CRC-16 as a single-bit error correction circuit). The framer then continues identifying the subsequent 4 bytes with CRC-16 as single-bit error correction circuit to see if they are correctable for the Length field. If not correctable, the framer goes back into “Hunt” state to perform shifting and checking process. If the framer finds out

the correctable CRC-16 sequence of the Length field, it goes into “Sync” state again and maintains the CRC-16 circuit in single-bit error correction status. This process continues until lost of frame detected and all the re-framing processes go over again. We analyze the performance of the proposed protocol as follows.

II. Probability of Loss of Frame

It is worth to note that at the receiver side, the framer is first delineated using <Label> and <CRC-16> to arrive at “Presync” state and then using <Length> and second <CRC-16> fields to get into “Sync” state. Probability of loss of frame (PLF) is directly a function of optical channel bit error rate (BER). Once the synchronization is achieved, single-bit error correction is activated. Thus, for a loss of frame, more than one error in the <Label> and <CRC-16> fields or less than one error in <Label> and <CRC-16> fields but more than one error in <Length> and second

<CRC-16> fields is required. Since all these fields are 2 bytes and the single-bit error correction is enabled, the probability of the frame loss due to the channel corruption in framing and delineation is:

$$PLF = PLF_{Label} + PLF_{Length}$$

where, PLF_{Label} is the probability that loss of frame happened in checking valid CRC-16 sequence of Label field and PLF_{Length} is the probability that loss of frame happened in checking valid CRC-16 sequence of Length field. These two probabilities are shown as follows:

$$PLF_{Label} = 1 - [(1 - P_e)^{32} + 32P_e(1 - P_e)^{31}]$$

$$PLF_{Length} = PLF_{Label}(1 - PLF_{Label})$$

where, P_e is the optical channel BER.

Fig. 4 shows the probability of loss of frame of the proposed IP over WDM protocol.

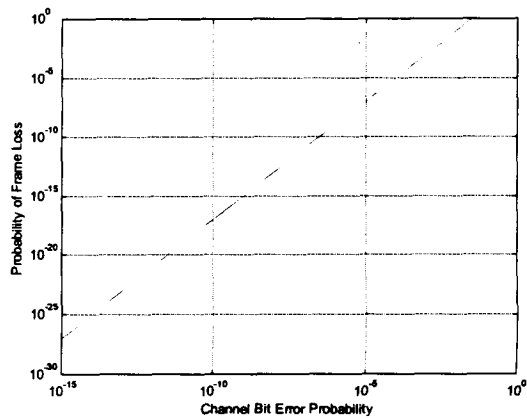


Fig 4. Probability of Frame Loss versus Channel BER.

After we obtain the PLF, we can roughly estimate the “Mean Time To Frame Loss” (MTTFL) which can be expressed as follows:

$$MTTFL = \frac{1}{PLF * FrameRate}$$

where, *FrameRate* is equal to *Channel Bit Rate / Frame Length*. However, since the transport is asynchronous and with variable packet length, we can’t get the accurate “Mean Time To Frame Loss”, unless we know the packet length distribution. We only can roughly estimate this parameter with a given packet length. Fig. 5 shows the graph of *MTTFL* (in Years) versus packet length (in Bytes) at optical channel bit rate 10Gbps and BER 10^{-12} . We can see *MTTFL* even at very small packet length distribution can still get almost a billion years!

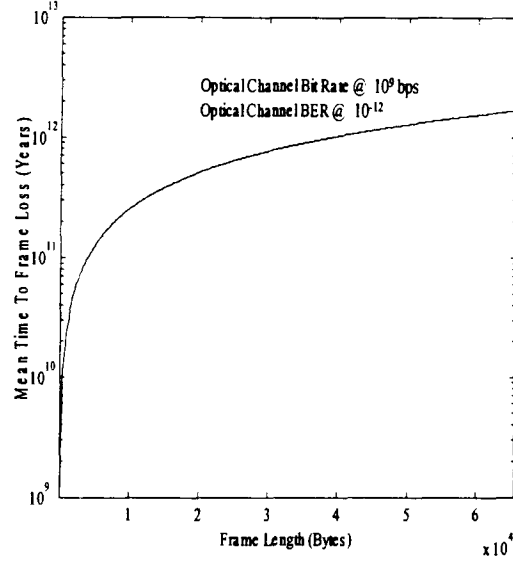


Fig. 5. Mean Time To False Frame versus Packet Length

III. Probability of False Frame

When parsing the frame byte-by-byte in search of a valid header, it is possible that the valid CRC-16 pattern is found inside the protocol data unit. This is the case that the CRC-16 header is simulated by the payload data. And the probability of false frame (*PFF*) is equal to the probability of two consecutive CRC-16 matches and is shown as follows:

$$PFF = \left(\frac{1}{2^{16}} \right)^2 = (2^{-32}) = 2.33 \times 10^{-10}$$

Even if “false framing” happens, the

framer does not stay in “Sync” state for the coming frames unless it finds more consecutive two CRC-16 matches at the right places. It could be an exact match or a match with only one bit in error. This means that the probability of falsely staying in “Sync” state is $Prob(\text{false sync}) = (33 \times 2^{-16})^2 = 25 \times 10^{-8}$. This makes the probability of rejecting false frame becomes $1 - 25 \times 10^{-8}$ and is equal to 0.99999975. From this, we can see that the probability of rejecting staying in false “Sync” state of the proposed protocol is nearly equal to 1.

IV. Mean Time To Frame

Re-frame refers to the transition from “Hunt” state to “Sync” state as shown in Fig. 3. Re-frame time, or mean time to frame (MTTF), is the duration of time required for a frame alignment circuit to re-establish frame alignment from being powered up or from a misalignment condition. A re-frame procedure consists of two separate processes: search and

confirmation. A search process sequentially searches the CRC-16 sequence for Label header to identify the beginning of each packet. If there is a match, the position is provisionally accepted as a candidate and a confirmation process is called in to test validity of the candidate position. If the confirmation criterion is met with the candidate position, frame alignment is declared. If not, a search process is reinitiated, followed by another confirmation process and so on.

1. MTTF Search Process

To consider the worst case, a search process is initiated just one byte after the correct frame alignment position. Therefore, all the n bytes in each packet, where n is variable for each packet, have to be checked. For the first $(n-1)$ byte position, a search process takes time to verify them wrong. If the payload simulates the CRC-16 sequence, a confirmation process is called in to detect its falsehood. After the falsehood

is detected, the search process is re-initiated and the next byte position is checked. At the n th byte position the search process accepts it. Using the technique of probability of generating function (PGF) in the generalized state transition diagram, we can derive the state transition diagram of search process as shown in Figure 6.

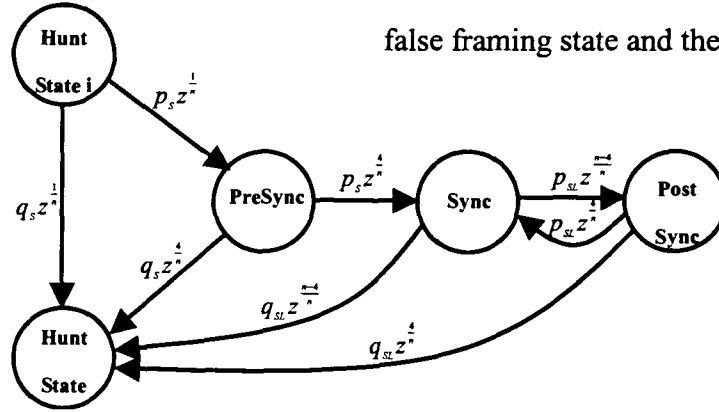


Fig. 6 Falsehood Detection Process in Search Process

The search process moves from the byte position i to $i+1$ in one byte duration. Since the proposed protocol is a packet based framing and delineation architecture, the more meaningful units in qualifying the MTTF duration is in packets. So, we denote one byte duration as $1/n$ packet, where n is the packet length in bytes. If the CRC-16 pattern of Label header is simulated, the

confirmation process takes $4/n$ packet to detect its falsehood. The probability of CRC-16 pattern is simulated by the payload is p_s , which is equal to $1/2^n$. The probability of not being simulated as the CRC-16 pattern is q_s , which is equal to $1 - p_s$. If the simulation continues happening, the framer will go into the false framing state and the probability of

staying in “Sync” state and “Post Sync” state is p_{st} which is equal to $33 \times p_s$ and the probability of back to “Hunt” state is q_{st} which is equal to $1 - p_{st}$. From Fig. 6, we then can show that the PGF of the state transition from state i to $i+1$ is given by:

$$q_s Z^{1/n} + p_s Z^{1/n} [q_s Z^{4/n} + p_s Z^{4/n} q_{st} Z^{(n-4)/n} + p_s Z^{4/n} \frac{p_{st} Z^{(n-4)/n}}{1 - p_{st} Z^{4/n}} q_{st} Z^{4/n}]$$

$$\cong q_s Z^{\frac{1}{n}} + P_s q_s Z^{\frac{4}{n}}$$

The maximum mean time to frame requires the falsehood detection for the first $(n-1)$ byte position and a correct CRC-16 pattern for the last n th byte position. Therefore if the correct CRC-16 pattern is not corrupted by optical channel bit errors, the search process takes the signal transfer function $\tau(Z)$ to the right position, where $\tau(Z)$ is given by:

$$\begin{aligned}\tau(Z) &= [q_s Z^{\frac{1}{n}} (1 + P_s Z^{\frac{4}{n}})]^{n-1} Z^{\frac{1}{n}} \\ &= q_s^{n-1} Z (1 + P_s Z^{\frac{4}{n}})^{n-1}\end{aligned}$$

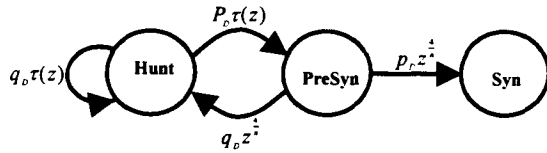


Fig. 7. Reframe Declaration Procedure

2. MTTF Confirmation Process

After the correct position of CRC-16 pattern is found, the confirmation process is called in. Fig. 7 shows the state transition diagram of the reframe declaration procedure. After the correct position is found, the probability that the

framer goes into “Presync” state is P_p .

And P_p is equal to the $(1-P_e)^n$, where P_e is the optical channel BER. The probability that the CRC-16 pattern is corrupted by the channel noise is q_o , which is equal to $1-P_p$.

The overall transfer function from the “Hunt” state to “Sync” state is given by

$$P_{RF}(Z) = \frac{P_D^2 \tau(Z) Z^{\frac{4}{n}}}{1 - q_o \tau(Z) [1 + P_D Z^{\frac{4}{n}}]}$$

which is the PGF of maximum average MTTF. Therefore, the maximum average MTTF, T_{RF} is given by

$$\begin{aligned}T_{RF} &= P'_{RF}(1) \\ &= \frac{1}{P_D^2} \left[\tau'(1) + \frac{4}{n} P_D \right]\end{aligned}$$

Fig. 8 (a) to (d) show the maximum MTTF with packet length 65535 bytes, 1500 bytes, 354 bytes and 576 bytes, respectively. The packet length of 1500 bytes is the maximum transfer unit (MTU) of Ethernet and the length of 576 bytes is the length that all routers must support.

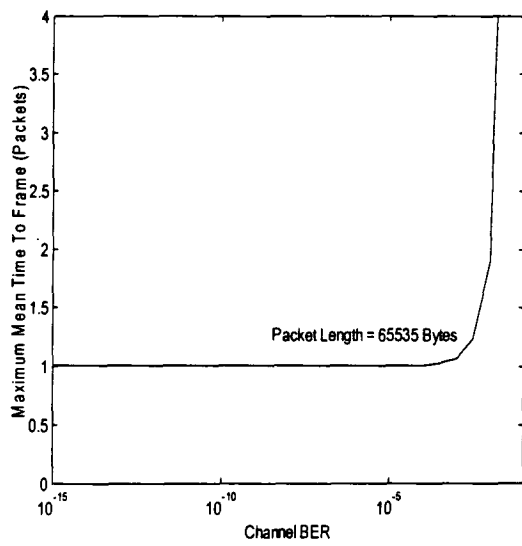


Fig. 8(a) Maximum Mean Time To Frame versus Channel BER with Packet Length 65535 Bytes

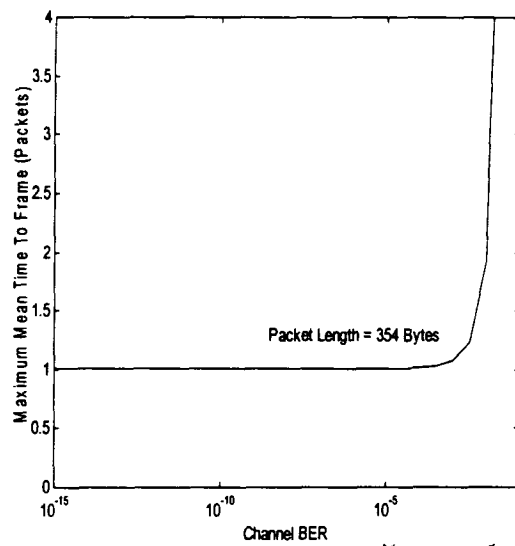


Fig. 8(c) Maximum Mean Time To Frame versus Channel BER with Packet Length 354 Bytes

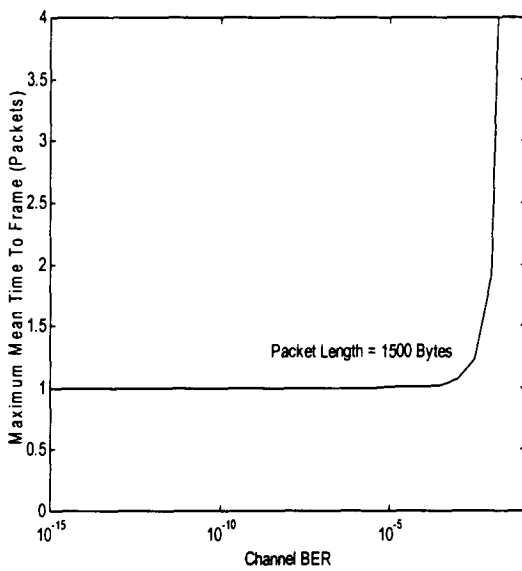


Fig. 8(b) Maximum Mean Time To Frame versus Channel BER with Packet Length 1500 Bytes

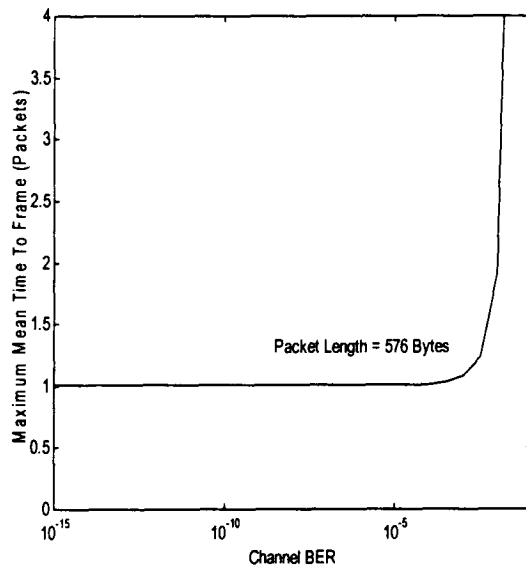


Fig. 8(d) Maximum Mean Time To Frame versus Channel BER with Packet Length 576 Bytes

From Fig. 8 (a) to (d), we can see that when optical channel BER is less than 10^{-5} , the maximum average MTTF is almost equal to one frame. This is quite straightforward, since the proposed protocol is actually in “Sync” state in less than one packet duration. The maximum average MTTF will be at most one packet to get into “Sync” state. This is much faster for the framing of variable packet length than in conventional one. Each different length packet arriving at “Sync” state is almost the same at channel BER less than 10^{-5} . We can see that although the packet length is a random variable and will affect the accuracy of actual MTTF results, we still can conclude that the MTTF will be almost less than one packet.

B. Performance Analysis of Asynchronous WDM Optical Router under Long Range Dependence Traffic

We first analyze the performance of the feed forward type asynchronous WDM optical routers under LRD traffic. In order to validate the impact of LRD traffic on the contention resolution techniques used in optical packet switching and the benefit of void filling algorithm, we have done programming in an ad-hoc discrete event simulator (DES) using raw C++ code to perform the analysis. In all the following simulations, we use an aggregated ON-OFF process modulated with a Pareto distribution to simulate LRD traffic. The Pareto distribution with cumulative distribution function (CDF), $P[X \leq x] = 1 - (b/x)^\alpha$, for $\alpha > 0$, has the degree of self-similarity, $H = (3 - \alpha)/2$, where $H \in (0.5, 1)$ and $\alpha \in (1, 2)$. All the traffic parameters for our simulations are Hurst parameter, $H = 0.9$, load of each source, $L_i = 0.8$, and the location parameter of the ON time Pareto distribution, $b = 400$ bytes. The switching parameters are 16×16

input/output ports, 16 wavelengths/port. The buffer depth is limited to 10 fiber delay lines for both of the main output buffers and common shared buffers. Wavelength conversion is activated in all input and re-circulation ports. The number of re-circulation allowed in the router is 30, which is a reasonable limit. The system void records for each output and re-circulation port are 40.

I. Feed Forward Type Asynchronous WDM Optical Router

We first validate the effect of excess load caused by asynchronous and variable length packets on optical buffers with fixed delay time, and the benefit of using void filling algorithm as shown in Fig. 9. We can see that switching performance is hardly improved by using wavelength conversion solely, which can improve the performance significantly for the synchronous and asynchronous optical packet switch under Bernoulli and bursty traffic. We also see the effect of

excess load caused by voids in the fiber delay line buffers. The longer delay line length for each fiber delay line is, the more packet contention occurs and more packets are dropped.

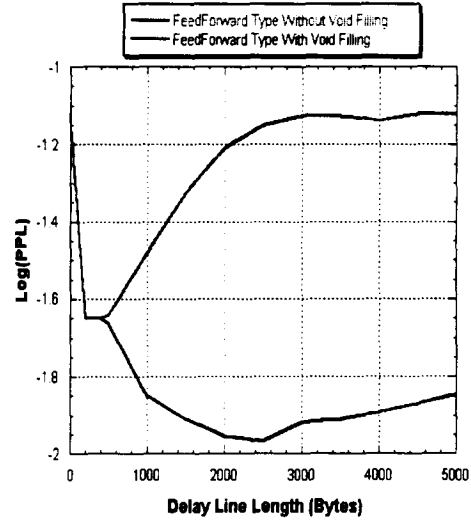


Fig. 9 Performance of feed forward type WDM optical router under LRD traffic

Fig. 10 is the performance analysis for the same router using multiple path routing (MPR). The three-category means that input packets are divided into three different types of traffic: type 1, 80% of the total traffic, not allowing to adopt MPR, type 2, 10% of the total traffic, allowing to have one chance to choose the second output port while the

original output buffers are full and type 3, 10% of the total traffic, allowing to have two chances to choose the alternate output port. On the other hand, the percentages of six-category traffic are $P_1 = 0.2\%$ for type 1, $P_2 = 30\%$ for type 2, $P_3 = 23\%$ for type 3, $P_4 = 20\%$ for type 4, $P_5 = 25\%$ for type 5 and $P_6 = 1.8\%$ for type 6. From Fig. 10, we believe that MPR is a good technique for asynchronous WDM optical router.

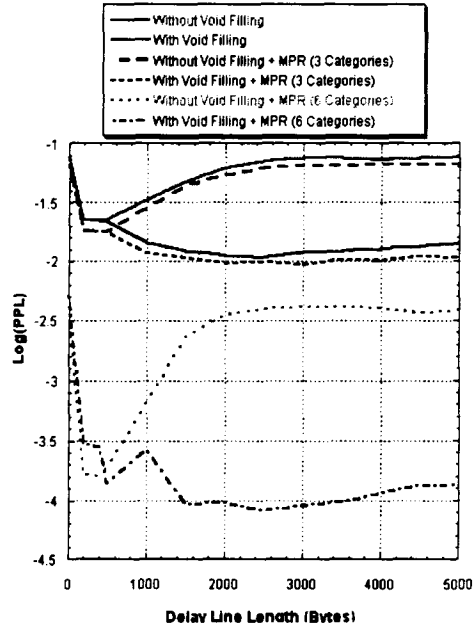


Fig. 10 Performance of feed forward type WDM optical router using MPR under LRD traffic

PSB Type Asynchronous WDM Optical Router

From our architecture of asynchronous WDM optical routers shown in Fig. 11, we force the delay line length of the re-circulation ports equal to the delay line length of the main output buffers in order to share the common fiber delay lines. In the following simulation we set the initial unit delay the same as the delay line length of fiber delay lines of the main output buffers. We simulate two extreme cases for the PSB type WDM optical routers. One is with single re-circulation port and the other is to use 16 re-circulation ports.

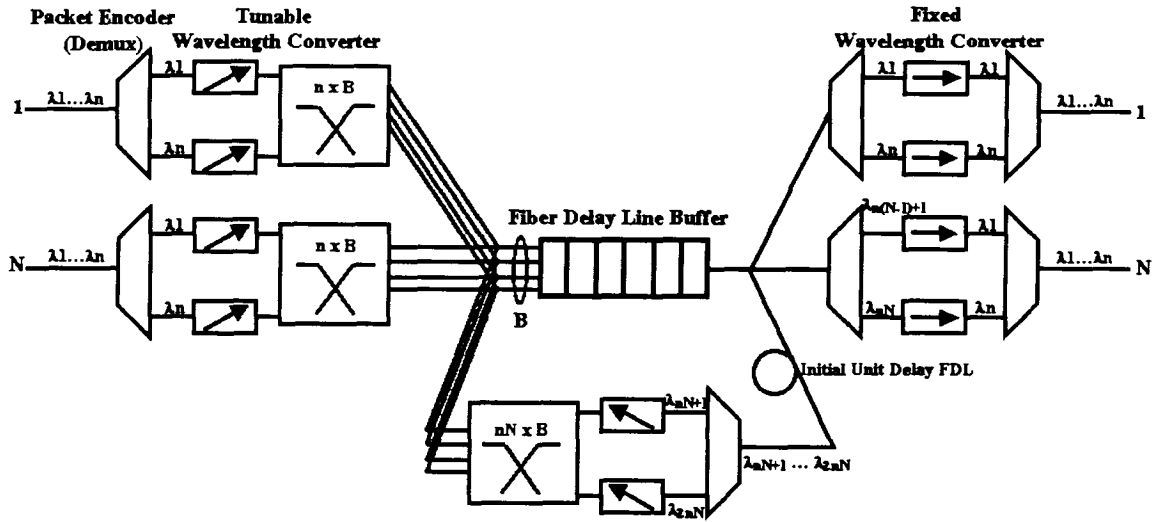


Fig. 11 Architecture of Asynchronous WDM router

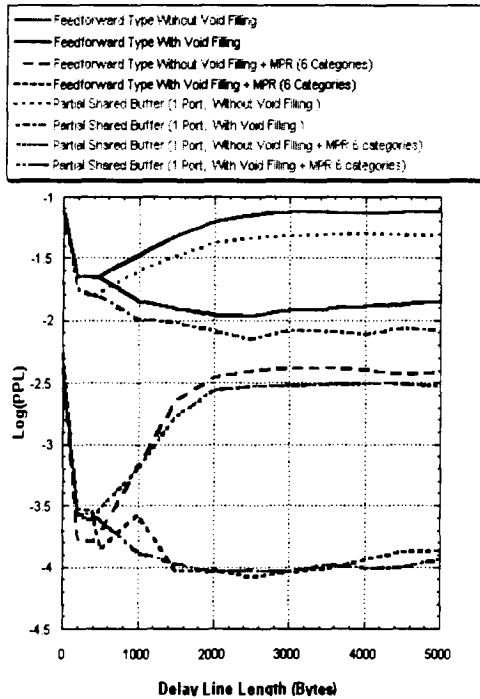


Fig. 12 Performance of feed forward type optical router versus PSB type optical router using 1 re-circulation port under LRD traffic

Fig. 12 shows the performance of feed

forward type WDM optical routers with and without activating void filling algorithm versus PSB type WDM optical routers using single re-circulation port operating at the same modes. We can see that the performance of PSB type routers using single re-circulation port looks like the performance of feed forward type routers in various operating modes. Although, the performance of PSB type routers using single re-circulation port is superior to the feed forward type routers, the improvement is only a little. The function of the re-circulated common shared buffers acts as buffers with larger

buffer depth for the feed forward type routers. The improvement of switching performance is regulated by LRD traffic. In this case, the PSB type routers using single re-circulation port can only play the role of buffer depth extension. It doesn't show any benefit of coping with asynchronous and variable length packet switching. However, during the period of packet-by-packet verification for our simulator, we find that there are two driving forces in the PSB type routers affecting the probability of packet loss. One is the increase of delay line length in the buffers. The longer the delay line length is, the higher probability of packet contention is. The other is traffic smoothing capability due to the common shared re-circulation buffer. The longer the delay line length is, the more powerful the traffic smoothing capability is. These two driving forces cancel the effect of each other on the probability of packet loss due to LRD traffic.

In order to compare this phenomenon

thoroughly, we analyze another extreme case of the PSB type routers using 16 re-circulation ports and without using void filling algorithm in Fig. 13. To compare with various operation modes in feed forward type routers, we depict the performance of 16 re-circulation ports PSB type routers without using void filling and with MPR activated in the same figure. We see that the PSB type WDM optical routers using 16 re-circulation ports without void filling can have comparable performance as the feed forward type routers without void filling but with MPR activated in the region of longer delay line length. We also find the same result in Fig. 13, for the performance comparison between the PSB type routers using 16 re-circulation ports with MPR activated but without void filling and the feed forward type routers with both void filling algorithm and MPR capability activated in the region of longer delay line length. Surprisingly, we find that the performance of the PSB type WDM

optical routers using 16 re-circulation ports has the similar properties with the feed forward type routers with void filling algorithm activated for the delay line length in both of with and without MPR activated modes.

This phenomenon is not straightforward as the void filling algorithm. We observe that during the period of packet-by-packet verification for our simulator, the driving force from the traffic smoothing is more powerful than the increase of packet contention probability resulted from the longer delay line length of the fiber delay line buffers. When there are more packets coming into the routers by using larger delay line length, it will cause more packets dropped in feed forward type routers. If we use shorter delay line length, the packet will come out from the re-circulation buffers more quickly. However, since the input traffic is LRD, it means that the traffic has a heavy tailed probability of having very long packet length. These long packets will

occupy the output buffers for a long time. Then, the packets coming out from the re-circulation buffers have to re-circulate in the routers again and again. This will cause higher packet loss probability. These results show that we can use the PSB type router with suitable re-circulation ports to handle asynchronous and variable length packets without using the computing-consuming void filling algorithm at the expense of system complexity.

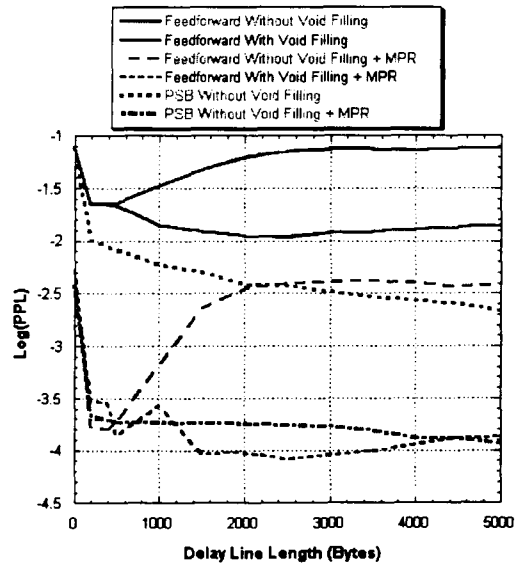


Fig. 13 Performance of PSB type WDM optical router with MPR activated under LRD traffic

A. Unicast Switch

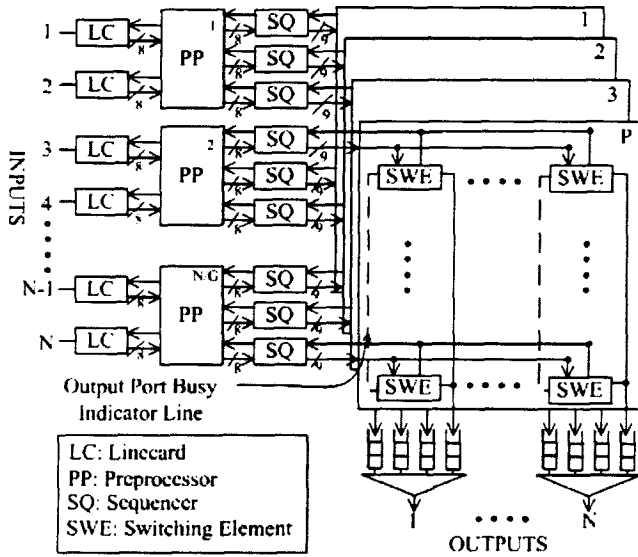


Fig. 1. The architecture of the NTU switch.

Fig. 1 depicts the architecture of the proposed optical switch, called NTU Switch. There are N input ports (output ports). The transmission speed of an input line and an output line is the same as that of the switching fabric. A header and a trailer will be attached to each incoming IP packet in linecards. The output port address, which is assigned according to the IP address, is placed in the header. The trailer acts as an ending delimiter. We let every G input ports of the switch group together to share a

preprocessor. The number of output lines of the preprocessor is $G \cdot R$, where R is an expansion ratio. An output line of the preprocessor is connected to an input line of the sequencer followed by the crossbar switching planes. We denote the number of planes and the number of sequencers of each plane by P and NR/P respectively. Each crossbar switching plane consists of $N^2 R/P$ switching elements, which form an $(NR/P) \times N$ matrix. The structures of all switching elements are identical except with differently stored output port addresses. The outputs of switching planes are connected to N multiplexers. Each multiplexer has P inputs and single output.

I. Preprocessor

Each preprocessor accepts the packets from a group of G input ports and distributes them to appropriate sequencers associated with the preprocessor. The bit stream of an incoming packet arrives on an eight-bit

word bus. The sequencer will append a control bit to this word to form a nine-bit word before sending this packet into the crossbar switching plane. The control bit of a header and a trailer is set to 1. For a data byte, it is set to zero. The packet format with nine-bit words is illustrated in Fig. 2. Note that the format of the trailer is identical with that of the header but it consists of an invalid output port address.

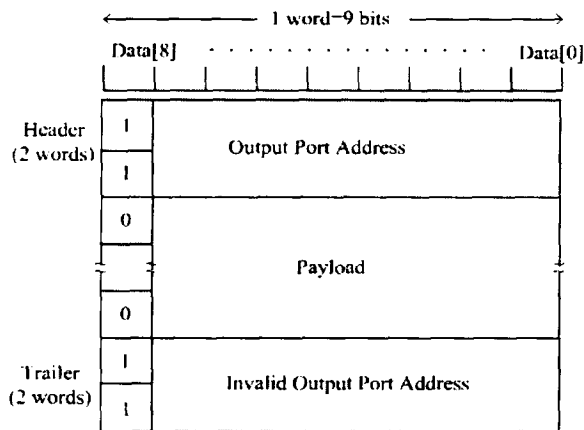


Fig. 2. The packet format of the NTU switch.

II. Sequencer

The sequencers associated with the same plane send the packets into the plane in a round robin fashion. All sequencers can start to send the packets

only at pre-assigned time points, which are illustrated in Fig. 3, where τ is the propagation delay of an output port busy indicator signal in the plane from the first switching element to the last switching element in the same column. When the packet is successfully transmitted to the destination port via a certain switching element, the switching element will send a feedback signal to inform the sequencer continuing the transmission until finished. Otherwise the sequencer will retransmit the packet at the next allowable entering time point. The sequencer may buffer more than one packet. If a packet is not successfully transmitted after several attempts, the sequencer may send the next packet waiting in the buffer to the switch plane at the following allowable entering time points.

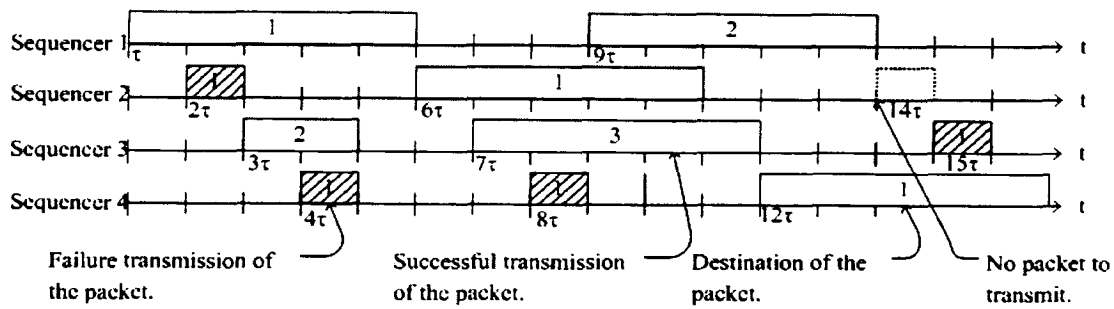


Fig. 3. The timing diagram of the sequencers in the same plane. We show the case with 4 sequencers.

III. Crossbar Switching Plane

The switching fabric is composed of P planes. For each plane, the switching elements in the same column corresponding to a single output port are connected to the output buffer via a vertical bus and an output port busy indicator line. The switching elements in the same row are connected to the horizontal bus and a feedback signal line associated with the sequencer. The schematic diagram of one switching element is shown in Fig. 4. It has two states: cross and bar. Initially, all the switching elements are in the bar state. The feedback signal and the output port busy indicator signal are idle. The switching element always listens to the

horizontal bus. The control bit 1 will activate the correlator to examine the address in the nondata word. If the two-word packet address matches the prestored output port address and the output port busy indicator signal is idle, then the switching element changes its state to cross and activates its connector to route the packet to the output port. At the same time it sends a feedback signal to the sequencer and sets the output port busy indicator. When any nondata words, which do not match the output port address, are detected, the switching element releases its connector and resets the output port busy indicator line and the feedback signal line idle. Simultaneously the state is changed back to its initial bar state. Because the trailer

of a packet is designed to consist of an invalid output port address, it will always reset the switching element to the initial state and therefore finish transmission of the current packet.

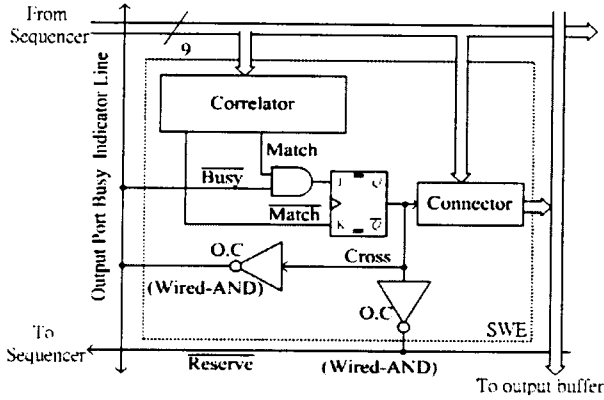


Fig. 4. The block diagram of the switching element.

IV. Analysis of the crossbar switching plane

We analyze a single crossbar switching plane first because it is the place where blocking of the packet streams occurs. Then we use the result of this analysis to evaluate the performance of the overall system.

1. Model Assumptions

In an $M \times N$ crossbar switching

plane, each sequencer has a buffer with storage capacity of K packets, including the transmitting one. The system time is divided into slots, and M slots are grouped into one frame. Each slot is equal to τ and set to one. Each sequencer is assigned permanently to a slot in the frame. The backlogged sequencer can send one packet in its assigned slot to the plane to perform self-routing switching. The packet arrival process of each sequencer is i.i.d. (independent and identically distributed)

Bernoulli process with rate σ and the length of each packet is i.i.d. geometric distributed with mean μ^{-1} slots. Each packet chooses one of the N output ports as its destination with equal probability. The late arrival model with delayed access is used in our analysis and simulation [8]. We use delayed replacement rule when a packet arrives in the last slot of a packet transmission time and the buffer is full, that is, the sequencer does not accept such a packet and packet loss occurs [9].

2. State Definition

The modes of a sequencer are: idle, backlogged, and transmission. The sequencer with empty buffer is in the idle mode, and it is in the transmission mode while transmitting a packet. The backlogged sequencer has packets in its buffer and is waiting for its assigned slot to start self-routing. In addition, the number of packets in the buffers is necessary for the sequencer to determine whether to accept the arriving packets or not. Moreover, after a packet departure, the mode of the sequencer is determined by its buffer status. Hence, we classify the sequencers as idle, backlogged or transmission with k packets in its buffer, where $k = 1, 2, \dots, K$.

Packet arrival and departure in each sequencer occur during one slot with the probabilities σ and μ respectively. However, at the beginning of one slot, only the assigned sequencer can start self-routing. The self-routing sequencer will start transmitting if it doesn't

collide with others, otherwise, it will enter backlogged mode. For convenience, we number all the sequencers from 1 to M and rotate the number with progress of slots, i.e., in the next slot, Sequencer i in the current slot will become Sequencer $i+1 \pmod{M}$. Without loss of generality, the current slot is assigned to Sequencer 1 which can start self-routing if it has packets waiting in its buffer. We observe the system at the beginning of slot t and define the state vector as $\mathbf{V}^{(t)} = (\mathbf{V}_1^{(t)}, \dots, \mathbf{V}_j^{(t)}, \dots, \mathbf{V}_M^{(t)})$, where $\mathbf{V}_j^{(t)} = (i_j^{(t)}; b_{j,1}^{(t)}, \dots, b_{j,K}^{(t)}; t_{j,1}^{(t)}, \dots, t_{j,K}^{(t)})$ and $i_j^{(t)} = \Pr[\text{Sequencer } j \text{ is idle in the time slot } t]$, $b_{j,k}^{(t)} = \Pr[\text{Sequencer } j \text{ is backlogged with } k \text{ packets in its buffer in the time slot } t]$, $t_{j,k}^{(t)} = \Pr[\text{Sequencer } j \text{ is transmitting packets with } k \text{ packets in its buffer in the time slot } t]$. For Sequencer j , we have

$$i_j^{(t)} + \sum_{k=1}^K b_{j,k}^{(t)} + \sum_{k=1}^K t_{j,k}^{(t)} = 1 \quad \forall j = 1, 2, \dots, M.$$

Note that we have defined a multi-dimensional Markov chain $\{\mathbf{V}^{(t)}, t = 1, 2, \dots, \infty\}$.

The stochastic behaviors of Sequencers 1, ..., M are identical and independent. Therefore, we can determine the state probabilities of Sequencer 2, ..., M in the next slot given the state probabilities of Sequencer 1, ..., $M - 1$ in the current slot:

$$\mathbf{V}_j^{(t+1)} = \mathbf{V}_{j-1}^{(t)} \cdot \mathbf{C}, \quad \forall j = 2, 3, \dots, M$$

where $\mathbf{C}$ is the routing probability matrix. Because self-routing is possible for Sequencer 1, and the probability of successful switching is determined by the modes of the transmitting sequencers, we have

$$\mathbf{V}_1^{(t+1)} = (1-s) \mathbf{V}_M^{(t)} \cdot \mathbf{D}_1 + s \mathbf{V}_M^{(t)} \cdot \mathbf{D}_0$$

where $s = \text{Pr}[\text{self-routing is successful}]$, and $\mathbf{D}_1(\mathbf{D}_0)$ is the routing probability matrix given that self-routing fails

(succeeds). The collision will occur if the self-routing packet is destined for an output port to which one of Sequencer 2, ..., M is transmitting. Based on the assumption of uniform traffic and using the law of total probability, we have

$$s = \sum_{m=0}^{M-1} T_m \left(1 - \frac{1}{N}\right)^m,$$

where $T_m = \text{Pr}[\text{There are } m \text{ transmitting sequencers among Sequencer } 2, \dots, M]$.

When the buffer of a sequencer can store only one packet, the corresponding routing matrices are given as follows.

$$\mathbf{C} = \begin{pmatrix} 1-\sigma & \sigma & 0 \\ 0 & 1 & 0 \\ \mu & 0 & 1-\mu \end{pmatrix},$$

$$\mathbf{D}_1 = \begin{pmatrix} 1-\sigma & \sigma & 0 \\ 0 & 1 & 0 \\ \mu & 0 & 1-\mu \end{pmatrix},$$

$$\mathbf{D}_0 = \begin{pmatrix} 1-\sigma & 0 & \sigma \\ 0 & 0 & 1 \\ \mu & 0 & 1-\mu \end{pmatrix},$$

$$\mathbf{C} = \begin{pmatrix} \mathbf{C}_{11}[p, q] & 0 \\ \mathbf{C}_{21}[p, q] & \mathbf{C}_{22}[p, q] \end{pmatrix},$$

$$\mathbf{D} = \begin{pmatrix} \mathbf{D}_{11}[p, q] & \mathbf{D}_{12}[p, q] \\ \mathbf{D}_{21}[p, q] & \mathbf{D}_{22}[p, q] \end{pmatrix}$$

where

$$\mathbf{C}_{11}[p, q] = \begin{cases} 1 - \sigma, & \text{if } p = q, 1 \leq p \leq K, \\ \sigma, & \text{if } p = q - 1, 1 \leq p \leq K, \\ 1, & \text{if } p = q = K + 1, \\ 0, & \text{otherwise.} \end{cases} \quad \mathbf{C}_{21}[p, q] = \begin{cases} \mu(1 - \sigma), & \text{if } p = q, 1 \leq p \leq K - 1, \\ \mu\sigma, & \text{if } p = q - 1, 1 \leq p \leq K - 1, \\ \mu, & \text{if } p = q = K, \\ 0, & \text{otherwise.} \end{cases}$$

$$\mathbf{C}_{22}[p, q] = \begin{cases} (1 - \sigma)(1 - \mu), & \text{if } p = q, 1 \leq p \leq K - 1, \\ \sigma(1 - \mu), & \text{if } p = q - 1, 1 \leq p \leq K - 1, \\ 1 - \mu, & \text{if } p = q = K, \\ 0, & \text{otherwise,} \end{cases}$$

$$\mathbf{D}_{11}[p, q] = \begin{cases} 1 - \sigma, & \text{if } p = q = 1, \\ R(1 - \sigma), & \text{if } p = q, 2 \leq p \leq K, \\ R, & \text{if } p = q = K + 1, \\ R\sigma, & \text{if } p = q - 1, 1 \leq p \leq K, \\ 0, & \text{otherwise.} \end{cases} \quad \mathbf{D}_{12}[p, q] = \begin{cases} \sigma(1 - R), & \text{if } p = q, 1 \leq p \leq K, \\ (1 - \sigma)(1 - R), & \text{if } p = q + 1, 2 \leq p \leq K, \\ 1 - R, & \text{if } p = K + 1, q = K, \\ 0, & \text{otherwise.} \end{cases}$$

$$\mathbf{D}_{21}[p, q] = \begin{cases} \mu(1 - \sigma), & \text{if } p = q = 1, \\ R\mu(1 - \sigma), & \text{if } p = q, 2 \leq p \leq K - 1, \\ R\mu, & \text{if } p = q = K, \\ R\mu\sigma, & \text{if } p = q - 1, 1 \leq p \leq K - 1, \\ 0, & \text{otherwise.} \end{cases}$$

$$\mathbf{D}_{22}[p, q] = \begin{cases} (1 - \sigma)(1 - \mu) + \mu\sigma(1 - R), & \text{if } p = q, 1 \leq p \leq K - 1, \\ 1 - \mu, & \text{if } p = q = K, \\ \mu(1 - \sigma)(1 - R), & \text{if } p = q + 1, 2 \leq p \leq K - 1, \\ \mu(1 - R), & \text{if } p = K, q = K - 1, \\ \sigma(1 - \mu), & \text{if } p = q - 1, 1 \leq p \leq K - 1, \\ 0, & \text{otherwise,} \end{cases}$$

where σ and μ are defined in the assumption of traffic model. According to the

definition, $\mathbf{D}_1(\mathbf{D}_0)$ is equal to $\mathbf{D}$ with $R = 1(0)$.

V. Limiting Probabilities

We assume the system is in the steady state and denote the steady state vector by $\mathbf{V}$. Using the above equations, we have

$$\mathbf{V}_j = \mathbf{V}_1 \cdot \mathbf{C}^{j-1}, \forall j = 2, \dots, M,$$

and

$$\mathbf{V}_1 = (1-s)\mathbf{V}_M \cdot \mathbf{D}_1 + s\mathbf{V}_M \cdot \mathbf{D}_0.$$

The steady state probabilities are in terms of $\mathbf{V}_1$, which satisfies a nonlinear equation

$$\mathbf{V}_1 = \mathbf{V}_1 \cdot \mathbf{C}^{M-1} \cdot [(1-s)\mathbf{D}_1 + s\mathbf{D}_0]$$

We define the generating function[10] to count the events that m sequencers are transmitting packets: $S(x)$

$$= \prod_{j=2}^M \left[1 + (x-1) \sum_{k=1}^K t_{j,k} \right].$$

In $S(x)$, we have counted the probabilities of transmission for Sequencer j , i.e., $t_{j,k}$ by associating them with x . Hence, T_m is equal to the coefficient of x^m in $S(x)$ and hence $S^{(m)}(0)/m!$. Because $\mathbf{C}$ is a stochastic matrix, we have

$$i_1 + \sum_{k=1}^K (b_{1,k} + t_{1,k}) = 1$$

Finally, we can solve for $\mathbf{V}_1$ by using Newton's Method for systems of nonlinear equations.

VI. Analysis of performance metrics

1. Definition

We denote the throughput of the switch by γ , which is defined as the average utilization factor per output port.

The generating function which counts the number of sequencers in the transmission mode is

$$\Gamma(x) = \prod_{j=1}^M \left\{ \psi_j + \sum_{k=1}^K (b_{j,k} + x \cdot t_{j,k}) \right\}.$$

The coefficient of x^m in $\Gamma(x)$ is the probability that there are m sequencers in

transmission mode. Hence the expected number of transmitting sequencers

is $\Gamma'(1)$. After using the normalization

$$\text{condition, we obtain } \gamma = \frac{\Gamma'(1)}{N} = \sum_{j=1}^M \sum_{k=1}^K \frac{t_{j,k}}{N}.$$

Because of conservation of the traffic

flow, we have $\frac{\sigma}{\mu}(1-\beta) = \gamma \frac{N}{M}$ in the

steady state, where we denote β as the

blocking probability. Hence,

$$\beta = 1 - (\gamma\mu\sigma/\sigma M)$$

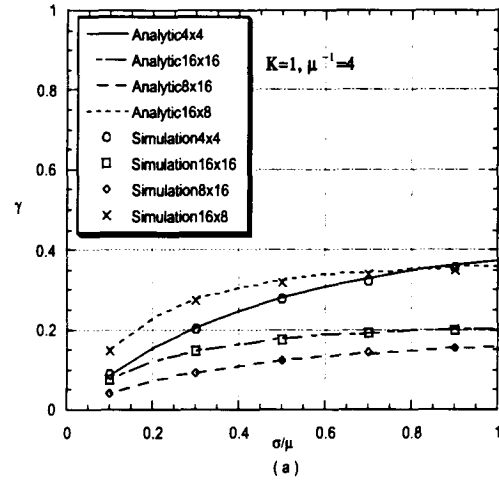
We define the delay experienced by each packet in the sequencer as the total time it spends from the instant it arrives at the sequencer buffer to the instant when it finishes transmission. The associated generating function for the system size is $Q(x) = \prod_{j=1}^M \left\{ i_j + \sum_{k=1}^K (b_{j,k} + t_{j,k}) x^k \right\}$. The coefficient of x^m in $Q(x)$ is the probability that there are m packets in the system in the steady state. Therefore the expected systems size is $Q'(1)$. Using the well-known Little's Law [10], we obtain the packet delay as follows

$$T = Q'(1) / N\gamma\mu = \sum_{j=1}^M \sum_{k=1}^K k(b_{j,k} + t_{j,k}) / N\gamma\mu$$

2. Validation of the Model

We use a stochastic self-driven discrete event simulation to verify the correctness of our analytic model. The assumption of simulation is identical with the analytic model. We measure the

time average utilization per output port as the throughput. The delay is measured by averaging over all the packets. Finally, the blocking probability is defined as the ratio of the total number of discarded packets due to overflow of the buffers to the total number of arriving packets. Fig. 5 plots the comparison of simulation results with the analytic model. Both the unbuffered case ($K=1$) and general case ($K=4$) are verified. It can be seen that the results obtained through the analytic model agree with simulation very well.



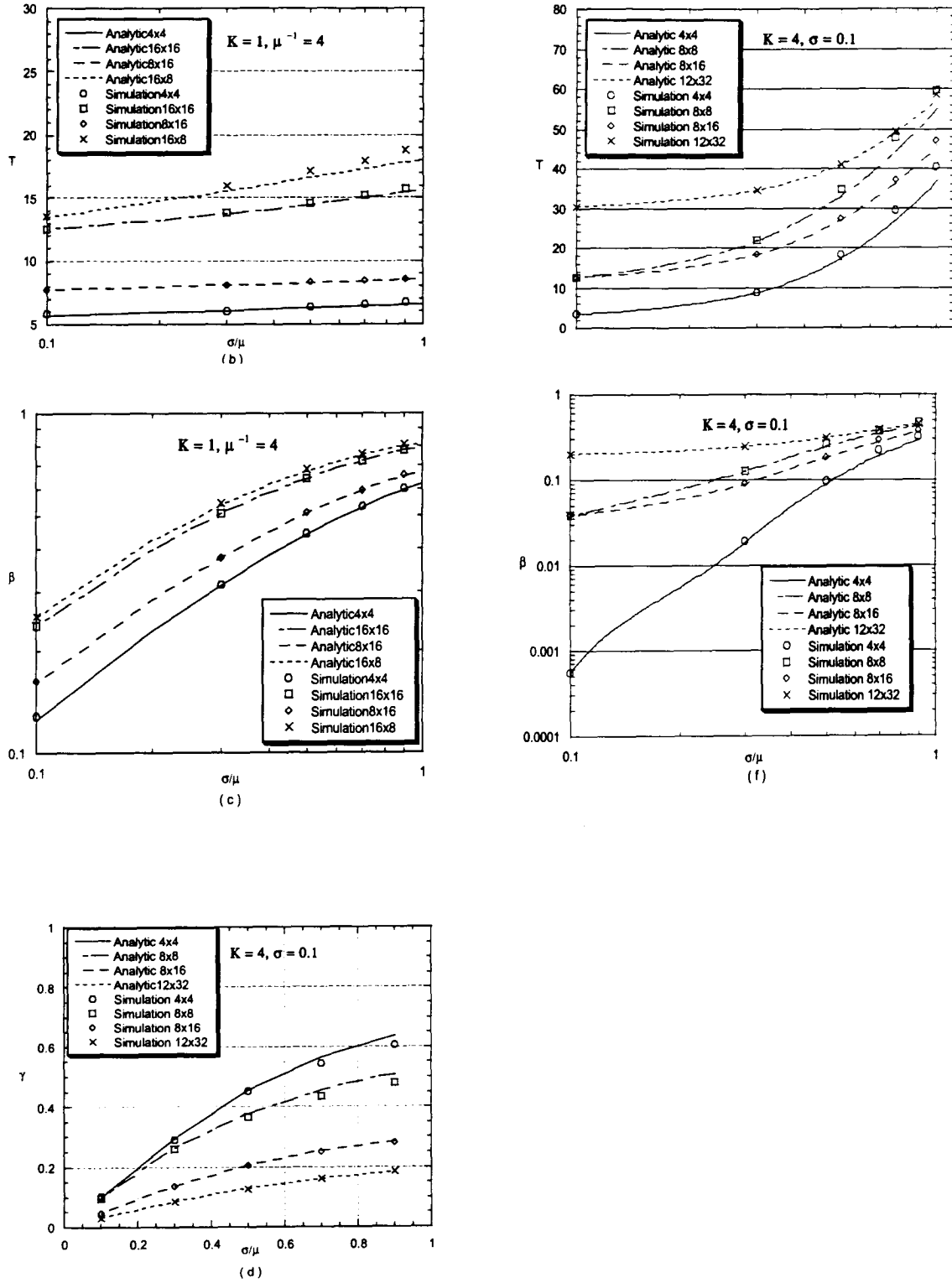


Fig. 5. Comparison of the simulation and analytic result for (a)throughput, $K=1, \mu^{-1}=4$, (b)delay, $K=1, \mu^{-1}=4$, (c)blocking probability, $K=1, \mu^{-1}=4$, (d)throughput, $K=4, \sigma=0.1$, (e)delay, $K=4, \sigma=0.1$, (f)blocking probability, $K=4, \sigma=0.1$.

VII. Performance of the multiplane switch

We assume that the packet arrival processes at the linecards are i.i.d. Bernoulli process with rate λ . The incoming packet will be sent to the connected preprocessor, where it chooses one of the connected sequencers with the same probability. Hence, the arrival rate of the Bernoulli process at each sequencer σ is equal to λ/R , where R is the expansion ratio.

1. Throughput for Infinite Buffer

Let us assume for the moment that the storage capacities of input and output buffers are infinite. Thus, we are considering a no-loss system. We denote the throughput per output port of the crossbar switching plane by γ . The overall switch throughput per output

port is the aggregate throughput of the outputs of the planes

$$\gamma_o = \sum_{l=1}^P \gamma_l = P\gamma. \text{ We are interested}$$

in the maximum throughput $\gamma_{\max}$ which occurs at saturation. When the system is at saturation, there are always packets waiting in the buffer of the sequencer to be sent. Thus, there are no idle modes and the state vector $\mathbf{V}$ equals $(b_{1,1}, \dots, b_{M,1}; t_{1,1}, \dots, t_{M,1})$. We have the

routing probability matrix

$$\mathbf{C}^{(s)} = \begin{pmatrix} 1 & 0 \\ \mu & 1 - \mu \end{pmatrix}$$

$$\text{and } \mathbf{D}^{(s)} = \begin{pmatrix} R & 1 - R \\ R\mu & 1 - \mu + \mu(1 - R) \end{pmatrix}. \text{ The}$$

associated generating function with

$$\text{throughput is } S(x) = \prod_{i=2}^M [1 + (x-1)t_{i,1}].$$

Then we can use the state transition equations and the normalization

$$\text{condition } \sum_{i=1}^M (b_{i,1} + t_{i,1}) = 1 \text{ to solve for}$$

the steady state vector $\mathbf{V}$. Finally, the

$$\text{maximum throughput is } \gamma_{\max} = \frac{P}{N} \sum_{j=1}^M t_j.$$

Fig. 6 shows the maximum throughput with different expansion ratio R and number of planes P . We see that the HOL blocking is alleviated by increasing the number of planes and expansion ratio. Moreover, the mean packet length will also affect the maximum throughput. For a 32×32 switch, we see that 100% throughput is attained with four planes and expansion ratio 1.5 if the mean packet length is larger than 20τ . This is because when the packet length is larger compared to the frame duration, the fraction of time the sequencer spends in waiting its turn to perform self-routing will be smaller. However, the large packet will also block the transmission of the other packets destined for the same output port. Hence, the gain of throughput will decrease as the packet length increases.

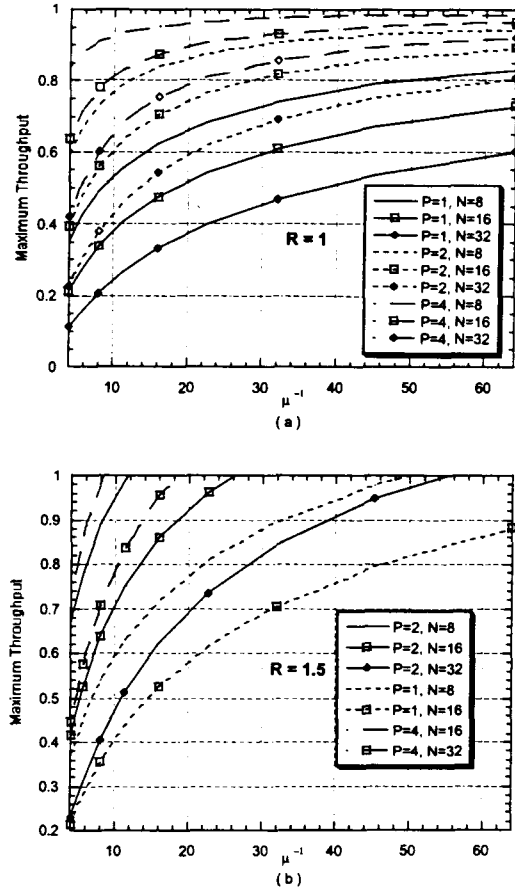


Fig. 6. The maximum throughput for infinite buffer for (a) $R=1$, (b) $R=1.5$.

2. Blocking Probability

In this and next subsection, we consider a loss system. Assume that the buffer capacity in each sequencer is K packets and the output buffer of each output port can store K_o packets. An output queue for each output port is shared by all P planes and no backpressure mechanism is used. We model the output buffer by a discrete time $Geom/Geom/1/K_o$ queueing

system. The offered load of the output buffer is $P \cdot \gamma$, and the departure rate is μ . We can use a Markov chain analysis to derive the state probability of the $Geom/Geom/1/K_o$ system [11]

$$P_n = \frac{P_0}{1 - \mu} \left(\frac{\lambda_o (1 - \mu)}{\mu (1 - \lambda_o)} \right)^n, \quad n = 1, \dots, K_o,$$

and

$$P_0 = (\mu - \lambda_o) / \left[\mu - \lambda_o \left(\frac{\lambda_o (1 - \mu)}{\mu (1 - \lambda_o)} \right)^{K_o} \right],$$

where $\lambda_o = P \cdot \gamma \cdot \mu$.

Because of the BASTA property of the arrival process [12], the blocking probability at the output buffer (P_{B2}) is equal to the probability of state K_o , that is, $P_{B2} = P_{K_o}$. Using conservation of packet flows for an $N \times N$ switch, we have $\lambda/\mu (1 - P_B) = P\gamma(1 - P_{B2})$, where P_B is the overall blocking probability. Therefore, we have
$$P_B = 1 - \frac{P \cdot \gamma \cdot \mu}{\lambda} (1 - P_{B2}).$$

Fig. 7 shows the blocking probability of a 16×16 switch, where we denote by L the mean packet length.

We find that the blocking probability is almost the same when the packet length is 10 or 100 for $R=1.5$ and $P=4$. This is because when packet length is larger than 8 the maximum throughput of a 16×16 switch has attained 100% if $R=1.5$ and $P=4$, and hence the performance of the switch doesn't change greatly with the packet length. The blocking probability is more sensitive to the change of output buffer capacity than that of input buffer capacity. Increasing the capacity of the output buffer will obtain more gain in blocking probability than increasing the capacity of the input buffer. However, the more the input buffer capacity is, the more the gain is obtained when we increase the output buffer capacity. There exists an upper bound beyond which the gain doesn't increase especially in the light load. Finally, we see that the maximum throughput with infinite buffer capacity can be attained if we increase both the capacities K and K_o . When $R=1$ and $P=2$, the

switch is input buffer limited and therefore the gain of the blocking probability doesn't increase as the output buffer capacity increases when the input buffer capacity is small. The blocking probability will be smaller when the mean packet length is 100. Comparing with the case $R = 1.5$ and $P = 4$, we see that the more alleviation of blocking effect using more planes and larger expansion ratio.

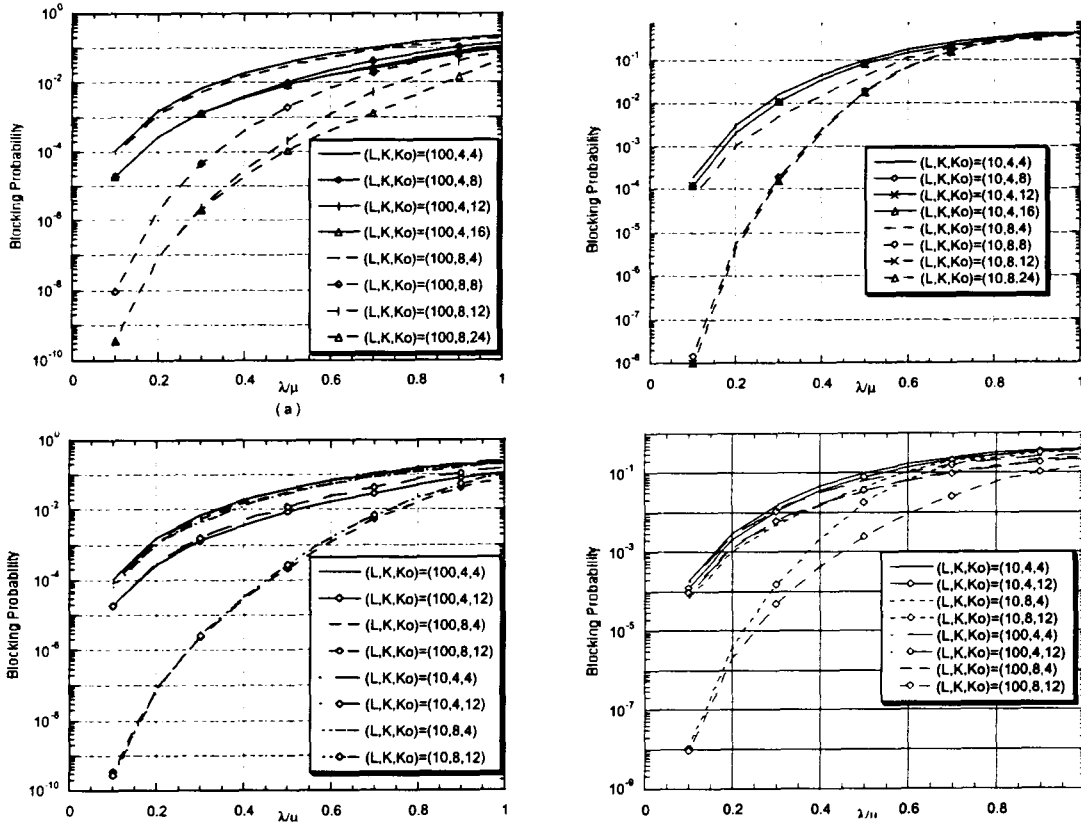


Fig. 7. The blocking probability of a 16×16 NTU switch for (a) $P=4, R=1.5, L=100$, (b) $P=4, R=1.5, L=100$ or 10 , (c) $P=2, R=1, L=10$, (d) $P=2, R=1, L=100$ or 10 .

3. Waiting Time

The waiting time is defined as the time interval from the time an arbitrary packet enters the buffer to the time the packet leaves the buffer.

Storing-and-forwarding switching is used in our model, hence the minimum waiting time is twice the packet length. We denote the mean waiting time of the switch by W . Clearly, $W = T + T_o$, where T is the mean waiting time spent in the sequencer buffer and T_o is the mean waiting time spent in the output buffer. T_o is obtained by using Little's Law:

$$T_o = \sum_{n=1}^{K_o} n P_n / \lambda_o (1 - P_{B2})$$

Fig. 8 shows the mean waiting time of packets in units of mean packet length in a 16×16 switch. We see that packets with mean length of 100 will spend less time in units of mean packet length waiting than to the packets with mean length of 10. The difference in the behavior of the waiting time is more

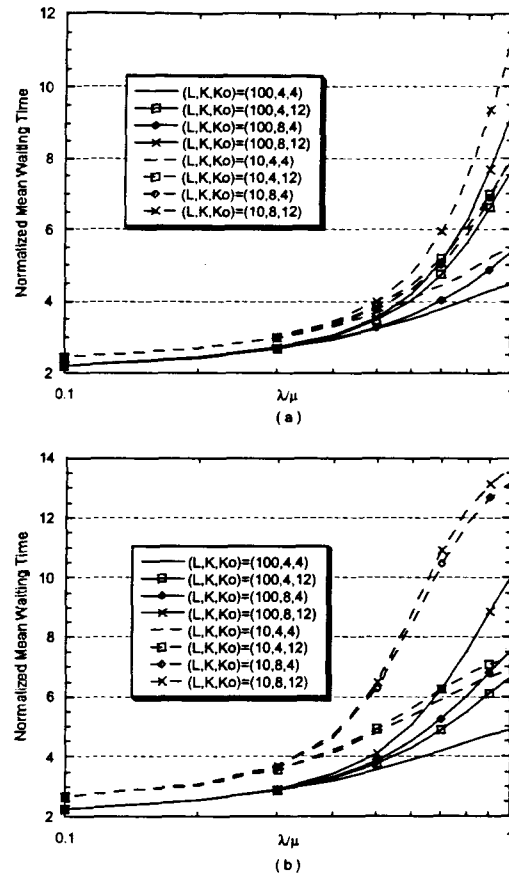


Fig. 8. The normalized mean waiting time of a 16×16 switch for (a) $P=4$, $R=1.5$, (b) $P=2$, $R=1$.

obvious when the switch is input buffer limited, i.e., $R = 1$ and $P = 2$. Therefore, it is more efficient to operate in longer packet length when the switch is input buffer limited. Note that when the switch is output buffer limited, that is, $R = 1.5$ and $P = 4$, increasing the input buffer capacity will obtain much gain in the blocking probability if the sufficient output buffer capacity is used. This is attained at the expense of increase in the mean waiting time.

VIII. CONCLUSION

We analyze the queueing system of the crossbar switching plane, where blocking of packet streams occurs. The maximum throughput, blocking probability, and waiting time of the overall system are obtained based on this analysis. As seen in the analysis, although the input port expansion and multiplane structure increase the switch complexity a little, it will significantly

improve the throughput. Our result shows that a reasonably sized NTU switch with four planes and 1.5 expansion ratio will have 100% maximum throughput, which matches OQ switches. Moreover, the result shows that the blocking probability and waiting time of the switch is output buffer limited when enough planes and expansion ratio are used.

B. Multicast Switch

There are many multicast switches proposed and analyzed in the literature [13]. However, a little is done about the multicast scheduling of variable length packets. In this report we assume that an incoming packet has multiple destinations and is of variable length without being fragmentized into cells. Hence a new architecture must be designed to solve output contention among all the multicast destinations, to replicate the packets and finally to route these copies to their destinations

respectively. The complication arises because the availability of outputs is of random nature.

Multicast queueing disciplines were discussed in [14] for fixed length packets. It was shown that the multicast service with Fanout splitting has better throughput and delay performance. However, the performance of a switch capable of handling variable length packets is not discussed. In this report, we assume a random HOL service with Fanout splitting. Multiple copies of an

incoming packet can be served at the same time, depending on the availability of the destinations. The HOL packet in an input is completely served if all its copies are transmitted. Then the next fresh packet in the queue can move into the HOL position. It will be shown that the maximum throughput depends on not only the mean of the Fanout distribution as expected, but also on a quantity termed Fanout Function, whose characteristics will be surveyed in detail. The mean waiting time will also be derived.

The HOL queueing discipline for multicast packets with variable length is more complicated than that for fixed length packets because of the asynchronous nature of output availability. Hence, it is difficult to schedule the routing of all the multicast copies by a central arbitrator. Instead, distributed packet replication, decentralized output contention resolution and self-routing are very

desirable. In this report we propose a multicasting switch based on the previous architecture to provide multicast switching function with random HOL service and Fanout splitting. This switch, which is a combined input output queue switch, can achieve 100% maximum throughput and hence emulates an Output Queue switch by using multiple planes, input grouping and expansion techniques.

I. Multicast Switch Operation

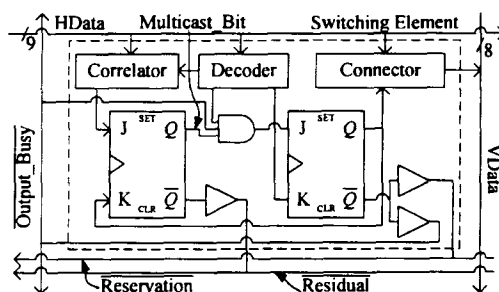


Fig.9. The switching element of multicast switch.

The architecture is Fig.9. For multicast switching, the linecard generates a multicast address packet and a data packet for each incoming packet. The multicast address packet consists of a truncated bit map of the incoming packet's multicast destination set, and

the multicast data packet carries the data unit of the incoming packet. Each output has a preassigned bit in the bit map. For example, in a bit map $\{b_i: i=1,2,\dots,N\}$, b_i is set to 1(0) if output i is(not) in the multicast destination set, for $i=1,2,\dots,N$. The transmission of the whole bit map is not necessary when output i are not in the multicast destination set for $i=I+1,\dots,N$, where I is the last output port number in the multicast destination set. Because both kinds of packets are of variable length, trailers are appended to indicate the ending.

The linecard sends the multicast address and data unit of its HOL packet to one of the shared sequencers via the coordination of the preprocessor. The sequencer can broadcast a packet to all the switching elements connected to it. The schematic diagram of the switching element is shown in Fig.9. A switching element consists of a multicast bit, which is represented by the output of a J-K Flip-Flop as shown in Fig.9. The

switching elements in the same column of a plane are connected to the same output and interconnected by an output busy indicator line. Once the multicast address packet is sent to a sequencer, the sequencer will broadcast this packet immediately. The connected switching element will check whether it is in the multicast destination set and if so, its multicast bit will be set to high, or else it is set to low. Note that the multicast address packet of an incoming packet is sent only one time to solve the multicast addressing problem.

The sequencers in the same plane must send a data packet in a round-robin fashion to prevent output conflict. The sequencer waits for its preassigned time point to broadcast the multicast data packet. When a data packet arrives, the switching element, whose multicast bit is high, will check the output busy indicator line output busy. If this line is high, the switching element will route the incoming data packet to the

connected output, reset its multicast bit to low, and set the connected output busy indicator line output busy to low. Otherwise, the switching element discards the data packet and sends a NACK signal to the sequencer through the signal Reservation. The switching elements must also send the updated multicast bit to the sequencer through the signal Residual.

Then, the sequencer checks signals Reservation and Residual. If Reservation is high, it means that there are no switching elements able to route the data packet and hence the sequencer will terminate this transmission. This data packet will be transmitted again at the next allowable time point. If Reservation is low, the sequencer will check Residual. If Residual is high, the data packet is already sent to all the multicast destinations and the sequencer completes the processing of the packet after this transmission. Otherwise, the sequencer will wait for the next

allowable time point to transmit the data packet again. This process is repeated until transmission completion, i.e. Residual is high.

II. Performance Analysis

Assume that the packet arrival process at each input is independent and identical Poisson process with rate λ_{in} . The length of each packet is independently and identically exponentially distributed with mean $1/\mu$. For multicast switching, a packet may request delivery to multiple destinations. This multiplicity denoted by F is termed Fanout, and its probability mass function is $p_f \triangleq \Pr[F=f], \forall f=1, \dots, N$. The multicast destination set of each packet is a random sample without replacement of all the N outputs. That is, the first copy chooses one of the N outputs randomly, the second copy chooses one of the remaining $N-1$ outputs randomly, and so on. When N is much larger than the mean Fanout, we can approximate the destinations of a multicast packet to

be independently and uniformly distributed for all outputs, since the resulting event of having two identical destinations in a packet becomes unlikely for large N . Therefore, the packet copy flows are spread evenly into all of the N outputs.

The buffer capacity of each linecard is assumed to be infinite. Ignoring the overhead of preprocessor, we can model the G linecards and the associated $G \cdot R$ sequencers by an $M/G/c$ queueing system, where the number of servers c is $G \cdot R$ and packet arrival rate is $G \cdot \lambda_{in}$. The service time is defined as the time each packet spent in the sequencer.

1. Service Time in Sequencers

The service time distribution of each multicast copy, denoted by $X_i, i=1, \dots, F$, is exponential with rate q :

$$q = \frac{\sum_{j=1}^N e^{-V_j(t)} V_j(t) \cdot \mu}{\sum_{j=1}^N V_j(t)} \approx \frac{E[e^{-V(t)}] \cdot \mu}{V} \quad (\text{as } N \rightarrow \infty)$$

$$= \frac{\lambda}{P \cdot V} = \mu = \frac{\lambda}{P}.$$

And an incoming packet spends time

X in the sequencer. The mean of X is

$$E[X] = M_X^{(1)}(0) = \frac{1}{q} \sum_{f=1}^N H_f \cdot p_f$$

$$= \frac{1}{q} E[H_F] = \frac{\eta \cdot E[F]}{\mu} \frac{P}{P - \rho_{in}},$$

where

$$\eta \triangleq \frac{E[H_F]}{E[F]}, \quad H_F \triangleq \sum_{i=1}^F i^{-1}$$

$\rho_{in} \triangleq (N\mu)$ is the input load, and η is the Fanout Function. The coefficient of variation of X depends only on the random variable F :

where

$$C_X^2 \triangleq \frac{\sigma_X^2}{E[X]^2} = \frac{E[H_F^2 - H_F^{(2)}]}{E[H_F]^2} - 1,$$

$$H_F^2 \triangleq (H_F)^2 \text{ and } H_F^{(2)} \triangleq \sum_{i=1}^F i^{-2}$$

2. Maximum Throughput

The maximum throughput per output, denoted by S , is attained when

$$\frac{G \cdot \lambda_{in}}{c} E[X] = 1.$$

We can solve λ_{in} under the saturation condition as

$$\lambda_m = \frac{R \cdot P \cdot \mu}{P \cdot E[H_F] + R \cdot E[F]}.$$

Then, we have the maximum throughput as

$$\begin{aligned} S &\triangleq \text{Min} \left(\frac{\lambda_m \cdot E[F]}{\mu}, 1 \right) \\ &= \text{Min} \left(\frac{R \cdot P}{R + P \cdot \eta}, 1 \right). \end{aligned}$$

3. Output Buffering Condition

For given multicast traffic with Fanout Function η and the 100% maximum throughput requirement, i.e. $S=1$, we can express the optimal values

$$P_o = \frac{R}{R - \eta} \text{ or } R_o = \frac{P \cdot \eta}{P - 1},$$

of R_o and P_o as

For multicast switching with finite mean Fanout, $0 < \eta < 1$, and without expansion, $R=1$, it is necessary to use multiple planes to obtain 100% throughput. Moreover, the smaller η is, more planes are required.

The required R_o or P_o is smaller in

the multicast switching than in the unicast switching. Therefore, a configuration of our system to emulate output buffering in unicast switching always also attains 100% maximum throughput in the multicast switching case.

4. Mean Waiting Time in Linecards

Because there is no exact solution of the M/G/c queueing system, we use the following approximations to evaluate the mean waiting time

$$E[W_q] = \left\{ (1 - \rho) \gamma \frac{c}{E[X]} + \rho \frac{1 + C_X^2}{2} \right\} E[W_q^{\text{exp}}],$$

where

$$E[W_q^{\text{exp}}] = \frac{(c\rho)^c \rho}{G\lambda_m c! (1 - \rho)^2} \left\{ \sum_{k=0}^{c-1} \frac{(c\rho)^k}{k!} + \frac{(c\rho)^c}{c! (1 - \rho)} \right\}^{-1},$$

$$\gamma = (1 - C_X^2) \frac{E[X]}{c+1} + C_X^2 \frac{E[X]}{c},$$

$$c = G \cdot R, \quad \rho = \frac{G\lambda_m}{c} E[X]$$

The normalized mean waiting time, $E[W_q] \cdot \mu$, is a function of the input load ρ_m and the Fanout distribution. The Fanout will affect the mean waiting time through C_X^2 .

C. Numerical Result

Assuming a 64×64 switch with P planes and R expansion ratio, the Fanout distribution can be deterministic or truncated Geometric distributed, both with the same mean Fanout $E[F]$. The deterministic Fanout is a fixed value of $E[F]$. The truncated Geometric Fanout has possible values of $1, 2, 3, \dots, 64$. We use these two distributions to evaluate the maximum throughput and optimal expansion ratio as shown in Tables 1 and 2 respectively. The simulation values in Table 1 are obtained by using deterministic Fanout and the bound of 95% confidence interval given as $(SimL, SimU)$. As a result, the deterministic distribution is a good candidate for Fanout distribution in performance evaluation because its Fanout Function is easy to evaluate and doesn't differ very much from other distributions.

A multicast switch with small expansion ratio and multiple planes will eliminate the HOL blocking, as shown in Table 2. As the mean Fanout $E[F]$ increases, it can lower the switch complexity to less than N^2 by using enough planes. For example, if we use three planes when the mean Fanout $E[F]$ is four, the switch complexity is $0.6931N^2$ for Geometric Fanout.

Although the number of linecards per preprocessor, G , doesn't affect the maximum throughput, it can lower the mean waiting time in the linecards as shown in Fig.10. When four linecards are grouped to share a preprocessor, the normalized mean waiting time is smaller than that when two linecards share a preprocessor.

$E \setminus F$	R	$P = 1$		$P = 2$	
		1	1.2	1	1.2
1	<i>All</i>	0.5000	0.5455	0.6667	0.7500
	<i>SimL</i>	0.4762	0.5150	0.6471	0.7156
	<i>Sim</i>	0.4920	0.5323	0.6699	0.7406
	<i>SimU</i>	0.5079	0.5497	0.6928	0.7657
	<i>Det.</i>	0.5714	0.6154	0.8000	0.8889
2	<i>Geo.</i>	0.5906	0.6339	0.8381	0.9280
	<i>SimL</i>	0.5557	0.5945	0.7984	0.8910
	<i>Sim</i>	0.5657	0.6056	0.8134	0.8975
	<i>SimU</i>	0.5758	0.6168	0.8284	0.9141
	<i>Det.</i>	0.6207	0.6626	0.9000	0.9910
3	<i>Geo.</i>	0.6455	0.6860	0.9530	1.0000
	<i>SimL</i>	0.6058	0.6390	0.9032	0.9759
	<i>Sim</i>	0.6124	0.6468	0.9150	0.9892
	<i>SimU</i>	0.6190	0.6546	0.9269	1.0000
	<i>Det.</i>	0.6575	0.6973	0.9796	1.0000
4	<i>Geo.</i>	0.6839	0.7220	1.0000	1.0000
	<i>SimL</i>	0.6442	0.6738	0.9759	1.0000
	<i>Sim</i>	0.6497	0.6807	0.9860	1.0000
	<i>SimU</i>	0.6551	0.6877	0.9960	1.0000

Table 1. The maximum throughput.

$E \setminus F$	P	2	3	4
1	<i>All</i>	2	1.5	1.3333
	<i>Det.</i>	1.5000	1.1250	1.0000
2	<i>Geo.</i>	1.3863	1.0397	0.9242
	<i>Det.</i>	1.2222	0.9167	0.8148
3	<i>Geo.</i>	1.0986	0.8240	0.7324
	<i>Det.</i>	1.0417	0.7813	0.6944
4	<i>Geo.</i>	0.9242	0.6931	0.6161

Table 2. The optimal expansion ratio.

D. Conclusion

We generalized the performance analysis of variable length packet unicast switching to random HOL multicast service with Fanout splitting. The Fanout F affects the maximum throughput through the Fanout Function, η . We formulate the maximum throughput analytically as a design guide.

The performance simulation results match the analytical calculation. The

optimal number of planes and expansion ratio are derived to emulate the output buffer switch. The input grouping with larger G can lower the mean waiting time. It is known that the condition of the switch for unicasting is more stringent than multicasting and we can lower the number of switching elements to provide multicasting. In particular, the complexity of the switch can be lowered to less than N^2 by using multiple planes in multicast switching.

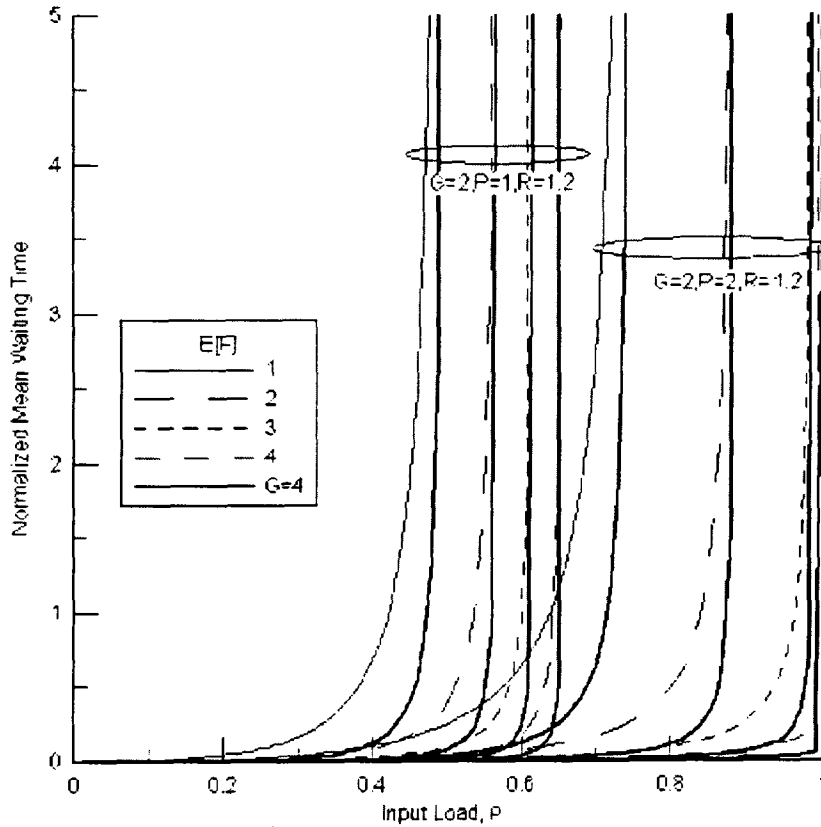


Fig. 10. The normalized mean waiting time.

【第三年】

A. Perfect Difference Code

Definition: Let W be the v -set of the integers $0, 1, 2, \dots, v-1$ modulo v . A set $D = \{d_1, d_2, \dots, d_k\}$ is a k -subset of W . For every $a \neq 0 \pmod{v}$, there are exactly λ ordered pairs (d_i, d_j) , $i \neq j$, such that

$$d_i - d_j \equiv a \pmod{v}.$$

(1)

A set fulfilling these requirements is called a *perfect difference set*. Since there are $k(k-1)$ pairs and the total number for $a \neq 0 \pmod{v}$ is $(v-1)$, the relation among v , k , λ is

$$\lambda = \frac{k(k-1)}{v-1}. \quad (2)$$

We can construct perfect difference codes $C_\tau = \{c_{\tau,0}, c_{\tau,1}, \dots, c_{\tau,(v-1)}\}$, $\tau \in W$, based on the cyclic difference set with the rule

$$c_{\tau,i} = \begin{cases} 1, & \text{for } i \in D_\tau \\ 0, & \text{otherwise.} \end{cases}$$

(3)

The code weight, code length, and total

number of codes are k , v , v , respectively,

where $v = k^2 - k + 1$. The cyclic-shifted property of perfect difference codes makes the design of a tunable transmitter or a tunable receiver simpler.

The application of the perfect difference sets in the SOCDMA systems is investigated first time. We also propose a new SOCDMA system using these sets as the signature codes. Based on the cyclic-shifted and unity cross correlation properties of these codes, we propose the transmitter and receiver structures as shown in Fig.1 and Fig. 2.

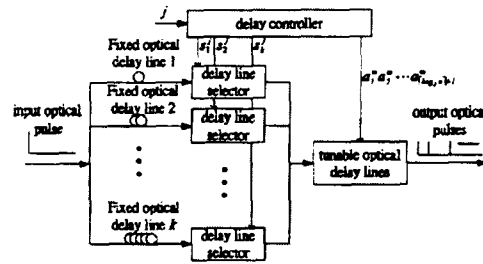


Fig. 1. Tunable transmitter structure.

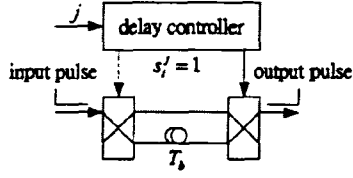


Fig. 2. Delay line selector in the i th branch.

An optimal divider shown in Fig. 3 is used in the receiver to enhance the system performance. The numerical result with consideration of shot noise, thermal noise, APD bulk, and surface leakage currents is shown in Fig. 4. PDC with the proposed receiver structure can effectively eliminate the MUI and the asymptotic error floor, which appears in the SOCDMA systems with modified prime sequence codes. Moreover, the proposed system can provide reliable communication even under heavy load without reducing the number of usable codes. It is believed that the proposed system performs better than other SOCDMA systems.

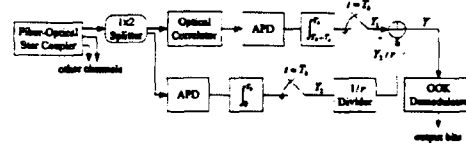


Fig. 3. The fixed receiver structure with optimal divider.

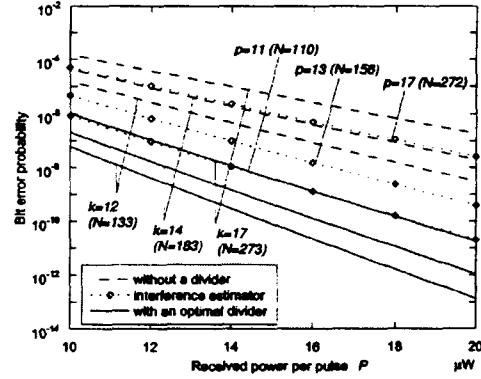


Fig. 4. Bit error probability comparisons under different code weights and full load.

B. Optical orthogonal codes with nonideal cross correlation

Recently, the construction and performance analysis of OOCs for optical code division multiple access (OCDMA) systems have been investigated widely. A $(v, w, \lambda_a, \lambda_c)$ -OOC is a family of $(0, 1)$ sequences with code length v , code weight w , the

maximum value of off-peak autocorrelation λ_a , and the maximum value of cross correlation λ_c . Most studies paid attention to $(v, w, 1)$ -OOCs with $\lambda_a = \lambda_c = 1$ for the sake of synchronization and minimizing interference. However, under the constraint of the ideal auto- and cross-correlation properties, the code size is upper bounded by $\lfloor (1/w) \lfloor (v-1)/(w-1) \rfloor \rfloor$, which is linear to the code length, where $\lfloor x \rfloor$ denotes the greatest integer not greater than x . Therefore, the code size is very sparse with respect to the code length. In order to obtain a larger code size, we should relax the constraint. We drop some marks appropriately from k marks of each code, such that the modified code weight becomes w and the code length is still equal to v . Moreover, the cross-correlation property should satisfy the constraint of $\lambda_c = 2$, even when the two codes are aligned with each other. For example, the set $\{0, 1, 3, 7, 15, 31, 36, 54, 63\}$ is a $(73, 9, 1)$ -perfect

difference set, where each element means the mark position. We can drop the first and the last five elements from the set, respectively, to form two subsets. Therefore, the two subsets form a family of $(73, 4, 1, 2)$ -OOCs. We can further use the follow algorithm to obtain the code with maximal code size.

- 1) Construct a perfect difference set with elements according to (A), where $k = q^r + 1$ and $v = q^{2r} + q^r + 1$, q is a prime number and r is a positive integer.
- 2) Each element of an extended field $F = \text{GF}(q^r) \cup \{\infty\}$, which also contains k elements, is related to one element of the perfect difference set.
- 3) Choose all the possible subsets with w elements from F and reserve the subsets fulfilling the property of a circle, where $w = q + 1$.
- 4) Each circle is related to w elements of the perfect difference set. The w elements form a code word with code length v and code weight w ,

where each element represents the mark position. The codes form a family of $(v, w, 1, 2)$ -OOCs.

The total number of codes is given by

$$T = q^{r-1} \cdot \frac{q^{2r} - 1}{q - 1} \quad (4)$$

Fig. 5 shows the bit error probability versus the received power per pulse of the systems using the four classes of OOCs under the number of simultaneous users $N=200$ and code length $v=6643$. When the received power P_w is small, these systems are power limited. When P_w is large, these systems are MUI limited. It also shows that the proposed codes with reversed codes perform better than ideal OOCs and Yang's codes. Moreover, the proposed codes have the largest code size. Fig. 6 shows the bit error probability versus the number of simultaneous users under the fixed received power $P_w = 0.5 \mu W$. At $BER \leq 10^{-9}$, the systems with the ideal OOCs have only about eight (with $w=3$) or 25 (with $w=4$) simultaneous users.

The systems with Yang's codes have about 52 (with $w=5$), 83 (with $w=6$), or 115 (with $w=7$) simultaneous users. On the other hand, the systems with the two proposed codes accommodate about 190 users and have more users than the conventional coding schemes.

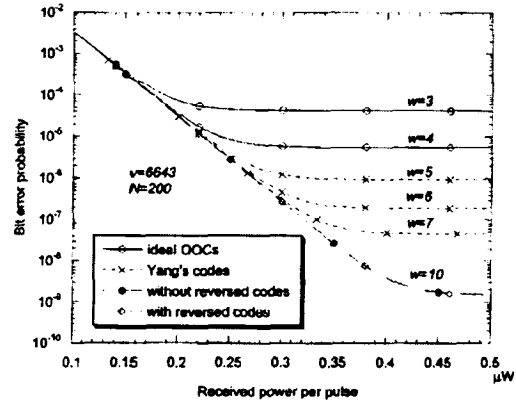


Fig. 5. The bit error probabilities versus the received power under $v=6643$ and $N=200$.

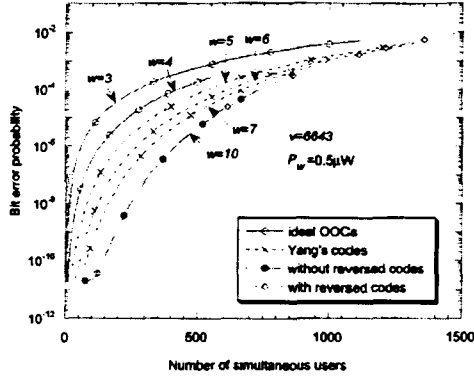


Fig. 6. The bit error probabilities versus the number of simultaneous users under $v=6643$ and $P_w = 0.5\mu W$.

We propose two new classes of OOCs. For the first time, we show that it is possible to construct a code family with code size larger than the code length. Compared with ideal OOCs and Yang's codes $(v, w, 1, 2)$ under the same code length and code weight, the performances of the systems with the four classes of codes are almost the same as one another. However, the proposed codes accommodate the users 10 times (which depends on the values of k and w) more than ideal OOCs for larger code length. To increase the number of simultaneous users for a given BER, we may increase the code

weight and reduce the code size. We show that it is possible to accommodate many more simultaneous users with the proposed codes than ideal OOCs and Yang's codes.

C. Jitter Performance Analysis of SOCDMA-Based EPON Using Perfect Difference Code

For the spread spectrum technique of wireless communication system, PN sequence is usually used as a signature pattern. However, the SOCDMA code structure is designed under some correlation constrain and we can't further assume it at random. Therefore, we discuss the correlation property of PDC, which has better performance than other codes in the literature. We have the following observations. The codes with weight k and length $v = k^2 - k + 1$ have $k-1$ codewords with first chip of value "1" and $k-1$ codewords with last chip of value "1" but not both. These $2k-2$ codes are called margin_1 codes. There

exists only one “special code” with both first chip and last chip of value “1”. The rest codes are named margin_0 codes as shown in Table. I. Note that the margin_1 codewords do not include the special code, which has chip “1” at both first and last chip position. We also observe that each codeword has two corresponding *burden codes*. For example, the desired receiver codeword “000001010000110010000” has corresponding burden codes “000000101000011001000” and “000010100001100100000” which are, respectively, the right shifted and left shifted versions of the desired code. We call it *burden code* because the cross-correlation between desired code and burden code is larger than the other codes. In the following analysis we will show that the burden code jitter has major impact on BER performance.

We can classify the receivers into three categories: the margin_1 receiver, the margin_0 receiver, and the special code receiver. A margin_1 receiver has

the margin_1 code as the signature codeword. When time error occurs, the correlator output of the last_chip_1 receiver is influenced by the next transmitted data bit of the interfering users. On the other hand, the first_chip_1 receiver is influenced by the previous transmitted data bit. The two kinds of receivers have the same correlation properties. Without loss of generality, we consider only the last_chip_1 receiver in the following analysis and the analysis of the other kinds is similar.

We suppose that the signal waveform of APD output is rectangular. Moreover, the output waveform after entering an ideal low pass filter with bandwidth much larger than the signal bandwidth (e.g. filter bandwidth is five times of signal bandwidth) is very closed to rectangular. The thermal noise and shot noise in this receiver are much smaller than MUI and will not significantly influence the system performance. Therefore, we can assume

a rectangular signal waveform in our analysis without losing the accuracy. We evaluate the jitter properties of three different receivers as the following.

A. The Receiver with Correlator

For the receiver using a correlator, if the desired signal has time error ε_d and the correlation error in each chip is Δ_d , the autocorrelation has the following property:

$$A = \begin{cases} (1-\Delta_d) \cdot k + y \cdot \Delta_d, & \text{if specialcode} \\ (1-\Delta_d) \cdot k + \Delta_d, & \text{otherwise} \end{cases} \quad (1)$$

where y is binomial distributed with equal probability that the desired signal has bit 1 or 0 in neighbor transmitted data bit.

The calculation of cross-correlation of PDC is more complicated than that of autocorrelation because the cross-correlations of codewords with first_chip_1 and last_chip_1 are different. Assume N_1 is the number of interfering users transmitting bit 1 in present bit duration and N_2 is the number of interfering users transmitting bit 1 in next bit

duration, then N_1 and N_2 are statistically independent. The cross-correlation between the i -th interfering user and the desired last_chip_1 receiver in present data bit duration C_1 is given by:

$$C_1 = \begin{cases} 1 & , \text{if } i \text{ has first_chip}0 \\ 1-\Delta_i x_i & , \text{if } i \text{ has first_chip}1 \\ 1+(k-2) \cdot \Delta_i x_i & , \text{if } i \text{ is a burdencode} \end{cases} \quad (2)$$

The cross-correlation of the last_chip_1 receiver with the i -th user in next transmitting bit is

$$C_{1n} = \begin{cases} 0 & , \text{if next bit transmits 0} \\ \Delta_i z_i & , \text{if next bit transmits 1} \end{cases} \quad (3)$$

where

$$x_i = \begin{cases} 1 & , \text{if late time error} \\ 0 & , \text{if early time error} \end{cases} \quad (4)$$

$$z_i = \begin{cases} 1 & , \text{if next bit 1} \\ 0 & , \text{if next bit 0} \end{cases} \quad (5)$$

Assuming that bit “1”, bit “0”, late time and early time errors have the probability equal to $\frac{1}{2}$. Let n and m be the number of interference users with first_chip_1 transmitting bit 1, then the total cross-correlation is given by:

$$C = \begin{cases} \sum_{i=1}^n (1 - \Delta_i x_i) + \sum_{i=1}^m \Delta_i z_i + (N_1 - n) \\ \quad , \text{if no burden code appears} \\ \sum_{i=1}^n (1 - \Delta_i x_i) + \sum_{i=1}^m \Delta_i z_i + (N_1 - n) \\ \quad + (k-2) \cdot \Delta_b x_b \\ \quad , \text{if one burden code appears} \\ \sum_{i=1}^n (1 - \Delta_i x_i) + \sum_{i=1}^m \Delta_i z_i + (N_1 - n) \\ \quad + (k-2) \cdot \Delta_b x_b + (k-2) \cdot \Delta_b x_b \\ \quad , \text{if two burden codes appear} \end{cases} \quad (6)$$

The special code receiver has the special code as its signature codeword. It has similar correlation property as (2) of the margin_1 receiver. The total cross-correlation of the special code receiver can also be derived as:

$$C = \begin{cases} \sum_{i=1}^n (1 - \Delta_i x_i) + 2 \cdot \sum_{i=1}^m \Delta_i z_i + (N_1 - n) \\ \quad , \text{if no burden code appears} \\ \sum_{i=1}^n (1 - \Delta_i x_i) + 2 \cdot \sum_{i=1}^m \Delta_i z_i + (N_1 - n) \\ \quad + (k-2) \cdot \Delta_b x_b \\ \quad , \text{if one burden code appears} \\ \sum_{i=1}^n (1 - \Delta_i x_i) + 2 \cdot \sum_{i=1}^m \Delta_i z_i + (N_1 - n) \\ \quad + (k-2) \cdot \Delta_b x_b + (k-2) \cdot \Delta_b x_b \\ \quad , \text{if two burden codes appear} \end{cases} \quad (7)$$

A margin_0 receiver has the margin_0 code as the signature codeword. The cross-correlation property is given by:

$$C_0 = \begin{cases} 1 + (k-1) \cdot \Delta_i x_i, & \text{if } i \text{ is a burden code} \\ 1, & \text{otherwise} \end{cases} \quad (8)$$

The total cross-correlation of margin_0 receiver can be derived as:

$$C = \begin{cases} N_1 \\ \quad , \text{if no burden code appears} \\ N_1 + (k-1) \cdot \Delta_b x_b \\ \quad , \text{if one burden code appears} \\ N_1 + (k-1) \cdot \Delta_b x_b + (k-1) \cdot \Delta_b x_b \\ \quad , \text{if two burden codes appear} \end{cases} \quad (9)$$

B. The Receiver using Simple Sampler

Here we analyze a receiver using simple sampler to reduce the jitter impact when the maximum jitter is less than one half of the chip duration. The receiver structure is shown in Fig. 4. The sampling point is at the center of the chip and the sampling rate is equal to the chip rate. In this receiver structure, the output is correct as long as the sampling instant falls within the corresponding chip duration. When the tracking loop is used in the receiver, the autocorrelation A will always be correct and $A=k$.

For the margin_1 receiver, we may follow the previous procedure and obtain the last_chip_1 receiver correlation value. The results of the first_chip_1 receiver and the special code receiver can be derived in the same way. The cross-correlation of the margin_1 receiver is given by:

$$C_1 = \begin{cases} 1 & , \text{if } i \text{ has first_chip } 0 \\ 1 - \alpha_i \beta_i & , \text{if } i \text{ has first_chip } 1 \\ 1 + (k-1 - \beta_b) \alpha_b & , \text{if } i \text{ is a burden code with first_chip } 1 \\ 1 + (k-1) \alpha_b & , \text{if } i \text{ is a burden code with first_chip } 0 \end{cases} \quad (10)$$

where α_b is the probability that the burden code has time error exceeding 0.5 chip and is binomial distributed with probability given by:

$$p(\alpha_b = 1) = \int_{-\varepsilon_{\max}}^{\varepsilon_{\max}} \int_{0.5 + \varepsilon_b}^{\infty} f(\varepsilon_i, \varepsilon_b) d\varepsilon_i d\varepsilon_b, \quad (11)$$

where ε_i is the jitter of tracking loop and ε_b is the jitter of the burden code. Assuming ε_i and ε_b are statistically independent, then their joint probability

density function is

$$\begin{aligned} f(\varepsilon_i, \varepsilon_b) &= f(\varepsilon_i) \cdot f(\varepsilon_b) \\ &= \frac{1}{\sqrt{2\pi\sigma_{\varepsilon_i}^2}} e^{-\frac{\varepsilon_i^2}{2\sigma_{\varepsilon_i}^2}} \cdot \frac{1}{2\varepsilon_{\max}} \end{aligned} \quad (12)$$

and β is binomial distributed with equal probability

$$\beta = \begin{cases} 1 & , \text{if next bit is 0} \\ 0 & , \text{if next bit is 1} \end{cases} \quad (13)$$

The total cross-correlation is given as:

$$C = \begin{cases} N_1 - n + \sum_{i=1}^n (1 - \alpha_i \beta_i) & , \text{if no burden code appears} \\ N_1 - n + \left(\sum_{i=1}^n (1 - \alpha_i \beta_i) \right) + (k-1 - \beta_b) \alpha_b & , \text{if only one burden code with first_chip } 1 \text{ appears} \\ N_1 - n + \left(\sum_{i=1}^n (1 - \alpha_i \beta_i) \right) + (k-1) \alpha_b & , \text{if only one burden code with first_chip } 0 \text{ appears} \\ N_1 - n + \left(\sum_{i=1}^n (1 - \alpha_i \beta_i) \right) + (k-1 - \beta_b) \alpha_b + (k-1) \alpha_b & , \text{if two burden codes appear} \end{cases} \quad (14)$$

The cross-correlation of the margin_0 receiver is given by:

$$C_0 = \begin{cases} 1 + (k-1) \alpha_b & , \text{if jitter } > 0.5 T_c \text{ and } i \text{ is burden code} \\ 1 & , \text{otherwise} \end{cases} \quad (15)$$

The total cross-correlation of the margin_0 receiver is

$$C = \begin{cases} N_1 & , \text{if no burden code appears} \\ N_1 + (k-1) \cdot \alpha_b & , \text{if one burden code appears} \\ N_1 + (k-1) \cdot \alpha_{b'} + (k-1) \cdot \alpha_{b''} & , \text{if two burden codes appear} \end{cases} \quad (16)$$

C. The Receiver using Joint Jitter Estimator and Canceller

We have defined two burden codes, which are right shifted and left shifted version of the desired code. Burden codes will cause additional multiple-user interference (MUI) as (2), (8) in the receiver with correlator and as (10), (15) in the receiver using simple sampler. Note that a left shifted burden code has additional MUI effect when it has early time error but no effect when late time error occurs. The case is similar for the right shift burden code. In order to improve the system performance, such additional MUI due to burden code jitter must be cancelled. We propose a joint jitter estimator and

canceller (JJEC) as shown in Fig. 5. The upper part is the data detection circuit with MUI canceller and the lower part is noncoherent delay-locked tracking loops. Two DLLs are employed in the lower part of the receiver, which act as two burden code jitter estimators. Let DLL1 track the time information of one of the burden codes $T_{b'}$, which is the right shift version of the receiver signature pattern. DLL2 tracks the time information of the other one $T_{b''}$. In the correlator receiver, JJEC estimates the time error and calculates the amount of jitter. Assuming that the burden code jitter is $\bar{\varepsilon}_\ell = T_b - T_\ell$, ($\ell = b', b''$) and correlation error is $\bar{\Delta}_\ell = \left| \bar{\varepsilon}_\ell \right|$. A decision circuit with zero threshold value is used to detect whether the burden code has the early or late time error, which may cause additional MUI. When additional MUI is induced, then the output of JJEC canceller is $\phi = \bar{\Delta}_\ell \cdot \xi$, where

$$\xi = \begin{cases} GT_c \frac{(k-1)}{k} \lambda_1 & , \text{if margin_0 receiver} \\ GT_c \frac{(k-2)}{k} \lambda_1 & , \text{if margin_1 receiver} \end{cases} \quad (17)$$

where G is the APD gain. In the simple sampling receiver, the average cross-correlation of a burden code to the desired receiver is k for the margin_0 receiver or $(k - \frac{1}{2})$ for the margin_1 receiver when the jitter exceeds $\pm 0.5T_c$. In order to detect and cancel the large cross-correlation of the burden code, the decision threshold of $\bar{\varepsilon}_b$ is set to $-0.5T_c$ and $\bar{\varepsilon}_{b^*}$ is set to $0.5T_c$. If $\bar{\varepsilon}_b < -0.5T_c$ or $\bar{\varepsilon}_{b^*} > 0.5T_c$, additional MUI will be induced. The amount of cancellation is set $\phi = GT_c \frac{(k-1)}{k} \lambda_1$ for the margin_0 receiver and $\phi = GT_c \frac{(k - \frac{1}{2} - 1)}{k} \lambda_1$ for the margin_1 receiver.

The BER of the receiver using correlator under various signal power and maximum peak jitter $\Delta_{\max}$ when

the system is full-loaded and code weight $k=12$ is shown in Fig.8. We observe that BER degrades with increasing $\Delta_{\max}$ and elevates as signal power increases. Figs. 9-10 compare the BER performance of the receiver using correlator and the receiver using simple sampler for various received signal power assuming $\varepsilon_{\max} = 0.5T_c$, $0.4T_c$ and $\sigma_{\varepsilon_i}^2 = 0.0025T_c$, respectively.

Numerical results show that the performance of the correlator receiver is power limited when no JJEC is used. This is because the MUI of burden codes dominates the performance degradation. If we increase the signal power, it also enhances MUI. The receiver with JJEC can significantly improve the BER and the performance depends on the accuracy of NC/DLL. The smaller the tracking error of NC/DLL is, the better performance JJEC can be achieved. These figures also show that the receiver using simple sampler has better BER performance

than the correlator receiver. The use of JJEC in the simple sampling receiver has good BER performance closed to the ideal case (no jitter) under the condition of small time error ε and NC/DLL performing well. This is because the sampling point falls within the chip boundary and the effect of MUI caused by burden code can be accurately detected and cancelled. Figs. 11-12 show the BER performance as functions of number of simultaneous users. The BER degrades when the number of simultaneous users increases. In addition, the receiver without JJEC has poor BER performance even in low user population. JJEC can indeed make the system robust against jitter.

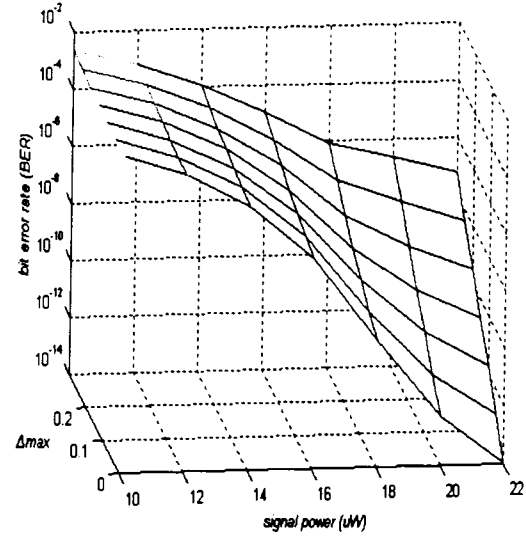


Fig. 8. Bit error rate (BER) of correlator receiver with various signal power and maximum peak-to-peak jitter ε_{max} when $k=12$.

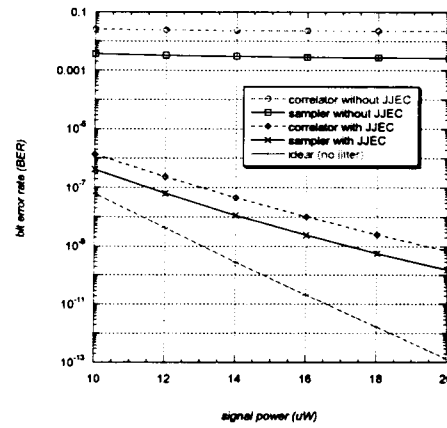


Fig. 9. Bit error rate (BER) with various signal power of correlator receiver and peak sampling receiver when $k=12$, $\varepsilon_{max} = 0.5 T_c$ and tracking error $= 0.0025 T_c$.

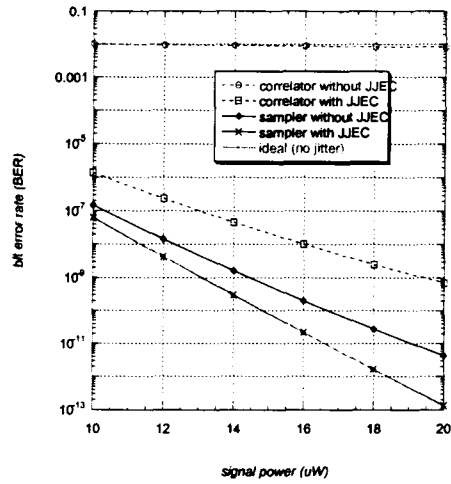


Fig. 10. Bit error rate (BER) with various signal power of correlator receiver and peak sampling receiver when $k=12$, $\varepsilon_{\max}=0.4T_c$ and tracking error $=0.0025T_c$.

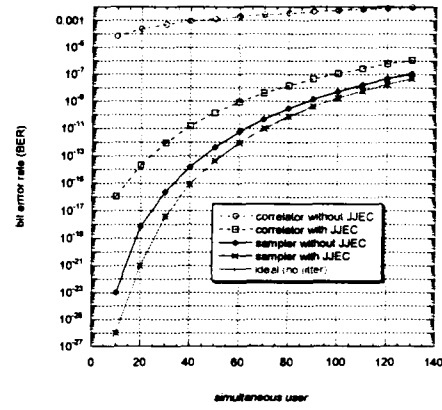


Fig. 12. Bit error rate (BER) with various simultaneous user of correlator receiver and peak sampling receiver when $k=12$, $\varepsilon_{\max}=0.4T_c$, signal power $=10\mu W$ and tracking error $=0.0025T_c$.

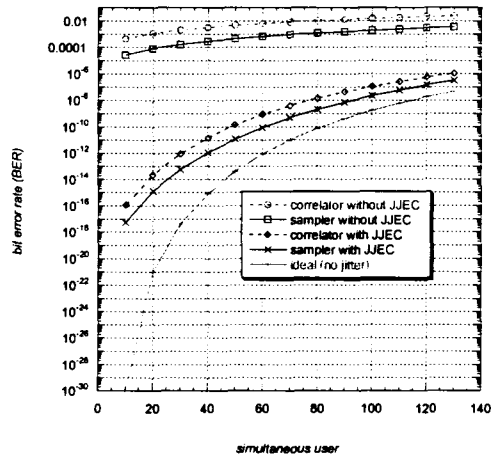


Fig. 11. Bit error rate (BER) with various simultaneous user of correlator receiver and peak sampling receiver when $k=12$, $\varepsilon_{\max}=0.5T_c$, signal power $=10\mu W$ and tracking error $=0.0025T_c$.

class		code sequence
Margin_0 codeword		010000000001010000110
		000000001010000110010
		000000010100001100100
		000000101000011001000
		000001010000110010000
		000010100001100100000
		000101000011001000000
		001010000110010000000
		010100001100100000000
		000110010000000001010
Margin_1 codeword		001100100000000010100
		011001000000000101000
	First	11001000000001010000
	chip_1	100000000010100001100
	code	101000011001000000000
		100001100100000000010
	Last	00100000000101000011
	chip_1	000000000101000011001
Special	code	010000110010000000001
		000011001000000000101
Special		code
100100000000010100001		

四、 計劃成果自評與結論

【第一年】

我們提出一個新的 IP over WDM 架構，利用標籤交換(label switching)與 WDM 波長資訊的觀念，設計一個可以有效簡化網路規約的封包格式，同時也將 MPLS 流量控制概念與 IP 傳輸與交換平面做有效整合。另一方面我們也分析了兩種可以處理非同步變動長度封包的高速光網路路由器：部份分享暫存器(PSB)與多重路徑路由(MPR)，模擬結果顯示 PSB WDM 光路由器比較適合使用在高速 IP 光纖網路。

【第二年】

本研究計畫設計一高速可非同步交換長度可變之封包的交換器。由於模組化的設計以及分散式的自我路由決定封包的交換排程，此新架構可以以簡單易行的硬體實現之。另外，我們經由此系統的理論分析與模擬，證明其可達到最佳化的效能，即百分之百的最高流通率以及低延遲時間。本交換

器可非同步地處理長度可變封包之多點發送交換，並且我們亦深入地分析可變長度封包的多點發送之效能，實為此領域之創新。本研究計畫發展之技術已申請美國和台灣專利，同時其成果亦發表於國際會議論文。

【第三年】

本研究計劃設計出兩組光分碼多工之編碼：完美差異碼與非理想交互關係碼。經過數值分析結果證明出這兩組編碼方式比傳統的同步碼與非同步碼有更佳的性能，此外、我們也提出相互時間偏差估測與消除電路可以有效改善完美相差碼對於時間誤差的敏感性；將此兩組編碼應用被動光網路可以真正達到寬頻接取的服務。

五、 參考文獻

【第一年】

- [1]EURESCOM Project P918-GI, "Integration of IP over Optical Networks: Networking and Management," Deliverable 1, Oct. 1999.

- [2]R. Callon, P. Doolan, N. Feldman,

- A. Fredette, G. Swallow and A. Viswanathan, "A Framework for Multiprotocol Label Switching," Internet Draft, Work in Progress, Sep. 1999.
- [3] E. C. Rosen, A. Viswanathan and R. Callon, "Multiprotocol Label Switching Architecture," Internet Draft, Work in Progress, Aug. 1999.
- [4] B. T. Doshi, S. M. Dravida, E. J. Hernandez-Valencia, W. A. Matragi, M. A. Qureshi, J. Anderson and J. S. Manchester, "A Simple Data Link Protocol for high-Speed Packet Networks", *Bell Labs Technical J.*, pp85-104, Jan.-Mar. 1999.
- [5] Bijan Raahemi, "Comparison of Frame Delineation Performance Between 10GE WAN PHY and 8B/10B," Nortel Networks' submission to IEEE802.3 HSSG, ver. 1.0, Nov. 1999.
- [6] R. D. Doverspike, S. Phillips, and J. R. Westbrook, "Future Transport Network Architecture," *IEEE Commun. Magazine*, vol. 37, issue 8, pp.96-101, Aug. 1999.
- [7] D. O. Awduche, Y. Rekhter, J. Drake, R. Coltun, "Multi-Protocol Lambda Switching: Combining MPLS Traffic Engineering Control with Optical Crossconnects," *Internet Draft*, Work in Progress, Nov. 1999.
- [8] Tom Worst, "General Switch Management Protocol," *Internet Draft*, Work in Progress, Oct. 1999
- [9] John Crowcroft, "IP over Photons: How not to Waste the Waist of the Hourglass," *IEEE Conference IWQoS '99, Seventh International Workshop on Quality of Service*, pp.9-11, April 1999.
- [10] ITU-T Recommendation I.432, "B-ISDN user-network interface – Physical layer specification: General characteristics", Aug. 1996.
- [11] DooWhan Choi, "Frame Alignment in a Digital Carrier System – A Tutorial," *IEEE Commun. Magazine*, vol. 28, issue 2, pp.47-54, Feb. 1990.
- [12] R. W. Sitter, "Systems Analysis of Discrete Markov Processes," *IRE Trans. on Circuit Theory*, Vol.CT-3, pp.257-266, Dec 1956.
- [13] Franco Callegati, Giorgio Corazza and Carla Raffaelli, "Exploitation of DWDM for optical packet switching with quality of service guarantees," *IEEE J. Select. Areas Commun.*, vol. 20, pp 190-201, Jan. 2002.
- [14] Lubo Tancevski, Siva Yegnanarayanan, Gerardo Castanon, Lakshman Tamil, Francesco Masetti and Tom McDermott, "Optical Routing of Asynchronous, Variable length packets," *IEEE J. Select. Areas Commun.*, vol. 18, pp 2084-2093, Oct. 2000.
- [15] W. E. Leland, M. S. Taqqu, W. Willinger and D. V. Wilson, "On the self-similar nature of Ethernet traffic (extended version)," *IEEE/ACM Trans. Networking*,

- vol. 2, no.1, pp. 1-15, Feb. 1994.
- [16] M. E. Crovella and A. Bestavros, "Self-similarity in World Wide Web traffic: Evidence and possible causes," *IEEE/ACM Trans. Networking*, vol. 5, no.6, pp. 835-846, Dec. 1997.
- [17] Meow C. Chia, David K. Hunter, Ivan Andonovic, P. Ball and I. Wright, "Optical packet switches: A comparison of designs," in *Proc. ICON'2000*, Singapore, Sept. 2000, pp. 365-369.
- [18] David K. Hunter, Meow C. Chia and Ivan Andonovic, "Buffering in optical packet switches," *IEEE/OSA J. Lightwave Technol.*, vol 16, no. 12, pp. 2081-2094, Dec. 1998.
- [19] David K. Hunter, Mohamed H. M., Meow C. Chia, Ivan Andonovic, Ken M. Guild, Anna Tzanakaki and Mike J. O'Mahony, John D. Bainbridge, Marc F. C. Stephens, Richard V. Penty and Ian H. White, "WAPNET: A Wavelength Switched Packet Network," *IEEE Commun. Mag.*, vol. 19, issue 9, pp. 120-129, March 1999
- [20] Wen De Zhong and Rodney, S. Tucker, "A Wavelength-routed photonic packet buffer combining travelling delay lines with delay loops," *IEEE/OSA J. Lightwave Technol.*, vol. 19, no. 8, pp. 1085-1092, Aug. 2001.
- [21] P. Gambini *et al.*, "Transparent optical packet switching: Network architecture and demonstration in the KEOPS project," *IEEE J. Select. Areas Commun.*, vol. 16, pp. 45-59, Sept. 1998.
- [22] Soeren L. Danielsen, Peter B. Hansen and Kristian E. Stubkjaer, "Wavelength conversion in optical packet switching," *IEEE/OSA J. Lightwave Technol.*, vol. 11, no. 12, pp. 2095-2108, Dec. 1998.
- [23] Gerardo Castanon, Lubo Tancevski and Lakshman Tamil, "Optical packet switching with multiple path routing," *Computer Networks.*, vol. 32, issue 5, pp. 635-662, May 2000.
- [24] Juan Diao and Pak. L. Chu, "Analysis of partially shared buffering for WDM optical packet switching," *IEEE/OSA J. Lightwave Technol.*, vol. 17, no. 12, pp. 2461-22469, Dec. 1999.
- [25] Lubo Tancevski, An Ge and Gerardo Castnon, "Optical packet switchi with partially shared buffers: Design Principles," in *Proc. OFC '2001*, Anaheim, March 2001, paper TuK3.
- [26] Franco Callegati, "Optical buffers for variable length packets," *IEEE Commun. Lett.*, vol. 4, pp. 292-294, Sept. 2000.

【第二年】

- [1] Kun-Tso Chen and Jingshown Wu, "An asynchronous IP switch for variable length packets," R.O.C. patent pending, U.S.A. patent pending.
- [2] N. Chandhok, A. Durrezi, R. Jagannathan, S. Seetharaman, K. Vinodkrishnan, "IP over Optical Networks: A summary of issues," Internet Draft draft-osu-ipo-mpls-issues-00.txt, Work in Progress, July 2000.
- [3] Lisong Xu, Harry Perros and George Rouskas, "A survey of optical packet switchin and optical burst switching," Technical Report, Department of Computer Science, North Carolina State University, 2000.
- [4] L. Ceuppens, et al., "Performance monitoring in photonic networks in support of MP λ S," Internet Draft draft-ceuppens-mpls-optical-00.txt, March 2000.
- [5] M.G. Hluchyj, and M. J. Karol, "Queueing in High-Performance Packet Switching," *IEEE J. Selected Areas Commun.*, Vol. 6, No. 9, pp. 1587-1597, Dec. 1988.
- [6] T. E. Anderson, S. S. Owicki, J. B. Saxe, and C. P. Thacker, "High-Speed Switch Scheduling for Local-Area Networks," *ACM Trans. Computer Systems*, Vol. 11, No. 4, pp. 319-352, Nov. 1993.
- [7] N. McKeown, A. Mekkittikul, V. Anantharam, and J. Walrand, "Achieving 100% Throughput in an Input-Queued Switch," *IEEE Trans. Commun.*, Vol. 47, No. 8, pp. 1260-1267, Aug. 1999.
- [8] J. J. Hunter, *Mathematical Techniques of Applied Probability. Volume 2. Discrete Time Models: Techniques and Applications*. New York: Academic Press, 1983.
- [9] M. S. Schmookler, "Limited capacity discrete time queues with single or bulk arrival," IBM technical report TR 00.2048, New York, June 18, 1970.
- [10] J. D. C. Little, "A simple proof of $L = \lambda W$," *Operations Research*, Vol. 9, pp. 383-387, 1961.
- [11] Hideaki Takagi, *Queueing Analysis, A foundation of Performance Evaluation, Volume 3: Discrete-Time Systems*. Elsevier Science Publishers B.V., Amsterdam, 1993.
- [12] S. Halfin, "Batch delays versus customer delays," the Bell System Technical Journal, Vol.62, No. 7. pp.2011-2015, Sep. 1983.
- [13] Guo and Ruay-Shiung Chang, "Multicast ATM switches: survey and performance evaluation," *Comput. Commun. Rev*, vol. 28, pp.98-131, Apr. 1998.
- [14] Joesph Y. Hui and Thomas Renner, "Queueing analysis for multicast packet switching," *IEEE Trans. Commun.*, vol. 42, pp.723-731, Feb. 1994.

【第三年】

- [1] Chi-Shun Weng and Jingshown Wu, "Perfect difference codes for synchronous fiber-optic CDMA communication systems," *J. Lightwave Technol.*, vol. 19, no. 2, pp. 186-194, Feb. 2001.
- [2] Chi-Shun Weng and Jingshown Wu, "Optical orthogonal codes with nonideal cross correlation," *J. Lightwave Technol.*, vol. 19, no. 2, pp. 1856-1863, Dec. 2001.
- [3] Jingshown Wu and Fong-Ray Gu, "Jitter performance analysis of SOCDMA-based EPON using perfect difference code," submitted to *J. Lightwave Technol.*
- [4] Chi-Shun Weng and Jingshown Wu, "Optical orthogonal codes with large cross-correlation and their performance bound for asynchronous optical CDMA systems," accepted by *J. Lightwave Technol.*
- [5] Che-Li Lin and Jingshown Wu, "Channel interference reduction using random Manchester codes for both synchronous and asynchronous fiber-optic CDMA systems," *J. Lightwave Technol.* Vol. 18, no. 1, pp.26-33, Jan. 2000.
- [6] Jingshown Wu and Che-Li Lin, "Fiber-optic code division add-drop multiplexers," *J. Lightwave Technol.* Vol. 18, no. 6, pp.819-824, June 2000.
- [7] Che-Li Lin and Jingshown Wu, "A synchronous fiber-optic CDMA system using adaptive optical hardlimiter," *J. Lightwave Technol.*, Vol. 16, no. 8, pp. 1393-1403, Aug. 1998.
- [8] Jyh-Horng Wu and Jingshown Wu, "Synchronous fiber-optic CDMA using hard-limiter and BCH codes," *J. Lightwave Technol.*, vol. 13, no. 6, pp. 1169 -1176, June 1995.
- [9] W. C. Kwong, P. A. Perrier, and P. R. Prucnal, "Performance comparisons of asynchronous and synchronous code-division multiple-access techniques for fiber-optic local area network," *IEEE Trans. Commun.* vol. 39, no. 11, pp. 1625-1634, Nov. 1991.
- [10] J. G. Zhang, "Improvement in bandwidth efficiency of asynchronous prime-code CDMA networks by error correction coding," *IEEE Electronics Letters*, vol. 30, no. 6, pp. 514-515, March 1994.
- [11] J. A. Salehi, "Code division multiple-access techniques in optical fiber networks- Part I: Fundamental principles," *IEEE Trans. Commun.* vol. 37, no. 8, pp. 824-833, Aug. 1989.
- [12] W. C. Kwong, G. C. Yang, and J. G. Zhang, "2ⁿ prime-sequence codes and coding architecture for optical code-division multiple-access," *IEEE Trans.*

- Commun.* vol. 44, no. 9, pp. 1152-1162, Sept. 1996.
- [13] H. M. H. Shalaby, "Synchronous fiber-optic CDMA systems with interference estimators," *J. Lightwave Technol.*, vol. 12, no. 11, pp. 2268-2275, Nov. 1999.
 - [14] W. C. Kwong and P. R. Prucnal, "All-serial coding architecture for ultrafast optical code-division multiple-access," in *Proc. ICC '93*, vol. 1, May 1993, pp. 552-556.
 - [15] J. Singer, "A theorem in finite projective geometry and some applications to number theory," *Trans. Amer. Math. Soc.*, vol. 43, pp. 377-385, 1938.
 - [16] J. Schönheim, "On maximal systems of k-tuples," *Studia Sci. Math Hungar.*, vol. 1, pp. 363-368, 1966.

國科會結案報告

智慧型寬頻光纖通信網路研究

子計畫二：應用於寬頻光通訊網路之波長轉換與波長選取技術研究

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執行單位：國立台灣科技大學電子工程系

計畫編號：NSC 91-2215-E-011-005

一、計畫摘要：

本計畫主要是針對高密度分波多工網路中的幾個關鍵問題進行理論與實驗上之探討，重點放在波長數目增加和通道間距縮小的趨勢下亟須解決的關鍵技術：波長選取、波長轉換和波長監控，以及因應此發展趨勢必須修正的網路設計概念。波長選取涉及窄帶寬、低串音且可調範圍大的濾波器和多工器的設計，波長轉換的研究重點則為提高轉換效率及降低雜訊。此外，我們也針對波長選取和波長轉換技術能實用性，對網路架構所造成的影響等，做適當的評估，研究重點在於善用使用波長路徑或虛擬波長路徑概念的網路系統中的波長資源。

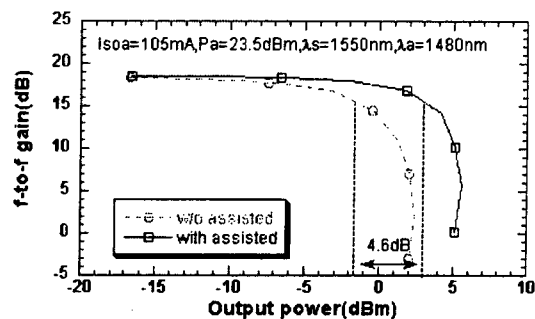
二、計畫成果：

(1)波長轉換：

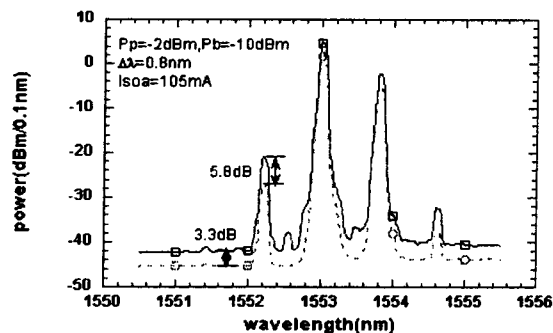
四波混合(Four Wave Mixing)波長轉換轉換效率之改善，以四波混合的方式來作波長轉換有許多優點，但其主要的缺點是轉換效率太低。本計畫的目的在於改善波長轉換效率。本計畫中，加入輔助光提高半導體光放大器的飽和輸出功率，轉換效率之最大值正比於飽和輸出功率的平方，進而改善整體波長轉換效率。

如圖(一)，加入輔助光後，半導體光放大器的飽和輸出功率提升了 4.6dB。從圖(二)可發現轉換訊號的強度在頻譜上，明顯上升了 5.8dB。經由

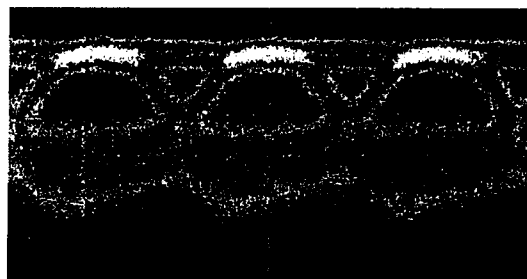
四波混合與輔助光技術，實際訊號傳輸及轉換效果可由圖(三)與圖(四)比較，在傳輸速率 10Gbps 下，加入輔助光，轉換訊號的振幅增加三倍多，相當有助於實際光訊號轉換的再傳輸。



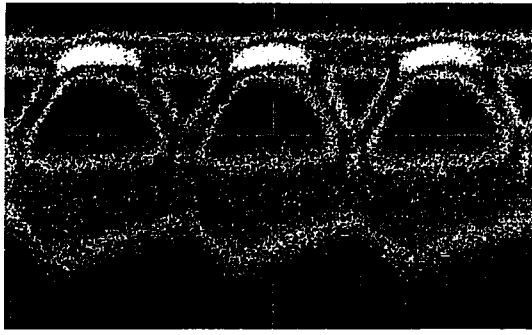
圖(一)以輔助光提高光飽和輸出功率



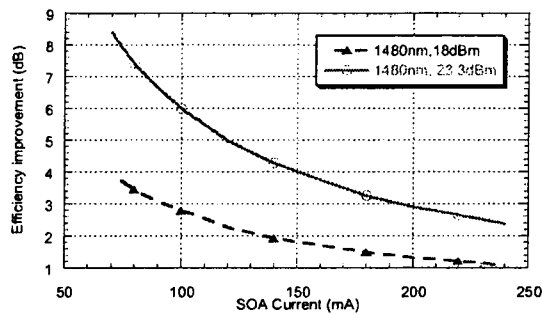
圖(二)加入輔助光影響四波混合的光譜



圖(三)未加入輔助光之四波混合波長轉換眼圖，速率 10Gbps，峰對峰值 221.7mV



圖(四)加入輔助光之四波混合波長轉換眼圖，速率 10Gbps，峰對峰值 702.3mV



圖(五)輔助光功率對波長轉換效率的影響

如圖(五)所示，當輔助光之光功率越大時，對於整體轉換效率增加更多，另外輔助光波長與半導體光放大器的操作電流也會影響轉換效率。因此，為有效運用此一嶄新技術，必須掌握材料之增益與透明點隨波長變化之關係，對此我們也已發展出一簡易理論，在進度報告及發表於 IEEE 的期刊中皆有詳細說明。

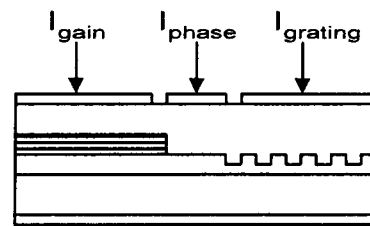
綜合上述成果，本計劃已提出了一波長轉換的解決方案，在實用性方面，設計光積體化波長轉換器，運用輔助光技術設計兩段式串接之光放大器結構，簡化操作方式，提供更高的轉換效率，實際成為光通訊網路中波長轉換器。

(2) 高速可調式雷射模組：

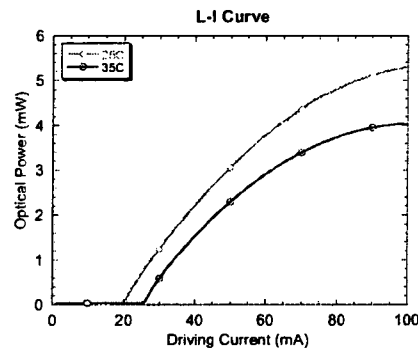
在光交換開關中，波長的切換速度及可調性相當重要，設計高速波長可調雷射模組配合波長轉換器，可完成高速可調波長轉換器。本計劃中，參考相關資料，自行設計一款分佈式布拉格反射鏡可調式半導體雷射 (DBR Laser)，其結構如圖六所示，調動光柵區的注入電流，改變反射鏡的

頻譜分佈，進而使雷射發光在特定選擇波長。增益區(gain)提供足夠的能量使雷射發光，最大輸出光功率大於 5mW，相位區則作為波長微調之用。

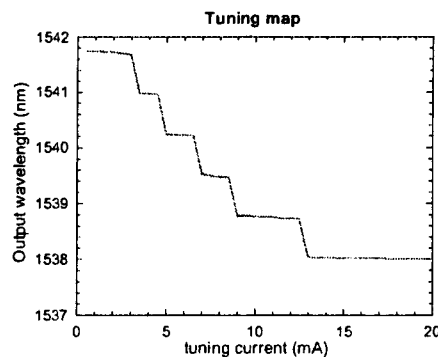
光柵區調動電流與發光波長之相對應關係如圖(八)，可選擇發光波長非連續式，在僅使用單一調動電流操作下，通道間距約為 0.8nm，可利用微調與熱調的技術，將發光波長對應在標準的 ITU 規範波長，階梯式的波長選擇性可避免在高速波長通道切換的過程中，產生額外不必要的光訊號，影響傳輸品質及速率。圖(九)為已封裝的 DBR 可調式雷射，內含溫控機制、檢光二極體及自製 DBR 雷射晶體。



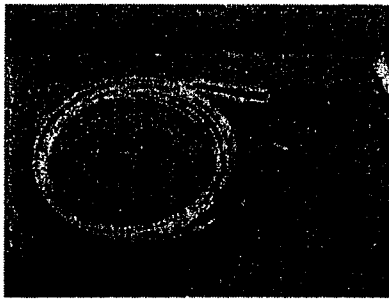
圖(六)DBR 可調式雷射



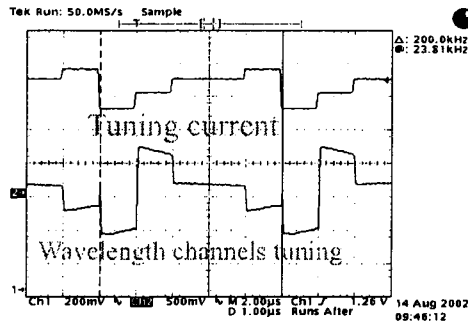
圖(七) DBR 雷射增益電流與輸出功率 L-I 曲線



圖(八)調動電流與發光波長



圖(九)已封裝之可調式雷射實體圖



圖(十)四通道高速調動電流與調動發光波長

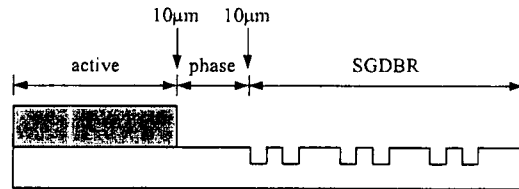
配合自行設計的可調式雷射驅動電路，目前擁有波長調變速率高於 10MHz 的可調式雷射模組，其高速調動圖如圖(十)所以，上曲線為調動電流，透過適當的波長對電壓轉換機制，得到下曲線為各調動電流之對應輸出波長。

(3)波長選取：

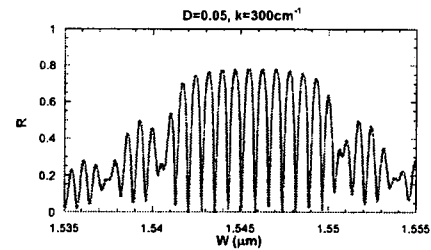
ITU 規定之波長通道：由於 DWDM 中波長位置及通道間隔已有標準規範，波長間隔為 0.8nm 甚至是 0.4nm。本計畫已設計光積體化光濾波之元件，用於選取各標準通道之訊號。

本計畫中，利用取樣光柵(SGDBR)做為預設波長的光濾波器，SGDBR 的反射率尖峰的間隔(ΔP)為 0.8nm，為了量測方便，我們在光濾波器前設計一段增益區，藉由注入電流至增益區使之發光，再經由光濾波器量測光譜，圖(十一)為光濾波器的結構圖。藉由兩段光柵之間的干涉，使之產生通道間隔為 0.8 nm，如圖(十二)為此固定通道光濾波器反射頻譜，可以發現此光濾波器不但有眾多且平坦的反射率頻譜，而且其相鄰通道的具有足夠的抑制效果，藉由

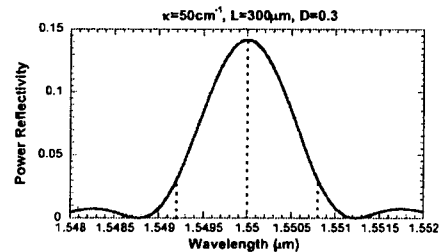
調整溫度或電流以改變波導的折射率，使得通道的位置符合 ITU 的規範。



圖(十一) 光濾波器的結構圖



圖(十二) 固定通道光濾波器



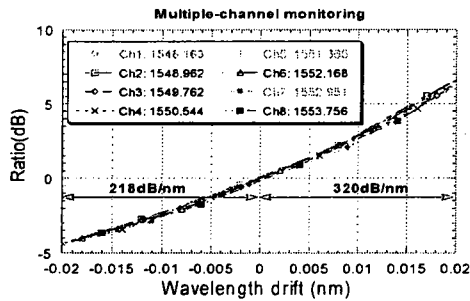
圖(十三) 可調式光濾波器

此外利用 SGDBR 光柵設計可調式光濾器，藉由調整電流以改變波導的折射率，可調動選取的濾波器的波長，如圖(十三)為可調選取單一通道波長訊號頻譜。

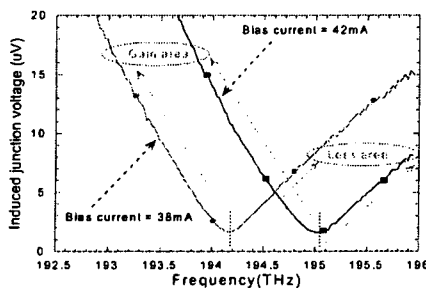
(4)波長監控：

對於高精密度波長監控技術，利用被動元件：FP etalon 共振腔濾波器搭配半導體光電元件，如半導體光放大器(SOA)或發光二極體(LED)。本計畫中所使用的被動 FP etalon 濾波器對溫度較不敏感，特性頻譜的漂移量約 0.011nm/°C，對於監控系統的穩定度與精確性有相當大的幫助，其共振週期(FSR)為 50GHz，利用溫控將濾波器的透射與反射頻譜的交叉點對準在標準 ITU 規範波長，觀察透射與反射量的差異，監控波長的細微偏移

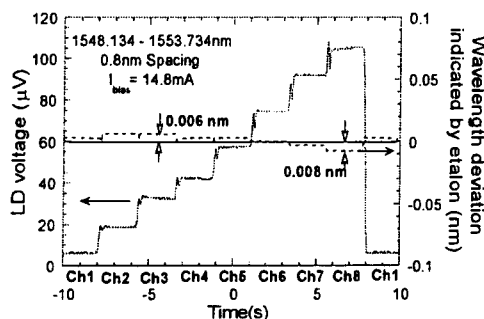
量，圖(十四)為 FP etalon 監控八個通道，各通道波長偏移與濾波器的響應變化，經光—電轉換器，不同漂移量轉換為不同電壓輸出，精確監控波長，且各通道特性相當一致且明顯。



圖(十四)使用 FP etalon 監控八個標準 ITU 波長



圖(十五)注入光頻率與 LED 感應介面電壓對應關係



圖(十六)動態監控 8 通道，通道間距 100GHz

通道監控方面，利用 LED 或 SOA 等半導體光電元件(SOD)的穿透特性，將 SOD 做為光感測器，光打入 SOD，SOD 的接面電壓會隨著注入光的波長不同，有不同的對應電壓，且對應區線可藉由偏壓電流調整至適當頻段，如圖(十五)所示，為利用 LED 作為感測器，LED 價格低廉，可節省系統成本，使監控系統具很高的

實用性。圖(十六)為利用上述技術，監控波長動態變化，包含通道跳躍與波長偏移，本技術可監控小於 0.008nm 以下的波長漂移現象，滿足 ITU DWDM 100GHz 通道間距規範的品質，並精密且快速監控並回應相關訊息給相關控制單元。

三、已發表之論文：

本計畫已投稿國際期刊及研討會論文多篇，有關波長轉換及光濾波器波長監控之成果。

1. S.-L. Lee, C.-T. Pien, and Y.-Y. Hsu, "Wavelength Monitoring with Low-Cost Laser Diodes for DWDM Applications," *Electron. Lett.*, vol.36, no.6, pp.545-546, 2000.
2. S.-L. Lee, C.-T. Pien, and Y.-Y. Hsu, "Operation Principles of Wavelength Sensing Using Transparent Properties of Semiconductor Optical Diodes," *J. Lightwave Technol.*, vol.19, no.85, pp. 655-665, May 2001.
3. S.-L. Lee, "Analytical Formula of Wavelength Dependent Transparent Current and its Implications for Designing Wavelength Sensors and WDM Lasers," *IEEE J. of Selected Topics in Quantum Electronics*, vol.7, no.2, pp. 201-209, 2001.
4. S.-L. Lee and Y.-Y. Hsu, "Study of effects of aging on transparent current stability of semiconductor lasers," *Electron. Lett.*, vol.37, no.12, pp. 767-769, 2001.
5. S.-L. Lee and Y.-Y. Hsu, C.-T. Pien, "High-resolution wavelength monitoring using differential detection of junction voltage across single diode laser," *IEEE Photonics Technol. Lett.*, vol. 13, no.8, pp.872-874, 2001.
6. S.-L. Lee, P.-M. Gong, and C.-T. Yang, "Performance Enhancement on SOA-Based Four-Wave-Mixing Wavelength Conversion Using an Assisted Beam," *IEEE Photon. Technol. Lett.*, 2002.
7. S.-L. Lee, Y.-Y. Hsu, and C.-T. Pien, "High-Resolution and Power Insensitive Wavelength Monitoring Using a Low Cost FP Laser for DWDM Applications," *CLEO/PR2001*, paper TuA2-4, pp. I-268-269.
8. Chin-Tien Yang, Pei-Miin Gong, San-Liang Lee, "Wavelength conversion using a semiconductor optical amplifier with an assisted beam," *Optics and Photonics Taiwan'01*, paper TC1-8, 2001.
9. Chun-Liang Yang and San-Liang Lee, "High-Resolution Wavelength Monitoring Technology for Tunable Laser Sources," *Optics and Photonics Taiwan'01*, paper FC3-5, 2001.
10. San-Liang Lee, Pei-Miin Gong, and Chin-Tien Yang, "Efficiency improvement of SOA-based four-wave-mixing wavelength converters by using an assisted beam," *OFC2002*, paper ThGG75, Anaheim, CA, USA, Mar. 2002.

11. Chun-Liang Yang and San-Liang Lee, "Wavelength monitoring of tunable DWDM sources using a FP etalon and a FP laser diode," OFC2002, paper ThC2, Anaheim, CA, USA, Mar. 2002.
12. Pei-Miin Gong, San-Liang Lee, and Chin-Tien Yang, "Wavelength conversion using the four-wave-mixing in a semiconductor optical amplifier assisted with an injection beam," OC2002, Taipei, Taiwan, Apr. 2001.
13. C.-Y. Chien, R. Xuan, P.-L. Jiang, and S.-L. Lee, "A fast wavelength-tunable transmitter for optical switching applications," *OPT'02*, 2002.

四、總結

本計畫之研究重點為波長轉換與波長選取技術，已研究出多項關鍵技術，並發表論文多篇。最顯著之成果為以適當波長輔助光改善半導體光放大器之性能，以實驗發現數項重要數據，成為首度以實驗完整探討此效應者，可搭配已製作之高速可調式雷射模組，完成高速波長可調波長轉換器。另外波長選取監控技術也有相當好的成果，可選擇標準 ITU 規範通道波長，精密監控波長漂移。本計畫擁有相當不錯的結果，波長轉換與波長選取等技術研究已有多篇文章刊登在國際知名期刊，學術性與實用性兼備。

Wavelength Monitoring with Low-Cost Laser Diodes for DWDM Applications

作者：李三良 (S.-L. Lee)，卞敬塘 (C.-T. Pien)，許毓倚 (Y.-Y. Hsu)

發表期刊：Electron. Lett., vol.36, no.6, pp.545-546, 2000.

內容成果：

半導體雷射二極體的操作工作區大致可分為三大項，分別為吸收區、透明點與增益區。操作電流大小在透明點以下為吸收區，當外部光源注入時，光子會被吸收，既會有損耗；反之操作電流在透明點以上，此時雷射二極體帶有增益，可放大光功率或是發光；而當操作電流在透明點，雷射二極體不吸收也不放大，光直接穿透過雷射二極體。經由實驗與理論發現，半導體雷射的透明點電流會隨注入的光源的波長的不同而有相對應的變化，利用此特性，可即時精確監控通道中光訊號波長的穩定度以及偏移量。本文中使用一半導體光放大器作為光感測器，搭配簡單的控制電路，達到波長監控的功能。

- 3 FREITAG, I., TÖNNERMANN, A., and WIELING, H.: 'Power scaling of diode-pumped monolithic Nd:YAG lasers to output powers of several watts', *Opt. Commun.*, 1995, 115, pp. 511-515
- 4 FREITAG, I., TÖNNERMANN, A., and WIELING, H.: 'Intensity stabilised Nd:YAG ring laser with 1.5 W single-frequency output power at 1.357 μm ', *Electron. Lett.*, 1997, 33, pp. 777-778
- 5 KARLSSON, H., and LAUREN, J.: 'Electric field poling of flux grown KTiOPO_4 ', *Appl. Phys. Lett.*, 1997, 71, pp. 3474-3476

Wavelength monitoring with low cost laser diodes for DWDM applications

San-Liang Lee, Ching-Tang Pien and Yu-Yi Hsu

A simple wavelength monitoring technique based on the transparency detection of low-cost laser diodes and very simple circuitry is demonstrated. The wavelength sensor can be operated at either fixed bias or in a sweeping mode. Wavelength monitoring using the proposed approach has been demonstrated to have sub-angstrom resolution and good stability.

Introduction: Owing to the very tight wavelength spacing used in dense wavelength division multiplexing (DWDM) networks, an effective approach for wavelength monitoring is required. Significant efforts have led to the realisation of several useful approaches for wavelength stabilisation and monitoring [1-4]. We have proposed a novel technique that is based on the detection of the transparent point of a semiconductor optical amplifier (SOA) for wavelength measurement and wavelength tracking [5]. In principle, the transparent condition of a semiconductor gain material is insensitive to facet reflection. Thus, a diode laser biased below threshold or an SOA with greatly relaxed requirements on the antireflection coatings can also be used for this purpose. A discrete component can be very useful for applications in external wavelength sensing or the wavelength stabilisation of diode lasers.

Operation principle: The transparency of a diode laser corresponds to the condition that the injected carrier density is large enough to make the induced separation between the quasi-Fermi levels of electrons and holes equal the incident photon energy [5]. This transparent current for incident light decreases as the wavelength increases. Thus, the wavelength of light incident on the diode laser can be resolved by characterising the variation of transparent current with wavelength for the diode. The applicable wavelength range falls between the bandgap wavelength and the lower bound determined by the condition that its corresponding transparent current must be below that for the threshold condition.

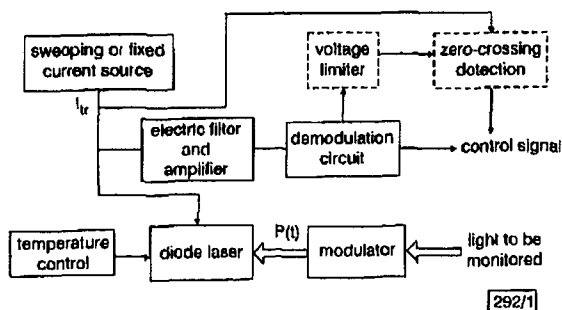


Fig. 1 Schematic diagram of wavelength sensing or stabilisation where laser diode acts as wavelength discriminator

Dashed blocks are for sweeping mode

Fig. 1 shows a schematic diagram of wavelength monitoring using a diode laser as a wavelength discriminator. Since a diode laser and simple detection circuitry are used, they can be integrated into a low-cost and compact module. The detection circuit can be operated in two modes: sweeping or fixed bias current. For the former case, the injection current, $I(t)$, to the laser diode is swept around the transparent current of the monitored wave-

length. The input light is usually modulated with a pilot-tone signal with a frequency that is much larger than the sweeping frequency of the diode current. Since the induced voltage is approximately proportional to $(I(t) - I_r)P(t)$, where I_r is the transparent current and $P(t)$ is the light power, the wavelength can be monitored by detecting the transparent current from the zero-crossing point of the demodulated signal of the induced voltage. For the latter case, the bias current is fixed at the transparent current for the incident wavelength. When the incident laser wavelength drifts, an induced voltage appears across the electrodes. The detected transparent current for the former case or the demodulated voltage for the latter case serves as the monitoring signal for the network control unit. For wavelength stabilisation of a WDM source, the signal can be fed back to adjust the source wavelength.

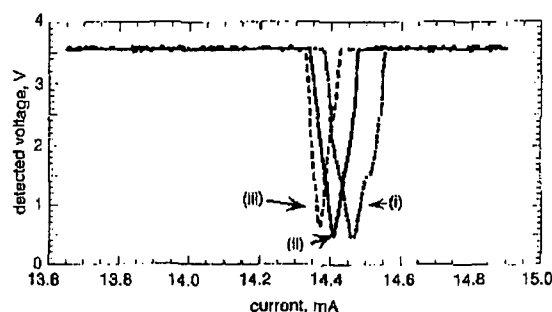


Fig. 2 Detected voltage (after demodulated circuit) against swept current for three wavelengths spaced by 0.1 nm

(i) 1550.1 nm
(ii) 1550.2 nm
(iii) 1550.3 nm

Experiments and results: A high-precision wavelength meter and a tunable laser source were used in the experiments as the wavelength reference and light source, respectively. Since the transparent condition is sensitive to temperature fluctuations, the laser diode that was used as the sensor was attached to a copper plate, the temperature of which was stabilised at 25°C. The laser diode emits light at wavelengths around 1550 nm if it is biased above its threshold current, 22.5 mA. The transparent current decreases monotonically from 20.6 mA at 1520 nm to 10.1 mA at 1580 nm.

For sweeping mode operation, a triangular signal was used to vary the bias current of the laser diode. The sweeping current must be below the threshold current. The induced voltage across the diode electrode was amplified, filtered, and demodulated. A voltage limiter was used for manifesting the transparent point. The dips in the induced voltage and the demodulated signal indicate the transparent condition. The demodulated signals for three wavelengths that differ by 0.1 nm are shown in Fig. 2 and their transparent currents can be clearly distinguished. Therefore, a sub-angstrom wavelength variation can be resolved with this approach.

The resolution of the wavelength discrimination is subject to the error in measuring the transparent current. A larger variation in transparent current for a given wavelength change will give rise to a higher resolution in wavelength sensing; and a small temperature variation rate for transparent current will relax the requirement on temperature control. The measured curve of transparent current against wavelength shows a slope of -0.2 mA/nm at a wavelength of 1550 nm. For this transparent characteristic, the error in measuring the transparent current must be kept within $2 \mu\text{A}$ to achieve 0.01 nm wavelength accuracy. The slope is higher for shorter wavelength and higher temperature [5]. Furthermore, the slope depends on the geometric size and material composition of the active layers and can be optimised for wavelength sensing applications.

For fixed-bias operation, the bias current was first adjusted to the transparent current for the incident wavelength. In this case, the sensitivity of the induced voltage with respect to a wavelength drift is of concern. Fig. 3 shows the induced voltage (before amplification) for three different bias currents when the wavelength varies. The slope is $> 50 \mu\text{V/nm}$ for all the curves. The

induced voltage will increase by $0.5\mu\text{V}$ for 0.01nm of wavelength change. This voltage change can be detected with the amplifying and filtering circuits shown in Fig. 1. For this measurement the light power coupled into the laser diode was estimated to be 0.3mW . Higher input powers will result in a larger slope and higher sensitivity to a given wavelength drift.

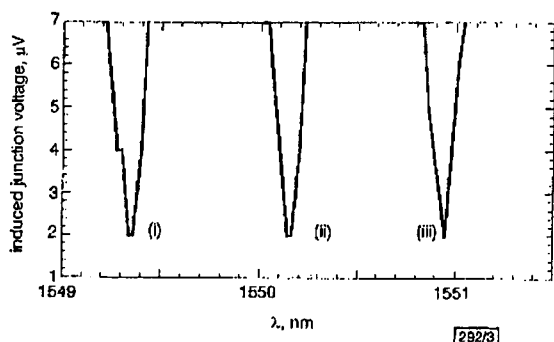


Fig. 3 Variation of induced voltage with wavelength change for different bias currents

- (i) 14.55mA
- (ii) 14.40mA
- (iii) 14.26mA

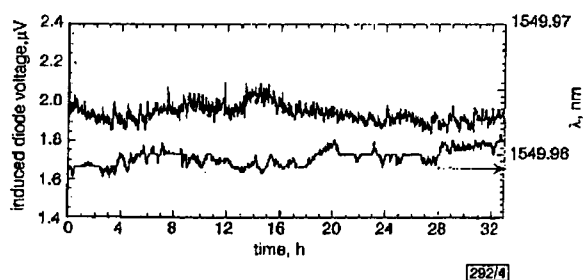


Fig. 4 Induced diode voltage for 33h continuous running when bias was fixed at transparency

Another critical issue for wavelength monitoring is the stability of the wavelength sensor. Since the transparency is a unique property for direct-bandgap optoelectronic devices, it should remain the same if the device temperature can be fixed. Furthermore, because of its low biasing current, compared with a diode laser operated above threshold, the degradation problem should be of less concern. Thus wavelength monitoring using transparency detection can be very stable and reliable. Fig. 4 shows that the induced voltage varies by $< 0.25\mu\text{V}$ over 33h of continuous measurement when the input wavelength is fixed. Referring to Figs. 2 and 3, this variation indicates $\sim 0.005\text{nm}$ change in the input wavelength or $1.0\mu\text{A}$ shift in the transparent current. This degree of variation is acceptable for most DWDM applications. Also shown in Fig. 4 is the input wavelength fluctuation measured with a commercial wavelength meter. Noting that the wavelength varies by at most 0.0035nm during the measurement, this is close to the value implied by the voltage variation.

Summary: We have demonstrated a novel approach for wavelength monitoring using a low-cost laser diode as a wavelength discriminator. The experimental results show that sub-angstrom wavelength resolution can be easily achieved. For applications in wavelength stabilisation, this approach can have very good sensitivity to wavelength drifts. We also demonstrated that the transparent property of a diode laser is stable and reliable.

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References

- 1 ILEGAMI, T., SUDO, S., and SAKAI, Y.: 'Frequency stabilization of semiconductor laser diodes' (Artech House, Boston, MA, 1995) Chap. 5
- 2 NELSON, L.E., CUNNING, S.T., and GILES, C.R.: 'Optical monitoring using data correlation for WDM systems', *IEEE Photonics Technol. Lett.*, 1998, 10, (7), pp. 1030-1032
- 3 COROY, T., MEASURES, R.M., WOOD, T.H., and HURRUS, C.A.: 'Active wavelength measurement system using an InGaAsP-InP quantum well electroabsorption filtering detector', *IEEE Photonics Technol. Lett.*, 1996, 8, (12), pp. 1686-1688
- 4 GUY, M., VILLENEUVE, B., IATRASSI, C., and TETU, M.: 'Simultaneous absolute frequency control of laser transmitter in both 1.3 and 1.55 μm band for multiwavelength communication systems', *J. Lightwave Technol.*, 1996, 14, (6), pp. 1136-1143
- 5 LEE, S.-L.: 'Wavelength measurement and tracking using semiconductor laser amplifiers for applications in photonic networks', *IEEE Photonics Technol. Lett.*, 1998, 10, (3), pp. 439-441

Decoding algorithm for DEC RS codes

Erl-Huei Lu, Chiou-Yng Lee and Ron-Lon Tsai

A step-by-step decoding algorithm is proposed for double-error-correcting Reed-Solomon codes of length $n = 2^m - 1$. Since the decoding algorithm can directly find the error value at the highest-order position in a received vector without determining the error location polynomial, the decoding procedure is very simple and regular; therefore, it is suitable for hardware implementation.

Introduction: A well-known decoding method, the step-by-step decoding algorithm [1-3], can decode a cyclic code in a serial manner with a low hardware complexity, particularly when the method is used for decoding binary cyclic codes. To decode binary cyclic codes, the basic principle of the method is to change one received bit temporarily and then check whether or not the number of error bits in the received vector is reduced by one. If it is, then the decoder corrects the corresponding bit. It then cyclically shifts the received vector one position so as to correct the next bit. Recently, a modified step-by-step algorithm was proposed [4] for decoding double-error-correcting (DEC) binary Bose-Chaudhuri-Hocquenghem (BCH) codes. The decoding algorithm directly determines whether or not the received bit at the highest-order position is correct, without requiring knowledge of the

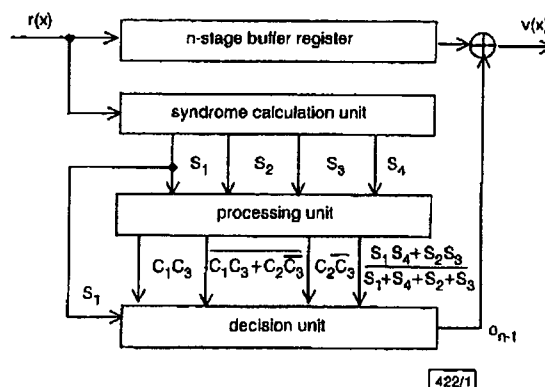


Fig. 1 Hardware structure for decoding DEC RS codes

Operation Principles of Wavelength Sensing Using Transparent Properties
of Semiconductor Optical Diodes

作者：李三良 (S.-L. Lee)，卞敬塘 (C.-T. Pien)，許毓倚 (Y.-Y. Hsu)

發表期刊：IEEE J. Lightwave Technol., vol.19, no.85, pp. 655-665, May
2001.

內容成果：

使用半導體光放大器作為波長感測元件，利用其透明點與入射光波長的關係式，設計波長監控系統，實驗量測監控效能，並與理論相互驗證。波長精密度可達 0.01nm，搭配 FP 雷射進行通道辨識，由於所用元件皆有可積體化的特質，未來可內建於半導體雷射光源內，擁有即時修正波長的功能，對於 DWDM 系統提供波長穩定高的光源，及波長通道品質監控系統。

Operation Principles of Wavelength Sensing Using Transparent Properties of Semiconductor Optical Diodes

San-Liang Lee, *Member, IEEE*, Ching-Tang Pien, and Yu-Yi Hsu

Abstract—The wavelength-sensitive transparent properties of direct-bandgap semiconductor optical diodes (DBSODs) can be utilized for wavelength sensing. A simplified theory is used to describe the operation principles of such wavelength sensing schemes. The principles can be applied to realize different measures, including sensing with or without pilot tone for single wavelength or multiple wavelengths. The simplified theory can also provide an easy way for estimating the wavelength resolution and accuracy. The wavelength resolution is demonstrated to be better than 0.01 nm by detecting the induced junction voltage for a fixed bias. The resolution is limited by the current resolution of the bias source to 0.03 nm when the wavelength is to be discriminated by transparent current detection. The feasibility and limitation of multiwavelength sensing using a single semiconductor optical amplifier are addressed both theoretically and experimentally. The limiting factors, including the power dependency and temperature stability, of the sensing schemes are also investigated.

Index Terms—Laser diodes, semiconductor optical amplifiers, transparent current, wavelength monitoring, wavelength sensing.

I. INTRODUCTION

WAVELENGTH sensing is indispensable for many applications, such as wavelength display for photonic instruments, wavelength calibration for optical characterization, optical monitoring in multiwavelength networks, wavelength control of single-wavelength lasers, and so on. The demand for accurate wavelength sensing increases dramatically due to the explosive applications of dense wavelength-division-multiplexing (DWDM) techniques in high-speed networks [1]–[3]. Since the channel spacing of DWDM systems is 1.6 nm or less, a wavelength sensor in the transmitter can help to control the laser wavelength [4]. Likewise, the sensor can also be placed inside an optical network to serve as a wavelength monitor or even a wavelength comparator for managing the optical channels [3], [5].

The most frequently used approaches for wavelength sensing utilize interferometer-based configurations [6]–[8]. Approaches using optical filters or double-detectors have also been demonstrated [9], [10]. A novel sensing scheme that utilizes the vari-

ation of transparent current with the incident wavelength in a semiconductor optical amplifier (SOA) or a low-cost laser diode (LD) has been demonstrated [11], [12]. The unique advantages of wavelength sensing using a SOA or LD are their compact size and ease of monolithic integration with semiconductor optical components. The key of wavelength sensing using an LD or SOA is the detection of the transparent condition from the induced junction voltage across its electrodes. The voltage characteristics in response to an incident light have been addressed in the literature for exploiting an SOA as an amplified detector or optical gain monitor [13]–[16]. Transparency detection was also used for determining the internal loss of a laser diode [17], [18]. The wavelength dependency of the transparent properties is seldom addressed in the literature. This paper will address in detail the operation principles and limitation of this scheme and compare the variety of detection techniques. The principles can be applied to various direct-bandgap semiconductor optoelectronic diodes (DBSODs), which can provide either gain or absorption onto the input optical signals.

This paper is organized as follows. In Section II, the operation principles of wavelength sensing using a DBSOD will be described in terms of a simplified theory for the transparent condition and the induced junction voltage. Different detection schemes for various cases of input light signals will be addressed. In Section III, we will demonstrate experimentally the performance of the proposed schemes. This is followed by discussion of the feasibility by addressing the sensing resolution and accuracy in Section IV. The required temperature stability and the effect of strong input power on the transparency are described in Section V. The conclusion is drawn in Section VI.

II. OPERATION PRINCIPLES

Wavelength sensing using a DBSOD is based on its wavelength-sensitive transparent characteristics, including the transparent current and the induced junction voltage near the transparency. Since the transparent current is also temperature dependent, similar principles may be applied for measuring the local temperature inside photonic integrated circuits. This paper focuses only on the applications for wavelength sensing. In order to clarify the sensing principles and discuss the feasibility of different approaches, we will derive simplified expressions for describing the transparent characteristics. The simplified approximations can consistently explain the experimental results. The theory that can rigorously describe the transparent properties will be addressed elsewhere [19].

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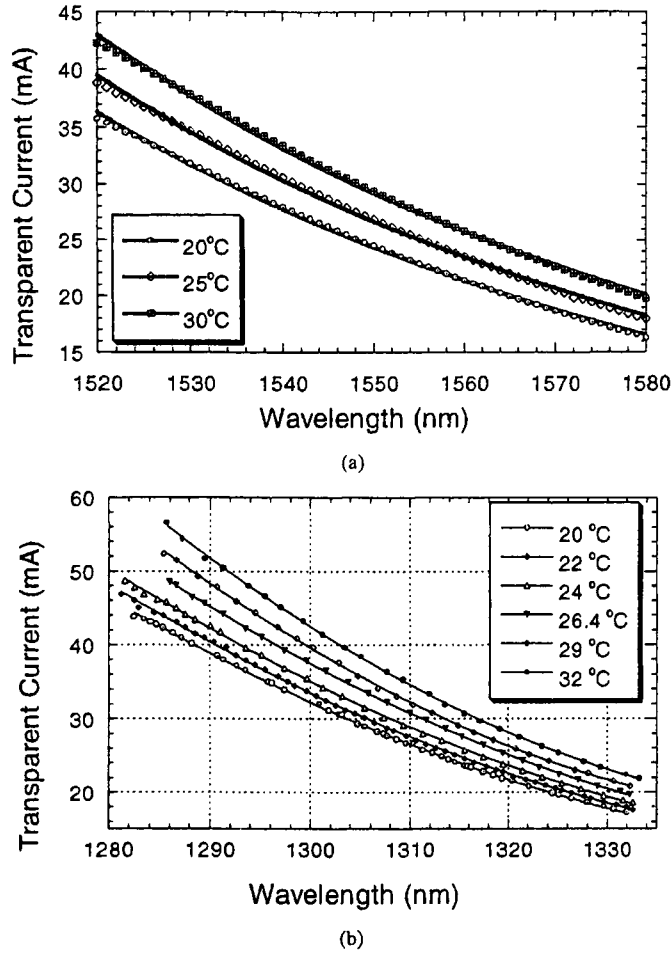


Fig. 1. Measured transparent currents versus wavelength for semiconductor optical amplifiers operating at (a) 1.55 μm and (b) 1.3 μm . The exponential function given by (2) was used for fitting the data in (a).

A transparent condition is defined as the situation that a DBSOD is biased with an injection current at which the net gain is zero. This corresponds to the condition that the separation of the quasi-Fermi energies is equal to the energy of the incident photon [20], so the transparent condition is a function of the incident wavelength λ . Fig. 1 shows the measured transparent characteristics for two examples of SOAs, operating at 1.3 and 1.55 μm , respectively. Such wavelength-dependent transparent characteristics are unique for a given DBSOD, so the incident wavelength can be determined once its corresponding transparent current is measured.

It can be shown that the transparent carrier density N_{tr} is related to the incident light wavelength λ by

$$\frac{hc}{\lambda} - E_g = 2kT \ln \left(\frac{N_{tr}}{N_t} \right) \quad (1)$$

where

- N_t $\sqrt{N_v N_c}$;
- N_c and N_v effective density of states at the conduction and valence band edges, respectively, for bulk materials;
- h Planck's constant;
- c speed of light in vacuum;

- E_g energy gap of the active material;
- k Boltzmann's constant;
- T absolute temperature.

For quantum-well structures, E_g has to be regarded as the energy separation between the lowest quantized states of the conduction band and valence band. Besides, N_c and N_v must be replaced with $N_c = kTm_e/\pi\hbar^2 L_z$, and $N_v = kTm_h/\pi\hbar^2 L_z$, where m_e and m_h are the effective mass of electrons and holes, respectively; and L_z is the thickness of quantum wells [21].

To make the following analysis tractable and to gain clear insight of the sensing principles, (1) was obtained with the following assumptions. First, Boltzmann's approximation is used to describe the relation between the carrier density and Fermi energy. Secondly, the active layer is undoped such that the electron density is the same as the hole density. Lastly, the position variation of carrier density and photon density inside the DBSOD is neglected. With further approximation of the transparent carrier density with a linear function of the transparent current as $N_{tr} = \eta_i I_{tr} \tau / qV_a$, the variation of transparent current with wavelength can be described as

$$I_{tr} = I_t \exp \left(\frac{E - E_g}{2kT} \right), \quad E \geq E_g \quad (2)$$

where

- $E = hc/\lambda$ photon energy;
- I_t $N_t q V_a / \eta_i \tau$;
- V_a volume of the active material;
- τ differential carrier lifetime;
- η_i internal quantum efficiency.

Equation (2) indicates that the transparent current increases exponentially with the photon energy. The sensor requires a larger injection current to achieve transparency for an incident light with shorter wavelength. In spite of the simplification in deriving (1) and (2), the equation can describe the functional dependency of the measured data very well. This can be observed from Fig. 1(a) by comparing the measured data with the curve fittings in the functional form of (2).

The transparent condition can easily be detected from the induced voltage across the diode junction, which is related to the separation between the quasi Fermi energies. The Fermi levels vary with the carrier density, which in turn varies with the photon density. When the diode is biased below transparency, the diode acts as an absorber and the carrier density increases with the light power due to the increasing generation of electron-hole pairs. On the other hand, when it is biased above transparency, the carrier density decreases with the light power from the increasing stimulated emission. Only at transparency are the carrier density and the induced voltage independent of the light power.

When the series resistance R_s of the diode is considered, the voltage between the diode electrodes can be written as

$$V_{ext} = IR_s + \frac{E_{fc} - E_{fv}}{q} \approx IR_s + \frac{E_g}{q} + 2V_T \ln \left(\frac{N}{N_t} \right) \quad (3)$$

where E_{fc} and E_{fv} are, respectively, the quasi-Fermi energies for electrons and holes; $V_T = kT/q$ is the thermal voltage. The second equality of (3) was obtained using the same assumptions

as those for (1). With a time-varying light traveling inside the active material, the carrier density can be determined from the rate equation [13]. By assuming the variation rate of carrier density to be much smaller than N/τ , the carrier density can be written implicitly as

$$N(t) = N_0 - [N(t) - N_{tr}(\lambda, T)]\bar{p}(t) \quad (4)$$

where

$$\begin{aligned} \bar{p}(t) &= P(t)/P_s && \text{normalized version of the input power} \\ &P(t); && \\ P_s &EA/\Gamma a\tau; && \\ N_0 &= \eta_i I(t)\tau/qV_a && \text{injected carrier density.} \end{aligned}$$

We have assumed a linear gain coefficient $g = a(N - N_{tr})$. Γ and A are the confinement factor and cross-sectional area of the active region, respectively. It is clear from (4) that the incident light depletes carrier density when the bias is above the transparency, while it adds carriers when it is below transparency.

The transparent characteristics of a DBSOD can be applied to the situation that the input light contains a single wavelength or multiple wavelengths. Furthermore, a pilot tone may or may not be used to modulate the input light, depending on the applications. Different cases of wavelength sensing will be discussed in the following.

Case 1—Single-Wavelength Signal Without Pilot Tone: Without pilot tone in the incident light, the transparent condition can be discriminated by subtracting the “dark” junction voltage, the voltage without light, from the induced junction voltage with light incidence. This voltage difference for a given injection level N_0 can be obtained from (3) and (4) as

$$\Delta V \equiv V_{ext} - V_{ext0} = 2V_T \ln \left(\frac{1 + \frac{N_{tr}}{N_0} \bar{p}}{1 + \bar{p}} \right) \quad (5)$$

where V_{ext0} is the dark junction voltage that can be measured in advance for a given DBSOD. The voltage difference vanishes when $N_0 = N_{tr}$, so the transparent current for a given input wavelength can be determined.

Wavelength sensing without using a pilot tone can be performed either by sweeping the bias current to locate the transparent condition or by measuring the junction voltage for a fixed bias current. For the former approach, the corresponding wavelength for an incident light can be determined from the current value that gives rise to a zero-crossing in the voltage difference, in response to the sweeping bias. Since it is relatively difficult to track precisely the dark voltage for a sweeping bias, the latter approach is usually favored for sensing without a pilot tone. For a fixed bias, the voltage difference given by (5) itself varies with the input wavelength, so wavelength discrimination can be performed from measuring the voltage difference. This approach requires fixing the input power at a designated value due to the power dependence of the voltage difference.

Case 2—Single-Wavelength Signal with Pilot Tone: When the incident light is modulated with a pilot tone of single frequency ω , the normalized optical power $\bar{p}$ in (5) is written as $\bar{p}(t) = \bar{p}_0(1 + m \cos(\omega t))$, where m is the modulation index

and $\bar{p}_0$ is the normalized average power. The transparent point can directly be detected from the induced ac voltage without the necessity of referring to the dark junction voltage. Two detection approaches can be used to extract the junction voltage: coherent detection and envelope detection [22]. For coherent detection, a local oscillator with angular frequency ω is used to demodulate the ac response. The resultant signal amplitude is proportional to

$$v_0 = \eta_v (N_{tr} - \bar{N}) m \bar{p}_0 \quad (6)$$

where the carrier density is expressed as the sum of a steady-state value $\bar{N}$ with a small-signal response $n(t)$; and

$$\eta_v \equiv \frac{1}{q} \frac{\partial(E_{fc} - E_{fv})}{\partial N} \approx \frac{2V_T}{N_0}. \quad (7)$$

The second equality in (7) exists for low-level injection. Noting that $\bar{N}$ approaches N_0 when the bias is close to the transparent condition, (6) shows that the induced voltage vanishes at transparency, so the transparent point can be detected from the zero-crossing point of the induced ac voltage. For envelope detection, a bandpass filter centered at ω is utilized to remove the dc and higher order harmonic components of the induced junction voltage. An envelope detector then detects the output signal. The resultant signal amplitude is similar to the expression in (6), but $N_{tr} - \bar{N}$ is replaced with $|N_{tr} - \bar{N}|$. Thus, the transparent condition corresponds to the bias current where the output voltage is minimal.

Similar to Case 1, wavelength sensing for an input light can be performed either by using a sweeping signal to locate the transparent current or by detecting the amplitude of the ac voltage for a fixed bias. For the latter approach, the ratio of the voltage amplitudes at two different biases can also be used for eliminating the power dependency. The logarithmic voltage ratio can easily be obtained from (6) as

$$\rho \equiv \ln \left(\frac{v_A}{v_B} \right) \approx \ln \left(\frac{1 - N_{tr}(\lambda)/\bar{N}_A}{1 - N_{tr}(\lambda)/\bar{N}_B} \right) \quad (8)$$

where the subscripts A and B refer to two bias levels of N_A and N_B , respectively. When the wavelength of the input light results in a transparent carrier density that is close to N_A , the voltage ratio approaches to $-\infty$. On the other hand, the voltage ratio reaches infinity if the corresponding transparent carrier density is equal to N_B . As the corresponding N_{tr} for the measured range of wavelengths falls between N_A and N_B , one of v_A and v_B increases while the other decreases as the wavelength becomes longer. Thus, there exists a monotonic change of the voltage ratio with the wavelength.

In practical applications, the voltage ratio can be obtained with one or two sensors. For the one-sensor case, a switching current source has to be used to bias the sensor at two current levels. When two sensors are used, the input power is coupled to the two sensors such that the voltage across each sensor can be measured.

Case 3—Multiple-Wavelength Signal with Same Modulation: The transparent detection principle can be applied to the cases that the input light contains multiple wavelengths, which may come from multiple single-wavelength lasers

or a Fabry–Perot (FP) laser. If the input light as a whole is modulated by the same single-tone signal, and the induced voltage across the diode junction is coherently detected, the output voltage will be proportional to

$$v_0 = \eta_v \sum_i (N_{tr,i} - \bar{N}) m_i \bar{p}_i. \quad (9)$$

It can be shown that the output voltage is zero as the injection level reaches the mean transparent carrier density for the input wavelengths, weighed by the optical power, i.e.,

$$N_0 = \frac{\sum m_i P_i N_{tr,i}}{\sum m_i P_i} \quad (10)$$

where the subscript i refers to the i th wavelength. Envelope detection can also be applied, and the resultant signal amplitude is similar to the expression in (9), but $\sum (N_{tr,i} - \bar{N})$ is replaced with $|\sum (N_{tr,i} - \bar{N})|$. Thus, the bias current where the output voltage is minimal again indicates the mean transparent carrier density. If all the input wavelengths have the same modulation index and fall within a small wavelength range such that the transparent carrier density can be approximated as a linear function of the wavelength, the detected average transparent current corresponds to the mean wavelength of the input signals [23].

Case 4—Multiple-Wavelength Signal with Separate Tone Modulation: For applications like wavelength tracking [3], [11], wavelength comparison, and wavelength monitoring of wavelength-division-multiplexing (WDM) signals [5], multiple light waves might simultaneously incident upon a sensor. Each light carries an individual wavelength and is modulated with a separate pilot frequency. In such a case, the transparent current for each wavelength can be recognized by using a separate bandpass filter centered at each pilot-tone frequency to extract the harmonic signals. The induced ac signal across the diode junction can be written as $v(t) = \sum_i v_i \cos(\omega_i t)$, where v_i is the signal amplitude at the i th pilot-tone frequency. This can be regarded as a general case of Case 2, so (6) can be used to represent the amplitude at each filter output, while subscript i has to be added to the symbols for identifying the wavelength. The steady-state carrier density can be written as

$$\bar{N} = \frac{N_0 + \sum N_{tr,j} m_j \bar{p}_j}{1 + \sum m_j \bar{p}_j}. \quad (11)$$

The bias level to zero the output voltage at the i th filter can be shown to be

$$N_0 = N_{tr,i} + \sum_j (N_{tr,i} - N_{tr,j}) m_j \bar{p}_j. \quad (12)$$

The bias level for zero voltage indicates the transparent condition for the i th wavelength when all the input light signals are small enough that $\bar{N}$ is approximately equal to the injection level N_0 . In this condition, the transparent current for the i th wavelength can be located by sweeping the bias current density such that the voltage output at the i th filter vanishes. Equation (12) reveals also that the detected transparency may be affected by the presence of other signals. This effect will be investigated in the next section.

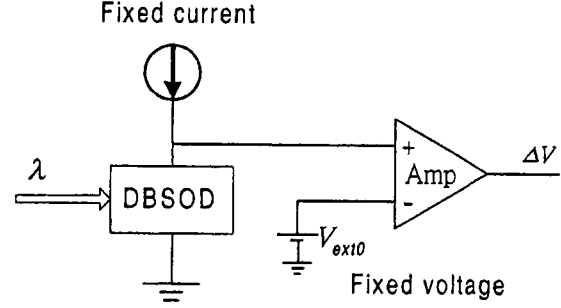


Fig. 2. Block diagram of wavelength sensing without using a pilot tone. V_{ex10} simulates the dark voltage for a given bias current.

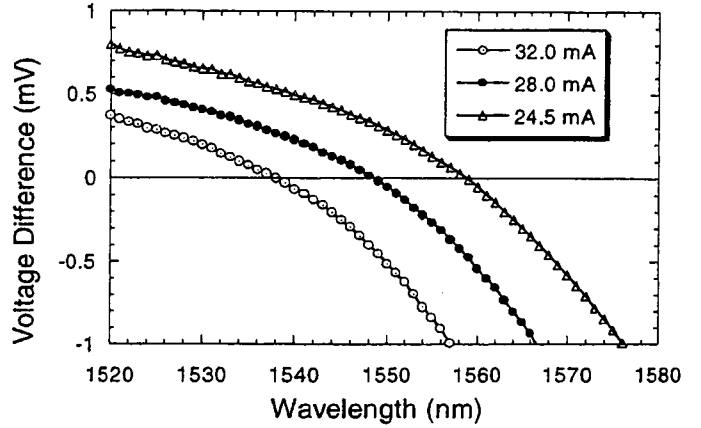


Fig. 3. The variation of the measured voltage difference ΔV with the incident wavelength for three different bias currents.

III. EXPERIMENTAL DEMONSTRATION

To demonstrate the performance and limitation of the above wavelength-sensing schemes, we present here the empirical results out of the SOA that has the characteristics shown in Fig. 1(a). We have performed the wavelength-sensing experiments using different types of sensors, including SOA, FP-LD, and distributed feedback (DFB)-LD, and the results show very similar behavior.

A. Sensing without Pilot Tone

To obviate the need to use an external optical modulator is a major advantage of the sensing without pilot tone. The light is directly incident upon the sensor and the measured junction voltage is compared with the dark voltage, which has to be measured in advance. Fig. 2 shows the schematic of the sensor, at which differential detection is exploited to manifest the voltage difference. The measured data shown in Fig. 3 indicate that the voltage difference decreases monotonically with the wavelength. The curves have larger absolute value of slope at the long wavelength side, where better resolution in wavelength sensing can be achieved. Different bias current to the SOA can be used to adjust the curve in order to cover different ranges of wavelengths.

In spite of potential larger sensing error arising from the $1/f$ noise and thermal fluctuation, sensing without pilot tone can still be useful for low-cost wavelength sensing and stabilization. For wavelength stabilization, the reference voltage is set as the

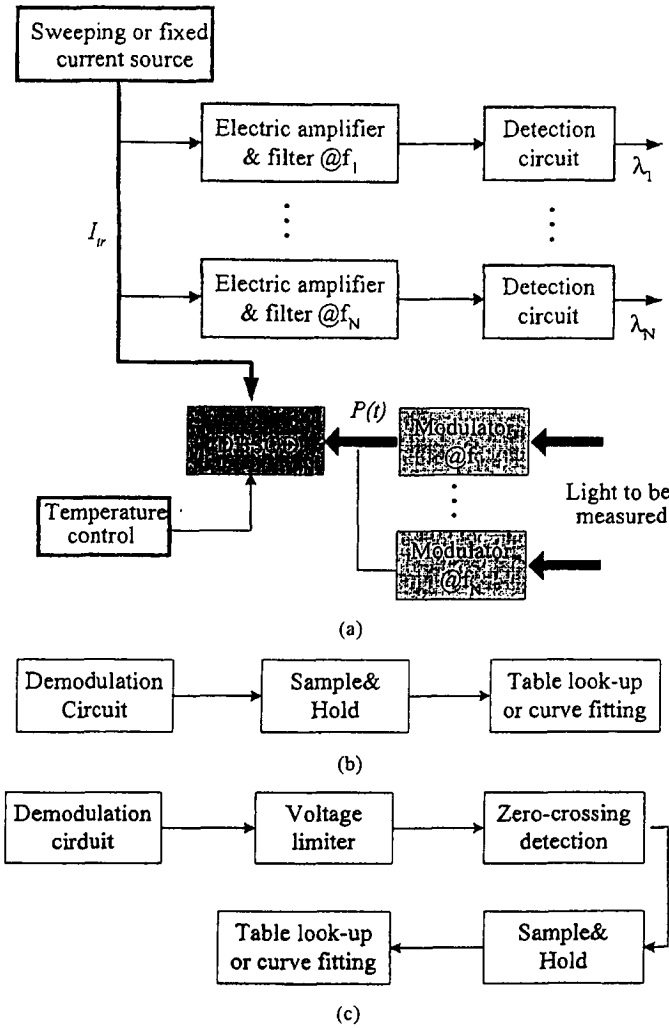


Fig. 4. Block diagram of wavelength sensing with pilot tones. (a) General diagram for multiple single-wavelength lights. Single wavelength input is a special case. (b) Detection circuit for sensing the ac voltage at a fixed bias current. (c) Detection circuit for locating the transparent point from a sweeping bias current.

junction voltage for a given wavelength such that the voltage at the differential output vanishes. As the wavelength drifts, a voltage will appear at the output that can be used for wavelength adjustment.

B. Single-Wavelength Light with Pilot Tone

The block diagram for wavelength sensing with pilot tone is depicted in Fig. 4(a). Again, the incident wavelength can be resolved from detecting either the transparent current or the junction voltage. The detection circuits for both approaches are illustrated in Fig. 4(b) and (c), respectively. For the former case, the injection current to the DBSOD sensor is swept over a range that covers the transparent current of the interested wavelengths. The input light is modulated with a pilot-tone signal of which the frequency is usually much larger than the sweeping frequency. The transparent current can be detected from the zero-crossing point of the demodulated signal of the induced voltage. The sensor can also be applied for wavelength stabilization, of which the bias is fixed at the transparent current for the incident wavelength.

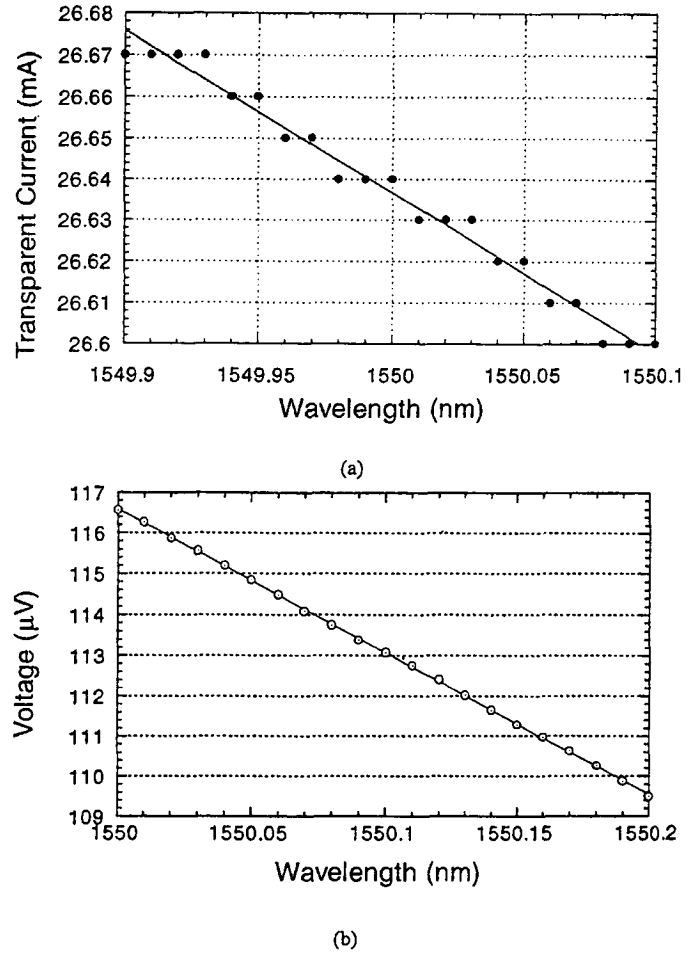


Fig. 5. Wavelength variation of (a) transparent current that was obtained with a sweeping bias and (b) induced junction voltage at a fixed bias (25 mA).

We have demonstrated subangstrom resolution of wavelength sensing using an LD as the sensor [12]. In order to investigate the resolution limit, the detected voltage across the SOA is measured for a set of wavelengths spaced by 0.01 nm. Fig. 5(a) shows that the current for minimal voltage can not be distinguished for certain adjacent wavelengths due to the resolution limitation of the current source (0.01 mA). The data shown in Fig. 1(a) indicate a slope of 0.33 mA/nm around 1550 nm for the I_{tr} versus λ curve at 25 °C. Therefore, this sensor can only resolve wavelengths in spacing larger than 0.03 nm for 0.01-mA current resolution, which is consistent with the data shown in Fig. 5(a). A current source with better resolution is required to improve the wavelength resolution.

When the wavelength is to be resolved from the induced junction voltage, the fixed current can be bias away from the transparent point for obtaining good signal-to-noise ratio. Fig. 5(b) demonstrates the voltage change, measured with a lock-in amplifier, for a fine scale of wavelength shift. The induced voltage changes by 0.35 μV in average for 0.01 nm of wavelength drift. The fluctuation in the measured voltage is less than 0.1 μV, so a wavelength resolution better than 0.01 nm can be achieved with this detection approach. Further measurement reveals that the fluctuation comes partly from the detection noise (around 0.03 μV in rms) and the fluctuation in the input power and polarization. Since the detected voltage is proportional to the input

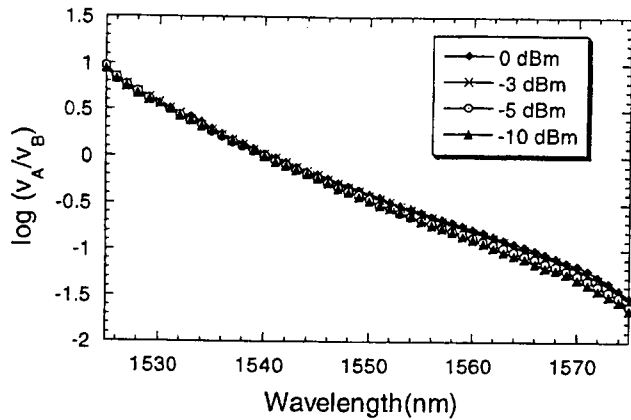


Fig. 6. The variation of the measured voltage ratio with the incident wavelength for different input power levels. The voltage ratio was obtained as the ratio of the induced voltage at 18.06-mA bias current to that at 39.25 mA.

power, this approach requires accurate monitoring of the input power.

In order to minimize the power dependency in the above approach, the ratio of the demodulated voltages at two different biases can be exploited for wavelength discrimination. The current source for biasing the sensor has to be replaced with a two-switching-level bias current; and the wavelength is determined from the quotient of the induced voltages for the two bias levels. Fig. 6 shows the curves of voltage ratio for two bias levels: 18.06 and 39.25 mA. The curves for different input powers are nearly identical at the short-wavelength side but deviate slightly at the long-wavelength side due to gain saturation. The denominator of the voltage ratio is the induced voltage for the bias of 39.25 mA, where the corresponding transparent wavelength is about 1519 nm. When the wavelength becomes larger than 1519 nm, the bias current of 39.25 mA results in amplification to the input signal. The amplification increases with the wavelength. As the input power is strong enough to cause gain saturation, the induced voltage for longer wavelength suffers from stronger gain suppression and gives rise to a larger voltage ratio. This tendency is consistent with the measured data. If the wavelength sensing requires the voltage ratio to be power independent, the input wavelength must fall within a limited range or the input power must be weak enough.

C. Mean Wavelength of Multiple Spectral Lines

The measurement was first performed using one tunable laser and a DFB laser of which the wavelength is 1550.1 nm. The two lasers were combined with an optical coupler, modulated by an optical modulator at 1-kHz frequency, and coupled into the SOA sensor. The input power to the SOA is about -7 dBm from both lasers. For the lone laser light, the transparent current is 26.7 mA for the DFB laser wavelength and 25.0 mA for the tunable laser, of which the wavelength was set to 1555 nm. When both lasers are present, the bias current to have minimal induced voltage lies between the transparent currents for the two lone lasers. The averaging effect is illustrated in Fig. 7(a). The curve for two lasers together has a slope twice that for a lone laser; and the voltage minimum occurs at the current where the voltages for the two lone lasers have equal amplitude. This can be

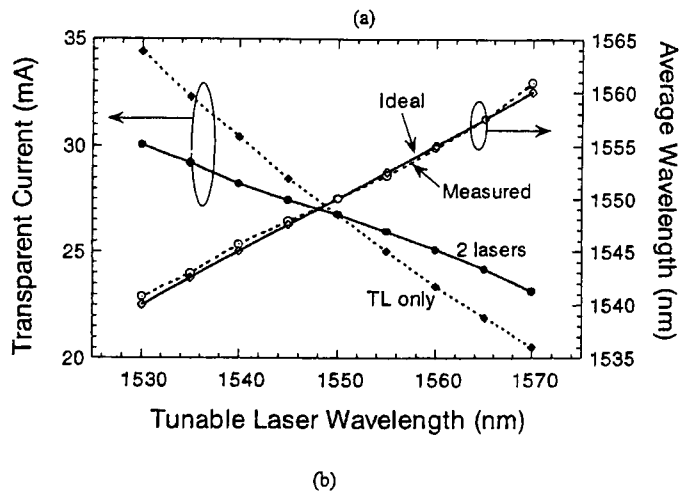
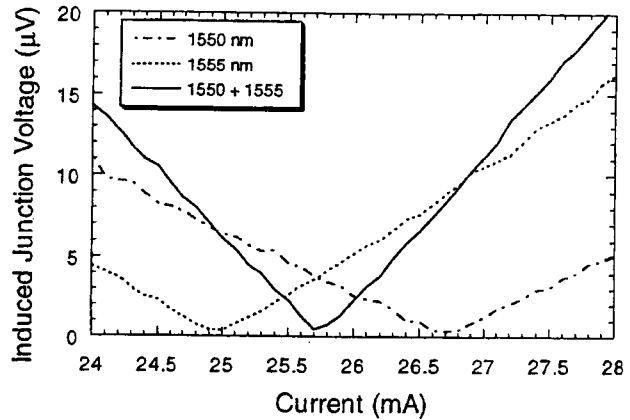


Fig. 7. (a) Induced junction voltage for two single-wavelength input lights that are modulated together at a pilot-tone frequency. (b) Resultant transparent currents and average wavelengths versus tunable laser wavelength.

clearly seen from (9). Upon varying the wavelength of the tunable laser, Fig. 7(b) summarizes the measured bias current that gives rise to minimum junction voltage and the corresponding mean wavelength. Also shown in the figure is the ideal average wavelength for the two lasers. It can be found that the measured average wavelengths are very close to the ideal values.

Next, this method was applied to measure the mean wavelength of the multiwavelength light from an FP laser. A useful application of this measurement is to use the wavelength sensor inside an optical power meter for wavelength calibration. With this built-in sensor, the wavelength-dependent responsivity of the photodiode can be automatically calibrated, no matter what type of incident light. In the experiment, the output of the FP laser was externally modulated with a pilot tone and coupled into the SOA. The corresponding wavelength was extracted from Fig. 1(a) once the transparent current is obtained. To investigate the sensing accuracy, a commercial wavelength meter was used to measure the mean wavelength simultaneously. Fig. 8 shows the measured mean wavelengths when the driving current of the diode source varies. It is well known that the lasing spectrum of an FP laser changes as its injection current varies. The mean wavelength measured with an SOA sensor is very close to the result obtained by the wavelength meter. The error bars are included in the figure to indicate the fluctuation in the measured transparent points. This fluctuation is believed

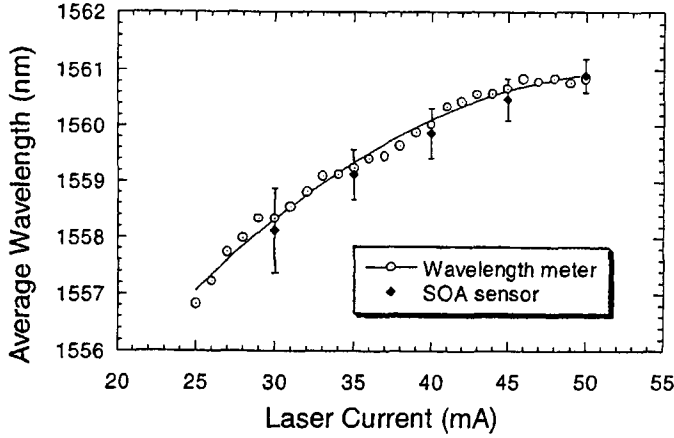


Fig. 8. Comparison of the mean wavelength for an FP laser, measured with an SOA sensor and a commercial wavelength meter.

to be caused by the mode competition associated with FP lasers and was also observed in the measurement with the commercial wavelength meter.

D. Multiwavelength Inputs with Separate Modulation Tone

When the light for each wavelength is modulated with a separate pilot tone, the wavelength-sensing approach can be applied for measuring multiple wavelengths at the same time. We demonstrated this using again a DFB laser and a tunable laser, which were modulated at 1.0 and 13.3 kHz, respectively. The induced ac voltage for each wavelength was detected using an individual bandpass filter centered at the corresponding pilot frequency. The transparent current for each wavelength was located where the junction voltage is minimal for a sweeping bias. The signals shown in Fig. 9(a) include the induced voltages for the lone lasers as well as for the two lasers together. This figure indicates the possibility of resolving the wavelengths of multiple lasers using a single SOA.

A critical concern for multiwavelength sensing is that the detected transparent current may be affected by the presence of the other signal, as pointed out in (12). When the bias of the sensor is set at the transparency of one wavelength, the light at the other wavelength is subject to amplification or absorption, depending on its wavelength. For example, if the other wavelength is longer than the transparent wavelength, this light will be amplified and hence depletes the carrier density. This decreases the steady-state carrier density such that the wavelength at transparency requires a larger bias current to reach transparency. Thus, crosstalk appears in the measured wavelength. This crosstalk is proportional to both the optical power and the wavelength difference between the signals. Fig. 9(b) shows the shift in the transparent current for the 1550.1-nm signal with the power and wavelength of the other incident light. The accuracy of using an SOA to measure multiple wavelengths is rather limited due to the crosstalk. The upper power limit of the input signals for achieving a given wavelength accuracy will be discussed next.

IV. WAVELENGTH ACCURACY AND RESOLUTION

We have demonstrated subangstrom resolution and good stability of wavelength sensing using the proposed technique [11],

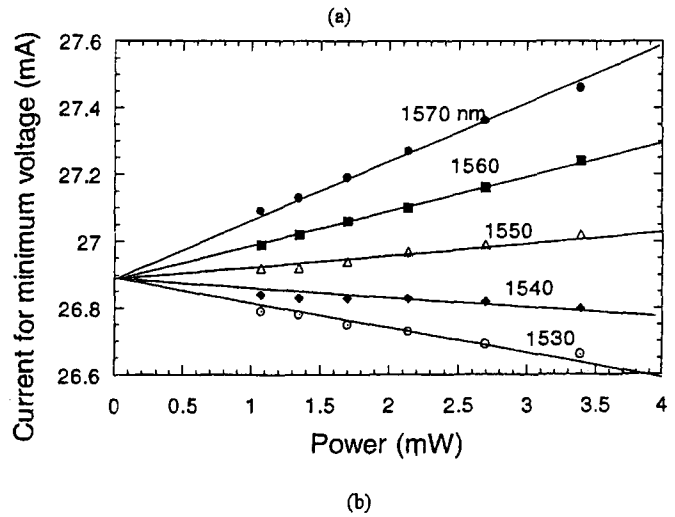
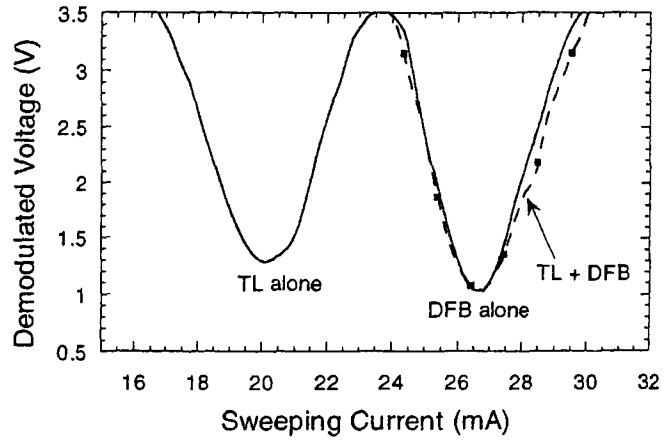


Fig. 9. (a) The response of the demodulated voltage to a sweeping current source for two lone lasers and for two lasers together and (b) crosstalk effects on the measured transparent current for the DFB laser versus input power and wavelength of the tunable laser. The DFB laser and tunable laser were modulated at 1 and 13.3 kHz, respectively.

[12]. Here, we would like to investigate the ultimate limitation of the sensing performance. Only the performance of sensing with pilot tone is discussed here due to the potential larger noise and fluctuation for sensing without a pilot tone.

A. Single Wavelength Sensing with Transparency Discrimination

For wavelength sensing by means of locating the transparency point, its resolution is limited by the resolution of the injection current and the slope of the I_{tr} versus λ curve. This can simply be described by

$$\Delta\lambda = \frac{\partial\lambda}{\partial I_{tr}} \Delta I \quad (13)$$

where ΔI is the resolution of the sweeping current. A larger slope of the I_{tr} versus λ curve will give rise to a better resolution in wavelength sensing for a given ΔI . The slope can be found from (2) to be proportional to I_{tr} , i.e.,

$$\frac{\partial I_{tr}}{\partial \lambda} \approx -\frac{E}{2kT} \frac{I_{tr}}{\lambda}. \quad (14)$$

Designing the DBSOD to have a larger transparent current will be beneficial to the resolution of wavelength sensing if the resolution is limited by the setting resolution of the current source. Equation (14) reveals also that wavelength sensing in shorter wavelength can have better resolution. For wavelength sensing around 1550 nm, the wavelength resolution can be estimated from $\Delta\lambda \approx 100 \text{ nm} \times \Delta I/I_{tr}$. That is, to obtain a resolution of 0.01 nm would require the current source to have a current resolution of 100 ppm. For most commercial current sources, the current resolution is limited at 0.01 mA. Thus, a transparent current as high as 100 mA is needed to achieve 0.01 nm of resolution. However, it should be mentioned that a current source operating at higher current values is usually subject to poor resolution.

Since $\delta\lambda$, the error in wavelength sensing, is proportional to the error in measuring the transparent current δI_{tr} , the worst case wavelength error can be described in a similar form as (13), with ΔI_{tr} being replaced by δI_{tr} , which can be described as

$$\delta I_{tr} = \delta I + \left(\frac{\partial v}{\partial I} \bigg|_{I_{tr}} \right)^{-1} \delta v + \frac{\partial I_{tr}}{\partial T} \delta T \quad (15)$$

where δI , δv , and δT are the fluctuations associated with the driving current, detected voltage, and temperature, respectively. Again, a larger slope in the I_{tr} versus λ curve results in a smaller wavelength error for a given error in measuring the transparent current. High-precision wavelength measurement can be achieved by using low-noise driving source and detection circuit, as well as using stable temperature control.

As the sensor is operated around its transparency, the noise comes mainly from the shot noise in the detection process and the thermal noise from the series and load resistance. For sensing with a pilot tone, a sufficiently narrow bandpass filter can be used to reject most of the noise associated with the induced voltage.

B. Single-Wavelength Sensing with Voltage Discrimination

When the incident wavelength is resolved from the detected ac voltage across a DBSOD instead of sensing the transparent current, the resolution and accuracy is determined by how precise the voltage measurement can be. That is

$$\Delta\lambda = \left[\frac{\partial v_0}{\partial \lambda} \right]^{-1} \Delta v_0. \quad (16)$$

Deriving from (6), the resolution in voltage measurement can be estimated from

$$\frac{\Delta v_0}{V_{bias}} \approx \frac{I_{tr}}{I_0} \frac{\Delta\lambda}{\lambda} m\bar{p}_0 \quad (17)$$

where V_{bias} is the bias voltage across the sensor for a bias current I_0 . It is clear that the sensing resolution is enhanced for a stronger light input. Fig. 10 shows the required resolution or accuracy on voltage measurement for achieving a given wavelength resolution. For a wavelength resolution of 0.01 nm, the voltage resolution must be better than 0.6 μV when $m\bar{p}_0 = 0.1$. This voltage resolution is achievable using a lock-in amplifier.

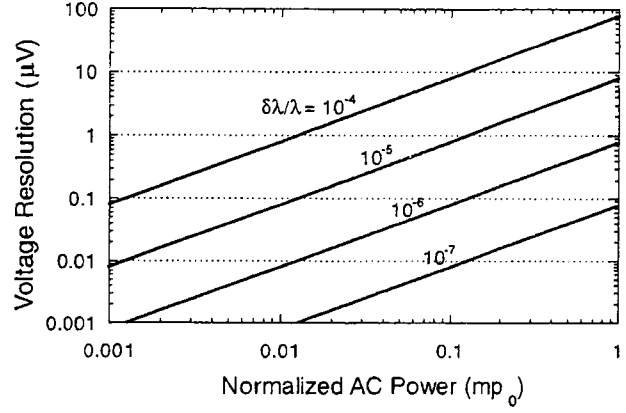


Fig. 10. Required voltage resolution versus input light power to achieve a given wavelength resolution ($I_{tr} = I_0$).

Following the description in Section II, the worst case error in the measured voltage δv_0 can in general be written as

$$\delta v_0 = \delta v + \frac{\partial v_0}{\partial I} \left(\delta I + \frac{\partial I_{tr}}{\partial T} \delta T \right) + \frac{\partial v_0}{\partial P} \delta P \quad (18)$$

where δP is the fluctuation in the incident power and δv is purely the noise-induced voltage fluctuation. One of the merits of resolving the wavelength from the induced voltage rather than from the transparent current is that the measurement is undertaken at better signal-to-noise ratio if the bias current is away from the transparency. However, at this operation condition, the voltage is sensitive to power fluctuation because of the linear proportionality between $\partial v/\partial P$ and $N - N_{tr}$. It can be shown that the wavelength error due to the power fluctuation can be estimated from

$$\frac{\Delta\lambda}{\lambda} \approx \frac{2kT}{E} \left(\frac{I}{I_{tr}} - 1 \right) \frac{\delta P}{P}. \quad (19)$$

To keep the wavelength error below 0.01 nm requires the power stability of the input light to be better than 0.004 dB when the bias level is shifted 10% away from the transparency. Such requirement in power stability is tough for a typical light source. It can be relaxed by biasing closer to the transparency or by differential detection using two sensors.

C. Multiwavelength Sensing

The principle of measuring multiwavelength signals with separate dithering on each wavelength is similar to the detection of single-wavelength signal, so (13) and (14) can be used to describe the resolution of multiwavelength sensing. However, the multiwavelength sensing is subjected to an additional degradation on the wavelength accuracy from the crosstalk effects. The wavelength deviation of the i th channel due to the presence of other channels can be estimated from

$$\frac{\delta\lambda}{\lambda} = \frac{2kT}{E} \sum_j \left(1 - \frac{N_{tr,j}}{N_{tr,i}} \right) m_j \bar{p}_j. \quad (20)$$

The worst cases happen when all the crosstalk channels contribute the same sign of wavelength deviation.

Equation (20) indicates that the multiwavelength sensing is only feasible for a few closed-spaced channels with low input power. For example, the worst case power limitation of each channel is about -14 dBm for measuring eight 0.8-nm spaced channels around 1550 nm with a wavelength accuracy of 0.01 nm, assuming 10% of modulation index and 10 mW of saturation power. If the wavelength accuracy can be relaxed to 0.1 nm, the crosstalk from a higher input power (-4 dBm) can be tolerated. This level of wavelength accuracy may find applications as channel monitors for recognizing the channel appearance or failure in managing DWDM networks. It is also applicable in wavelength comparison or wavelength tracking where only the relative difference in wavelength is relevant and a larger amount of crosstalk can be tolerated.

V. DISCUSSIONS

As stated in previous sections, the transparent current is sensitive to temperature variation. How stable the temperature has to be controlled for achieving a certain degree of wavelength accuracy depends on the change rate of the transparent current with the temperature. For instance, the thermal change rate of the transparent current shown in Fig. 1(a) is about 0.5 mA/°C at 1550 nm and 25 °C. Wavelength sensing around this condition requires better than 0.006 °C temperature stability to keep the wavelength fluctuation smaller than 0.01 nm. Such temperature stability can be obtained with commercial temperature controller.

Due to the thermal effect, the sensing scheme can have better accuracy in the environment where the temperature is well conditioned. Otherwise, the temperature gradient between the SOA active region and the heat sink may cause fluctuation in the measured wavelength. If the sensor is to be operated over a wide range of ambient temperature, the sensor must be packaged with separate temperature control on the SOA chip and the case. This can be done by attaching the SOA, which is hermetically packaged with internal thermoelectric coolers, on another temperature-controlled plate, as illustrated in Fig. 11(a). Thermal isolation and forced convection of heat may have to be pursued as well. A double layer of temperature control and thermal isolation can keep the SOA chip from the variation of ambient temperature. The data shown in Fig. 11(b) indicate that the transparent current varies 0.012 mA/°C if the case temperature is allowed to change with the ambient temperature. When the case is temperature controlled and isolated from the outside temperature, the transparent current remains nearly fixed (with less than 0.01-mA fluctuation) over the temperature range from 25 to 70 °C.

Another critical concern of the sensing approach is the effect of the incident power on the transparent current. In principle, at transparent condition, the active material is balanced between stimulated amplification and absorption and is thus insensitive to the incident power. However, the light may be absorbed due to the free-carrier absorption as well as the nonuniformity in the material composition and carrier distribution. The absorption of light will raise the temperature around the active region and increase the transparent current. This effect will set a upper limit

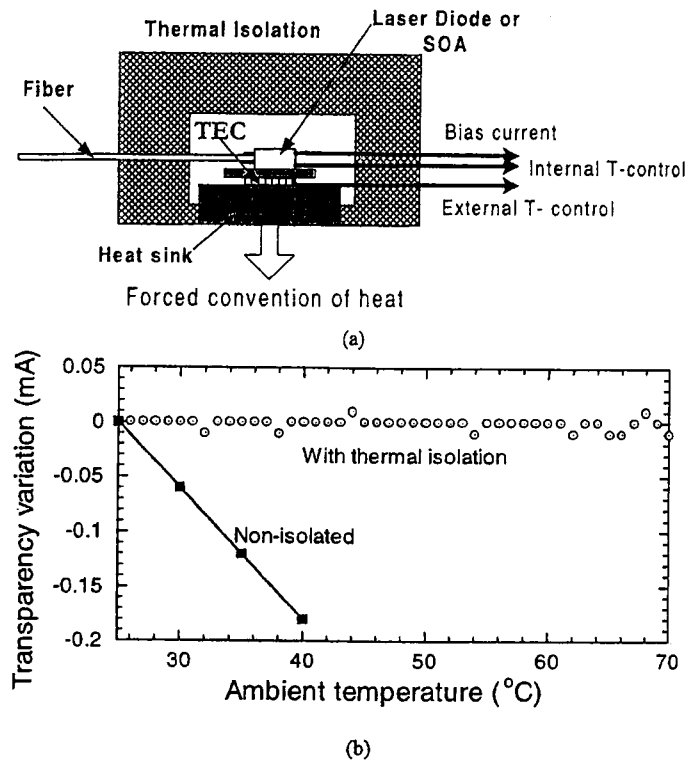


Fig. 11. (a) Schematic of double-layer temperature control for stabilizing the sensor temperature and (b) transparent current stability when the case temperature is controlled or allowed to vary with the ambient temperature. The input wavelength is 1550 nm.

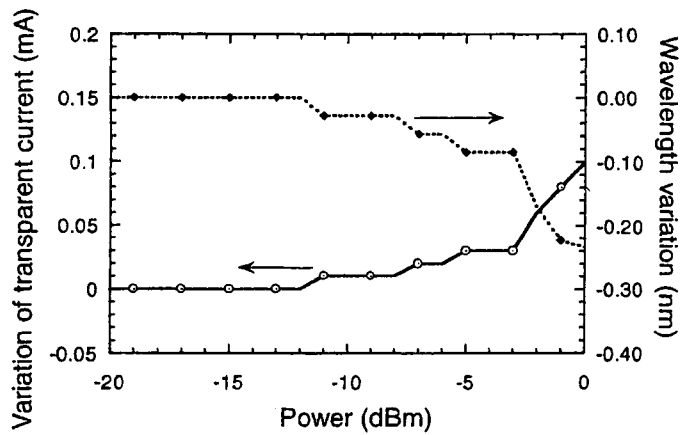


Fig. 12. Variation of measured transparent current and the corresponding wavelength shift with the input light power. The input wavelength is 1550 nm.

on the allowable incident power when the input power is not fixed.

Fig. 12 depicts the variation of the transparent current with the incident light power. The transparency remains unchanged for a power level below -12 dBm and deviates about 0.03 mA, which corresponds to about 0.1 nm wavelength shift, for up to -3 dBm input power. The minimum detectable power is about -20 dBm for the SOA sensor due to the detection noise and the fluctuation of the incident light. The wavelength sensing technique can still be applied for the case of high input power if the input light power is monitored or even stabilized.

VI. CONCLUSION

We reported the operation principles of several different modes of wavelength sensing using DBSODs as the sensors. By the help of a simplified theory describing the wavelength dependent transparency in a DBSOD, the performance and limitations for each mode of wavelength sensing can be realized; and the measures for improving the sensing resolution and accuracy can be implemented accordingly.

For single-wavelength input, we demonstrated empirically a resolution as good as 0.03 nm using an SOA biased with a commercial current source. To achieve better than 0.01-nm wavelength resolution requires optimization of the sensor structure and utilization of higher resolution current source. The wavelength resolution can also be improved to be better than 0.01 nm by detecting the induced junction voltage for a fixed bias. The ratio of induced voltages at two different currents can further be used to remove the power dependency such that the wavelength sensing can be immune to the power fluctuation.

For multiple wavelength input, a DBSOD can measure the average wavelength if the inputs are modulated together. We used two single-wavelength lasers to demonstrate accurate measurement of the average wavelength. This approach was also applied to measure the mean wavelength of an FP laser. The results show the feasibility of this scheme for applications in optical power meters as wavelength calibrators. When each wavelength of the input light comes from a single-wavelength light source and is modulated at a separate frequency, then a DBSOD may be applicable for resolving all the wavelengths. However, the experiments and analysis indicate that the accuracy of this scheme is limited by the crosstalk from the presence of other wavelengths.

The wavelength-sensing schemes are simple and versatile for many applications. The required stability in temperature control may limit their accuracy and applicability. A double layer of temperature control can help to stabilize the sensor over a wide range of ambient temperature. The transparent current could also be shifted by strong incident power. Thus, the approach is only insensitive to the input power over a limited range.

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REFERENCES

- [1] C. A. Brackett, "Dense wavelength division multiplexing networks: Principles and applications," *IEEE J. Select. Areas Commun.*, vol. 8, pp. 948-964, Aug. 1990.
- [2] A. J. Keating and A. J. Lowery, "Wavelength stabilization in packet-switched WDM networks," *J. Lightwave Technol.*, vol. 15, no. 1, pp. 76-85, 1997.
- [3] M. Guy, B. Villeneuve, C. Latrasse, and M. Tetu, "Simultaneous absolute frequency control of laser transmitters in both 1.3 and 1.55 mm bands for multiwavelength communication systems," *J. Lightwave Technol.*, vol. 14, no. 6, pp. 1136-1143, 1996.
- [4] T. Ikegami, S. Sudo, and Y. Sakai, *Frequency Stabilization of Semiconductor Laser Diodes*. Boston, MA: Artech House, 1995.
- [5] L. E. Nelson, S. T. Cundiff, and C. R. Giles, "Optical monitoring using data correlation for WDM systems," *IEEE Photon. Technol. Lett.*, vol. 10, pp. 1030-1032, 1998.

- [6] J. P. Monochalin, M. J. Kelly, J. E. Thomas, N. A. Kunit, A. Szoke, P. H. Lee, and A. Javan, "Accurate laser wavelength measurement with a precision two-beam scanning Michelson interferometer," *Appl. Opt.*, vol. 20, no. 5, pp. 736-757, 1981.
- [7] S. J. Bennett and P. Gill, "A digital interferometer for wavelength measurement," *J. Phys.*, pt. E, vol. 13, pp. 174-177, 1980.
- [8] M. B. Morris, T. J. McIlrath, and J. J. Snyder, "Fizeau wavemeter for pulsed laser wavelength measurement," *Appl. Opt.*, vol. 23, pp. 3862-3868.
- [9] T. Coroy, R. M. Measures, T. H. Wood, and C. A. Burrus, "Active wavelength measurement system using an InGaAs-InP quantum-well electroabsorption filtering detector," *IEEE Photon. Technol. Lett.*, vol. 8, pp. 1686-1688, 1996.
- [10] A. Densmore and P. E. Jessop, "A quantum-well waveguide photodetector for high-precision wavelength monitoring about 1.55 μm ," *IEEE Photon. Technol. Lett.*, vol. 11, no. 12, pp. 1653-1655, 1999.
- [11] S.-L. Lee, "Wavelength measurement and tracking using semiconductor laser amplifiers for applications in photonic networks," *IEEE Photon. Technol. Lett.*, vol. 10, no. 3, pp. 439-441, 1998.
- [12] S.-L. Lee, C.-T. Pien, and Y.-Y. Hsu, "Wavelength monitoring in DWDM networks using low cost semiconductor laser diode/amplifiers," in *Proc. Conf. Optical Fiber Communications 2000 (OFC)*, Baltimore, MD, Mar. 2000, WK1.
- [13] D. Marcuse, "Heterodyne detection with an injection laser—Part I: Principle of operation and conversion efficiency," *IEEE J. Quantum Electron.*, vol. 26, no. 1, pp. 615-627, 1990.
- [14] M. Gustavsson, A. Karlsson, and L. Thylen, "Traveling wave semiconductor laser amplifier detectors," *J. Lightwave Technol.*, vol. 8, no. 4, pp. 610-617, 1990.
- [15] R. Hui, I. Maio, and I. Montrosset, "Analysis of the all-electrical automatic gain control of traveling-wave semiconductor laser amplifiers," *Proc. Inst. Elect. Eng. Optoelectron.*, vol. 141, no. 1, pp. 65-68, 1994.
- [16] R. M. Fortenberry, A. J. Lowery, and R. S. Tucker, "Up to 16 dB improvement in detected voltage using two-section semiconductor optical amplifier detector," *Electron. Lett.*, vol. 28, no. 5, pp. 474-476, 1992.
- [17] P. A. Andrekson, N. A. Olsson, T. Tanbun-Ek, R. A. Logan, D. Coblenz, and H. Temkin, "Novel technique for determining internal loss of individual semiconductor lasers," *Electron. Lett.*, vol. 29, no. 2, pp. 171-172, 1992.
- [18] S.-L. Lee, M. E. Heimbuch, D. A. Cohen, L. A. Coldren, and S. P. DenBarrs, "Integration of semiconductor laser amplifiers with sampled grating tunable lasers for WDM applications," *IEEE J. Select. Topics Quantum Electron.*, vol. 3, no. 2, pp. 615-627, 1997.
- [19] S.-L. Lee, "Analytical formula of wavelength dependent transparent current and its implications for designing wavelength sensors and WDM lasers," *IEEE J. Select. Topics Quantum Electron.*, submitted for publication.
- [20] S.-L. Chuang, *Physics of Optoelectronic Devices*. New York: Wiley, 1985, ch. 10.
- [21] T. Makino, "Analytical formulas for the optical gain of quantum wells," *IEEE J. Quantum Electron.*, vol. 32, no. 3, pp. 493-501, 1996.
- [22] A. B. Carlson, *Communication Systems*, 3rd ed. New York: McGraw-Hill, 1986, pp. 220-225.
- [23] D. Derickson, Ed., *Fiber Optic Test and Measurement*. Englewood Cliffs, NJ: Prentice-Hall, 1998, p. 121.



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Analytical Formula of Wavelength Dependent Transparent Current and its
Implications for Designing Wavelength Sensors and WDM Lasers

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內容成果：

半導體雷射二極體的操作工作區大致可分為三大項，分別為吸收區、透明點與增益區。操作電流大小在透明點以下為吸收區，當外部光源注入時，光子會被吸收，既會有損耗；反之操作電流在透明點以上，此時雷射二極體帶有增益，可放大光功率或是發光；而當操作電流在透明點，雷射二極體不吸收也不放大，光直接穿透過雷射二極體。經由實驗與理論發現，半導體雷射的透明點電流會隨注入的光源的波長的不同而有相對應的變化，利用此特性，可即時精確監控通道中光訊號波長的穩定度以及偏移量。本文中使用一半導體光放大器作為光感測器，搭配簡單的控制電路，達到波長監控的功能。

Analytical Formula of Wavelength-Dependent Transparent Current and Its Implications for Designing Wavelength Sensors and WDM Lasers

San-Liang Lee, *Member, IEEE*

Abstract—The wavelength dependency of transparent current is investigated theoretically and empirically for bulk and quantum-well (QW) materials. A new analytical formula is presented for obtaining accurate Fermi levels of QW structures under current injection. The transparent current can be described with an exponential function of the wavelength using two fitting parameters. The simple expression can clearly indicate the fundamental limit of the wavelength resolution when the wavelength dependency is applied for wavelength sensing. The formula was also exploited to obtain the condition of achieving uniform threshold current for wavelength division multiplexing (WDM) lasers that may cover a wide range of wavelength.

Index Terms—Differential gain, laser diodes, quantum wells (QWs), semiconductor optical amplifiers (SOAs), transparent current, wavelength sensing.

I. INTRODUCTION

THE TRANSPARENT condition is a unique property that determines if an active material is under amplifying or absorptive condition. This property is thus strongly correlated with the characteristics of semiconductor lasers and amplifiers, such as the threshold current, thermal effects, and gain. The theory for the transparent property was developed concurrently with the laser theory [1]. In most of the previous literature, the transparent properties of semiconductor lasers and amplifiers were addressed as the bias condition for zero peak gain [2]–[7]. This type of transparent condition is not appropriate for describing the characteristics of single-wavelength lasers, like distributed-feedback (DFB) lasers, or semiconductor optical amplifiers (SOAs). For these devices, the transparency for a specific wavelength is of major concern. Moreover, the investigation on the transparency aimed at achieving low threshold current or reducing the thermal effects of laser diodes (LDs).

In this paper, we will focus on the wavelength dependency of the transparent condition. Investigating such a wavelength

dependency is important for designing the material structure of wavelength-division-multiplexing (WDM) lasers in order to cover a wide range of wavelength [8]. It is also important for investigating the wavelength resolution when the wavelength-dependent transparency in an SOA or LD is used for wavelength sensing [9], [10]. Fast evolution in fiber optic networks has resulted in strong demand for laser sources that can provide dozens to hundreds of densely spaced wavelength channels, either simultaneously or in a tunable way [11]. There is also increasing interest in using a few coarse WDM (CWDM) channels to expand the information capacity for access networks [12], [13]. The preferred channel spacing for such type of applications is 20 nm or more. If these WDM lasers are to be fabricated out of a single wafer, the wavelength dependency of the transparent condition must be optimized in order to minimize the variation of the threshold current for different wavelength channels. On the other hand, for applications as wavelength sensors, the transparent current is favored to be as sensitive to wavelength change as possible.

In order to investigate and optimize the wavelength dependent characteristics for different applications, analytical expressions for the transparent current will be derived in this paper. Simple formulas will help to delineate the relations between the device performance and the design parameters. We will show that the variation of transparent current with wavelength can be roughly described with an exponential function. This analytical formula will be applied to discuss the wavelength resolution of the sensors and the differential gain of WDM laser sources.

The remaining part of this paper will be organized as follows. The transparent condition and the ideality factor for indicating the wavelength dependency will be addressed in Section II. In Section III, an approximate expression was derived to describe the quasi-Fermi levels for quantum-well (QW) structures. The analytical formula for calculating the corresponding wavelength for a specified transparent current will be addressed for both bulk and QWs. Numerical calculation to show the accuracy of the approximation will be presented in Section IV. Experimental transparent characteristics for commercial SOAs will also be presented for comparison. The implications for designing DWDM sources and wavelength sensors will be addressed in Section V. A brief conclusion will be drawn in Section VI.

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II. WAVELENGTH-DEPENDENT TRANSPARENCY

For a direct-bandgap semiconductor material, it can provide either gain or loss onto the input optical signals, depending on its bias condition. The material is operated at its transparent condition when its net gain is zero. This corresponds to the condition that the occupation probability for the conduction band is equal to that for the valence band. Following the Bernard–Duraffourg inversion condition [1], the transparent carrier density N_{tr} can be obtained from

$$\mu_n(N_{tr}) + \mu_p(N_{tr}) = \varepsilon - \varepsilon_g \quad (1)$$

where

- μ_n and μ_p normalized quasi-Fermi levels for electrons and holes, measured from the band edge and positive into the bands, respectively;
- ε normalized photon energy;
- ε_g normalized energy gap between the lowest and highest subbands for electrons and holes, respectively.

Throughout this paper, all the energy parameters are normalized with respect to the thermal energy, kT . The transparent current corresponds to the injection current that makes the induced separation between the quasi-Fermi levels of electrons and holes equal the incident photon energy. Thus, to reach transparency the required injection current is higher for an incident light with a shorter wavelength.

The variation of transparent current with wavelength can be characterized by

$$\frac{\partial J_{tr}}{\partial \lambda} = \frac{qd}{\eta_i} \frac{\partial R(N)}{\partial N} \bigg|_{N_{tr}} \frac{\partial N_{tr}}{\partial \lambda} = -\frac{qd}{\eta_i} \frac{\varepsilon}{\lambda} \frac{\partial R(N)}{\partial N} \bigg|_{N_{tr}} \frac{\partial N_{tr}}{\partial \varepsilon}. \quad (2)$$

In this equation, the transparent current density is related to the transparent carrier density by $J_{tr} = qdR(N_{tr})/\eta_i$, where q is the electron charge; d is the thickness of the active layer; η_i is the internal quantum efficiency; and $R(N)$ is the carrier recombination rate. $R(N)$ is frequently written as: $R(N) = AN + BN^2 + CN^3$, where A , B , and C are the carrier recombination coefficients [7], [8]. Both N_{tr} and $\partial N_{tr}/\partial \varepsilon$ depends on the material structures.

In order to provide a unified measure for comparing different material structures, we define an ideality factor ξ for the wavelength dependency as

$$\xi \equiv \frac{\partial \varepsilon}{\partial J_{tr}} J_{tr} = \frac{1}{kT} \left[\frac{\partial J_{tr}/\partial \varepsilon}{J_{tr}} \right]^{-1}. \quad (3)$$

ξ is constant if the transparent current is an exponential function of the photon energy. The transparent current density can thus be expressed with the ξ -parameter as:

$$J_{tr} = J_t \exp \left(\int_{E_r}^E \frac{dz}{\xi(z)kT} \right) \approx J_t \exp \left(\frac{E - E_r}{\xi kT} \right) \quad (4)$$

where E_r is a reference energy and J_t stands for the transparent current density at the reference energy. The second equality exists when ξ remains almost constant within an interested wavelength range. For this case, (3) indicates that the variation rate

of the transparent current density is linearly proportional to the transparent current density. That is, a device that has a larger transparent current density will give rise to a more sensitive transparency. A smaller value of ξ further indicates a stronger dependency. The role of the ξ -parameter in (4) is like the ideality factor in describing the current–voltage characteristics of a diode.

III. TRANSPARENT CARRIER DENSITY AND FERMI ENERGY

We will derive in the following the explicit formulas for the wavelength-dependent transparency. Only the bulk and unstrained QW structures are discussed here but the formulation can be extended to other material structures. It is assumed throughout this paper that all subbands are parabolic with effective mass approximation. Usually, the transparent carrier density does not vary significantly over the interested wavelength range, so the carrier recombination rate will be simply expressed as $R(N) \approx bN^x$. The exponent x ranges between 2 and 3 for long-wavelength lasers. In addition, the bandgap shrinkage effect is approximated as $\Delta \varepsilon_g = -cN^k$, where c and k are parameters depending on the material structure [14].

The wavelength dependency is related to how the Fermi levels vary with the carrier density. For bulk materials, the carrier density and Fermi level are governed by the Fermi–Dirac integral that can be approximated by the Joyce–Dixon expansion [8]. The simplest approximation would be the Boltzmann's approximation, but it is oversimplified in describing the transparency where at least one of the electron and hole concentrations is under the degenerate condition.

For QW structures, the expression for obtaining the Fermi level with respect to a given carrier density is often written as [3]:

$$\mu_{n,p} = \ln [\exp(r_{n,p}) - 1] \quad (5)$$

where $r_n = N/N_e$, $r_p = P/N_h$ with N and P being the electron and hole concentration in the active region, respectively. N_e and N_h are defined as

$$N_\gamma = \frac{kTm_\gamma}{\pi \hbar^2 d}, \quad \gamma = e, h. \quad (6)$$

Equation (5) was frequently used to explain the gain as a logarithmic function of carrier concentration [3]–[6], but it is exact only if a single subband is occupied. A better formula for obtaining the Fermi levels of electrons is derived in Appendix A as:

$$\mu_n = \ln \left(\frac{-D_{1n} + \sqrt{D_{1n}^2 + 4D_{2n} [\exp(r_n) - 1]}}{2D_{2n}} \right) \quad (7)$$

where:

$$D_{1n} = \sum_i \exp(\varepsilon_{c0} - \varepsilon_{ci}) \quad (8a)$$

$$D_{2n} = \frac{1}{2} \left\{ D_{1n}^2 - \sum_i \exp [2(\varepsilon_{c0} - \varepsilon_{ci})] \right\}. \quad (8b)$$

The quantized energy levels, ε_{ci} , can be obtained once the QW structure is given. The Fermi level for holes can be written in a

similar way as (7) by substituting the subscript n with p ; and (8) can be used for holes by changing the subscript c to v .

Equation (7) is exact for conduction and valence bands with two quantized subbands. For QWs with more than two quantized levels, e.g., for thick QWs, this expression can usually provide excellent accuracy. In fact, for single subband, (7) approaches (5) after taking the limit of $D_{2n} = 0$ and $D_{1n} = 1$. It will be shown later that, as the QW supports more than one bound states, calculating the transparent carrier density with (5) may result in considerable error.

The closed-form expressions for the wavelength-dependent transparency can then be written as

$$\epsilon - \epsilon_g = \frac{2}{x} \ln \left(\frac{\eta_i J_{tr}}{J_{b0}} \right) + \sum_i A_i \left(\frac{\eta_i J_{tr}}{J_{bi}} \right)^{i/x}, \quad \text{for bulk} \quad (9)$$

$$\epsilon - \epsilon_g = \ln(D_{rn} D_{rp}), \quad \text{for QWs} \quad (10)$$

where A_i s are the Joyce–Dixon coefficients; and

$$D_{rn} = \frac{-D_{1n} + \sqrt{D_{1n}^2 + 4D_{2n}(e^{y_n} - 1)}}{2D_{2n}} \quad (11a)$$

$$D_{rp} = \frac{-D_{1p} + \sqrt{D_{1p}^2 + 4D_{2p}(e^{y_p} - 1)}}{2D_{2p}} \quad (11b)$$

for two or more bound states or

$$D_{rn} = e^{y_n} - 1 \quad (11c)$$

$$D_{rp} = e^{y_p} - 1 \quad (11d)$$

for single bound states. The associated parameters are defined as

$$y_n \equiv r_n |_{N=N_{tr}} = N_{tr}/N_e = (\eta_i J_{tr}/J_{n0})^{1/x}, \quad (12a)$$

$$y_p \equiv r_p |_{N=N_{tr}} = N_{tr}/N_h = (\eta_i J_{tr}/J_{p0})^{1/x} \quad (12b)$$

$$J_{b0} = qbd(N_c N_v)^{x/2} \quad (12c)$$

$$J_{bi} = qbd [N_c^i N_v^i / N_v^i (N_c^i + N_v^i)]^{x/i} \quad (12d)$$

$$J_{n0} = qBdN_e^x \quad (12e)$$

$$J_{p0} = qBdN_h^x. \quad (12f)$$

The analytical expressions, (9) to (11a)–(11d), can be used to obtain the corresponding wavelength for a given transparent current with good accuracy. For the extreme case of low-level injection the transparent current density can simply be written as

$$J_{tr} \approx J_t \exp \left[\frac{x}{2} (\epsilon - \epsilon_g) \right], \quad \epsilon \geq \epsilon_g. \quad (13)$$

J_t stands for J_{b0}/η_i and $\sqrt{J_{n0} J_{p0}}/\eta_i$ for bulk and QW, respectively. This suggests that the relation between the transparent current and wavelength may be written as an exponential function, as given by (4).

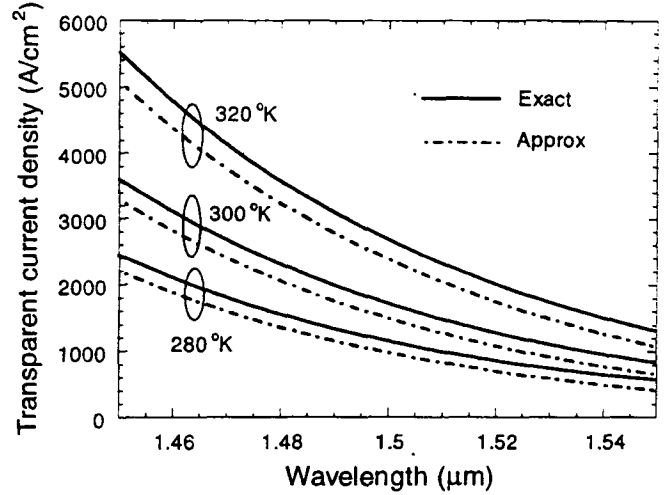


Fig. 1. Calculated variation of transparent current density with wavelength for bulk InGaAsP at different temperature. The exact results were calculated from searching the current density for zero gain using standard gain calculation. The approximate results refer to the calculation using (9) without including lineshape broadening.

The analytical formulas for the ideality factor can also be derived from (3), (9), and (10) as

$$\xi = \frac{2}{x} + \sum_i A_i \left(\frac{i}{x} \right) \left(\frac{\eta_i J_{tr}}{J_{bi}} \right)^{i/x} + \frac{k \Delta \epsilon_g}{x}, \quad \text{for bulk materials} \quad (14)$$

and

$$\xi = \xi_n + \xi_p + \frac{k \Delta \epsilon_g}{x}, \quad \text{for QWs} \quad (15)$$

where

$$\begin{aligned} \xi_{n,p} &\equiv \left(J \frac{d\mu_{n,p}}{dJ} \right)_{J=J_{tr}} \\ &= \frac{1}{x} \frac{y_{n,p} e^{y_{n,p}}}{D_{rn,p} \sqrt{D_{1n,p}^2 + 4D_{2n,p}(e^{y_{n,p}} - 1)}} \end{aligned} \quad (16a)$$

for the cases of more than one bound states, and

$$\xi_{n,p} = \frac{1}{x} \frac{y_{n,p} e^{y_{n,p}}}{e^{y_{n,p}} - 1} \quad (16b)$$

for the cases of single bound state.

For bulk materials, the terms inside the summation of (14) are larger for higher transparent current density (or shorter wavelength). The wavelength dependency of ξ for QWs can easily be delineated for the single-subband case. It can be proven from (16b) that ξ_n is larger than $1/x$, so is ξ_p . Furthermore, both ξ_n and ξ_p increase monotonically with J_{tr} , so ξ is again larger for shorter wavelength. The bandgap shrinkage effect decreases ξ slightly and is also wavelength dependent since $|\Delta \epsilon_g|$ increases with carrier density.

IV. NUMERICAL AND EXPERIMENTAL RESULTS

Fig. 1 shows the calculated transparent current density for the InGaAsP bulk material. The material parameters are listed in

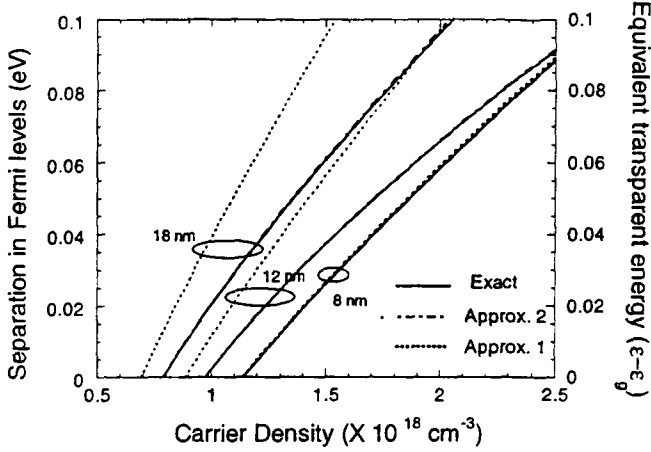


Fig. 2. Calculated separation in Fermi levels ($\mu_n + \mu_p$) versus carrier density for InGaAs-InGaAsP QWs of different well widths. Approx. 1 and 2 refer to the calculation using (5) with modified N_r and (7), respectively.

Appendix B [15]. The “exact” solution was obtained from locating the carrier density for zero gain following the standard gain calculation procedures. The effects of bandgap shrinkage and lineshape broadening as well as the temperature dependency of the recombination coefficients were included in the calculation. It is clear that the transparent current density is larger for shorter wavelengths and increases with the temperature. This figure also indicates that the approximate results calculated using (9) without considering lineshape broadening can lead to reasonably close results with the exact calculation.

For QW structures, the numerical results for lattice-matched InGaAs-InGaAsP wells are shown as examples. The material parameters for the calculation are also listed in Appendix B [16]. Fig. 2 depicts the calculated separation in quasi-Fermi levels for different well widths. In Fig. 2, Approx. 1 refers to the calculation using (5) with modified N_r (Makino’s approximation [6]). This approximation is equivalent to (5) for wells with a single bound state, which is the case for the conduction band of this specific example. In Fig. 2, Approx. 2 refers to the calculation with (7). For well width of 120 Å and 180 Å, (7) can provide excellent accuracy, but the error from Approx. 1 is significant. The error comes mainly from the calculation of the electron Fermi level, since for these two well widths the conduction band has two subbands. This reveals that the accuracy of Approx. 1 is questionable for the cases of multiple subbands. It can also be observed from this figure the influence of well width on the transparent current density. When the incident light energy is close to the bandedge transition, $\epsilon - \epsilon_g$, the transparent carrier density decreases with the well width, d . This can be easily predicted from (10) in the limit of low-level injection and single bound state.

Fig. 3 shows the calculated variation of the transparent current density with the incident wavelength for QWs of 80- and 120-Å well width. The exact solution again includes the bandgap shrinkage and lineshape broadening effects. Moreover, the Fermi energies are obtained from (A.1) and N_{tr} was obtained from searching the zero-gain condition. For the approximation, N_{tr} was calculated directly from (1) and

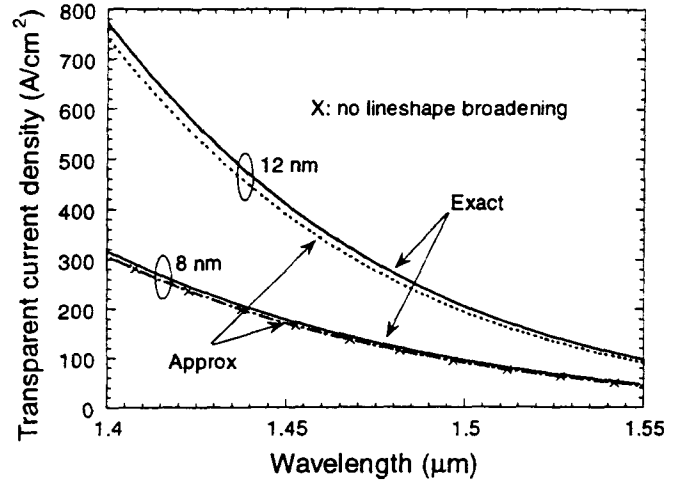


Fig. 3. Calculated wavelength-variation of the transparent current density for QWs of 80- and 120-Å well width. The approximation refers to the calculation using (1) and (7) with the inclusion of bandgap shrinkage. Marker “x” indicates a similar calculation as the exact one except for the lineshape broadening.

(7) with the inclusion of bandgap shrinkage. Equation (5) was used in obtaining μ_n for the 80-Å well, because it contains only one bound state. For this specific example, the transparent current of the 80-Å well is smaller than that of the 120-Å well. Fig. 3 clearly indicates that the analytical approximations can provide really good accuracy; and the error is mainly caused by the neglect of the lineshape broadening. To prove this, the results obtained from locating the zero-gain condition without considering the lineshape broadening is also shown in the figure for comparison. They are almost identical to the results of analytical approximation.

It has to be pointed out that the Bernard-Duraffourg condition, (1), does not account for the lineshape broadening effect. The exact calculation of the gain coefficient involves the integration of the product of the reduced density of states, Fermi functions, and the lineshape functions [8]. Zero gain does not necessarily occur at the Bernard-Duraffourg condition. The gain coefficient can be written as (C.1) in case that the lineshape function is narrow enough that the Fermi functions are approximately constant over the range of energies for which the lineshape function is significant. In such case, transparency occurs at the Bernard-Duraffourg condition.

Fig. 4 shows the calculated ξ -parameter for the bulk material. The parameter varies slightly versus the wavelength and its value ranges from 1.3 to 1.6 for most of the wavelength range. The material tends to have a slightly smaller ξ -parameter at higher temperature. For comparison, the results from the analytical approximation are also shown in Fig. 4. It indicates that the rising of ξ at the long wavelength side (near the bandedge) is due to the linewidth broadening effect. Though the analytical approximation can provide acceptable accuracy in calculating the transparent current, the error in calculating ξ is noticeable.

The ξ -parameter for QW structures of two different well widths is depicted in Fig. 5. In contrast to the bulk case, the analytical expression (15) can be really accurate in calculating ξ for QWs. Lineshape broadening is again the main cause of the difference between the exact and approximate results.

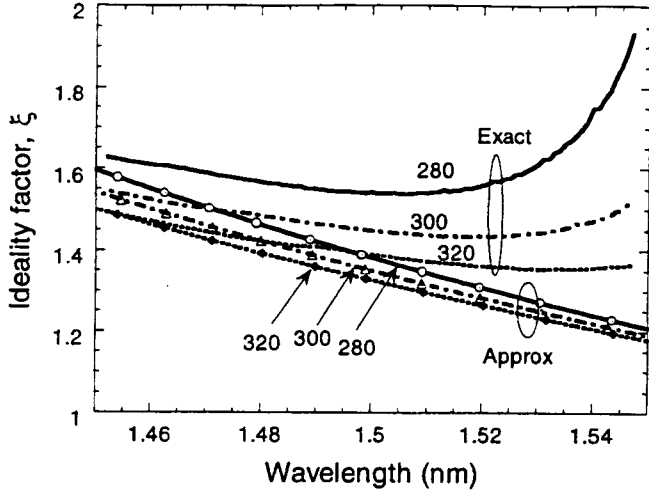


Fig. 4. Calculated ξ -parameter for bulk material under the same condition as Fig. 1.

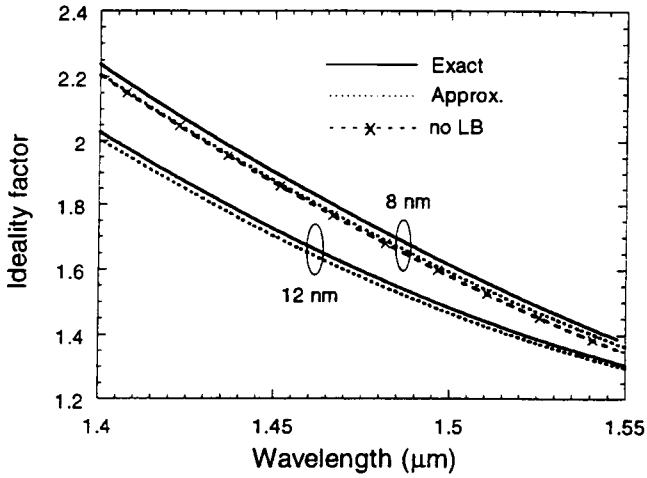


Fig. 5. Calculated ξ -parameter for QWs under the same condition as Fig. 3.

In order for comparison with the theory, we measured the ξ -parameter for two commercial SOAs operating around 1.55 μm (Alcatel 1901) and 1.3 μm (Phillips CQF882/E). Figs. 6 and 7 show the results that were obtained by calculating the derivative numerically from the measured transparent current versus wavelength. The transparent current was determined from minimizing the induced junction voltage across the diode [9]. In measuring these data, the light from a tunable laser was modulated with an external modulator and coupled into the SOA. The wavelength was monitored with a high-precision wavelength meter and the induced junction voltage was detected with a lockin amplifier. The SOA was biased with a low-noise current source and a temperature controller was used to stabilize its temperature. The tunable laser used for measuring the 1.55- μm SOA has a resolution of 1 pm and very good wavelength stability, while the 1.3- μm tunable laser has a relatively poor resolution and stability.

It is shown that the measured ξ -parameters for both SOAs fall within the reasonable range predicted by the theory. Matching the experimental data with numerical results is not of major concern here because the detailed device parameters of the SOAs

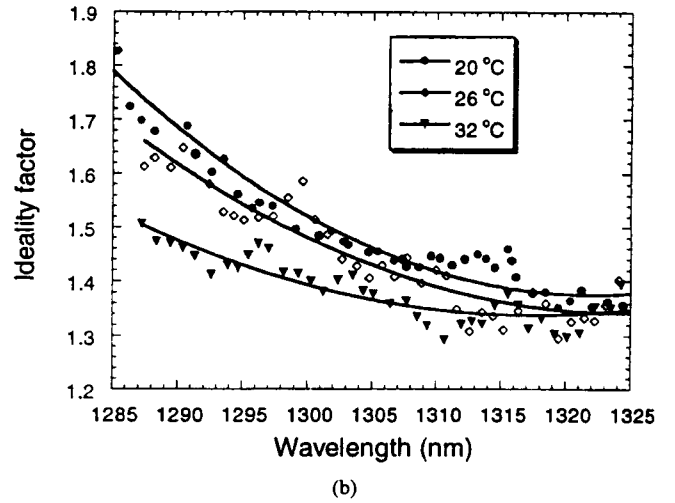
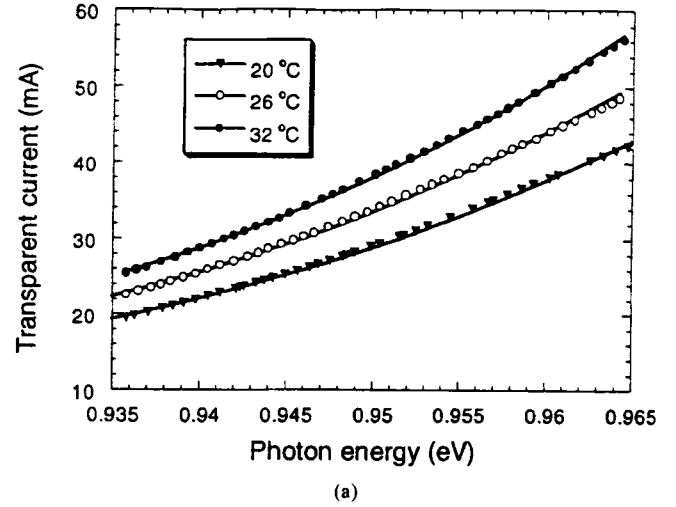


Fig. 6. (a) Measured transparent current and (b) ideality factor for the 1.3- μm SOA. The transparent current is curve-fitted using (4). Choosing the reference energy as 0.95 eV, the curve-fitted ideality factor is (1.49, 1.43, 1.37) and the reference transparent current is (28.8, 33.6, 38.0) mA for the temperature of (20, 26, 32) $^{\circ}\text{C}$, respectively.

are not known exactly. However, the variation of the ξ -parameter with the wavelength shows very consistent tendency with the theoretical analysis. The measured transparent current can be well represented using the exponential form of (4) with a constant ξ . Let the reference energy be 0.8 eV for the transparent current of the 1.55- μm SOA, the curve-fitted ξ -parameter is 1.505, 1.504, 1.504, and the reference transparent current is 24.37, 26.70, 29.49 mA, for the temperature of 20, 25, and 30 degree C, respectively. The curve fitting achieves less than 1% error from the measured data for the wavelength range between 1530 and 1580 nm. Though the ξ -parameter shows larger fluctuation for the 1.3- μm SOA due to poor performance of the measurement apparatus, the curve fitting using a constant ξ -parameter can still provide really good accuracy.

V. IMPLICATIONS FOR WAVELENGTH SENSORS AND WDM LASERS

It has been shown that the wavelength dependency of transparent current can be represented in an exponential form with two fitting parameters, J_{tr} and ξ . This analytical expression can

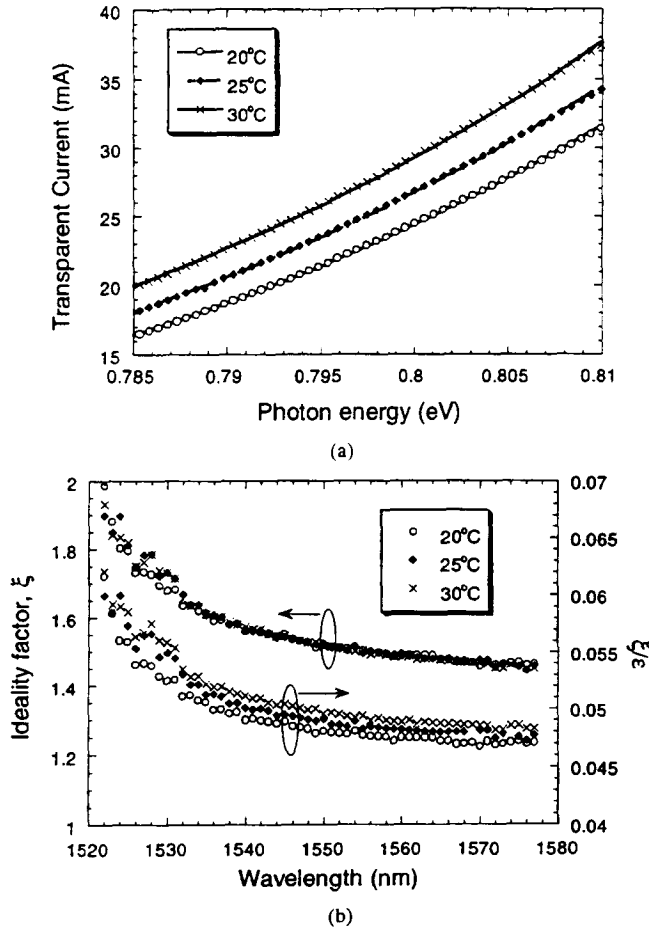


Fig. 7. (a) Measured transparent current and (b) ideality factor for the 1.55- μm SOA. The transparent current is curve-fitted using (4). Choosing the reference energy as 0.8 eV, the curve-fitted ideality factor is (1.505, 1.504, 1.504) and the reference transparent current is (24.37, 26.70, 29.49) mA for the temperature of (20, 25, 30) $^{\circ}\text{C}$, respectively.

help to discuss and optimize the material structures of wavelength sensors and WDM lasers. For applications of wavelength sensing, a stronger wavelength dependency for the transparent current is desired to achieve better wavelength resolution and accuracy. Due to the roughly exponential dependency, $\partial J_{\text{tr}}/\partial \lambda$ is larger at a shorter wavelength whereat higher wavelength resolution can be achieved for a given current resolution. Increasing the active volume of the SOA can also improve the slope of I_{tr} vs. λ . However, for high-precision measurements, the current resolution is usually limited by a specified $\Delta I/I$. From (4), the wavelength resolution $\Delta \lambda$ for a given current resolution ΔI is given by

$$\frac{\Delta \lambda}{\lambda} = \frac{\xi}{\epsilon} \frac{\Delta I}{I_{\text{tr}}} \quad (17)$$

This equation indicates that a smaller ξ gives a better resolution in terms of $\Delta \lambda/\lambda$ for a given $\Delta I/I$. For the 1.55- μm SOA, Fig. 7 indicates that the resolution around 1550 nm can be roughly estimated from $\Delta \lambda \approx (75\text{nm}) \Delta I/I_{\text{tr}}$. That is, to achieve 0.01-nm wavelength resolution requires a $\Delta I/I_{\text{tr}}$ of 1.3×10^{-4} . $\Delta \lambda/\lambda$ is larger (worse) at shorter wavelengths.

It is clear that exploiting the ξ -parameter is very helpful in evaluating the inherent wavelength resolution when an SOA or

LD is used as a wavelength sensor. From the theoretical and experimental results, the typical value of this parameter ranges between 1 and 2 within the interested wavelength range. Therefore, the ultimate resolution is bounded by (17) with $\xi = 1$. Modifying the material structure might change ξ but not in a significant way. To better the resolution will require improving the current resolution of the bias source.

For realizing WDM laser sources out of an active material, the gain may have to cover a wide range of wavelength. A small ξ -parameter gives rise to strong wavelength variation of the transparent characteristics. Therefore, the value of the ξ -parameter can affect the threshold characteristics significantly. The gain coefficient is frequently written as a logarithmic function of the current density [18], [19]

$$g = \beta J_{\text{tr}} \ln \left(\frac{J}{J_{\text{tr}}} \right) \quad (18)$$

where β is the differential gain at transparency. The wavelength-dependent transparent current itself tends to raise the threshold current at shorter wavelength. Equation (18) indicates that the differential gain can be larger for shorter wavelength. To achieve a less wavelength-dependent threshold current requires optimization of the material structure and bias current to have the wavelength dependency of transparent current and differential gain compensate each other.

The logarithmic functional form of gain coefficient as given in (18) has been claimed to well describe the gain of QW lasers, but it usually refers to the peak gain. In order to simplify the discussion, we will assume here that (18) can be used for describing the wavelength-dependent gain coefficient. It can be shown that (see Appendix C) the differential gain at transparency can be written as

$$\beta = \frac{\chi K_0}{2\epsilon J_{\text{tr}}} \left[1 + \cosh \left(\frac{m_r}{m_c} (\epsilon - \epsilon_g) - \mu_n \right) \right]^{-1} \cdot \sum_i \left[\frac{1}{2} + \frac{1}{\pi} \tan^{-1} \left(\frac{\epsilon - \epsilon_{gi}}{\delta\epsilon/2} \right) \right] \quad (19)$$

where $\epsilon = hc/\lambda kT$, and ϵ_{gi} is the transition energy between the i th subbands of the conduction and valence bands. The parameter χ has very similar wavelength dependency as ξ does, and is defined as

$$\chi \equiv J \frac{\partial(\epsilon - \epsilon_g)}{\partial J} \bigg|_{J=J_{\text{tr}}} = \xi - \frac{k \Delta \epsilon_g}{x} \quad (20)$$

For a strained QW that has equal effective mass for electrons and holes, the term inside the hyper-cosine of (19) is zero, so it has a larger β than unstrained QWs, provided that the other factors are the same.

Equation (19) reveals the factors relating to the wavelength variation of differential gain. Roughly speaking, the last two terms have relatively weak dependency except for the lineshape broadening effect at the band edge. Furthermore, both χ and ϵ increases as the wavelength decreases. Let $g_0 = \beta J_{\text{tr}}$, then g_0 increases slightly as the wavelength moves to shorter wavelength. Due to the exponential dependence of J_{tr} on the wavelength, (19) indicates that β becomes smaller for shorter wavelength.

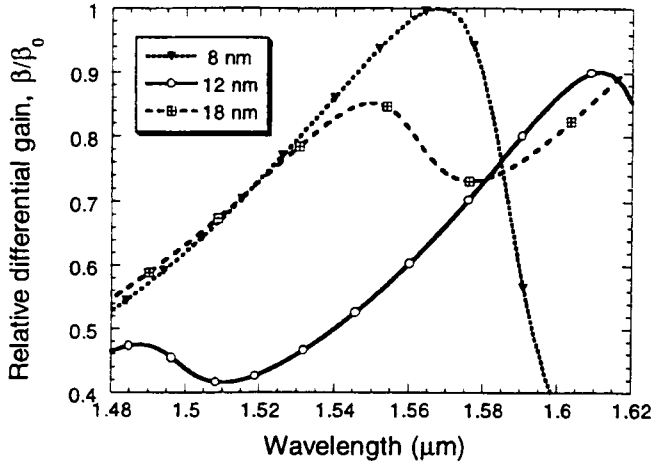


Fig. 8. Calculated differential gain at transparency for QWs of different widths. The parameters are the same as Fig. 2. β_0 is the maximal differential gain for the 80-Å well.

Fig. 8 depicts the differential gain at transparency versus wavelength for QWs of different well thickness. For the 80-Å well that has a single quantized level in the conduction band, the value of β drops at the short-wavelength side as the wavelength is shortened. On the other hand, β also decreases with the wavelength at the long-wavelength side due to the lineshape broadening effect. The wavelength variation is similar for the 120 and 180-Å wells except that the second peak appears at the wavelength corresponding to the transition of the second subbands. For comparison, Fig. 9 shows the measured gain and the differential gain at transparency for the 1.55- μm SOA. The gain was measured by comparing the input and output power of the SOA. Narrow-band optical filters were placed before and after the SOA to minimize the effects of spontaneous emission. Thus accurate gain coefficient can easily be obtained with the known transparent current. The results show very similar tendency as the calculated data of the 180-Å well. This can also reveal the transition energies of the quantized subbands for the SOA.

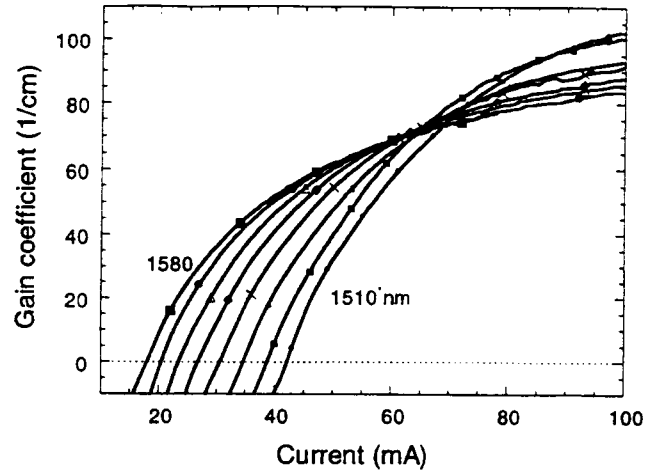
The differential gain at a fixed bias J is $\beta J_{\text{tr}}/J = g_0/J$, which becomes larger at shorter wavelength in spite of the decreasing β . Thus, as the bias increases from the transparent point, the difference in the gain coefficient among different wavelengths can decrease and there is a chance that the gain coefficients for different wavelengths become very close at some bias. This can be observed from the measured gain curves at the bias of around 65 mA, as shown in Fig. 9(a). It is thus viable to achieve a uniform threshold current for WDM lasers by optimizing the number of QWs and the mirror loss.

From (18), the wavelength variation of the required threshold current density, J_{th} , can be written as

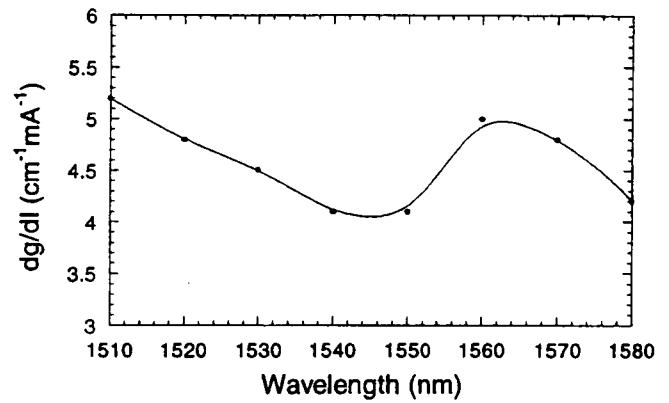
$$\frac{\partial J_{\text{th}}}{\partial \lambda} = -\frac{\epsilon J_{\text{tr}}}{\lambda \xi} \left[1 - \frac{g_{\text{th}}}{\beta J_{\text{tr}}} \right] \quad (21)$$

where g_{th} is the threshold gain coefficient. Therefore, the optimal choice of the threshold gain coefficient to minimize the wavelength variation is

$$g_{\text{th}} = \beta J_{\text{tr}}. \quad (22)$$



(a)



(b)

Fig. 9. (a) Measured gain coefficient and (b) differential gain at transparency for the 1.55- μm SOA.

This is again verified from the measured gain curves for the SOA, as shown in Fig. 9(a). Noting that the length of a laser source is typically shorter than that of the SOA (about 1 mm), so the optimal current can be much smaller than 65 mA if the same material is used. Furthermore, βJ_{tr} needs to be almost constant over the wavelength range in order to satisfy the condition of (22) for a wide range of wavelengths.

VI. CONCLUSION

This paper addresses the wavelength dependency of the transparent current in bulk and QW materials, aiming at designing material structures for applications in wavelength sensing and WDM lasers. An analytical formula was first derived for QWs to obtain the Fermi levels from a given carrier density. This formula was shown to provide much improved accuracy in calculating the Fermi levels when the QW contains multiple bound states. We then described the wavelength dependency with an exponential function of the photon energy. The ideality factor for indicating how sensitive the transparent current varies with the wavelength was defined and calculated with a simple expression. The calculated and measured ideality factor varies slowly with wavelength and ranges between 1 and 2 for the wavelength range of typical applications. This suggests that, over a limited wavelength range, the variation of transparent current with wavelength can be described in an exponential form using a

constant ideality factor. We applied such a functional form to curve-fit the experimental data and obtain really good consistency.

The fundamental limitation on the resolution of wavelength sensing using the wavelength dependency can easily be described with the help of the analytical formula. It is shown that the ultimate resolution (about 0.01 nm around 1.55 μm) of this sensing technique may be satisfactory for DWDM applications. However, it requires improvement on the resolution of the current source if higher wavelength resolution is desired. The exponential function form can also help to provide a simple formula for describe the differential gain and to find the condition for achieving uniform threshold current over a wide range of wavelength. The measured gain curves showed the feasibility of designing WDM lasers with uniform threshold for different wavelengths.

APPENDIX A

For QW structures, the relation between the carrier densities and the Fermi energies can be described by [8]

$$r_{n,p} = \sum_i \ln [1 + \exp(\mu_{ni,p})] \quad (\text{A.1})$$

where μ_{ni} and μ_{pi} are the normalized Fermi levels measured from the band edge of the i th quantized subband for electrons and holes, respectively. The summation is over all the subbands. We will derive the equations for electrons here, and those for holes can be obtained similarly. Equation (A.1) can be rewritten as

$$(e^{r_n} - 1) = D_{1n}Y + D_{2n}Y^2 + D_{3n}Y^3 + \dots \quad (\text{A.2})$$

where

$$Y = \exp(\mu_{n0}) \equiv \exp(\epsilon_F - \epsilon_{c0}) \quad (\text{A.3})$$

$$D_{1n} = \sum_i \exp(\epsilon_{c0} - \epsilon_{ci}) \quad (\text{A.4})$$

$$D_{2n} = \frac{1}{2} \sum_i \sum_j \exp(\epsilon_{c0} - \epsilon_{ci}) \exp(\epsilon_{c0} - \epsilon_{cj}), \quad i \neq j \quad (\text{A.5})$$

$$D_{3n} = \frac{1}{6} \sum_i \sum_j \sum_k \exp(\epsilon_{c0} - \epsilon_{ci}) \exp(\epsilon_{c0} - \epsilon_{cj}) \cdot \exp(\epsilon_{c0} - \epsilon_{ck}), \quad i \neq j \neq k. \quad (\text{A.6})$$

After some derivations, (A.5) can be expressed as (8b) and (A.6) can be written as

$$D_{3n} = \frac{1}{6} \left[D_1^3 - 3D_1^2 \sum_i \exp(2(\epsilon_{c0} - \epsilon_{ci})) + 2 \sum_i \exp(3(\epsilon_{c0} - \epsilon_{ci})) \right]. \quad (\text{A.7})$$

Under small or moderate injection condition and when the QW has sparsely spaced quantum states, the terms after Y^2 can be neglected in (A.2). Thus, the closed-form solution given by

(7) can be used to obtain the quasi-Fermi levels for a given bias condition. This solution is exact if the QW has two bound states. For higher injection condition or wide wells, the third-order term must be included. Such an equation can still have closed-form solutions. The solution is exact when the QW has three bound states.

APPENDIX B MATERIAL PARAMETERS

A. Bulk Structures: InGaAsP (Bandgap Wavelength = 1.551 μm at 300 K) [8], [15]

$$\eta_i = 0.9, d = 0.2 \mu\text{m}, b = 0.2776 \text{ cm}^2/\text{A}.$$

$$\text{Lineshape broadening: } \delta\epsilon = 0.0131 \text{ eV}.$$

$$\text{Bandgap shrinkage: } k = 0.3841, c = 2.7884 \times 10^{-9} \text{ eV}/(\text{cm}^3)^{0.3841}.$$

$$B = 4.49 \times 10^{-15} \exp(-T/77) \text{ m}^3/\text{s}.$$

$$C = 4.17 \times 10^{-39} \exp(-1123.9/T) \text{ m}^6/\text{s}.$$

$$\epsilon_g(T) = \epsilon_g(300 \text{ K}) - a_T(T - 300), \text{ with } a_T = 0.0141 \text{ K}.$$

The other material parameters and transition matrix are adopted from [8]

B. QW Structures: Unstrained InGaAs-InGaAsP (1.25 Q) QW[6], [16]

$$B = 7 \times 10^{-17} \text{ m}^3/\text{s}, C = 6 \times 10^{-41} \text{ m}^6/\text{s}.$$

The QW parameters are the same as [6]. The lineshape broadening and bandgap shrinkage are modeled with the same parameters as bulk cases.

APPENDIX C

The gain coefficient can be written in the following form to clarify the wavelength dependency [6], [8]:

$$g = \frac{K_0}{\epsilon} (f_c - f_v) \sum_i \left[\frac{1}{2} + \frac{1}{\pi} \tan^{-1} \left(\frac{\epsilon - \epsilon_{gi}}{\delta\epsilon/2} \right) \right] \quad (\text{C.1})$$

where

K_0 gain scaling factor that includes the transition matrix element and other material parameters;

f_c and f_v Fermi functions for electrons and holes, respectively;

ρ_r reduced density of state;

$\delta\epsilon$ full-width at half-maximum (FWHM) of the Lorentzian lineshape.

To clarify the wavelength dependency, (19) was derived with the following assumptions: 1) all the quantized subbands have the same effective mass; 2) the transition matrix element is wavelength insensitive [8]; and 3) $\delta\epsilon$ is assumed to be small enough that the Fermi functions are approximately constant over the range of energies for which the lineshape function is significant.

The differential gain at transparency can be written as

$$a \equiv \left. \frac{dg}{dN} \right|_{N=N_{tr}} = \frac{K_0}{\epsilon} \left. \frac{d(f_c - f_v)}{dN} \right|_{N_{tr}} \cdot \sum_i \left[\frac{1}{2} + \frac{1}{\pi} \tan^{-1} \left(\frac{\epsilon - \epsilon_{gi}}{\delta\epsilon/2} \right) \right]. \quad (\text{C.2})$$

From the fact that $df_c/d\mu_n = -df_v/d\mu_p$ at transparency, we can obtain

$$\left. \frac{d(f_c - f_v)}{dN} \right|_{N_{tr}} = \left[\frac{df_c}{d\mu_n} \frac{d(\mu_n + \mu_p)}{dN} \right]_{N=N_{tr}}. \quad (C.3)$$

It can easily be shown that

$$\frac{df_c}{d\mu_n} = \left[2 + 2 \cosh \left(\frac{m_r}{m_c} (\epsilon - \epsilon_g) - \mu_n \right) \right]^{-1} \quad (C.4)$$

$$\frac{d(\mu_n + \mu_p)}{dN} = \frac{d(\epsilon - \epsilon_g)}{dN} = \frac{xJ}{N} \frac{d(\epsilon - \epsilon_g)}{dJ} = \frac{x\chi}{N}. \quad (C.5)$$

Therefore, the differential gain at transparency becomes

$$a = \frac{K_0}{\epsilon} \frac{x\chi}{N_{tr}} \left[2 + 2 \cosh \left(\frac{m_r}{m_c} (\epsilon - \epsilon_g) - \mu_n \right) \right]^{-1} \sum_i \left[\frac{1}{2} + \frac{1}{\pi} \tan^{-1} \left(\frac{\epsilon - \epsilon_{gi}}{\delta\epsilon/2} \right) \right]. \quad (C.6)$$

The differential gain at transparency can also be obtained from (18) as

$$a = \beta \frac{xJ_{tr}}{N_{tr}}. \quad (C.7)$$

Comparing (C.6) and (C.7), β is given by (19).

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REFERENCES

- [1] M. G. A. Bernard and G. Duraffourg, "Laser conditions in semiconductors," *Phys. Status Solidi*, vol. 1, pp. 699–703, 1961.
- [2] G. H. B. Thompson, *Physics of Semiconductor Laser Devices*. New York: Wiley, 1980, p. 79.
- [3] K. J. Vahala and C. E. Zah, "Effects of doping on the optical gain and the spontaneous noise enhancement factor in quantum well amplifiers and lasers studied by simple analytical expressions," *Appl. Phys. Lett.*, vol. 52, no. 23, pp. 1945–1947, 1988.
- [4] K. Y. Lau, "Ultralow threshold quantum well lasers," in *Quantum Well Lasers*, P. S. Zory, Jr., Ed. San Diego, CA: Academic, 1993, ch. 4.
- [5] K. Kikuchi, M. Amano, C. E. Zah, and T. P. Lee, "Measurement of differential gain and linewidth enhancement factor of 1.5- μ m strained quantum-well active layers," *IEEE J. Quantum Electron.*, vol. 30, pp. 571–577, Feb. 1994.
- [6] T. Makino, "Analytical formulas for the optical gain of quantum wells," *IEEE J. Quantum Electron.*, vol. 32, pp. 493–501, Mar. 1996.
- [7] S. L. Chuang, *Physics of Optoelectronic Devices*. New York: Wiley, 1995, ch. 10.
- [8] L. A. Coldren and S. W. Corzine, *Diode Lasers and Photonic Integrated Circuits*. New York: Wiley, 1995.
- [9] S.-L. Lee, "Wavelength measurement and tracking using semiconductor laser amplifiers for applications in photonic networks," *IEEE Photon. Technol. Lett.*, vol. 10, pp. 439–441, Mar. 1998.
- [10] S.-L. Lee, C.-T. Pien, and Y.-Y. Hsu, "Wavelength monitoring in DWDM networks using low cost semiconductor laser diode/amplifiers," in *Proc. OFC 2000*, Baltimore, MD, Mar. 2000, paper WK1.
- [11] T. P. Lee, C. E. Zah, and B. Bhat, "Multiwavelength DFB laser array transmitters for ONTC reconfigurable optical network testbed," *J. Lightwave Technol.*, vol. 14, pp. 967–976, June 1996.
- [12] B. E. Lemoff, "Technology alternatives for 10 Gbits/s LAN's," in *Proc. Optical Communication Conf. 2000*, Baltimore, MD, Mar. 2000, paper W11.
- [13] H. Kawata, H. Nakamura, and T. Sugie, "Service multiplex systems with wide pass band WDM (WWDM) technology for access networks," in *Proc. Fifth Optoelectronics Communications Conf.*, Japan, July 2000, paper 11D1-4.
- [14] S. H. Park, J. I. Shim, K. Kudo, M. Asada, and S. Arai, "Band gap shrinkage in GaInAs/GaInAsP/InP multi-quantum well lasers," *J. Appl. Phys.*, vol. 72, no. 1, pp. 279–281, 1992.
- [15] M.-S. Lin, A. B. Piccirilli, Y. Twu, and N. K. Dutta, "Temperature dependence of polarization characteristics in buried facet semiconductor laser amplifiers," *IEEE J. Quantum Electron.*, vol. 26, pp. 1772–1778, Oct. 1990.
- [16] M. Silver and E. P. O'Reilly, "Optimization of long wavelength InGaAsP strained quantum-well lasers," *IEEE J. Quantum Electron.*, vol. 31, pp. 1193–1200, July 1995.
- [17] P. J. A. Thijs, L. F. Tiemeijer, J. J. M. Binsma, and T. V. Dongen, "Progress in long-wavelength strained-layer InGaAs(P) quantum-well semiconductor lasers and amplifiers," *IEEE J. Quantum Electron.*, vol. 30, pp. 477–499, Feb. 1994.
- [18] P. W. A. McIlroy, A. Kurobe, and Y. Uematsu, "Analysis and application of theoretical gain curves to the design of multi-quantum-well lasers," *IEEE J. Quantum Electron.*, vol. 21, pp. 1958–1963, Dec. 1985.
- [19] J. Z. Wilcox, G. L. Peterson, S. Ou, J. J. Yang, M. Jansen, and D. Schechter, "Gain and threshold-current dependence for multiple-quantum-well lasers," *J. Appl. Phys.*, vol. 64, pp. 6564–6566, Dec. 1988.

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論文四

Study of effects of aging on transparent current stability of semiconductor lasers

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內容成果：

經由實驗觀察半導體光電元件的介面電壓與透明點電流的關係，並經由長時間的穩定度測試驗證其可行性與正確性。透明點為半導體光電元件操作在增益為零的狀態下，操作電流值，會隨入射光的波長的變化，而有相對應的關係。利用透明點電流的特性除了可以作為波長檢測器外，當元件操作在透明點時，以半導體光放大器為例，可減少串因雜訊，並提高其響應速度，在半導體光電元件中是一項相當重要的參數。

epitaxy (MBE). The cavity design had two focal points, one on the gain structure and the other one on the SESAM. The beam diameter on the gain structure was $\sim 200\mu\text{m}$ while it was significantly smaller ($\sim 18\mu\text{m}$) on the SESAM. The output coupler had a transmission of 1.5%.

Results: A slowly scanning autocorrelator with lock-in amplification was used to measure the pulse shape (Fig. 2). The autocorrelation was recorded for a maximum output power of 213 mW, showing that the pulses have a Gaussian shape and are free of any pedestals down to -30 dBc . The optical bandwidth (inset of Fig. 2) was $\sim 0.5\text{ nm}$, close to the transform limit for Gaussian pulses. The peak power was 30 W. The pulse duration as well as the stability of the modelocking was strongly dependent on the cavity setup. We observed that the pulses became shorter when the mode size on the SESAM was reduced, while the mode size on the gain structure was kept constant [12].

With photoluminescence microscopy we detected dark line defects in the gain structure. This might explain the low efficiency of $\sim 10\%$ for continuous wave operation, and its observed degradation: the output power significantly degrades within a few hours. With refined designs (e.g. including strain compensation for the quantum wells) we expect further substantial improvements.

Conclusion: We have demonstrated what is to our knowledge the highest power directly from a passively modelocked semiconductor laser oscillator. It is based on a vertical-external-cavity surface-emitting laser passively modelocked with an SESAM. We obtained 213 mW with 3.2 ps pulses at a repetition rate of 2.06 GHz. The basis for this performance is the gain structure design, which allows for low thermal impedance and a smooth gain spectrum. We believe that such lasers will soon deliver multi-watt average powers at several GHz repetition rates.

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References

- HO, P.T., GLASSER, L.A., IPPEN, E.P., and HAUS, H.A.: 'Picosecond pulse generation with a cw GaAlAs laser diode', *Appl. Phys. Lett.*, 1978, 33, p. 241
- GEE, S., ALPHONSE, G., CONNOLLY, J., BARTY, C., and DELFYETT, P.: 'Ultrashort pulse generation by intracavity spectral shaping and phase compensation of external-cavity modelocked semiconductor lasers', *IEEE J. Quantum Electron.*, 2000, 36, p. 1035
- ARAHIRA, S., MATSUI, Y., and OGAWA, Y.: 'Mode-locking at very high repetition rates more than terahertz in passively mode-locked distributed-Bragg-reflector laser diodes', *IEEE J. Quantum Electron.*, 1996, 32, p. 1211
- KUZNETSOV, F., HAKIMI, F., SPRAGUE, R., and MOORADIAN, A.: 'Design and characteristics of high-power ($> 0.5\text{-W}$ CW) diode-pumped vertical-external-cavity surface-emitting semiconductor lasers with circular TEM₀₀ beams', *IEEE J. Sel. Top. Quantum Electron.*, 1999, 5, p. 561
- XIANG, W.H., FRIEBERG, S.R., WATANABE, K., MACHIDA, S., SAKAI, Y., IWAMURA, H., and YAMAMOTO, Y.: 'Sub-100 femtosecond pulses from an external-cavity surface-emitting InGaAs/InP multiple quantum well laser with soliton effect-compression', *Appl. Phys. Lett.*, 1991, 59, p. 2076
- JIANG, W. and BOWERS, J.: 'Ultrafast vertical cavity semiconductor lasers' in DULUNG, I.N., III (Ed.): 'Compact sources of ultrashort pulses' (Cambridge University Press, New York, 1995), pp. 208–273

- HOLM, M.A., CUSUMANO, P., BURNS, D., FERGUSON, A.I., and DAWSON, M.D.: 'Mode-locked operation of a diode-pumped, external-cavity GaAs/AlGaAs surface emitting laser'. Conf. Lasers and Electro-Optics. CLEO'99, 1999 OSA Technical Digest Series (Optical Society of America, Washington DC, 1999), 1999, Paper CTuK63
- HOOGLAND, S., DHANJAL, S., TROPPER, A.C., ROBERTS, S.J., HÄRING, R., PASCHOTTA, R., and KELLER, U.: 'Passively mode-locked diode-pumped surface-emitting semiconductor laser', *IEEE Photonics Technol. Lett.*, 2000, 12, p. 1135
- KELLER, U., WEINGARTEN, K.J., KÄRTNER, F.X., KOPF, D., BRAUN, B., JUNG, I.D., FLUCK, R., HÖNNINGER, C., MATUSCHEK, N., and AUS DER AU, J.: 'Semiconductor saturable absorber mirrors (SESAMs) for femtosecond to nanosecond pulse generation in solid-state lasers', *IEEE J. Sel. Top. Quantum Electron.*, 1996, 2, p. 435
- KELLER, U.: 'Semiconductor nonlinearities for solid-state laser modelocking and Q-switching' in GARMIRE, E. and KOST, A. (Eds.): 'Nonlinear optics in semiconductors' (Academic Press, Inc., Boston, 1999), pp. 211–286
- SO, W.W., and LEE, C.C.: 'Fluxless process of fabricating In-Au joints on copper substrates', *IEEE Trans. Comp. Packag. Technol.*, 2000, 23, p. 377
- HÄRING, R., PASCHOTTA, R., MORIER-GENOUD, F., KELLER, U., GARNACHE, A., OESTERLE, U., ROBERTS, J.S., HOOGLAND, S., DHANJAL, S., and TROPPER, A.C.: 'Passively mode-locked diode-pumped surface-emitting semiconductor laser'. Conf. Lasers and Electro-Optics (Optical Society of America, Washington DC, 2000), 2000

Study of effect of aging on transparent current stability of semiconductor lasers

San-Liang Lee and Yu-Yi Hsu

Wavelength sensing using the wavelength-dependent transparency of laser diodes requires long-term stability of the transparent current. An accelerated aging test has been performed that demonstrates that the transparent current becomes stable over a wide range of wavelengths after 300 h of aging.

Introduction: A direct-bandgap semiconductor optoelectronic device is operated at its transparent condition when its material gain is zero. This unique property is strongly correlated with the characteristics of semiconductor lasers and amplifiers, such as the threshold current, thermal effects, and gain. Many research groups have made great efforts towards investigating the transparent properties for achieving low threshold current or reducing the thermal effects of laser diodes [1, 2]. The transparent condition is also important for operating a semiconductor optical amplifier as a low-crosstalk gate switch [3] or for achieving fast response and high saturation power with optical speed-up at transparency [4]. Besides, the wavelength dependency of the transparency can be exploited for wavelength sensing or monitoring [5, 6].

One critical concern of the transparent property for the above applications is the long-term stability. The most problematic situation relates to wavelength-sensing applications. To obtain a wavelength accuracy of 0.01 nm around the 1.55 μm fibre window, wavelength discrimination by detecting the transparent current would need a current accuracy of 100 ppm [6]. Such current accuracy imposes a critical requirement on the reliability of the transparent current. In this Letter, we report the results of accelerated aging tests on commercial laser diodes and investigate how the transparent current changes as a result of aging.

Experiment: To investigate the wavelength dependency of the transparency stability, the transparent current was measured with detection of the induced voltage across the diode junction, in response to input-modulated light signals [5, 6]. This technique requires stable optical coupling, so the measurements were carried out with pigtailed lasers to eliminate time-consuming optical coupling. Four Fabry-Perot (FP) lasers from the same vendor were tested at 85°C with an injection current of 150 mA, which is about one order higher than the transparent current. The transparent current was measured at 25°C after each aging period. For the experiment, a high-performance tunable laser with optical modula-

tion was used as the probe signal, and the induced voltage across the diode junction was measured with a lockin amplifier. The modulation frequency was 1 kHz. Owing to the fact that the transparent current depends on the input polarisation, a polarisation controller was placed between the tunable laser and the diode laser under test. The transparent current corresponds to the bias current where the induced voltage vanishes, and it can be measured to a resolution of 0.01 mA with a low-noise current source.

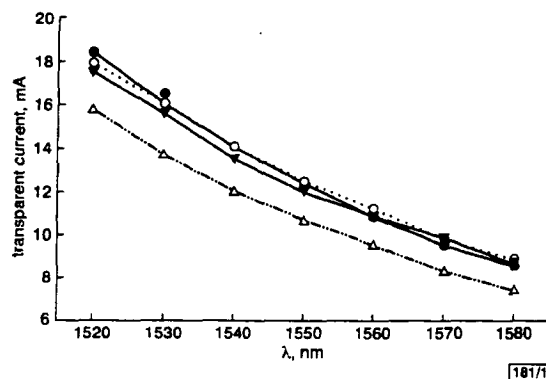


Fig. 1 Wavelength dependent transparent current for lasers before aging test

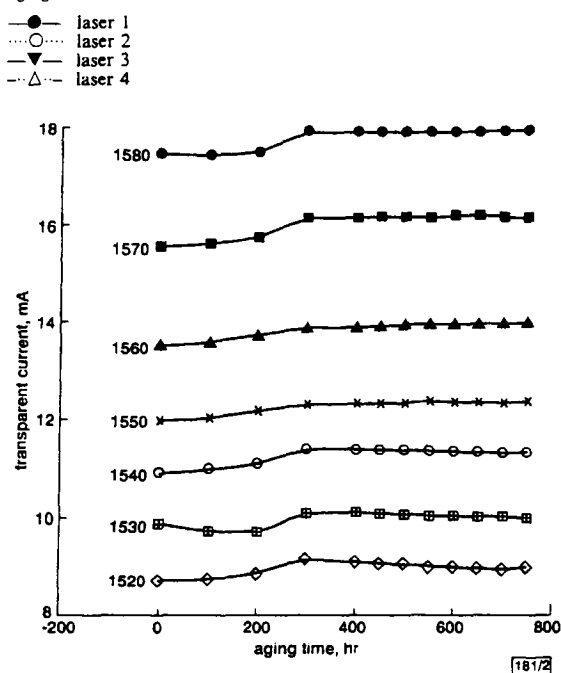


Fig. 2 Transparent current variation during aging for laser 3

Results: Fig. 1 shows the transparent current before aging test for the four lasers over a wavelength range from 1520 to 1580 nm. The transparent characteristics are similar except that the transparent current of laser 4 is considerably lower. After accelerated aging, all lasers show apparent drift in the transparent current over the first 300 h. The transparent current remains almost fixed after 300 h. Fig. 2 depicts the evolution of transparent current against aging time observed at different wavelengths for laser 3. All the wavelength channels show a very similar tendency in terms of transparent current variation. To compare the difference between the lasers, the drift in the transparent current of a laser is represented by averaging the drift over all the wavelengths. The average wavelength drift is summarised in Fig. 3. Lasers 1 and 2 show relatively small wavelength drift and their transparent conditions also become stable after ~300 h of aging. This aging characteristic is similar to the results observed in the aging analysis of DBR lasers [7]. Since the transparent current is much smaller than the aging current, these data indicate that sufficient stability for

wavelength sensing can be generated by resolving the transparent current if the sensors first pass several hundred hours of burn-in test.

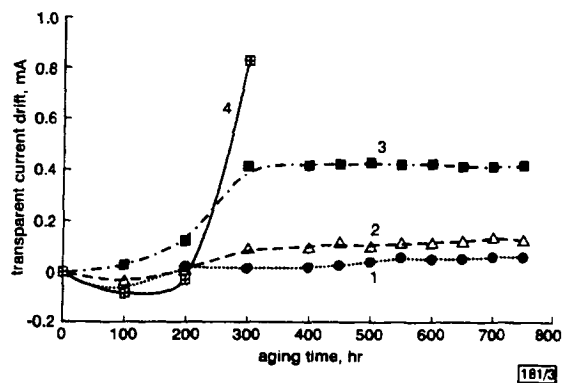


Fig. 3 Comparison of average transparent current drift among lasers

The aging of laser 4 results in a significant increase in the transparent current for all wavelengths, in spite of the relatively low transparency before aging. The increase in transparent current is accompanied by an obvious rise in the threshold current and decrease in the output power. This result suggests that the performance degradation of this laser is most likely due to the increase in the non-radiative recombination or leakage current. Facet degradation alone will not change the transparent current.

All the wavelengths show the same stability against aging time but a slightly different amount of transparency drift after 750 h of accelerated aging. Further analysis of the aging test data, as shown in Fig. 4, reveals that the transparent current drift shows resonance-like variation with wavelength. This can be attributed to the effect of the FP resonance of the laser diode on the measurement of the transparent current. Such an effect arises from the slight dependence of the transparent current on the input light power due to free-carrier absorption and nonuniform carrier distribution [6]. The drift in the transparent condition during aging leads to a shift in the resonant wavelength of the FP cavity because of the change in the refractive index for a fixed bias current. This changes the light intensity inside the cavity for a given wavelength. Thus, the amount of transparency drift depends on the relative position of the incident wavelength with respect to the resonant peak. Such resonance-like variation can be eliminated when the facets are coated with antireflection film.

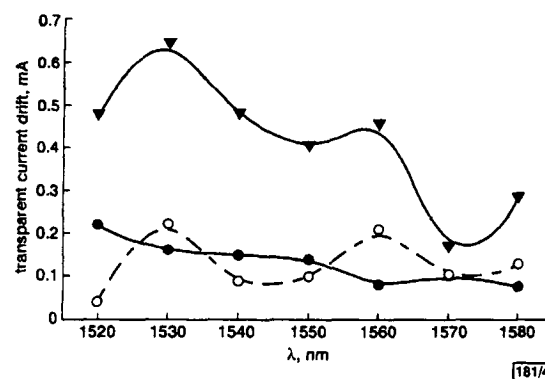


Fig. 4 Transparent current drift against wavelength

—●— laser 1
—○— laser 2
—▲— laser 3

Conclusion: The results of reliability tests indicate that the transparent current of laser diodes becomes stable after 300 h of burn-in test. Although the transparent current drift is slightly different for different wavelengths, good transparency stability can be observed for a wide range of wavelengths. This result is important for applying wavelength-dependent transparent current in wavelength sensing and monitoring.

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References

- 1 VAHALA, K.J., and ZAH, C.E.: 'Effects of doping on the optical gain and the spontaneous noise enhancement factor in quantum well amplifiers and lasers studied by simple analytical expressions', *Appl. Phys. Lett.*, 1988, **52**, (23), pp. 1945-1947
- 2 LAU, K.Y.: 'Ultralow threshold quantum well lasers' in ZORY, P.S., JR. (Ed.): 'Quantum well lasers' (Academic Press, San Diego, CA, 1993), Chap. 4
- 3 SHIBATA, Y., YAMADA, Y., HABARA, K., and YOSHIMOTO, N.: 'Semiconductor laser diode optical amplifier/gates in photonic packet switching', *J. Lightwave Technol.*, 1998, **16**, (12), pp. 2228-2235
- 4 DUPERTUIS, M.A., PLEUMEEKERS, J.L., HESSLER, T.P., SELBMANN, P.E., DEVEAUD, B., DAGENS, B., and EMERY, J.Y.: 'Extremely fast high-gain and low-current SOA by optical speed-up at transparency', *IEEE Photonics Technol. Lett.*, 2000, **12**, (11), pp. 1453-1455
- 5 LEE, S.-L.: 'Wavelength measurement and tracking using semiconductor laser amplifiers for applications in photonic networks', *IEEE Photonics Technol. Lett.*, 1998, **10**, (3), pp. 439-441
- 6 LEE, S.-L., PIEN, C.-T., and HSU, Y.-Y.: 'Operation principles of wavelength sensing using transparent properties of semiconductor optical diodes', *J. Lightwave Technol.*, 2001, **19**, pp. 655-665
- 7 DELORME, F., PIERRE, P., DE BAILLIENCOURT, H., TEROL, G., and DEVOLDERE, P.: 'Wavelength aging analysis of DBR lasers using tuning section IM frequency response measurements', *IEEE Photonics Technol. Lett.*, 1999, **11**, (3), pp. 310-312

Efficiency improvement of broadband microwave amplifiers using current sharing concept

A.S. Virdee and B.S. Virdee

The authors present experimental results of a 2 to 18 GHz high-gain dual-stage feedback amplifier employing a novel concept of current sharing that drastically reduces the current consumption by 50% without compromising the amplifier performance and hence produces a more than two-fold improvement in the power-added efficiency.

Introduction: The relentless drive to increase the packaging density of microwave sub-systems and hence the reduction in the sub-system size inevitably puts great demand on the miniaturisation of components and devices that are employed in the sub-system realisation. This demand also, however, puts severe constraint on the current consumption requirements in order to minimise the power dissipation of the components and devices. Amplifier devices employed in microwave sub-systems are the major source of current consumption and, hence, power dissipation. Reduction in the power dissipation of such devices is essential to minimise adverse thermal effects and hence provide significant improvement in system reliability.

In this Letter, the concept of current sharing is introduced and applied to a dual-stage feedback amplifier that is employed in the realisation of high-gain broadband feedback amplifiers operating across the 2 to 18 GHz frequency range. This device is employed in highly demanding systems such as electronic warfare systems. The DC bias arrangement conventionally employed in broadband amplifier designs results in poor efficiency performance due to excessive current consumption. In this Letter, it is demonstrated

that a novel current sharing concept can be successfully implemented onto a dual-stage broadband feedback amplifier to realise a substantial reduction in current consumption and an improvement in the efficiency performance. For comparative purposes, an identical specification broadband amplifier was designed and fabricated using a conventional biasing technique.

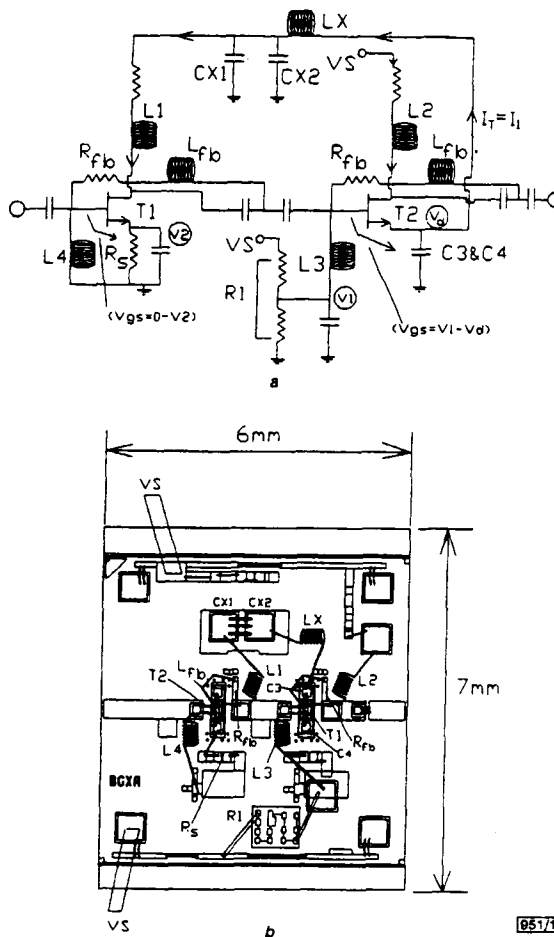


Fig. 1 2-18 GHz two-stage feedback amplifier in current sharing mode
a Schematic diagram
b Layout

Current sharing concept: In conventional feedback amplifiers, the DC bias [1] is applied using the technique of 'self-bias'. The advantage of this biasing technique is that the devices can be operated from a single positive supply without the need to have a sequencing circuit [2] that provides separate gate and drain bias. However, the major problem associated with amplifier designs using this type of biasing mode is that each active device requires a separate biasing network, hence an increased current requirement.

In this Letter, an alternative technique is introduced that employs the concept of current sharing, as illustrated in Fig. 1a, where the same current drawn by one active device is also used to bias another active device. The total current ($I_T = I_1$) drawn by the dual-stage feedback amplifier is set by the source resistor R_S of the first device T1, which is biased in the self-bias mode. Biasing of the second device T2 (voltage V_1) is set by the potential divider R1 connected to the gate of device T2. The drain voltage of T2 (voltage V_d) is connected to the source decoupling capacitors C3 and C4, that enables the correct gate-source voltage ($V_{gs} = V_1 - V_d$) to forward bias the device T2. The current flows from VS through T2, T1 and R_S to ground. The supply voltage is used to bias T2, T1 and the potential divider R1. Inductor LX provides RF isolation between the source of device T2 and the drain of device T1. The selection of components C3, C4, CX1, CX2, and the isolating inductor LX are critical in ensuring stable DC operation of the amplifier across the entire (2-18 GHz) multi-octave

論文五

High-resolution wavelength monitoring using differential detection of
junction voltage across single diode laser

作者：李三良 (S.-L. Lee)，許毓倚 (Y.-Y. Hsu)，卞敬塘 (C.-T. Pien)

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內容成果：

利用便宜的 FP 雷射，並使用差動量測技術，提高波長監控的解析度與精確度。半導體雷射的介面電壓會受入射光的偏振狀態、光功率、波長等因素而改變，作為波長檢測器必須抑制因偏振與光功率導致的誤差，並提高波長解析度。本文中提出利用差分方式，將週期性些微的偏壓電流變化注入雷射二極體，使其中交互工作在對稱的正波長鑑別曲線與負波長鑑別曲線，扣除因偏振與功率變化所造成的介面電壓變化值，再利用波長變化比例的關係，提高波長鑑別的解析度。

High-Resolution Wavelength Monitoring Using Differential/Ratio Detection of Junction Voltage Across a Diode Laser

San-Liang Lee, *Member, IEEE*, Yu-Yi Hsu, and Ching-Tang Pien

Abstract—High-resolution, compact, and power-insensitive wavelength monitoring can be realized by using a low-cost diode laser as the wavelength sensor. The wavelength resolution can be improved by more than an order with the help of the Fabry–Perot resonance inside the cavity of a diode laser, as well as by differential or ratio detection. Wavelength resolution of 0.01 nm can be obtained with much relaxed requirement on the detection circuitry. The wavelength monitoring with such resolution can tolerate up to 15 dB of power variation on the incident light power.

Index Terms—Differential detection, semiconductor lasers, transparent current, wavelength monitoring, wavelength sensing.

I. INTRODUCTION

DENSE wavelength-division-multiplexing (DWDM) networks require effective schemes for wavelength monitoring in order to control the source wavelength or manage the wavelength channels. Most existing monitoring techniques use interferometer-type apparatus [1]–[7]. Interferometers made of Fabry–Perot (FP) etalons are especially popular for wavelength control of DWDM lasers [5]–[7]. The detection of the transparent condition of a semiconductor optical amplifier (SOA) or laser diode was proposed for wavelength measurement or monitoring [8], [9]. This approach has the advantages of compactness and possibility for integration with other optoelectronic devices. However, the wavelength resolution is limited by the resolution of the sweeping current for locating the transparent condition.

In this letter, we demonstrate simple schemes of high-resolution wavelength monitoring using the transparent properties of a diode laser. Significant improvement on the monitoring resolution is achieved by utilizing a combination of cavity effects and differential detection, so wavelength resolution of 0.01 nm can easily be obtained. Using ratio detection can further relax the requirement on stabilizing the input power. This technique can be tuned to monitor multiple wavelength channels by simply changing the bias condition.

II. OPERATION PRINCIPLE

The transparent current of a diode laser is roughly inversely proportional to the incident light wavelength, similar to the characteristic of an SOA [10]. However, it was found in experiments that the resonant cavity of a diode laser causes an oscillating characteristic in the wavelength variation of transparent current. Such characteristic can improve the resolution of wavelength monitoring for fixed channels, because the resonant effect can enhance the variation rate of the transparency with the wavelength.

For using as a wavelength monitor, the diode laser is usually biased around the transparent condition for the monitored wavelength. The induced junction voltage in response to a dithered input light can simply be demodulated with envelope detection [11], which does not require phase-sensitive detection. The output signal can then be written as $v = \eta|I - I_{tr}(\lambda)|P$, where I and I_{tr} are the bias current and transparent current, respectively. P is the input power and η is a scaling constant. The output voltage appears as a V-shape around the transparency in varying the bias current, as shown in Fig. 1(a). As the wavelength changes, the V-curve will shift to the right or left side, depending on the sign of wavelength change. The wavelength change can be monitored from differential detection of the junction voltages for a two-level switching bias. The output voltage after the differential amplifier vanishes for a monitored wavelength. When the input wavelength shifts toward the long-wavelength side, the V-curve moves to the left, and the differential voltage becomes positive. On the other hand, a negative voltage at the differential output indicates a blue shift in the wavelength. Therefore, the output voltage can clearly indicate the sign of wavelength drift and can serve as a control signal for managing the incident wavelength.

Fig. 1(b) shows the schematic of such differential wavelength monitoring technique. The output differential voltage for $\Delta\lambda$ of wavelength drift can be written as

$$\Delta v = A(\eta_+ + \eta_-)P\Delta\lambda\partial I_{tr}/\partial\lambda \quad (1)$$

where A is a scaling factor. The subscript associated with the η -parameter is used to designate the difference in the material gain for the two bias levels. Equation (1) indicates that the differential detection roughly doubles the variation of the output voltage with the wavelength drift, comparing to the wavelength discrimination without differential detection. This doubles the wavelength resolution for the same voltage resolution.

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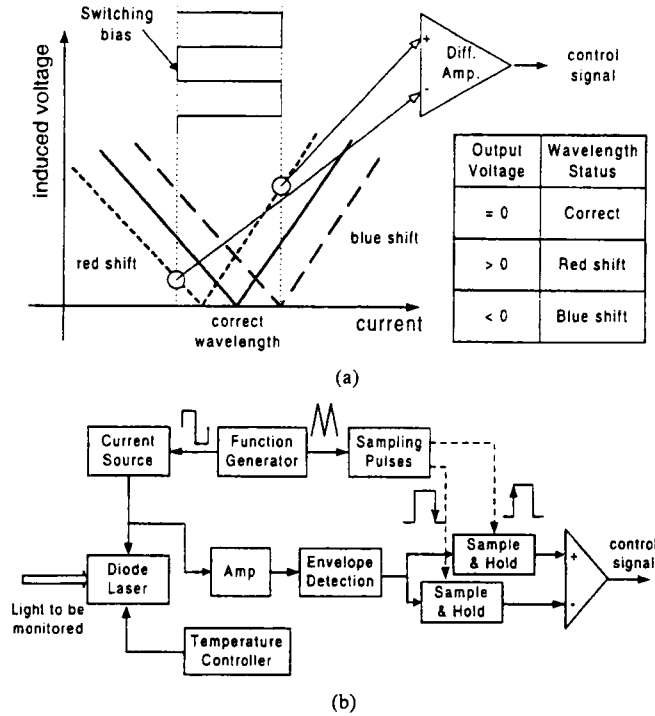


Fig. 1. (a) Illustration of the concept. (b) Block diagram of wavelength monitoring using differential detection of the induced junction voltage across a diode laser.

An alternative scheme is to utilize the voltage ratio of the two bias levels for wavelength discrimination. The differential voltage depends on the incident power, so its accuracy is affected by any input power fluctuation. Besides, the induced voltage may be polarization sensitive due to different coupling and confinement between the TE and TM polarizations. Detection with voltage ratio can achieve better wavelength stability without the necessity of power monitoring. It can also eliminate the polarization sensitivity since the transparent points of the polarizations are the same. In addition, for applications in wavelength stabilization, ratio detection can generate power-independent control signal for adjusting the source wavelength. This idea can be realized by simply changing the differential amplifier at the output stage of Fig. 1(b) into a voltage divider. The change in voltage ratio (in decibels) for $\Delta\lambda$ of wavelength drift can be written as

$$\Delta\rho = 4.34(1 + \eta_+/\eta_-)(\Delta\lambda/\Delta I_-)\partial I_{tr}/\partial\lambda \quad (2)$$

where ΔI_- is the difference between the lower bias level and the transparent point. Monitoring with ratio detection can also take advantage of resolution enhancement, as indicated by the $1 + \eta_+/\eta_-$ factor.

III. EXPERIMENTS AND RESULTS

Fig. 2 shows that the transparent current of the tested laser diode has a resonant behavior for a fine scale of wavelength change. The period of the resonance was found to be identical to the cavity mode spacing of the diode laser. Such resonant effect can give rise to improvement on the wavelength resolution by about an order, due to the enhancement of the differential slope from 0.21 to 2.0 mA/nm. Since changing the diode tem-

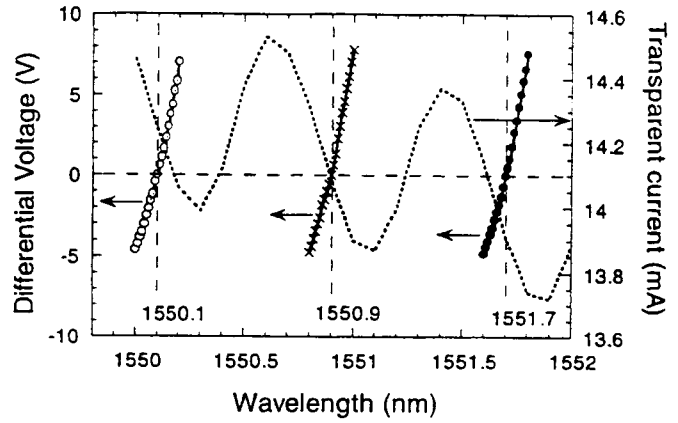


Fig. 2. Variation of transparent current and output differential voltage with the wavelength. The differential voltages for three different biases were measured in a wavelength step of 0.01 nm.

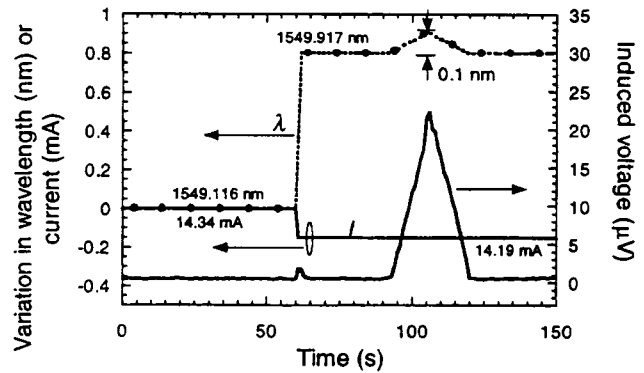


Fig. 3. Demonstration of tunable wavelength monitoring using the transparent characteristic of an FP laser. When the wavelength is tuned by 0.8 nm, the bias current is changed accordingly such that the induced voltage remains at the minimum. As the wavelength drifts slightly, the induced voltage increases to indicate the drift.

perature can shift the $I_{tr}-\lambda$ curve, the monitored wavelength can always be aligned to where the slope is maximal. If the cavity mode spacing is designed to be roughly equal to the channel spacing for a DWDM system, one diode laser can be tuned to monitor multiple channels by simply changing the bias current. Such multiwavelength monitoring capability is shown in Fig. 2, where the differential voltages for wavelength variation around three wavelength channels spaced by 0.8 nm are depicted.

The tunable monitoring was demonstrated using an external-cavity tunable laser to simulate the channel switching and wavelength drift. Fig. 3 shows that as the wavelength is switched to another channel, the bias current can be tuned to the corresponding transparent current for the new wavelength such that the induced voltage remains minimal. When the wavelength drifts due to the change in ambient condition or aging, the rise of the induced voltage indicates this drift, and an appropriate countermeasure can be applied to bring the wavelength back to the correct value. Such tunable monitoring is especially advantageous when it is applied to monitor fast tunable lasers.

The voltage difference shown in Fig. 4(a) increases with the input power, so raising the input power can improve the monitoring resolution. The required resolution in voltage measurement is 1.2 μV for 0.01 nm of wavelength resolution if the

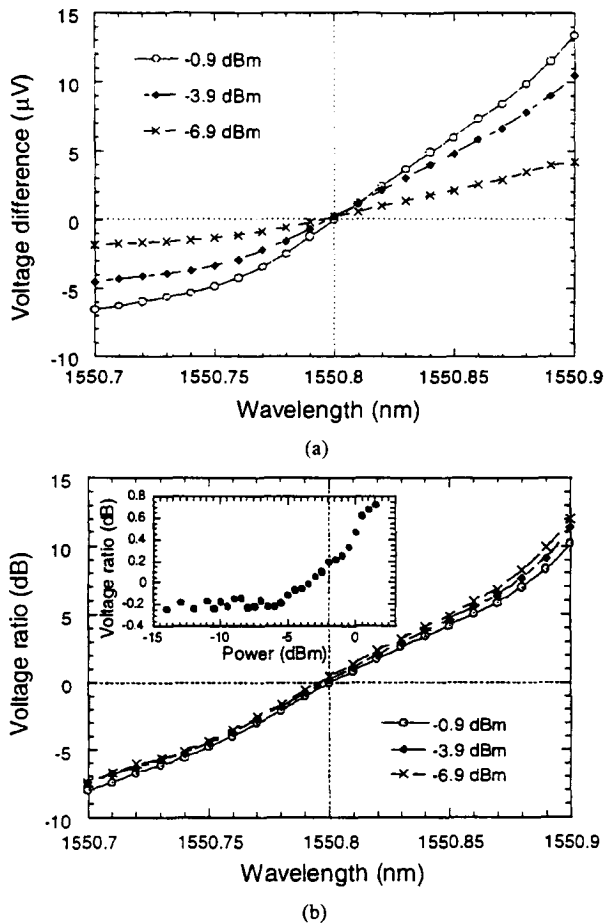


Fig. 4. Output variation versus wavelength drift for (a) differential detection and (b) ratio detection. The two bias levels are 13.99 and 14.48 mA, respectively; and the transparent current for the monitored wavelength is 14.31 mA. The inset shows the variation of voltage ratio with input power.

input power is -0.9 dBm. This resolution is more achievable with simple circuitry than that of wavelength monitoring using the induced voltage of an SOA [8]. Fig. 4(b) depicts the measured voltage ratio for the two bias levels in response to different input power. The slope of the curves, 0.9 dB/0.01 nm, is consistent with the calculation using (2) for $\Delta I_{tr} = 0.17$ mA, and $\partial I_{tr}/\partial \lambda = 2.0$ mA/nm. This slope is almost independent of the power level. The data shown in the inset of Fig. 4(b) indicate that the change in the voltage ratio for 15 dB of power change is less than that for 0.01-nm wavelength drift. Therefore, wavelength monitoring with ratio detection can achieve better than 0.01-nm resolution with high tolerance on power variation. Further measurements indicate that the voltage fluctuation during half an hour of continuous operation is much less than the voltage difference for 0.01 nm of wavelength drift.

Temperature sensitivity and long-term reliability for the transparent current are the critical concerns in practical applications. To achieve wavelength stability of 0.01 nm would require a temperature stability of about 0.01 °C, which is achievable but tough for existing temperature control techniques. Opti-

mization of diode parameters may relax the requirements on the stability and reliability of temperature controllers. The transparent point has been shown to be stable over 33 h of continuous operation [9]. It was also shown that the transparent current becomes stable after 300 h of accelerated aging [12]. Therefore, reliable and accurate wavelength monitoring with transparency detection is achievable.

IV. CONCLUSIONS

We have demonstrated high-resolution wavelength monitoring schemes using a low-cost diode laser as the sensor. The FP resonance inside the cavity of a diode laser was shown to significantly enhance the slope of the I_{tr} - λ curve. Differential or ratio detection can further improve the monitoring performance. The wavelength monitoring technique is tunable, such that a diode laser can monitor many wavelength channels. The demonstrated 0.01-nm wavelength resolution with much relaxed requirement on the detection circuitry indicates that such schemes can be very attractive for application in DWDM systems.

REFERENCES

- [1] A. J. Keating and A. J. Lowery, "Wavelength stabilization in packet-switched WDM networks," *J. Lightwave Technol.*, vol. 15, pp. 76–85, Jan. 1997.
- [2] J. P. Monochalin, M. J. Kelly, J. E. Thomas, N. A. Kunitz, A. Szoke, P. H. Lee, and A. Javan, "Accurate laser wavelength measurement with a precision two-beam scanning Michelson interferometer," *Appl. Opt.*, vol. 20, pp. 736–757, Mar. 1981.
- [3] M. Teshima, M. Koga, and K.-I. Sato, "Performance of multiwavelength simultaneous monitoring circuit employing arrayed-waveguide grating," *J. Lightwave Technol.*, vol. 14, pp. 2277–2285, Oct. 1996.
- [4] G. E. Shtengel, R. F. Kazarinov, and L. E. Eng, "Simultaneous laser wavelength locking and spectral filtering using fiber Bragg grating," in *Int. Semiconduct. Laser Conf.*, Nara, Japan, 1998, pp. 269–270.
- [5] B. Villeneuve, H. B. Kim, M. Cyr, and D. Gariepy, "A compact wavelength stabilization scheme for telecommunication transmitters," in *WDM Comp. Technol., IEEE/LEOS Summer Top. Meetings*, 1997, pp. 19–20.
- [6] S. Matsumoto and M. Fujise, "Frequency stabilization of $\lambda/4$ -shifted DFB laser to stabilized optical resonator," *Electron. Lett.*, vol. 25, pp. 814–815, June 1989.
- [7] D. A. Ackerman, K. F. Dreyer, U. Koren, J. W. Stayt, S. L. Broutin, W. A. Asous, J. E. Johnson, L. J.-P. Ketelsen, K. K. Kamath, S. O'Brien, W. J. Shakespeare, M. A. Esheman, M. M. Meyers, D. A. Snyder, and E. S. Mak, "Wavelength, modal, and power stabilization of tunable electro-absorption modulated distributed Bragg reflector lasers," in *Int. Semiconduct. Laser Conf.*, Monterey, CA, 2000, pp. 49–50.
- [8] S.-L. Lee, "Wavelength measurement and tracking using semiconductor laser amplifiers for applications in photonic networks," *IEEE Photon. Technol. Lett.*, vol. 10, pp. 439–441, Mar. 1998.
- [9] S.-L. Lee, C.-T. Pien, and Y.-Y. Hsu, "Wavelength monitoring in DWDM networks using low-cost semiconductor laser diode/amplifiers," in *Conf. Opt. Fiber Commun. 2000*, Baltimore, MD, Mar., pp. 168–170.
- [10] G. E. Shtengel and D. A. Ackerman, "Internal optical loss measurements in 1.3 μ m InGaAsP lasers," *Electron. Lett.*, vol. 31, pp. 1157–1158, July 1995.
- [11] A. B. Carlson, *Communication Systems*, 3rd ed, New York: McGraw-Hill, 1986, pp. 220–225.
- [12] S.-L. Lee and Y.-Y. Hsu, "Aging study on transparent current stability of semiconductor lasers," *Electron. Lett.*, submitted for publication.

論文六

Performance Enhancement on SOA-Based Four-Wave-Mixing Wavelength Conversion Using an Assisted Beam

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內容成果：

波長轉換器為實現高速光通訊網路中一個相當重要的元件。本文使用半導體光放大器作為波長轉換的元件，利用半導體光放大器的非線性效應：四波混合技術，注入一幫浦光源與訊號光源，產生第三道對應的轉換波長。然而目前使用此方式有波長轉換效率不高的缺點，因此提出再多加一道輔助光，提高轉換效率。經由實驗，當加入輔助光之後，波長轉換器的轉換效率可提升 5dB，且訊號功率與背景功率比也提升了 3dB，對於轉換訊號的功率級訊號品質，增加了不少幫助，提高使用四波混合製作全光波長轉換器的可行性。

Performance Enhancement on SOA-Based Four-Wave-Mixing Wavelength Conversion Using an Assisted Beam

San-Liang Lee, *Member, IEEE*, Pei-Miin Gong, and Chin-Tien Yang

Abstract—We propose to use an additional injection beam of short wavelength to enhance the wavelength conversion that utilizes the four-wave-mixing effect in a semiconductor optical amplifier (SOA). The experiments demonstrated larger than 5-dB improvement on conversion efficiency, as the SOA is biased at the corresponding transparent condition for the assisted beam. At such a condition, an assisted beam can increase the saturation intensity without sacrificing the gain of an SOA, and this leads to an increase in conversion efficiency. The assisted beam also improves the signal-to-background-noise ratio in wavelength conversion. The effects of different assisted wavelengths and the measures for further improvement are discussed.

Index Terms—Assisted beam, four-wave mixing, optical networks, semiconductor optical amplifiers, wavelength conversion.

I. INTRODUCTION

ALL-OPTICAL wavelength converters can be used in optical networks to solve the wavelength-blocking problems and provide wavelength routing and switching. There exist several wavelength conversion techniques using semiconductor optical amplifiers (SOAs), including cross-gain modulation, cross-phase modulation, and four-wave mixing (FWM) [1], [2]. The conversion using the FWM effect has the merits of transparency to data rate and format and providing chirp compensation to the input signals. Though high conversion efficiency has been demonstrated with SOAs of very high gain [3], [4], the FWM conversion using typical SOAs generally suffers from poor efficiency and noise figure. Significant improvement on the conversion efficiency and noise figure is required for practical applications.

The FWM conversion efficiency increases approximately with the unsaturated gain and the square of the saturation power of an SOA [5]. Therefore, increasing the gain and/or saturation power of an SOA is essential to improve the conversion efficiency. The saturation intensity is inversely proportional to the carrier lifetime, so speeding up the carrier recovery rate of an SOA can enhance the conversion efficiency. Several research groups have proposed the schemes to change carrier dynamics and gain characteristics of an SOA by injecting an

additional light, accompanying the signal light [6]–[10]. The additional light is referred as an assisted light hereafter. In principle, there are three scenarios between the SOA bias and the role of the assisted light. For the first scenario, the SOA is operated at the gain regime for the assisted light [6]–[8]. The presence of the assisted light raises the saturation intensity but decreases the amplifier gain. The second scenario is that the SOA is operated at a bias close to the transparent condition for the assisted beam [9], [10]. This technique can increase the saturation intensity without sacrificing the gain, and it was recently proposed to speed up the gain recovery of an SOA [10], [11]. The third one is to use an assisted light for which the SOA is biased at the absorptive regime. The assisted light can increase both the gain and saturation power of an SOA, so significant improvement on the conversion efficiency should be achievable by using an assisted light. However, the SOA current for this scenario is usually too small to provide enough gain for wavelength conversion. In this letter, we will demonstrate the improvement on the conversion efficiency and signal-to-noise ratio for FWM-based SOA wavelength converters by injecting an assisted beam.

II. GAIN CHARACTERISTICS

The transparent condition of an SOA corresponds to the condition that the material gain of the SOA is zero for the incident wavelength. The transparent current for an assisted beam can be determined from comparing the gain-versus-current curves at a given wavelength between the presence and absence of the assisted beam. The transparency locates at the intersection of the two curves, since the gain is not changed for the presence of the assisted beam. Fig. 1(a) depicts the gain curves at 1550 nm for the cases with and without a 1480-nm assisted beam for a commercial 1.55- μm SOA used in our experiments. The two curves intersect at 105 mA of bias current, which corresponds to the transparent current for the 1480-nm light. The transparent current is larger for a shorter wavelength.

The gain of an SOA can simply be increased by raising the bias current to the SOA, but the ultimate gain is limited by the thermal effect and amplified spontaneous emission (ASE) at high-injection level. On the other hand, adding an assisted beam can further increase the gain and/or the saturation output power. Fig. 1(b) shows that the assisted beam does not affect the unsaturated gain at the transparent condition and raises the saturation power by 4.4 dB. Also shown in Fig. 1(b) are the gain saturation characteristics when the SOA is biased at 200 mA. At such

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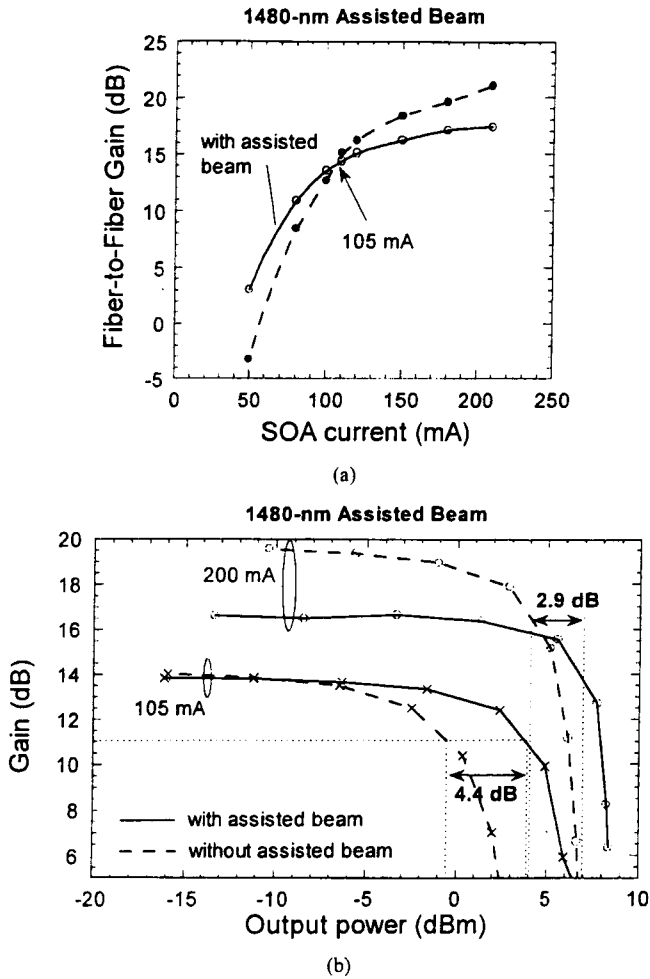


Fig. 1. Comparisons of (a) small-signal gain versus current and (b) gain versus output power of an SOA for a 1550-nm signal between the absence and presence of an assisted beam. The solid and dashed curves represent the gain with and without an assisted beam. The assisted power is 23 dBm.

a high-injection bias, the amplified assisted beam causes about 2.8 dB of reduction in the gain but 2.9 dB of increase in the output saturation power.

III. WAVELENGTH CONVERSION

To investigate the improvement on wavelength conversion efficiency with an assisted beam, the measurement was carried out using a 1480- or 1430-nm assisted beam. From both simulated and empirical results, the assisted beam can have best effects when it is incident contradirectionally with the probe and pump signals. The pump and probe beams out of two commercial tunable lasers with built-in isolators were combined using a 50/50 coupler and inputted to the SOA. A 1.48/1.55- μm multiplexer was used at the output end to separate the input assisted beam from the output conjugated signal. The multiplexer has a broad passband that provides low insertion loss to both the 1480- and 1430-nm assisted beams. The conjugated output and noise level were measured with an optical spectrum analyzer.

The measurement was first performed with a 1480-nm assisted beam of 23.3-dBm power. The probe power was fixed at -10 dBm, and the pump wavelength was fixed at 1553 nm.

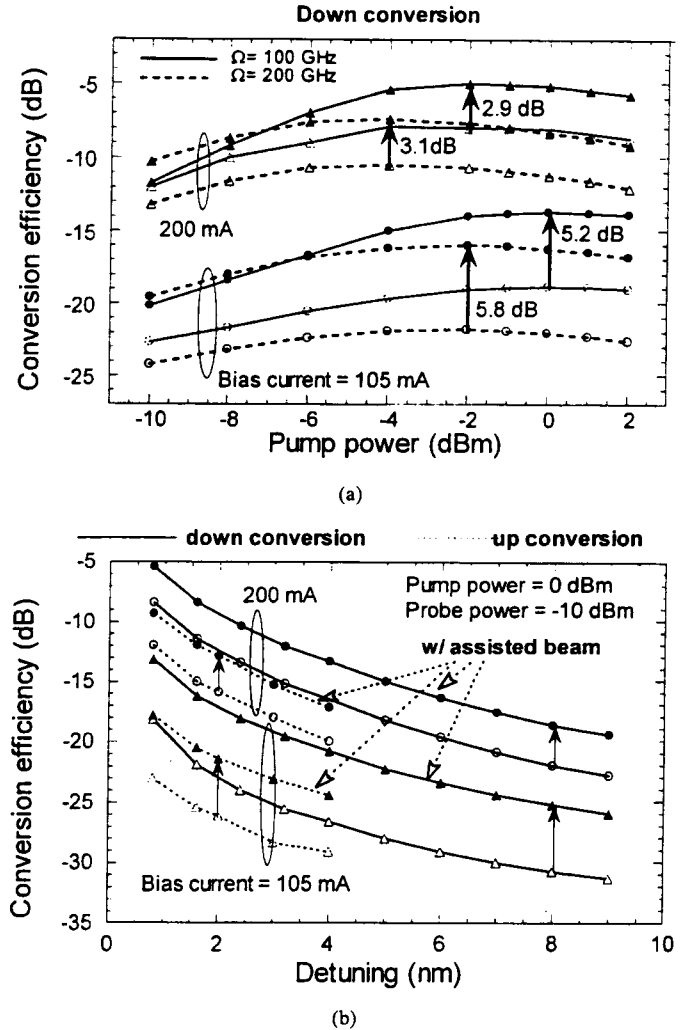


Fig. 2. Effects of 1480-nm assisted beam on (a) wavelength conversion efficiency versus pump power and (b) conversion efficiency versus detuning. The curves marked with solid circles or triangles represent the results with an assisted beam. The dashed lines in part (b) indicate the results for up conversion. The assisted power is 23 dBm.

Wavelength conversion was performed at two different bias currents, one at transparency (105 mA) and the other at high injection (200 mA). The comparison of the conversion efficiency with and without the assisted beam is shown in Fig. 2. Biasing at the transparent current for the assisted beam, the conversion efficiency can be enhanced by 5.2 dB when the detuning (Ω) between the pump and probe beams is 100 GHz. The enhancement increases slightly as the detuning becomes larger. When the SOA is biased at high-injection level to obtain higher gain and higher conversion efficiency, the enhancement is smaller. This is due to the fact that the assisted beam increases the saturation power but decreases the available gain. Around 3 dB of improvement on the conversion efficiency can still be obtained at 200 mA of bias current. The assisted beam provides a similar amount of improvement for both the up and down conversions, as shown in Fig. 2, though the efficiency for the up conversion is lower.

In addition to the improvement on conversion efficiency, the assisted beam can increase the signal-to-background-noise ratio (SBR) by about 3 and 1.5 dB at the bias levels of 105 and

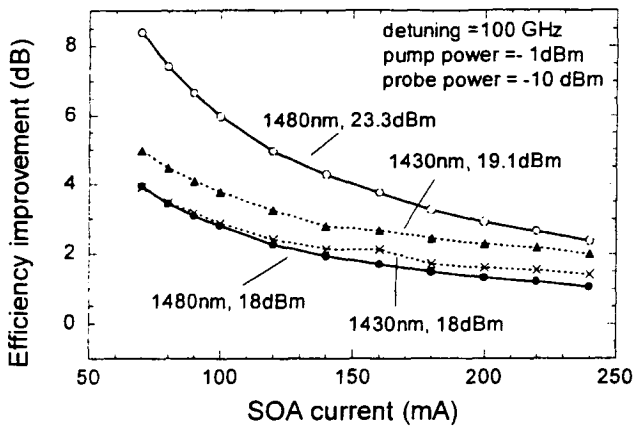


Fig. 3. Comparison of improvement on conversion efficiency between using a 1480- and 1430-nm assisted beam. The results with a 1480- and 1430-nm assisted beams are represented with solid and dashed curves, respectively.

200 mA, respectively. With biasing at the transparent condition, the measured spectra after conversion reveals that the assisted beam raises the conjugated signal and noise level by around 6 and 3 dB, respectively [12]. The pump and probe beams experience a larger gain due to an increase in the saturation power by the assisted beam. This helps to increase the conversion efficiency. The rise of amplified spontaneous emission with an assisted beam results from the induced carriers by the assisted beam.

For wavelength conversion, both the SOA gain and saturation power need to be as high as possible. When a shorter-wavelength assisted beam is used, the SOA can be operated at a higher bias level and still provide transparency to the assisted beam. Therefore, it is possible to obtain both high conversion efficiency and large improvement from using a shorter-wavelength assisted beam. Fig. 3 shows the improvement on conversion efficiency from using a 1430- or 1480-nm assisted beam. The efficiency improvement is determined from the difference in the conversion efficiency between the cases with and without an assisted beam. The improvement for using a 1430-nm beam with 19.1-dBm power is 2.3 dB at 200-mA bias where it is close to the transparent current for the assisted beam. Though the difference is not significant, the 1430-nm assisted beam provides larger improvement than the 1480-nm assisted beam for the same power (18 dBm). The difference is more pronounced at a higher SOA bias, where the 1480-nm beam depletes more carriers than the 1430-nm beam does.

IV. DISCUSSION AND SUMMARY

The improvement on the conversion efficiency and SBR by using an assisted beam is also verified in the theoretical analysis that accounts for the wavelength-dependent transparency, carrier-dependent internal loss, and the spatial distribution of the signals and ASE. The analysis will be addressed elsewhere. The analysis indicates that the intrinsic loss inside the SOA has a strong effect on this approach. For an SOA operating at the 1.55- μm wavelength region, the intrinsic loss arises mainly from the free carrier absorption and intervalence-band absorption. Both absorption mechanisms increase with the carrier density. As the SOA is biased at the transparent current

for the assisted beam, the beam experiences the intrinsic loss as it propagates along the SOA. Thus, the loss at transparency is larger for a shorter-wavelength assisted beam. This loss limits the improvement on the saturation power and the conversion efficiency. Experimental results reveal that the fiber-to-fiber loss at transparency increases by about 10 dB as the wavelength of the assisted beam changes from 1480 nm to 1430 nm. This corresponds to an increase in the intrinsic absorption coefficient by about 23 cm^{-1} . In order to increase the conversion efficiency by using an assisted beam, efforts must be made to reduce the effects of intrinsic loss. One of the alternatives is to inject the assisted beam laterally or transversally instead of longitudinally such that the beam intensity is not reduced by the intrinsic loss.

In summary, we demonstrated that the saturation intensity and wavelength conversion efficiency of an SOA could be increased with an assisted beam. The conversion efficiency can be improved by larger than 5 dB when the SOA is biased at the transparent current for 23 dBm of 1480-nm assisted beam. The assisted beam can also improve the SBR by about 3 dB. Using a shorter-wavelength assisted beam can provide the improvement at higher gain if the effects of carrier-induced loss can be reduced.

REFERENCES

- [1] T. Durhuus, B. Mikkelsen, C. Joergensen, S. L. Danielsen, and K. E. Stubkjaer, "All-optical wavelength conversion by semiconductor optical amplifiers," *J. Lightwave Technol.*, vol. 14, pp. 942–954, June 1996.
- [2] S. J. B. Yoo, "Wavelength conversion technologies for WDM network applications," *J. Lightwave Technol.*, vol. 14, pp. 955–966, June 1996.
- [3] I. Tomkos, I. Zacharopoulos, D. Syvridis, Th. Sphicopoulos, C. Caroubalos, and E. Roditi, "Performance of a reconfigurable wavelength converter based on dual-pump-wave mixing in a semiconductor optical amplifier," *IEEE Photon. Technol. Lett.*, vol. 10, pp. 1404–1406, Oct. 1998.
- [4] A. D'Ottavi, F. Martelli, P. Spano, A. Mecozzi, and S. Scotti, "Very high efficiency four-wave mixing in a single semiconductor traveling-wave amplifier," *Appl. Phys. Lett.*, vol. 67, pp. 2753–2755, 1995.
- [5] A. D'Ottavi, F. Girardin, L. Graziani, F. Martelli, P. Spano, A. Mecozzi, S. Scotti, R. Dall'Ara, J. Eckner, and G. Guekos, "Four-wave mixing in semiconductor optical amplifiers: A practical tool for wavelength conversion," *IEEE J. Select. Topics Quantum Electron.*, vol. 3, pp. 522–528, Apr. 1997.
- [6] M. Yoshino and K. Inoue, "Improvement of saturation output power in a semiconductor laser amplifier through pumping light injection," *IEEE Photon. Technol. Lett.*, vol. 8, pp. 58–59, Jan. 1996.
- [7] M. Tsurusawa, M. Usami, and Y. Matsushima, "New method for reduction of carrier lifetime in semiconductor optical amplifier using assisted light," *Jpn. J. Appl. Phys.*, vol. 38, pp. 1265–1268, 1999.
- [8] G. Soulage, P. Doussiere, A. Jourdan, and M. Sotom, "Clamped gain traveling wave semiconductor optical amplifier as a large dynamic range optical gate," in *Proc. ECOC'94*, vol. 1, Florence, Italy, 1995, pp. 451–454.
- [9] K. Inoue and M. Yoshino, "Gain dynamics of a saturated semiconductor laser amplifier with 1.47- μm LD pumping," *IEEE Photon. Technol. Lett.*, vol. 8, pp. 58–59, Apr. 1996.
- [10] M. A. Dupertuis, J. L. Pleumeekers, T. P. Hessler, P. E. Selbmann, B. Deveaud, B. Dagens, and J. Y. Emery, "Extremely fast high-gain and low-current SOA by optical speed-up at transparency," *IEEE Photon. Technol. Lett.*, vol. 12, pp. 1453–1455, Nov. 2000.
- [11] J. L. Pleumeekers, M. Kauer, K. Dreyer, C. Burrus, A. G. Dentai, S. Shunk, J. Leuthold, and C. H. Joyner, "Acceleration of gain recovery in semiconductor optical amplifiers by optical injection near transparency wavelength," *IEEE Photon. Technol. Lett.*, vol. 14, pp. 12–14, Jan. 2002.
- [12] S.-L. Lee, P.-M. Gong, and C.-T. Yang, "Efficiency improvement of SOA-based four-wave-mixing wavelength converters by using an assisted beam," in *Proc. OFC 2002*, Anaheim, CA, Mar. 2002, Paper ThGG75.

High-Resolution and Power Insensitive Wavelength Monitoring Using a Low Cost FP Laser for DWDM Applications

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內容成果：

半導體雷射二極體的操作工作區大致可分為三大項，分別為吸收區、透明點與增益區。操作電流大小在透明點以下為吸收區，當外部光源注入時，光子會被吸收，既會有損耗；反之操作電流在透明點以上，此時雷射二極體帶有增益，可放大光功率或是發光；而當操作電流在透明點，雷射二極體不吸收也不放大，光直接穿透過雷射二極體。經由實驗與理論發現，半導體雷射的透明點電流會隨注入的光源的波長的不同而有相對應的變化，利用此特性，可即時精確監控通道中光訊號波長的穩定度以及偏移量。本文中使用一半導體光放大器作為光感測器，搭配簡單的控制電路，達到波長監控的功能。

TuA2-4

High-Resolution and Power Insensitive Wavelength Monitoring Using a Low Cost FP Laser for DWDM Applications

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Abstract:

High-resolution, compact, and power-insensitive wavelength monitoring using a low-cost FP laser can achieve better than 0.01nm wavelength resolution. The resolution is significantly improved with the help of the FP resonance and ratio detection.

Most existing wavelength monitoring techniques use interferometric apparatus [1,2]. We proposed to use the detection of the transparent condition of a semiconductor optical amplifier (SOA) or laser diode for wavelength sensing or monitoring [3]. This approach has the advantages of compactness and integrability with other optoelectronic devices. However, the sweeping bias current for locating the transparent condition must have a resolution of 100 ppm in order for obtaining a wavelength resolution of 0.01 nm. In this paper, we demonstrate simple schemes of high-resolution wavelength monitoring using a FP laser diode as the sensor. Significant improvement on the monitoring resolution is achieved by use of the cavity effects and ratio detection, so wavelength resolution of 0.01 nm can easily be obtained.

For using as a wavelength monitor, the diode laser is usually biased around the transparent condition for the monitored wavelength. When the induced diode voltage in response to a modulated input light is demodulated with envelope detection, the output can be written as $v = \eta |I - I_{tr}(\lambda)| P$, where I and I_{tr} are the bias current and transparent current, respectively. P is the input power and η is a constant depending on the diode parameters. The output voltage appears as a V-shape around the transparency in varying the bias current, as shown in Fig. 1. As the wavelength changes, the V-curve will shift to the right or left side, depending on the sign of wavelength change. The wavelength change can be monitored from ratio of the junction voltages for a two-level switching bias. The output after the ratio detector vanishes for the monitored wavelength. When the wavelength shifts toward the long-wavelength side, the V-curve moves to the left, and the output becomes positive. On the other hand,

a negative signal at the ratio detector output indicates a blue shift in the wavelength. Therefore, the output voltage can clearly indicate the sign of wavelength drift and can serve as a control signal for managing the incident wavelength. Since the voltage ratio is independent of the input power level, it can be immune to power fluctuation without the necessity of power monitoring. For applications in wavelength stabilization, ratio detection can generate power-independent control signal for adjusting the source wavelength.

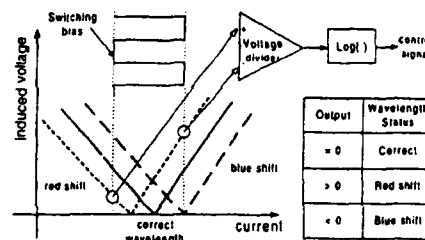


Fig. 1 Concept of wavelength monitoring using ratio detection of the induced voltage across a FP laser.

The change in voltage ratio (in dB) for $\Delta\lambda$ of wavelength drift can be written as

$$\Delta\rho = 4.34 \left(1 + \eta_+ / \eta_- \right) \left(\Delta\lambda / \Delta I_- \right) \partial I_{tr} / \partial \lambda \quad (1)$$

where ΔI_- is the difference between the lower bias level and the transparent point. The subscript associated with the η -parameter is used to designate the difference in the material gain for the two bias levels. Eq. (1) indicates that the ratio detection can enhance wavelength resolution via the $1 + \eta_+ / \eta_-$ factor.

It was found in experiments that the resonant cavity of a FP laser causes oscillating characteristic in the wavelength variation of transparent current. This is due to the variation of transparent current with the input power. Such transparent property can improve the wavelength resolution of wavelength monitoring at fixed channels for its enhancement on $\partial I_{tr} / \partial \lambda$.

Fig. 2 shows the transparent characteristic of the

FP laser used as the sensor. The laser was temperature controlled at 25 °C. The transparent current shows resonant characteristic for a fine scale of λ change. The period of the resonance was found to be identical to the cavity mode spacing of the diode laser. Such resonant effect can give rise to improvement on the resolution of wavelength monitoring by about an order, due to the enhancement of differential slope from 0.21 mA/nm to 2.0 mA/nm. Since changing the diode temperature can shift the I_{tr} - λ curve, the monitored wavelength can always be aligned to the middle of a slope. If the cavity mode spacing is designed to be equal to the channel spacing for a DWDM system, one diode laser can be used to monitor multiple channels by simply changing the bias current.

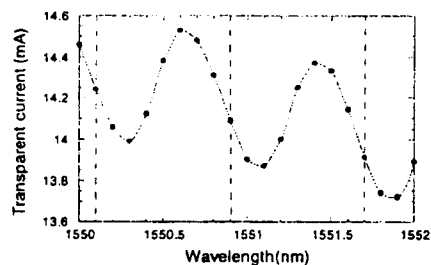


Fig. 2 Variation of transparent current with wavelength.

Fig. 3 depicts the measured voltage ratio for the two bias levels in response to different input power. The slope of the curves, 0.9 dB/0.01 nm, is consistent with the value calculated from Eq. (1). This slope is almost independent of the power level. The data shown in the inset of Fig. 3(b) indicate that the change in the voltage ratio for 15 dB of power change is less than that for 0.01-nm wavelength drift. Therefore, wavelength monitoring with ratio detection can achieve better than 0.01 nm resolution with high tolerance on power variation. The results shown in Fig. 4 indicate that the ratio fluctuation during half an hour of measurements is much less than the ratio change for 0.01 nm of wavelength drift. The monitoring technique can have really good long-term stability.

In summary, we have demonstrated simple and high-resolution wavelength monitoring schemes using a low-cost diode laser as the sensor. The FP resonance inside the cavity of a diode laser was shown to significantly enhance the slope of the I_{tr} - λ curve. Ratio detection can further improve the monitoring performance. The wavelength

monitoring technique is tunable such that single diode laser can monitor many wavelength channels. The demonstrated 0.01-nm wavelength resolution with much relaxed requirement on the detection circuitry indicates that such schemes can be very attractive for application in DWDM systems.

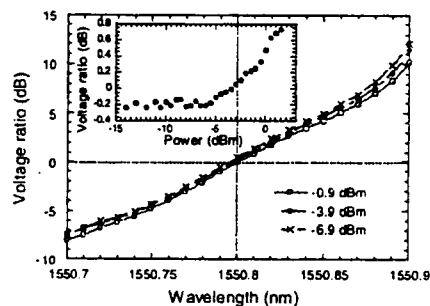


Fig. 3 Output variation versus wavelength drift for ratio detection. The two bias levels are 13.99 and 14.48 mA, respectively; and the transparent current for the monitored wavelength is 14.31 mA. The inset shows the variation of voltage ratio with input power.

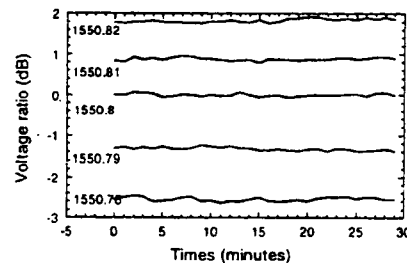


Fig. 4 Wavelength monitoring stability for wavelengths of 0.01-nm spacing.

Acknowledgment: This work was supported in part by the National Science Council under Contract NSC89-2215-E011-014, and by the Ministry of Education, Taiwan, R.O.C., under Contract 89-E-FA06-2-4.

References

- [1] J. P. Monochalin, *et al.*, *Applied Optics*, vol.20, no. 5, pp. 736-757, 1981
- [2] M. Teshima, *et al.*, *JLT-14*, no. 10, pp. 2277-2285, 1996.
- [3] S.-L. Lee, *et al.*, *OFC 2000*, paper WK1, March, 2000, Baltimore, USA.

Wavelength conversion using a semiconductor optical amplifier with an assisted beam

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發表論文：Optics and Photonics Taiwan' 01, paper TC1-8, 2001.

內容成果：

波長轉換器是實現全光網路中不可或缺的元件，透過波長轉換達到光封包訊號的路由功能，將光纖中的有限波長通道有效的利用，提高交換效率與網路傳輸效率，而全光式波長轉換器對於日漸提升的通道傳輸速率提供暢行無阻的通道切換。目前已有多種波長轉換的技術被提出，本實驗室以光半導體放大器中的非線性效應-四波混合作為基礎。四波混合的技術可以同時多波長轉換，轉換效率不受原始光訊號的調變方式影像，有較高的光熄滅比，並具有色散補償功能等優點；相對的，轉換效率低為其主要缺點，經過研究並改善，提出多項實驗數據與操作方式，製作具有高轉換效率的波長轉換器。

Wavelength conversion using a semiconductor optical amplifier with an assisted beam

Chin-Tien Yang, Pei-Miin Gong, San-Liang Lee

Abstract—We demonstrated experimentally the improvement on the wavelength conversion efficiency by using an assisted beam to enhance the saturation power of an SOA. The effects of intrinsic loss and the countermeasures for further improvement are also discussed.

Index Terms—Assisted beam, SOA, saturation power, conversion efficiency, FWM, wavelength conversion.

I. INTRODUCTION

All-optical wavelength converters will play important roles in the future optical networks. It can solve the wavelength blocking problems and provide wavelength routing and switching to enable optical internets. There exist several frequency conversion techniques using semiconductor optical amplifiers (SOAs), including cross gain modulation (XGM), cross phase modulation (XPM), and four-wave mixing (FWM). The wavelength conversion using the FWM effect is transparent to data rate and format and can provide chirp compensation, but it suffers from poor conversion efficiency. All-optical conversion can be applied in practical systems only if the conversion efficiency and noise figure can be significantly improved.

It is well known that, as the saturation power of the SOA increases, the FWM conversion efficiency increases [1]. Ref. [1] demonstrated experimentally that the saturation power can be enhanced by using an assisted beam, which also enhances the carrier recovery rate [2]. The gain recovery of a SOA is limited by the long carrier-recovery time, but it can be improved by increasing the applied current or the device length. However both approaches will decrease the optical bandwidth, increase amplified spontaneous emission (ASE), and increase the electrical consumption. When the wavelength of the assisted beam is chosen such that the bias current of the SOA corresponds to the transparent current for the assisted beam, the carrier recovery rate and saturation power can be improved without sacrificing the gain. Thus, using an assisted beam at transparency may improve the wavelength conversion efficiency.

II. PRINCIPLE

The effects of using an assisted beam can be observed from the rate equation [3]:

$$\frac{dN}{dt} = \frac{I}{qV} - \frac{N}{\tau} - \frac{\Gamma g_m P_a}{AE_a} - \frac{\Gamma g_m P_s}{AE_s} \quad (1)$$

where the subscripts "a" and "s" denote the assisted

beam and the signal, respectively; E is the photon energy; N is the carrier density; I is the injection current; τ is the carrier lifetime; q is the electric charge; V is active volume; Γ is the confinement factor; g_m is the material gain. P is the incident power; and A is cross-sectional area of the active region. From (1), we can get [3]

$$\frac{1}{\tau_{eff}} = \frac{1}{\tau} + \frac{P_a}{\tau P_a^{sat}} \quad (2)$$

$$P_s^{sat} = \frac{AE_s}{\Gamma a_s \tau_{eff}} \quad (3)$$

where P_a^{sat} is the saturation power for the assisted beam and τ_{eff} is the effective carrier lifetime. Eq.(2) indicates that the effective carrier lifetime decrease as the assist beam power increases. Thus, the signal saturation power in (3) can be enhanced with the reduced effective carrier lifetime.

Fig. 1 depicts the operation principle using an assisted beam in a SOA. If the SOA is operated at the current small than the transparency for the assisted beam, the assisted beam will be absorbed and offer the excess carriers to increase the signal gain. On the other hand, if the SOA is operated above the transparency for the assisted beam, the assisted beam will deplete the carriers and decrease the signal gain. Only when the SOA is operated at the transparency current for the assisted beam, the assisted beam can increase the saturation power while keep the gain unchanged.

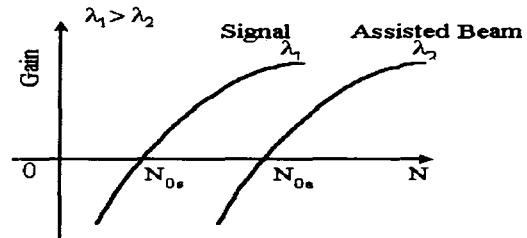


Figure 1 Gain curves for the signal and assisted beam

The maximum conversion efficiency using the FWM mechanism can be given by [5]

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$$\eta_{max} \propto G_0 P_{sl}^2 \quad (4)$$

where G_0 is the unsaturation gain and P_{sl} is internal saturation power of the SOA. Thus, the conversion efficiency can be increased by using an assisted beam to raise the saturation power..

III. EXPERIMENT RESULTS

Fig. 2 depicts the measured transparent currents against the light wavelength for a commercial SOA. The transparent current is higher for shorter wavelength. Thus, it may be beneficial to use an assisted beam of shorter wavelength such that the SOA can be biased at higher current to provide higher signal gain.

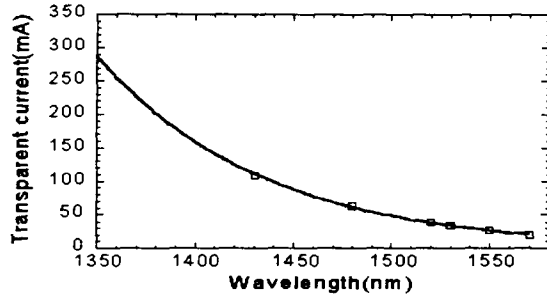
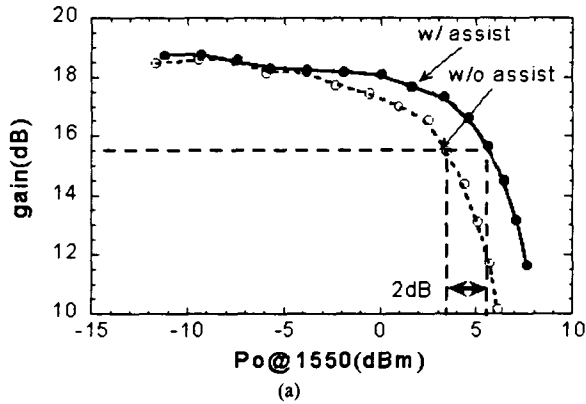
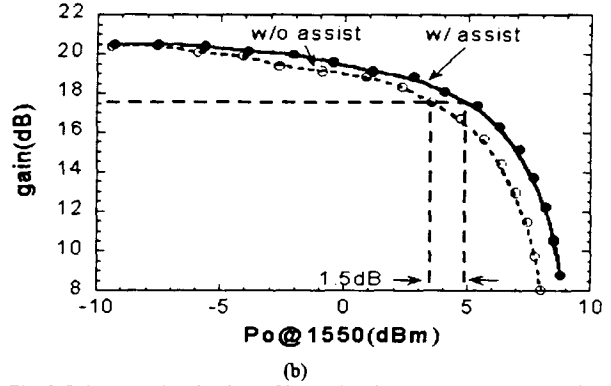


Fig. 2 Measured transparent currents against the light wavelength

In our experiments, we use assisted beams of 1480nm and 1430nm. Fig. 3 shows the gain saturation characteristics of the SOA when the assisted beam is used. The assisted power is 16dBm for both wavelengths; and the 1550-nm signal power is -25dBm. The transparent currents are 114mA and 125mA for the 1480-nm and 1430-nm assisted beam, respectively. The saturation power is enhanced by 2.5dB when the 1480-nm assisted beam is used, while 1.5dB enhancement is observed for the 1430-nm assist beam.



(a)



(b)

Fig. 3 Gain saturation for the 1550-nm signal when the wavelength of the assisted beam is (a) 1480-nm (b) 1430-nm

To prove the improvement on wavelength conversion efficiency with an assisted beam, the measurement setup shown in Fig. 4 is used. Since the SOA is long, the assisted beam is incident contradirectional with the signal and pump signals to have best effects.

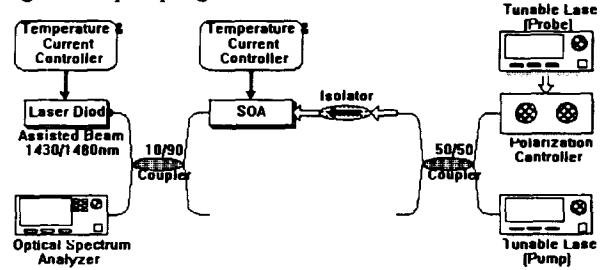


Fig. 4 Measurement setup

The measurement is performed at 25°C with 4dBm of 1550-nm pump power and 0dBm of probe power. The comparison of the conversion efficiency with and without the assisted beam are shown in Fig. 5, and summarized in Table I and II. Biasing at its corresponding transparent current for the assisted beam, the conversion efficiency can be enhanced by 4 and 3 dB when the assisted wavelength is 1430-nm and 1480-nm, respectively. The improvement reduces to 1.5dB for both assisted wavelengths when the SOA is biased at 200mA, well above the transparent current, to have larger signal gain and conversion efficiency. If the efficiency is required to be above -20dB, the wavelength conversion range ($2\Delta\lambda$) is improved from 14nm to 18nm by using the assisted beam. The signal-to-background noise ratio (SBR) is also enhanced slightly due to the improvement on the conversion efficiency.

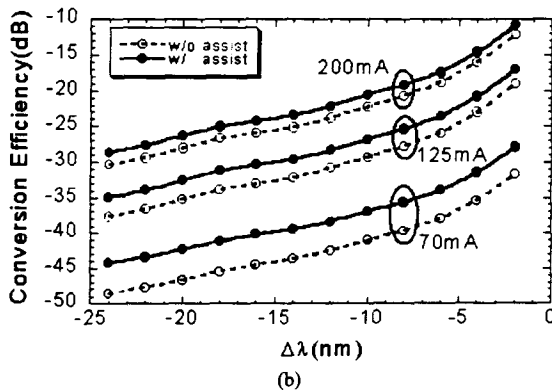
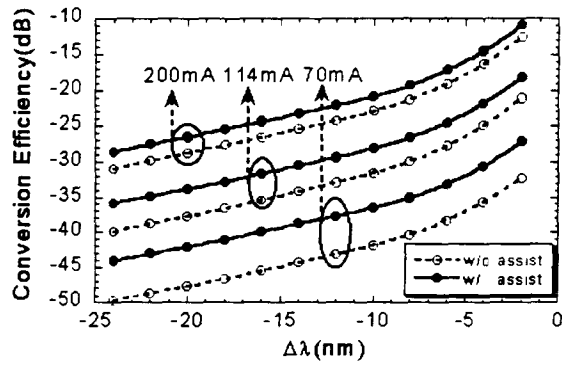


Fig. 5 Conversion efficiency vs. $\Delta\lambda$ when the wavelength of the assisted beam is (a) 1480-nm (b) 1430-nm

Table I Conversion efficiency and SBR for using 1480-nm assisted beam

Biased current (mA)	Conversion efficiency (dB)		SBR (dB)	
	No assisted beam	With assisted beam	No assisted beam	With assisted beam
70	-32	-27	12	15
114(transparency)	-21	-18	20.5	22
200	-12.5	-11	26.5	27.5

Table II Conversion efficiency and SBR for using 1430-nm assisted beam

Biased current (mA)	Conversion efficiency (dB)		SBR (dB)	
	No assisted beam	With assisted beam	No assisted beam	With assisted beam
70	-31.5	-28	12	13.5
125(transparency)	-21	-17	21.5	22.5
200	-12	-10.5	26.5	27

IV. DISCUSSION AND SUMMARY

The improvement on the conversion efficiency was expected to be greater than the experimental results as the assisted beam is used. Especially, using a 1430-nm assisted beam was expected to offer better performance than using the 1480-nm one. It was found that the intrinsic loss inside the SOA has strong effect on this

approach. Fig. 6 shows the measured intrinsic loss (loss at transparent condition) for various assisted wavelengths. The loss increases to a significant value as the wavelength moves to a shorter wavelength. This reduces the improvement by the assisted beam. This can be clearly seen in the simulation, as shown in Fig. 7.

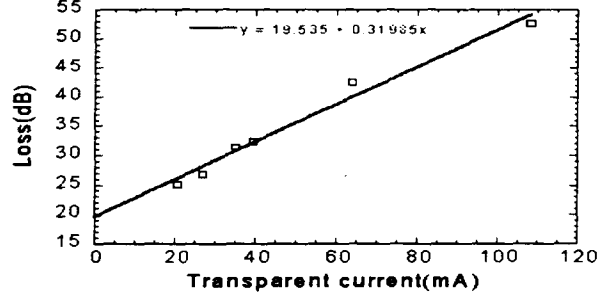


Fig. 6 Insertion loss at the transparent condition for different wavelengths

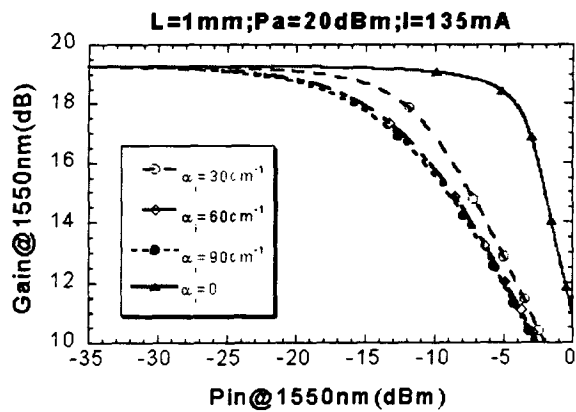


Fig. 7 Gain curves for different loss

In order to take advantages of using assisted beams, two countermeasures can be performed to reduced the effects of intrinsic loss. One is to reduce the intrinsic loss by employing strain into the active material to suppress the intervalence band absorption. The other is to use a shorter and high-gain SOA for the wavelength conversion. Our simulation has proved that the assisted beam can provide greater improvement when the SOA length is reduced to 500um from 1000um.

REFERENCE

- [1] A. D'Ottavi, F. Girardin, L. Graziani, F. Martelli, P. Spano, A. Mecozzi, S. Scotti, R. Dall'Ara, J. Eckner, G. Guekos, *IEEE J. Select. Topics Quantum Electron.*, vol. 3, pp.522-528, 1997.
- [2] M. Yoshino, and K. Inoue, *IEEE Photonics Technology Letters*, vol.8, no.1, pp.58-59, January 1996.
- [3] R. J. Manning, and D.A.O. Davies, *Optical Letters*, vol.19, no.12, June 1994.
- [4] M. Usami, et al., *App. Phys. Lett.* vol. 72, num. 21, May 1998.
- [5] M. A. Dupertuis, et al, J. Y. Emery, *IEEE Photonics Technology Letters*, vol.12, no.11, pp1453-1455, November 2000.

論文九

High-Resolution Wavelength Monitoring Technology for Tunable Laser Sources

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發表論文：Optics and Photonics Taiwan' 01, paper FC3-5, 2001.

內容成果：

波長可調式元件在光通訊網路架構中所佔有的地位越來越重，然而要確保可調式元件其對波長的穩定度及精確度，對於用在 DWDM 系統中，通道間距相當小，對於光訊號波長要求的更嚴苛，因此需要一精密且即時的波長監控系統，監控波長的變化，適時送出回受控制訊號或或警示訊號，以提供良好的通訊品質。針對波長監控系統，我們提出了數種方案，具有低成本、高精確度、速度快、標準 ITU 波長通道鑑別等功能。

High-Resolution Wavelength Monitoring Technology for Tunable Laser Sources

Chun-Liang Yang and San-Liang Lee

Abstract — High-resolution and tunable monitoring technology can be realized using a FP etalon to detect the wavelength drift and a FP diode laser for channel discrimination. It can achieve 0.01nm wavelength resolution and require a temperature stability of about $\pm 0.2^\circ\text{C}$.

Index Terms — FP Etalon, DWDM, transparent current, wavelength monitoring, FSR.

I. Introduction

DWDM networks require effective schemes for wavelength monitoring in order to control the source wavelength or manage the wavelength channels. Many research institutes and companies are developing tunable lasers for applications as spare sources or fast wavelength switching components for DWDM networks. Wavelength control for tunable lasers is essential and it requires fast tuning speed. Interferometers made of Fabry-Perot (FP) etalons are the popular choices for wavelength control of DWDM lasers [1]-[3]. However, a FP etalon filter can not distinguish among different channels since it has periodic wavelength characteristic [3]. The detection of the transparent condition of a semiconductor optical amplifier (SOA) or laser diode was proposed for wavelength measurement or monitoring [4]-[5]. This approach has the advantages of compactness and possibility for integration with the lasers. However, the wavelength resolution is limited by the required stability (about 0.01°C) in temperature control.

In this paper, we propose a high-resolution tunable wavelength monitoring which combines advantages of stable temperature properties of a FP etalon filter and the fast tunability of the transparent properties of a FP diode laser. Significant relief to the required temperature stability (about $\pm 0.2^\circ\text{C}$) is achieved by utilizing the FP etalon filter, so wavelength resolution of 0.01nm can easily be obtained. The ambiguity among different channels for the etalon can be overcome with the transparent detection. Our solution can be applied for high-resolution wavelength discrimination and tunable channel recognition.

II. Operation Principle

Fig. 1 shows the measured transmission and reflection characteristics of the FP etalon filter used in the experiments. The characteristics are periodic in wavelength with a free-spectral range (FSR) of 0.4nm. This value of FSR is ideal for DWDM channels of 0.4 or 0.8-nm spacing. The channel wavelength to be monitored is usually located at the crossover of the reflection and transmission spectra, as shown in Fig. 2. The crossover can be aligned to any DWDM channel by adjusting the substrate temperature. On the other hand, Fig. 2 indicates that the spectra is shifted by about 0.011nm per degree of temperature change, so the temperature control stability of about

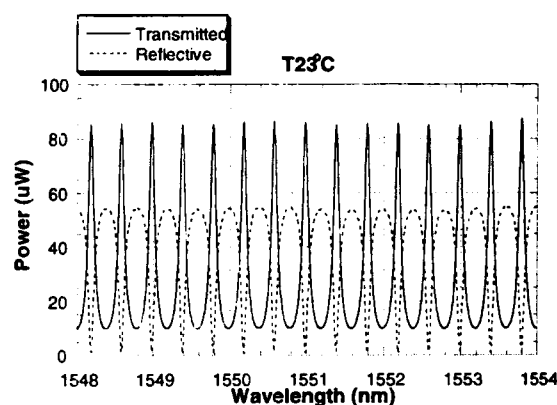


Fig. 1. FP etalon filter with the periodic wavelength characteristic of 0.4nm spacing

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0.1°C would produce only 0.001nm of wavelength change. Therefore, the etalon can serve as the reference standard of a fixed channel. As the wavelength drifts, the transmitted and reflected power would be different. The ratio of transmitted and reflected power can indicate the amount of wavelength drift. Fig. 3 shows that eight fixed

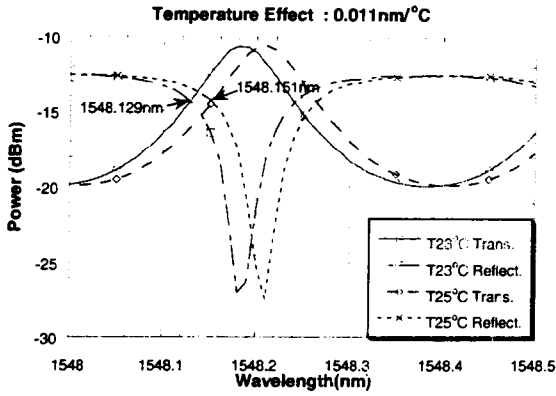


Fig. 2. The fixed channel wavelength is located on the crossover and the temperature effect is 0.011nm/°C.

channels have the same characteristic so these channels can be monitored on the same way. If the FSR of the etalon is identical to the DWDM channel spacing, one etalon can be used to monitor many channels. In order to recognize the channels, every channel must be aligned to the center of the slope at the I_r - λ curve of the FP diode laser, as shown in Fig. 4 The action is achievable by controlling the diode temperature.

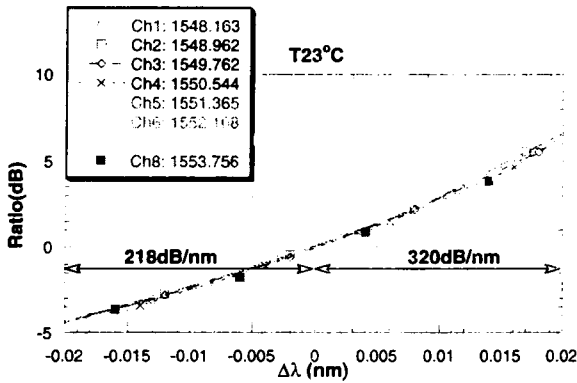


Fig.3. The ratio of transmitted and reflected power versus a channel wavelength drift

Fig. 5 shows the schematic of the complete wavelength monitoring system, and its operation flowchart is shown in Fig. 6. For using as a wavelength stabilizer for tunable lasers, the output

wavelength is first monitored by FP etalon to a wavelength accuracy with $\pm 0.02\text{nm}$. Due to the periodicity, the etalon can not tell which channel the laser is operating. Thus, the channel wavelength has to be determined by the transparent detection circuitry. Owing to the stabilization with the etalon, the wavelength will fit in the transparent current window for a specific channel. The channel can be

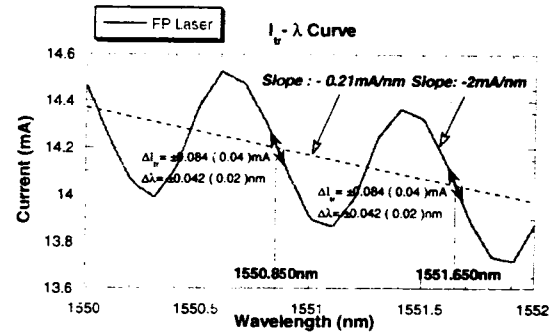


Fig. 4. The transparent current window has the margin of about $\pm 0.084\text{mA}$

recognized with a reference to the wavelength drift data supplied by the etalon. The transparent current window has a margin of about $\pm 0.084\text{mA}$, as shown in Fig.4, which is equal to the product of the coarse slope of the I_r - λ curve (-0.21mA/nm) and the 0.8nm channel spacing. The corresponding temperature margin is about $\pm 0.218^\circ\text{C}$ for the window since the transparent current of the FP diode laser varies with

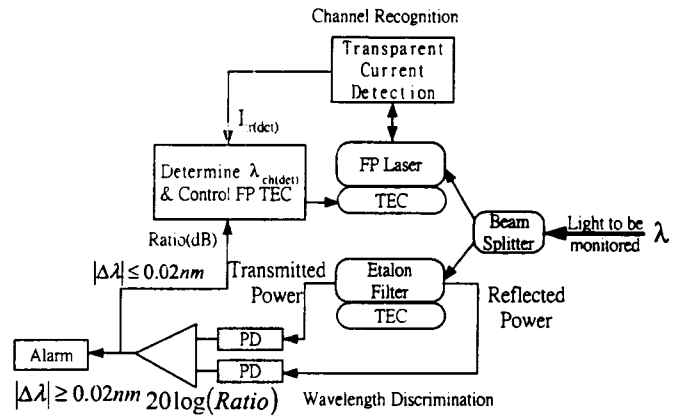


Fig.5. The schematic of wavelength monitoring system

the temperature at a rate of about $0.385\text{mA/}^\circ\text{C}$. Therefore, the wavelength for a tunable laser source can be recognized as long as its wavelength does not deviate from the correct channel by more than

$\pm 0.02\text{nm}$. This number is limited by the 0.4-nm spaced FP etalon filter of which the slope is linear within $\pm 0.02\text{nm}$. From Fig. 2, the transmitted or reflected spectral characteristic is monotonic over a wavelength rang larger than 0.1nm . Therefore, the locking range of our approach is larger than 0.1nm .

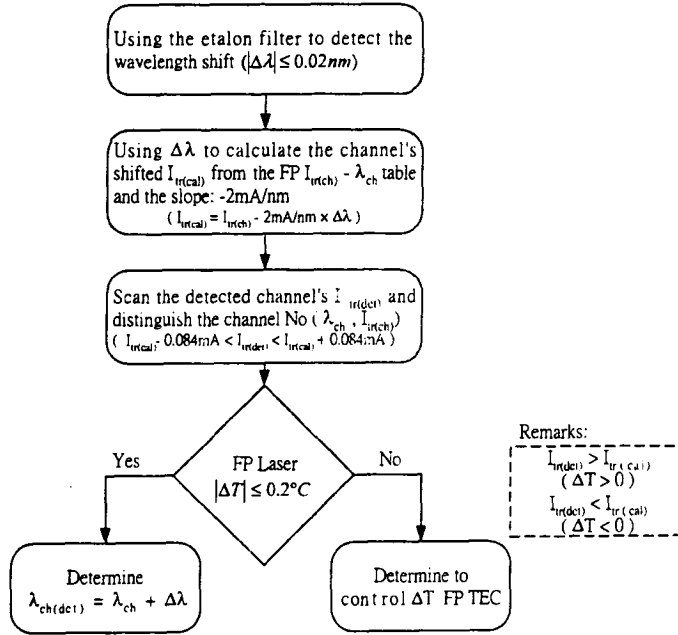


Fig.6. The operation principle flowchart

III. Conclusions

We have proposed a high-resolution and tunable wavelength monitoring schemes using a combination of a FP etalon filter and a diode laser as the sensors. The system uses the FP etalon filter as a accurate wavelength reference since it can have a large tolerance of $\pm 0.2^\circ\text{C}$ on the temperature control unit. The exact channel wavelength is then recognized by the fast tunable transparent characteristics of a FP diode laser. The circuitry and control flowchart are proposed. 0.01nm of wavelength resolution can easily be achieved with much relaxed requirement on the stability of temperature control.

References

- [1] B. Villeneuve, H. B. Kim, M. Cry, and D. Garipey, "A compact wavelength stabilization scheme for telecommunication transmitters," in WDM Comp. Technol., IEEE/LEOS Summer Top. Meetings, 1997, pp. 19-20.
- [2] S. Matsumoto and M. Fujise, "Frequency stabilization of $\lambda/4$ -shifted DFB laser to stabilized optical resonator," Electron. Lett., vol. 25, pp. 814-815, June 1989.
- [3] D. A. Ackerman, K. F. Dreyer, U. Koren, J. W. Stayt, S. L. Broutin, et. al, "Wavelength, modal, and power stabilization of tunable electro-absorption modulated distributed Bragg reflector lasers," in Int. Semiconduct. Laser Conf., Monterey, CA, 2000, pp. 49-50.
- [4] S.-I. Lee, "Wavelength measurement and tracking using semiconductor laser amplifiers for applications in photonic networks," IEEE Photon. Technol. Lett., vol. 10, pp. 439-441, Mar. 1998.
- [5] S.-L. Lee, Y.-Y. Hsu, and C.-T. Pien, "High-resolution wavelength monitoring using differential/ratio detection of junction voltage across a diode laser," IEEE Photon. Technol. Lett., vol. 13, pp. 872-874, August 2001.

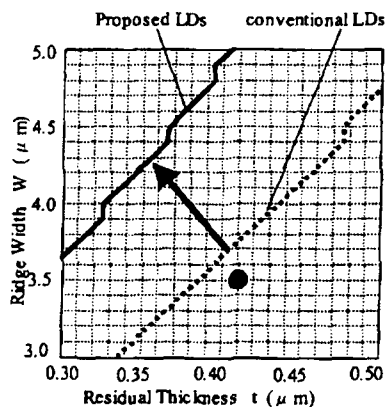
Efficiency improvement of SOA-based four-wave-mixing wavelength converters by using an assisted beam

作者：李三良(S.-L. Lee), 龔佩敏(P.-M. Gong), 楊金添(C.-T. Yang)

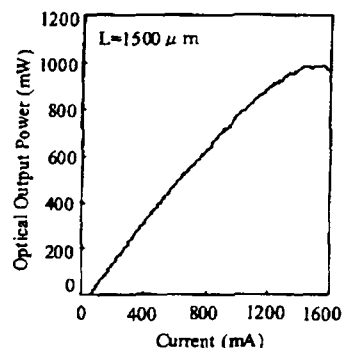
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內容成果：

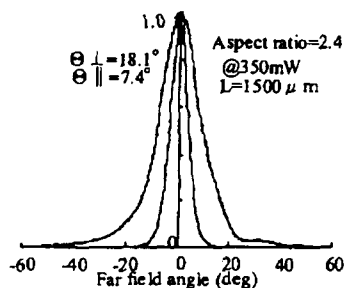
波長轉換器為實現高速光通訊網路中一個相當重要的元件。本文使用半導體光放大器作為波長轉換的元件，利用半導體光放大器的非線性效應：四波混合技術，注入一幫浦光源與訊號光源，產生第三道對應的轉換波長。然而目前使用此方式有波長轉換效率不高的缺點，因此提出再多加一道輔助光，提高轉換效率。經由實驗，當加入輔助光之後，波長轉換器的轉換效率可提升 5dB，且訊號功率與背景功率比也提升了 3dB，對於轉換訊號的功率級訊號品質，增加了不少幫助，提高使用四波混合製作全光波長轉換器的可行性。



(a) Cutoff conditions of first order mode

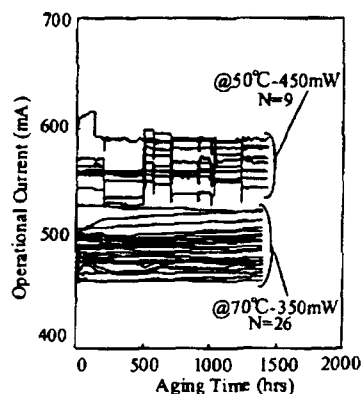


(b) Power versus current (P-I) characteristics



(c) Far field patterns (FFPs)

ThGG74 Fig. 2. Characteristics of proposed LDs.



ThGG74 Fig. 3. Aging test results.

asymmetrical structure does not affect the aging performances in particular.

5. Conclusion

The new ridge waveguide LD consisted of layers to support asymmetrically expanded optical field normal to the active layer has been proposed in order to satisfy the increase of kink free output power and the reduction of aspect ratio. More than 650 mW kink free output power and less than 2.5 aspect ratio have been achieved by the proposed LD. Good aging performances are also expressed in the proposed LD. It is believed that the LDs are suitable for realizing 980 nm modules with high coupling efficiency and high power EDFAs.

References

1. N. Taketa et al., Electron. Lett., vol. 34, pp. 381-382, 1998.
2. S. Yamamura et al., Proc. Opt. Fiber Commun. Conf. (OFC 2000), Baltimore, Maryland, paper ThK3-1, pp. 162-164, 2000.

ThGG75

5:30 pm

Efficiency enhancement of SOA-based four-wave-mixing wavelength converters by using an assisted beam

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1. Introduction

The wavelength conversion using the four-wave mixing (FWM) effect in a semiconductor optical amplifier (SOA) can provide the needed functions for optical networks,¹ but it suffers from poor conversion efficiency. Significant improvement on the conversion efficiency and noise figure is required for practical applications. The FWM conversion efficiency increases approximately with $G_0 P_p^2$, where G_0 is the unsaturated

gain and P_p is the saturation intensity of the SOA.¹ Therefore, increasing the gain and/or saturation power of a SOA is the key to improve conversion efficiency. The saturation intensity is inversely proportional to the carrier lifetime. The carrier recovery rate can be enhanced by injecting a holding light of which the wavelength is in the gain region of the SOA.² This method raises the saturation intensity but decreases the amplifier gain. Dupertuis et al. proposed to use an assisted beam at transparency to speed up a SOA.³ This technique can increase the saturation intensity without sacrificing the gain. In this paper, we will demonstrate that using an assisted beam can improve the efficiency and signal-to-noise ratio of FWM wavelength conversion.

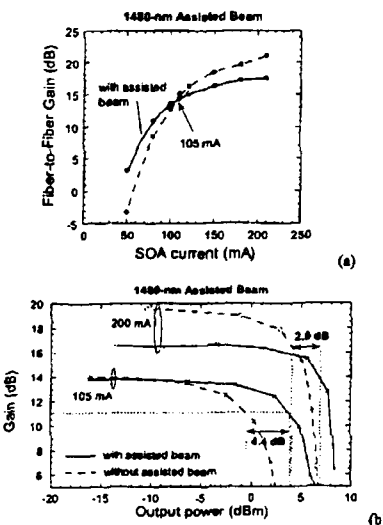
2. Gain characteristics

The transparent condition of a SOA corresponds to the condition that the material gain of the SOA is zero for the incident wavelength. Fig. 1(a) compares the gain curves at 1550 nm for the 1.55-μm SOA used in our experiments with and without a 1480-nm assisted beam. The two curves intersect at the current that corresponds to the transparent condition for the assisted beam since the gain is not changed for the presence of the assisted beam. An alternative for detecting the transparent condition is to locate the bias current where the junction voltage across the SOA junction vanishes. However, the background power of the strong assisted beam will induce background junction voltage and then affect the accuracy in determining the transparency. Therefore, we use Fig. 1(a) to determine the transparent current, which is 105 mA for the 1480-nm assisted beam. Fig. 1(b) shows that the assisted beam does not affect the unsaturated gain at this bias and raises the saturation power by 4.4 dB. Also shown in this figure are the gain saturation characteristics when the SOA is biased at 200 mA. At such a high-injection bias, the amplified assisted beam depletes the carriers in the SOA. Thus, the measured data show that the assisted beam causes about 2.8 dB reduction in the gain but 2.9 dB increase in the output saturation power.

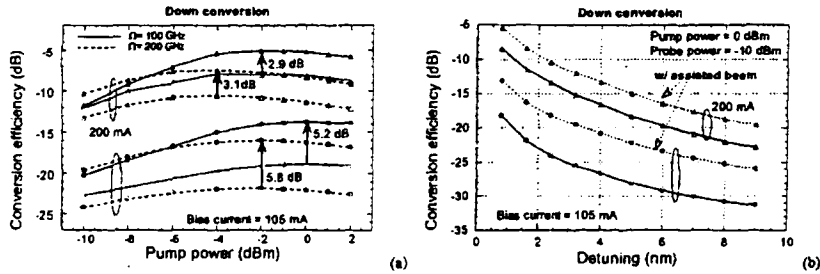
3. Wavelength conversion

To prove the improvement on wavelength conversion efficiency with an assisted beam, the measurement was carried out using a 1480 or 1430-nm assisted beam. From both simulated and empirical results, the assisted beam can have best effects when it is incident contra-directionally with the signal and pump signals. The pump and probe beams out of two commercial tunable lasers with built-in isolators were combined using a 50/50 coupler and inputted to the SOA. A 1.48/1.55-μm multiplexer was used at the output end to combine the input assisted beam and output conjugated signal. The multiplexer has a broad passband that provides low insertion loss to both the 1480- and 1430-nm assisted beams. The conjugated output was measured with an optical spectrum analyzer.

The measurement was first performed with a 1480-nm assisted beam of 23.3-dBm power. The probe power was fixed at -10 dBm and the pump wavelength was fixed at 1553 nm. Wavelength conversion was performed at two different bias currents, one at transparency (105 mA) and the other at high injection (200 mA). The comparison of the conversion efficiency with and without the assisted beam is shown in Fig. 2. Biasing at the



ThGG75 Fig. 1. Effects of assisted beam on the 1550-nm signal gain. The assisted power is 23 dBm.

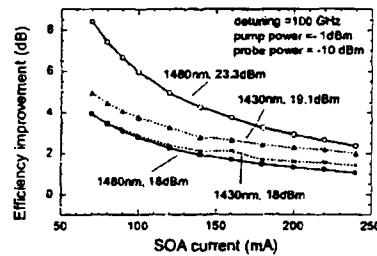


ThGG75 Fig. 2. Effects of 1480-nm assisted beam on wavelength conversion efficiency. The assisted power is 23 dBm.

transparent current for the assisted beam, the conversion efficiency can be enhanced by 5.2 dB when the detuning (Ω) between the pump and probe beams is 100 GHz. The enhancement increases slightly as the detuning becomes larger. When the SOA is biased at high injection level to obtain higher gain and higher conversion efficiency, the enhancement is smaller. This is due to the fact that the assisted beam increases the saturation power but decreases the available gain. Around 3 dB of improvement on the conversion efficiency can still be obtained at 200 mA of bias current.

In addition to the improvement on conversion efficiency, the assisted beam can increase the signal-to-background-noise ratio (SBR) by about 3 dB and 1.5 dB at the bias levels of 105 and 200 mA, respectively. The spectra shown in Fig. 3 indicate the improvement on SBR when the bias current is 105 mA. It also indicates that the strong pump and probe beams experience larger gain due to the increase in the saturation power by the assisted beam. This helps to increase the conversion efficiency. The rise of amplified spontaneous emission with an assisted beam might attribute to the broad spectrum of the assisted beam. Since both the efficiency and SBR are improved, the wavelength conversion range (2Ω) can be increased by using an assisted beam.

For an assisted beam of shorter wavelength, its corresponding transparent current is higher. Therefore, it is possible to obtain both high conversion efficiency and large improvement from using a shorter-wavelength assisted beam. Fig. 4 compares the improvement from 1430-nm and 1480-nm assisted beams. The improvement for using a 1430-nm beam with 19.1-dBm power is 2.3 dB at 200-mA bias, which is close to the transparent current for the assisted beam. Though the difference is not significant, the 1430-nm assisted beam provides larger improvement than the 1480-nm assisted beam for the same power (18



ThGG75 Fig. 4. Comparison of improvement on conversion efficiency.

dBm). The difference is more pronounced at higher SOA bias, where the 1480-nm beam depletes more carriers than the 1430-nm beam does. It is expected that the 1430-nm assisted beam can achieve the same improvement as the 1480-nm beam with less power.

The improvement on the performance of wavelength conversion by using an assisted beam can be clearly seen from the eye-diagram, as shown in Fig. 5, when the probe beam was modulated at 2.5 Gbps data rate. The 1480-nm assisted beam increases the signal amplitude by >5 dB and enhances the eye opening. After wavelength conversion, the conjugated signal was selected with an optical filter, amplified by an EDFA, and received with a high-speed detector, followed by an electronic amplifier and OC-48 filter.

4. Discussion and summary

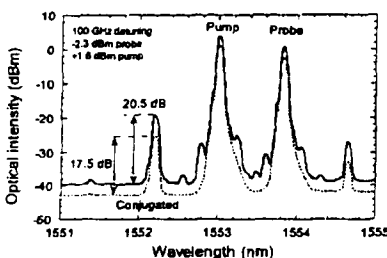
From numerical simulation, the improvement on the conversion efficiency was expected to be greater than the experimental results as the assisted beam is used. Especially, using a 1430-nm

assisted beam at high bias was expected to offer much better performance than using the 1480-nm one. The simulation indicates that the intrinsic loss inside the SOA has strong effect on this approach. For a SOA operating at 1.55- μ m wavelength region, the intrinsic loss arises mainly from the free carrier absorption and intervalence-band absorption. Both absorption mechanisms increase with the carrier density. As the SOA is biased at the transparent current for the assisted beam, the beam experiences the intrinsic loss as it propagates along the SOA. Thus, the loss at transparency is larger for a shorter-wavelength assisted beam. This loss limits the improvement on the saturation power and the conversion efficiency. Experimental results reveal that the fiber-to-fiber loss at transparency increases by about 10 dB as the wavelength of the assisted beam changes from 1480 nm to 1430 nm. This corresponds to an increase in the intrinsic absorption coefficient by about 23 cm^{-1} . In order to increase the conversion efficiency by using an assisted beam, efforts must be made to reduce the effects of intrinsic loss. For example, the improvement can be larger when a shorter and higher-gain SOA is used as the wavelength converter. Another approach is to inject the assisted beam laterally instead of longitudinally such that the beam intensity is not reduced by the intrinsic loss. This approach has been used to realize a linear optical amplifier by integrating a SOA with a vertical-cavity surface-emitting laser.⁴

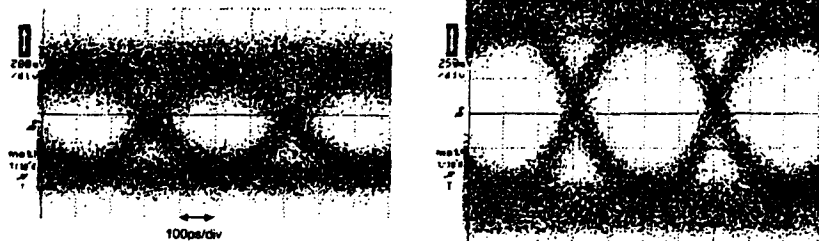
In summary, we demonstrated that the saturation intensity and wavelength conversion efficiency of a SOA could be increased with an assisted beam. The conversion efficiency can be improved by larger than 5 dB when the SOA is biased at the transparent current for 23-dBm of 1480-nm assisted beam. The assisted beam can also improve the SBR by about 3 dB.

References

1. A. D'Ottavi, et al., "Four-wave mixing in semiconductor optical amplifiers: a practical tool for wavelength conversion," *IEEE J. Select. Topics in Quantum Electron.*, vol. 3, pp. 522-528, 1997.
2. R.J. Manning, and D.A.O. Davies, "Three-wavelength device for all-optical signal processing," *Opt. Lett.*, vol. 19, pp. 889-891, 1994.
3. M.A. Dupertuis, et al, *IEEE Photon. Technol. Lett.*, "Extremely fast high-gain and low current SOA by optical speed-up at transparency," vol. 12, no. 11, pp. 1453-1455, 2000.



ThGG75 Fig. 3. Optical spectra showing 3 dB improvement on SBR.



ThGG75 Fig. 5. Eye patterns after wavelength conversion with (right) and without (left) a 1480-nm assisted beam. The SOA bias current is 105 mA. The noise is not as severe as what is shown in the figures since the original signal from the pattern generator is not perfect, showing amplitude fluctuation for the data bits.

4. D.A. Francis, S.P. Djaili, and J.D. Walker, "A single-chip linear optical amplifier," OFC 2001 Postdeadline paper 13.

ThGG76

5:30 pm

Investigation of Electroabsorption Modulator for 40 Gb/s Transmitter Application

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1. Introduction

Direct 40 Gb/s electrical modulation as in electrical time division multiplex (ETDM) has advantages over optical time division multiplex (OTDM) with respect to the mechanical size, cost and volume production for 40 Gb/s transmission system. The progress of 40 Gb/s component technologies, especially in the electrical multiplexer/demultiplexer and the electro-optic modulator (EOM), has made the development to deployment of 40 Gb/s ETDM transmission system within reach. In the choice of EOM, elec-

troabsorption modulator (EAM) has the advantage over LiNbO₃ modulator (LNM), on its small size, compatibility to integrate with DFB lasers, low driving voltage, and bias-drift insensitivity. To our knowledge, whereas there are a number of publications on EAM design,¹⁻³ there are few publications on EAM modeling and the application of such a model on the design optimization. In this paper, a large signal model for EAM is presented on the insertion loss (IL) and extinction ratio (ER) prediction with respect to different bias and RF driving signal levels. A figure of merit (FOM) for EAM is proposed to compare EAM performances. Furthermore, two system performance parameters are introduced to evaluate EAM IL and ER on end-to-end system performance.

2. Large Signal Model

In EAM, the most important characteristics are its electro-optic responses. Fig. 1(a) shows a typical curve (bottom one, measured DC loss curve) of an EAM insertion loss versus the bias voltage in the absence of a modulation signal. It is apparent from Fig. 1(a) that the EAM absorption is nonlinear. Furthermore, the IL and ER of EAM under a large signal modulation are quite different from

the DC condition. Predicting the IL and ER under AC condition is an important task for the transmitter design and system performance analysis.

Fig. 1(b) shows the two curves of measured s21 for an EAM under two different biases. Since EAM absorption is nonlinear, theoretically, only a small signal can be used to characterize its dynamic electro-optic response. In other words, the measured s21 cannot be directly used to simulate EAM output. Fortunately, it can be seen from Fig. 1(b) that the electro-optic response is almost unaffected with EAM bias within most of IL range. This strongly indicates that under first order approximation an EAM could be modeled as an E-E filtering block followed by an E-O transform block, and the filtering functionality is independent on modulation signal.

In the model, the RF signal is filtered by a filter basing on s21 characteristics first, and then transformed by the measured insertion loss vs. bias curve. The optical output of the modulator is then calculated. The physics behind this model is that the electroabsorption effect is almost instantaneous and the EAM bandwidth comes from the parasitic capacitances, which are almost static for the kind of EAM in question.

3. Modeling and Experimental Result Comparison

Based on the proposed model, EAM insertion losses in presence of various driving signal levels and at different bias voltages have been simulated. It has been found that they agree with the measurement data very well. In Fig. 1(a), the simulated and measurement results with a 2.2 Volts swing range are plotted together for comparison. Within our simulation range (-0.5 ~ -3 Volts on bias, and 1.0 ~ 3.0 Volts on swing range), a maximum AC insertion loss prediction error of smaller than 0.3 dB, has been obtained.

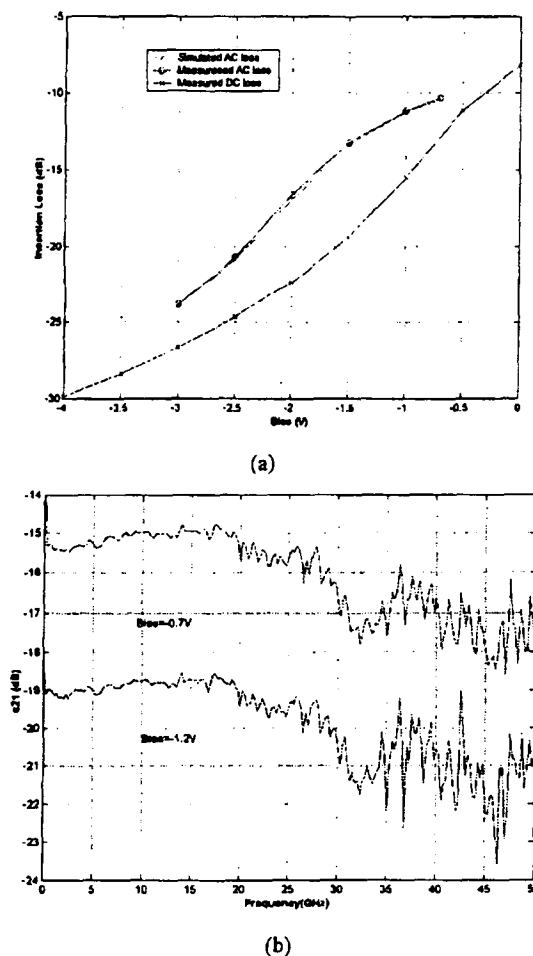
In order to examine the model on the signal waveform and extinction ratio, the PRBS (2¹⁷-1) signal waveform after EAM using digital communication analyzer (DCA) has been recorded. The EAM model based 40 Gb/s NRZ simulation waveform matches well the measured waveform. It is realized that the high frequency signal difference observed on level ones and zeros are attributable to measurement deviations. The difference of extinction ratio between the calculations from the experimental measurement and our simulation is only about 0.5 dB. The optical eye diagrams from the model based simulation and from the experimental measurement at EAM output have been plotted and recorded, respectively, which are shown in Fig. 2(a). In Fig. 2(b), based on the modeled EAM characteristics the calculated extinction ratios vs. bias with two different swing ranges are plotted.

4. Evaluation of EAM Parameters and Optimization of EAM Settings in 40 Gb/s Transmitter Design

In order to evaluate the EAM parameters and optimize the EAM settings for the transmitter module development, a figure of merit (FOM) and two EAM system performance evaluation parameters are proposed in this section. The EAM FOM is proposed in equation (1):

$$FOM_{EAM} = \frac{ER_{EAM}}{IL_{EAM}} \quad (1)$$

where IL_{EAM} is EAM AC insertion loss (in dB), and ER_{EAM} is the corresponding extinction ratio



ThGG76 Fig. 1. (a) Typical EAM DC insertion loss curve (bottom curve) and Comparison between simulated and measured EAM insertion loss; (b) Measured s21 for an EAM under two different biases.

Wavelength monitoring of tunable DWDM sources using a FP etalon and a FP laser diode

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發表論文：OFC2002, paper ThC2, Anaheim, CA, USA, Mar. 2002.

內容成果：

在 DWDM 架構化的高速光交換網路，光訊號波長快速的在通道中切換，除判斷精確的波長，對於高速的監控技術方面，使用便宜的 FP 雷射二極體作為監控通道的方法，經由實驗量測得 FP 雷射的介面電壓對於注入光訊號波長會有週期性且遞增或遞減的變化，利用此對應的關係，可快速判別出光訊號位於的標準光通道波長的所在位置；對於細微的波長誤差方面，使用被動式 FP etalon 作為波長鑑別器，經由 etalon 頻譜響應特性，轉換注入 etalon 光訊號的穿透與反射光功率，將光訊號功率變化引起的誤判雜訊抑制，精確判斷波長的偏移量。且 etalon 對於環境溫度變化的敏感度低，老化現象慢，適合做為穩定的波長漂移辨識元件。配合 FP 雷射二極體與 FP etalon 架設具有高速且精確的波長監控系統。

and found that, even in a temperature-controlled chamber, the unit required a ~2 h warmup before it stabilized. Some newer OSA units contain an internal wavelength reference such as an acetylene absorption cell.

Wavelength meters that measure a signal's wavelength relative to a built-in fundamental reference (often a helium-neon laser operating at the 632.991 nm vacuum wavelength) usually don't require frequent calibration. The most convenient way to check a wavelength meter is to measure a laser that has a known wavelength, such as the 1523.488 nm helium-neon laser (~1 part in 10^6 absolute stability) or a laser that is stabilized to a fundamental reference. Alternatively, a tunable laser can be tuned over an absorption line while measuring the wavelength with a wavelength meter; the measured line center can then be compared with the true line center.

In all of these cases, it is important to use the wavelength of light in vacuum, rather than the wavelength in air, when calibrating equipment. The wavelength of light in air can vary substantially due to its dependence on the atmospheric pressure, temperature, and humidity.

5. Conclusions

There are a variety of convenient wavelength calibration standards in the 1500 nm region, and other standards are under development for the WDM L-band. Fundamental references based on atomic and molecular absorption or emission lines provide the highest accuracy, but they are not available in all wavelength regions. Wavelength division multiplexing will likely expand into the 1300 and 1400 nm regions, and it will be very difficult to find absolute references for this wide wavelength range. Artifacts such as etalons or fiber Bragg gratings can provide references at arbitrary wavelengths, but they can suffer from large sensitivity to temperature, strain, and pressure. Passive or active thermal stabilization can substantially reduce this variability, but the artifact references need to be checked periodically against a fundamental reference. Further in the future, stabilized frequency combs may become commercially available to provide frequency (wavelength) markers throughout the WDM regions.

References

- S.L. Gilbert, "Frequency Stabilization of a Fiber Laser to Rubidium: a high accuracy 1.53 μm wavelength standard," in *Frequency Stabilized Lasers and Their Applications*, Proc. SPIE 1837, 146–153 (1993).
- Y.C. Chung, "Frequency-Locked 1.3- and 1.5- μm Semiconductor Lasers for Lightwave Systems Applications," *J. Lightwave Technol.* 8, 869–876 (1990).
- U. Fischer and C. von Helmolt, "Absorption Spectra of Excited Kr 84 States between 1.5 and 1.58 μm and Their Use for Absolute Frequency Locking," *J. Lightwave Technol.* 14, 139–143 (1996).
- A. Bruner et al., "Frequency stability at the kilohertz level of a rubidium-locked diode laser at 192.114 THz," *Appl. Opt.* 37, 6410–14 (1998).
- D.A. Humphreys, "Accurate wavelength calibration for optical spectrum analyzers," in *Technical Digest—Symposium on Optical Fiber Measurements*, Natl. Inst. Stand. Technol. Spec. Publ. 905, pp. 97–100 (1996).
- W.C. Swann and S.L. Gilbert, "Pressure-induced shift and broadening of 1510–1540-nm acetylene wavelength calibration lines," *J. Opt. Soc. Am. B* 17, 1263–1270 (2000); S.L. Gilbert and W.C. Swann, "Acetylene $^{12}\text{C}_2\text{H}_2$ Absorption Reference for 1510 nm to 1540 nm Wavelength Calibration—SRM 2517a," Natl. Inst. Stand. Technol. Spec. Publ. 260–133 (2001 Edition).
- S.L. Gilbert, W.C. Swann, and C.M. Wang, "Hydrogen Cyanide $\text{H}^{13}\text{C}^{14}\text{N}$ Absorption Reference for 1530–1560 nm Wavelength Calibration—SRM 2519," Natl. Inst. Stand. Technol. Spec. Publ. 260–137 (1998).
- Y. Kokubun et al., "Athermal Waveguides for Temperature-Independent Lightwave Devices," *Photon. Technol. Lett.* 5, 1297–1300 (1993).
- T. Iwashima et al., "Temperature compensation technique for fibre Bragg gratings using liquid crystalline polymer tubes," *Electron. Lett.* 33, 417–419 (1997).
- W.C. Swann et al., "Hybrid multiple wavelength reference using fiber gratings and molecular absorption," in *Proc., Bragg Gratings, Photosensitivity, and Poling in Glass Waveguides*, OSA Technical Digest (Optical Society of America, Washington DC, 1999), pp. 63–65.
- V. Bertogalli et al., "Testing passive DWDM components: uncertainties in swept-wavelength measurement systems," in *Proc., 2001 Optical Fiber Communication Conference* (Optical Society of America, Washington DC, 2001), paper THB6.
- M. Maleki and J.C. Petersen, "Wavelength calibration of a fiber-optical spectrum analyzer: temperature dependence," *Appl. Opt.* 36, 4451–4455 (1997).

ThC2

9:00 am

Wavelength monitoring of tunable DWDM sources using a FP etalon and a FP laser diode

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1. Introduction

Tunable laser sources are critical components for DWDM networks as spare sources or fast wavelength-switching devices. The wavelength monitor for controlling a tunable laser diode must cope with multiple output wavelengths and the required tuning speed. Fabry-Perot (FP) etalons have been commercially used for wavelength con-

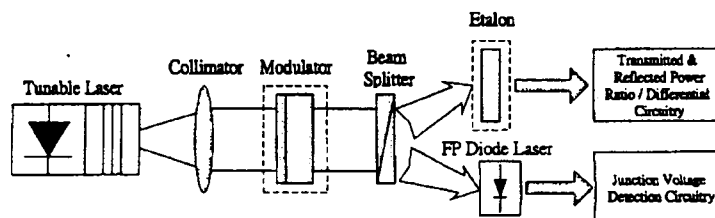
trol of DWDM lasers. However, a FP etalon cannot distinguish among different channels since it has periodic wavelength characteristic. This makes it difficult to monitor the mode-hopping problem that might happen in tuning a tunable laser.¹ Etalons with special tuning mechanism or array waveguide gratings were also proposed to monitor tunable lasers,^{2,3} but they may have limited response speed or require expensive components. On the other hand, wavelength monitoring by detecting the transparent current or the induced junction voltage across a FP laser diode (FPLD) was demonstrated to provide 0.01-nm wavelength resolution.⁴ This approach is very attractive since the wavelength sensor can be monolithically integrated with the tunable laser. Moreover, a FPLD can easily be tuned to monitor multiple channels by adjusting its bias current. However, to achieve 0.01-nm resolution with this approach requires tough stability on temperature control.

We propose here a high-resolution tunable wavelength monitoring technique that combines the advantages of stable temperature properties of a FP etalon filter as well as fast tunability of the transparent properties of a FPLD. Significant relief to the required temperature stability, about $\pm 0.2^\circ\text{C}$, is achieved, so wavelength resolution of 0.01 nm can easily be obtained for many channels. The ambiguity among different channels for the etalon can be resolved from the induced voltage across the LD. This technique can also be applied in optical networking units to monitor the wavelength paths in a scanning mode.

2. Operation principles

Fig. 1 shows the schematic of the proposed wavelength monitoring technique for tunable laser module. The laser output is first collimated and dithered at a specific frequency with an optical modulator. Then, the signal is divided into two paths, one to an FP etalon, and the other to a FPLD. Both the transmitted and reflected signals of the etalon are measured and the resultant differential voltage or voltage ratio for these two signals are used for detecting the wavelength drift from the wavelength grid. The FPLD can resolve the wavelength channel from either the corresponding transparent current or the junction voltage.

The tunable laser light is dithered in order to improve the detected signal quality for both monitoring paths. For easy packaging with the tunable laser in a compact module, a surface-normal type of modulator is favored to simplify light coupling. Since the required dithering frequency is between kHz and MHz, modulators made of liquid crystal or micromechanical silicon modulators⁵ are among the good choices for such applications.

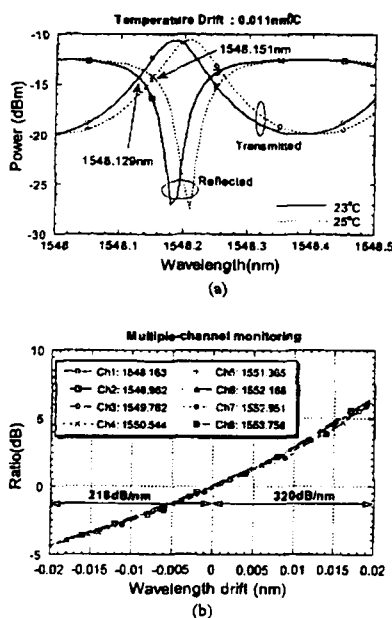


ThC2 Fig. 1. Schematic of wavelength monitoring for a tunable laser.

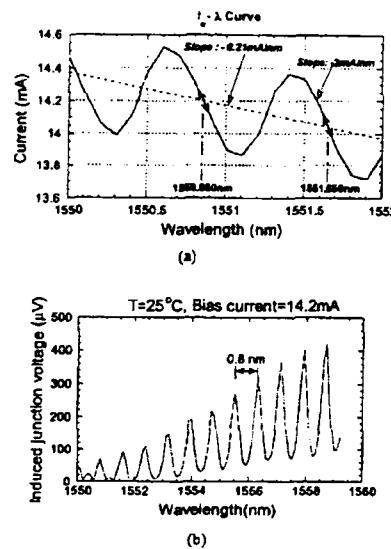
3. Experimental demonstration

The proposed method exploits the FP etalon as both a wavelength locker and a stable wavelength reference because of its stable response to ambient temperature variation. Fig. 2(a) depicts the reflected and transmitted characteristics of an etalon of which the cavity length is specially designed to give rise to a free-spectral range (FSR) of 50 GHz. The intersection point of the transmitted and reflected curves is frequency used as the monitoring point. Fig. 2 shows that the point varies at a rate of 0.011 nm/°C as the temperature changes. If the tolerance on wavelength deviation is 0.01 nm, this etalon can tolerate up to 0.9°C of temperature change. The etalon was temperature-controlled to align one of the ITU wavelengths to a specific crossing point of the two curves. Fig. 2 shows the response for the etalon to eight wavelengths of 100-GHz channel spacing at the same temperature, 26°C. Since the channel spacing matches twice the FSR, the responses to wavelength drift are almost identical among the channels. Therefore, the monitoring circuit needs to memorize only one of these curves. The maximal deviation among the curves results in 0.005 nm of wavelength deviation, which is much smaller than the channel spacing.

As the tunable laser is tuned to output a given DWDM wavelength, the etalon must monitor the wavelength deviation from the wavelength grid and feedback the deviation to the laser control unit to adjust the wavelength to the closest wavelength grid. Then, the FPLD will recognize the channel and detect if mode hopping occurs. Two modes of operation for the FPLD can be performed for channel recognition. One is to sweep the bias current and locate the transparent current (I_T) for the incident wavelength (λ), and the other is to detect the induced junction voltage across the diode for a fixed bias current. Fig. 3 shows the wavelength-dependent characteristics



ThC2 Fig. 2. Characteristics of the FP etalon: (a) at different temperature and (b) around eight ITU wavelengths.

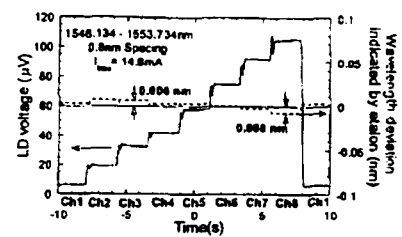


ThC2 Fig. 3. Wavelength variation of transparent current (a) and junction voltage (b) for a FPLD.

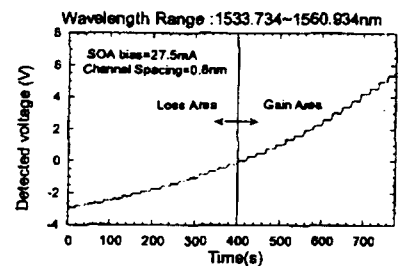
for a 1.55-μm FPLD. For a SOA, the transparent current decreases as the wavelength moves to the long-wavelength side. On the other hand, due to the facet reflection, resonant effects appear in the wavelength variation of transparent current and induced voltage. The wavelength spacing between the resonant peaks is equal to the cavity mode spacing of the LD. Here, we chose a FPLD of which the cavity mode spacing is very close to 100 GHz (0.8 nm).

Although it is desired to have a monotonic I_T -vs- λ curve, which can be obtained from a SOA, the resonant effects can enhance the wavelength resolution. It was also found in the experiments that the induced junction voltage across a FPLD is larger than that across a SOA. For monitoring a DWDM tunable laser, the resonant characteristics can be used if the cavity mode spacing is identical to the channel spacing. After the wavelength is aligned to the wavelength grid, only the discrete points spaced by the channel spacing will be used for channel recognition. The curve can easily be adjusted by changing the diode temperature to align the wavelength grid to the middle of the slope, as indicated in Fig. 3. The transparent current differs by 0.16 mA between the adjacent channels. The tolerance on temperature control of the FPLD is relaxed to $\pm 0.2^\circ\text{C}$ because the transparent current varies by 0.385 mA/°C with the temperature. This benefits again from the use of the FP etalon as a wavelength reference. Moreover, the deviation of the transparent current from the value for a wavelength grid can indicate the temperature shift and serve as the control signal for stabilizing the temperature.

Detecting the junction voltage may be favored for monitoring fast tunable lasers since it has a faster response than locating the transparent current. Fig. 4 demonstrates a tunable wavelength monitoring with a combination of a FP etalon and a FPLD. A commercial tunable laser was tuned to eight DWDM channels in a step of 0.8 nm. The junction voltage of the FPLD changes in a stair way in response to the wavelength tuning.



ThC2 Fig. 4. Tunable monitoring of eight 100-GHz channel. The monitored light is modulated at 1 kHz.



ThC2 Fig. 5. Output voltage of SOA for 35 channels. The induced voltage was amplified by 2×10^5 .

Tiny signal appears at the output of the etalon circuit since the wavelengths are allocated at the wavelength grid. The etalon signal indicates a maximal wavelength deviation of 0.008 nm, which is acceptable for DWDM channels of 100-GHz spacing.

4. Discussion

For monitoring a widely tunable laser that can output more than 30 channels, a FPLD might suffer from the resonant characteristics if the cavity mode spacing is not exactly equal to the channel spacing. In such cases, a SOA with monotonic wavelength variation of transparent current and junction voltage is needed. Fig. 5 shows the output signal as the tunable laser is tuned over 35 channels. The channel wavelengths can be clearly resolved from the junction voltage. In the experiments, the SOA is biased at the transparent current for the 1548.134-nm channel. Phase-sensitive detection is employed to obtain the voltage such that the channels reside in the gain region of the SOA can be distinguished from those in loss region. A SOA for this application can be low-cost since it does not require tough anti-reflection coating as required for conventional applications. Noting that, as a SOA is used, the mode hopping or wavelength channels can be determined from the junction voltage without waiting for the etalon to bring the wavelength to the closest wavelength grid.

References

1. D.A. Ackerman, et al., "Wavelength, modal, and power stabilization of tunable electro-absorption modulated distributed Bragg reflector lasers," in *Int. Semiconductor Laser Conf.*, pp. 49-50, Monterey, CA, 2000.
2. K. Tatsuno et al., "50 GHz spacing, multi-wavelength tunable locker integrated in a

transmitter module with a monolithic-modulator and a DFB laser," *OFC 2001*, Anaheim, USA, paper TuB5, 2001.

3. K.R. Tamura et al., "32 wavelength tunable mode-locked laser with 100 GHz channel spacing using an arrayed waveguide grating," *OFC 2001*, Anaheim, USA, paper TuJ5, 2001.
4. S.-L. Lee, et al., "High-resolution wavelength monitoring using differential/ratio detection of junction voltage across a diode laser," *IEEE Photon. Technol. Lett.*, vol. 13, pp. 872-874, 2001.
5. K.W. Goosen, et al., "Silicon modulator based on mechanically-active anti-reflection layer with 1 Mbit/sec capability for fiber-in-the-loop applications," *IEEE Photon. Technol. Lett.*, vol.6, pp. 1119-1121, 1994.

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Birefringent Gires-Tournois Interferometer (BGTI) for DWDM Interleaving

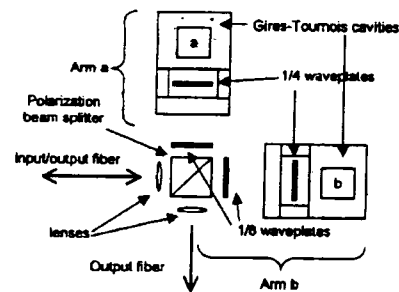
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1. Introduction

Interleavers enable the practical implementation of dense wavelength division multiplexing (DWDM) systems by transforming the task of separating or combining a single set of very closely spaced channels into separating or combining multiple sets of widely spaced channels. Regardless of the approach, be it thin-film filter, fiber-Bragg grating, arrayed waveguide grating, Echelle grating, or bulk diffraction grating, mux/demux modules can be less expensive and easier to make if the channel spacing is wide relative to the required passband width. Interleavers allow the application of these wide channel spacing solutions to the harder problem of more closely spaced channels. Most interleaver designs are based on cascaded Mach-Zehnder (M-Z) interferometers. Researchers have constructed these cascaded M-Z interleavers out of fused fibers and planar lightwave circuits for example.¹⁻² Other, non-M-Z and hybrid-M-Z approaches exist as well.³

We propose a Birefringent Gires-Tournois Interferometer (BGTI) that combines the benefits of the nonlinear phase response of Gires-Tournois (G-T) cavities, free space propagation, and low sensitivity to physical layout. This new structure is assembled in a Michelson configuration except that the beam splitter is replaced by a polarization beam splitter (PBS) and the mirrors are replaced by Gires-Tournois (G-T) cavities. Furthermore, 1/8 waveplates are placed between the polarization beam splitter and the G-T cavities in each arm of the interferometer and 1/4 waveplates are placed within the G-T cavities (see figure 1).

The BGTI interleaver possesses four desirable features: low stage count, temperature insensitivity, frequency based periodicity, and fabrication simplicity. The highly nonlinear phase response of G-T cavities enables the BGTI interleaver to accomplish with a single stage what other M-Z designs require two or even three stages to achieve.¹⁻² The temperature stability and frequency, as opposed to wavelength, periodicity are derived from the predominantly free-space de-



ThC3 Fig. 1. Birefringent Gires-Tournois Interferometer (BGTI) interleaver.

sign. Most of the propagation through the interleaver is through air whose index of refraction dependence on temperature and wavelength is weak. Designs dependent on wavelength periodicity suffer from channel center walk-off at the extreme channels in a DWDM system since the channel grid is defined by frequency. The fabrication simplicity stems from the use of thin, birefringent plates to create the path length differential between the two beams that interfere to produce the interleaver response.

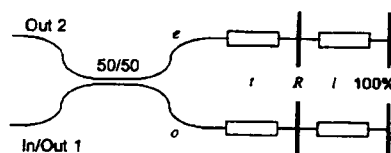
2. Operation of device

Fundamentally, the two outputs of the BGTI interleaver are the sum and difference of the equally weighted outputs of two otherwise identical all-pass filters whose frequency responses are shifted by 1/2 period. A phase delay is added in one arm to provide phase offset adjustment (see figure 2). Madsen provides a description of arbitrary filters based on this type of basic building block.⁴

In the BGTI, the *s* and *p* polarizations defined by the PBS are rotated 45° with respect to the ordinary (*o*) and extraordinary (*e*) axes of the waveplates. The function of the PBS is two-fold. The PBS segregates the incoming *s* and *p* polarizations for separate treatment in the two arms of the BGTI. In conjunction with the waveplates, it also functions as a 50/50 splitter.

Considering only the *p* polarized light, half the power travels along the *e* axes of the two waveplates and the other half travels along the *o* axes due to the 45° rotation of the *p* light with respect to those axes. Upon return to the PBS, the *e* and *o* beams from the waveplates are combined as sum and difference beams and are output in either the *s* or *p* plane defined by the PBS. The operation is similar for the *s* polarized light traveling through the other arm.

The two arms of the interferometer operate independently so that there is no interference between light of the same wavelength from the two arms. The light of a given wavelength traveling back to the PBS from arm a is always orthogonal in polarization to the light of the same wave-



ThC3 Fig. 2. All-pass filter decomposition of a single arm in the BGTI interleaver.

length from arm b (see figure 1). For this reason, the lengths of arms a and b are inconsequential to the operation of the device and therefore temperature or vibration induced deviation is harmless. In fact, the only reason two arms are included is to accommodate arbitrary input polarization. This does mean that inequality between the lengths of the two arms will produce polarization mode dispersion. However, this effect is small and easy to control.

3. Theory

As stated previously, each arm of the BGTI interleaver operates independently of the other and is mathematically identical to the filter shown in figure 2. The phase response of each arm in the filter shown in figure 2 is:

$$\phi_i = 2k_i t + 2 \tan^{-1} \left[\frac{1 + \sqrt{R}}{1 - \sqrt{R}} \tan(k_i l) \right]$$

where the subscript *i* denotes either the extraordinary or ordinary arm of the filter, *k_i* is the polarization dependent wavevector, *t* is the thickness of the waveplate outside the cavity, *R* is the power reflectivity of the cavity, and *l* is the cavity length including the waveplate inside the cavity.

Within the cavity, the round trip phase difference between the *e* and *o* arms due to the 1/4 waveplate is π . This ensures a symmetric response in the two output ports. The selection of *R* determines the depth of phase modulation. Large *R* will produce rapid change in the phase response causing a sharp transition from the passband to the stopband. Large *R* will also produce ripple in the passband and chromatic dispersion. Outside the cavity, the round trip phase difference between the *e* and *o* arms due to the 1/8 waveplate provides a phase difference bias to produce full range in the response of the two arms.

By manipulating the values of *R*, *t*, and *l*, different filtering functions can be achieved. The BGTI is a basic building block that can be used to build general IIR optical filters as described by Madsen.⁴ By adding mirrors and waveplates, higher order all-pass filters can be constructed.

4. Results

The two devices discussed in this paper were built out of discrete, micro-optic components. The G-T cavities were air gap cavities constructed with low thermal expansion spacers. The first device uses *R* = 18.5% mirrors. As shown in figure 3, the calculated and measured responses of the two ports match very well over the passband. However, the two outputs have different isolation. The isolation of port 1 is worse than that of port 2. Moreover, the isolation is worse than calculated in both cases. These features are caused by the imperfect and asymmetric segregation of the *s* and *p* polarizations by the PBS cube. Since the extinction ratios of the transmission and reflection operations of the PBS are unequal, different amounts of crosstalk will leak into each port since the path to each port involves different numbers of transmission and reflection interactions with the PBS. Practically, PBS isolation is typically limited to 30 dB at best. Therefore, the isolation floor of the BGTI is similarly restricted. The 1 dB passband of the device shown in figure 3 is 77 GHz and the insertion loss is less than 1 dB.

In order to improve the cross-port isolation, additional BGTI stages can be added. In figure 4, the output spectra of a two stage BGTI interleaver with *R* = 22.0% for all cavities is shown. Since the

Wavelength conversion using the four-wave-mixing in a semiconductor optical amplifier assisted with an injection beam.

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內容成果：

波長轉換器為實現高速光通訊網路中一個相當重要的元件。本文使用半導體光放大器作為波長轉換的元件，利用半導體光放大器的非線性效應：四波混合技術，注入一幫浦光源與訊號光源，產生第三道對應的轉換波長。然而目前使用此方式有波長轉換效率不高的缺點，因此提出再多加一道輔助光，提高轉換效率。經由實驗，當加入輔助光之後，波長轉換器的轉換效率可提升 5dB，且訊號功率與背景功率比也提升了 3dB，對於轉換訊號的功率級訊號品質，增加了不少幫助，提高使用四波混合製作全光波長轉換器的可行性。

Wavelength conversion using the four-wave-mixing in a semiconductor optical amplifier assisted with an injection beam

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Abstract

We demonstrated experimentally >5dB improvement on wavelength conversion efficiency and >4dB improvement on saturation power of a semiconductor optical amplifier by using an assisted beam at transparency. The effects of intrinsic loss and the measures for further improvement are discussed.

1.Introduction

All-optical wavelength converters will play important roles in the future optical networks. It can solve the wavelength blocking problems and provide wavelength routing and switching to enable optical internet. The wavelength conversion using the four-wave mixing (FWM) effect in a semiconductor optical amplifier (SOA) can provide the needed functions for optical networks [1], but it suffers from poor conversion efficiency. Significant improvement on the conversion efficiency and noise figure is required for practical applications. Ref. [2] demonstrated experimentally that the saturation power can be enhanced by using an assisted beam, which also enhances the carrier recovery rate [3]. So the conversion efficiency of four-wave mixing (FWM) can be improved as the saturation power is increased [1]. Another way to improve the conversion efficiency is to increase gain by increasing applied current. It will increase amplified spontaneous emission (ASE) and increase the electrical consumption. However, the assisted beam will also deplete the signal gain by injecting an assisted beam of which the wavelength is in the gain region of the SOA. Dupertuis *et al.* proposed to use an assisted beam at transparency to speed up a SOA [4]. This technique can increase the saturation intensity without sacrificing the gain and improve the conversion efficiency of FWM. In this paper, we will demonstrate that using an assisted beam can improve the efficiency and signal-to-noise ratio of FWM wavelength.

2.Principle

From rate equation the effects of using an assisted beam can be observed in the following relationship:

$$\frac{1}{\tau_{eff}} = \frac{1}{\tau} + \frac{P_a}{P_a^{sat}} \dots\dots\dots(1) \quad P_s^{sat} = \frac{AE_s}{\Gamma a_s \tau_{eff}} \dots\dots\dots(2)$$

where P_a is the assisted beam power; E_s is the signal photon energy; Γ is the confinement factor; A is cross-sectional area of the active region; P_a^{sat} is the saturation power for the assisted beam and τ_{eff} is the effective carrier lifetime. Eq.(2) indicates that the effective carrier lifetime decrease as the assist beam power increases. Thus, the signal saturation power in (3) can be enhanced with the reduced effective carrier lifetime. If the SOA is operated at the current small than the transparency for the assisted beam, the assisted beam will be absorbed and offer the excess carriers to increase the signal gain. On the other hand, if the SOA is operated above the transparency for the assisted beam, the assisted beam will deplete the carriers and decrease the signal gain. Only when the SOA is operated at the transparency current for the assisted beam, the assisted beam can increase the saturation power while keep the gain unchanged

The FWM conversion efficiency increases approximately with $G_0 P_s^2$, where G_0 is the unsaturated gain and P_s is the saturation intensity of the SOA [1]. Thus, the conversion efficiency can be increased by using an assisted beam to raise the saturation power..

3.Experiment

To prove the improvement on wavelength conversion efficiency with an assisted beam, the measurement setup for SOA gain characteristics and conversion efficiency shown in Fig. 1 was used. Since the SOA is long, the assisted beam is incident contra-directionally with the signal and pump signals to have the best effects. The pump and probe beams out of two commercial tunable lasers with built-in isolators were combined using a 50/50 coupler and inputted to the SOA. A 1.48/1.55- μ m multiplexer was used at the output end to combine the input assisted beam and output conjugated signal. The multiplexer has a broad passband that provides low insertion loss to both the 1480- and 1430-nm assisted beams. The signal output and conjugated output was measured with an optical spectrum analyzer.

3.1 Gain characteristics:

The transparent condition of a SOA corresponds to the condition that the material gain of the SOA is zero for

the incident wavelength. Fig. 2 compares the gain curves at 1550 nm for the 1.55- μm SOA used in our experiments with and without a 1480-nm assisted beam. The two curves intersect at the current that corresponds to the transparent condition for the assisted beam since the gain is not changed for the presence of the assisted beam. An alternative for detecting the transparent condition is to locate the bias current where the junction voltage across the SOA junction vanishes. However, the background power of the strong assisted beam will induce background junction voltage and then affect the accuracy in determining the transparency. Therefore, we use Fig. 2 to determine the transparent current, which is 105 mA for the 1480-nm assisted beam. Here the assisted beam power in Fig.2 is 23.3dBm.

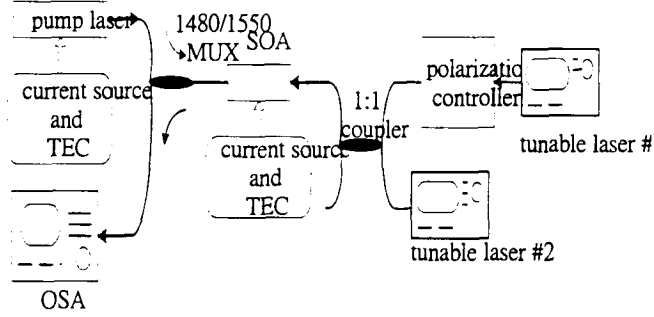


Fig.1: Measurement setup

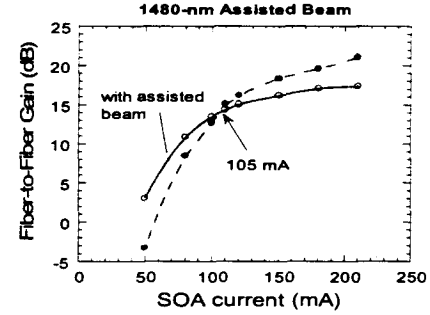


Fig.2: Measured transparent current for 1480 assisted beam

Fig. 3 shows that the assisted beam does not affect the unsaturated gain at 105mA and raises the saturation gain by 4.4 dB. Also shown in this figure are the gain saturation characteristics when the SOA is biased at 200 mA. At such a high-injection bias, the amplified assisted beam depletes the carriers in the SOA. Thus, the measured data show that the assisted beam causes about 2.8 dB of reduction in the gain but 2.9 dB increase in the output saturation power. Fig. 4 shows the measured intrinsic loss (loss at transparent condition) for various assisted wavelengths. The loss increases to a significant value as the wavelength moves to a shorter wavelength.

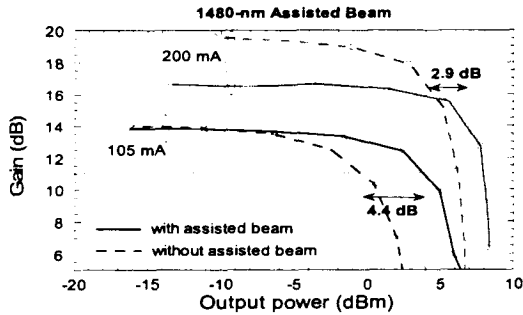


Fig.3: Gain curves when the wavelength of the assisted beam is 1480-nm and the assisted power is 23.3dBm

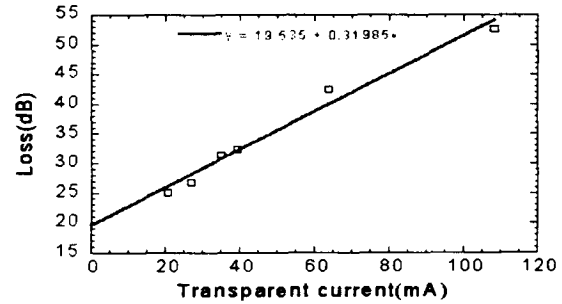


Fig.4: The internal loss at transparency for different wavelength

3.2 The conversion efficiency of FWM:

The conditions in Fig. 5(a) are 1480-nm assisted beam of 23.3 dBm power, -10dBm probe power, 1553nm pump wavelength and performed at two different bias current, one at transparency (105 mA) and the other at high injection (200mA). The conversion efficiency can be enhanced by 5.2dB at the transparent current of 1480nm assisted beam when the detuning(Ω) between the pump wavelength and probe wavelength is 100Gbps. The enhancement increases slightly as the detuning becomes larger. This situation can also be observed at a high injection level (200mA). There is only 2.9dB enhancement of conversion efficiency at 200mA. This is due to the fact that the assisted beam increases the saturation power but decreases the available gain. Around 3.1 dB of improvement on the conversion efficiency can still be obtained at 200 mA of bias current. And as the detuning increases, the conversion efficiency decreases primarily Fig.5 (b). That enhancement increases slightly as the detuning becomes larger can also be seen in Fig. 5(b). For an assisted beam of shorter wavelength, its corresponding transparent current is higher. Therefore, it is possible to obtain both high conversion efficiency and large improvement from using a shorter-wavelength assisted beam. The 1430-nm assisted beam provides larger improvement than the 1480-nm assisted beam for the same power (18dBm). The difference is more pronounced at higher SOA bias, where the 1480-nm beam depletes more carriers than the 1430-nm beam does. Here the transparency current of 1430nm assisted beam is close to 200mA.

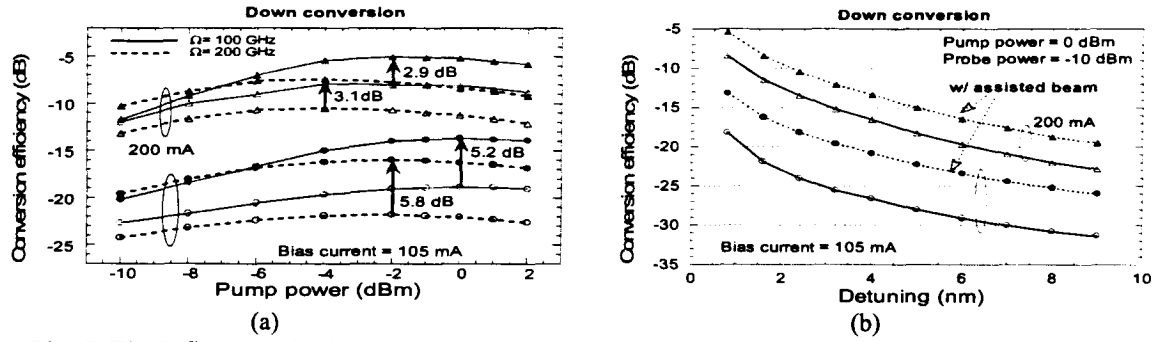


Fig. 5: The influence of using 1480nm assisted beam on wavelength conversion efficiency. The assisted power is 23 dBm.

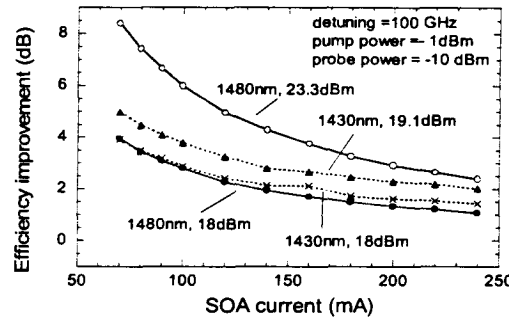


Fig. 6: The improvement of conversion efficiency for different assisted wavelength

4. Discussion and summary

In addition to the improvement on conversion efficiency, the assisted beam can increase the signal-to-background-noise ratio (SBR) by about 3 dB and 1.5 dB at the bias levels of 105 and 200 mA, respectively. The spectrum of FWM indicates that the strong pump and probe beams experience larger gain due to the increase in the saturation power by the assisted beam. Thus, it is helpful to increase the conversion efficiency. But in our simulation using an assisted beam at the contra-direction to raise carrier density can result in a slight increase in the amplified spontaneous emission (ASE). The SBR can be improved by injecting the assisted beam laterally instead of longitudinally such that the beam intensity is not reduced by the intrinsic loss. This approach has been used to realize a linear optical amplifier by integrating a SOA with a vertical-cavity surface-emitting laser [5].

Fig. 4 indicates that the internal loss of 1430nm wavelength in SOA is higher than 1480nm wavelength. If a shorter-wavelength assisted beam is used and the SOA is biased at the transparency corresponding to the assisted beam, efforts must be made to reduce the effects of intrinsic loss. One method is to reduce the intrinsic loss by employing strain into the active material to suppress the intervalence band absorption. The other one is to use a shorter and high-gain SOA for the wavelength conversion.

In summary, we demonstrated that the saturation power and wavelength conversion efficiency of a SOA could be increased with an assisted beam. The saturation power can be improved by 4.4dB when the SOA is biased at the transparent current for 23-dBm of 1480-nm assisted beam. The conversion efficiency can be improved by larger than 5 dB

5. Reference

- [1] A. D'Ottavi, et al., "Four-wave mixing in semiconductor optical amplifiers: a practical tool for wavelength conversion," *IEEE J. Select. Topics in Quantum Electron.*, vol. 3, pp.522-528, 1997.
- [2] M. Yoshino, and K. Inoue, "Improvement of Saturation Output Power in a Semiconductor Laser Amplifier through Pumping Light Injection," *IEEE Photonics Technology Letters*, vol.8, no.1, pp.58-59, January 1996
- [3] R. J. Manning, and D.A.O. Davies, "Three-wavelength device for all-optical signal processing," *Opt. Lett.*, vol.19, pp.889-891, 1994.
- [4] M. A. Dupertuis, et al., "Extremely fast high-gain and low current SOA by optical speed-up at transparency," *IEEE Photon. Technol. Lett.*, vol.12, no.11, pp. 1453-1455, 2000.
- [5] D.A Francis, S.P. DiJaili, and J.D. Walker, "A single-chip linear optical amplifier," OFC 2001 Postdeadline paper 1.

A fast wavelength-tunable transmitter for optical switching applications

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內容成果：

未來的高速光網路強調可調性，可調性可提供網路擁有較高的傳輸效能。
本文利用波長可調式 DBR 雷射，設計具有高速波長調動功能的波長可調式光發射器，調動速度可達 10MHz，5nm 以上的波長調動範圍，適合應用於光分封網路(OPS)與光脈衝式交換網路，作為光發送器、光路由器或可調式光塞取器，實現高速光網路架構。透過簡單的控制電路，將不同波長轉換為數位控制代碼，容易操作與應用，並具有低成本的優點。

A fast wavelength-tunable transmitter for optical switching applications

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Abstract: High-speed wavelength-tunable transmitters are demonstrated by using DBR lasers and high-speed electronics. The DBR lasers are designed to have 100-GHz cavity mode spacing for easy tuning. We obtained <100 ns switching time and > 5mW output power.

NSC project code: NSC 91-2215-E-011-005

1. Introduction

Nowadays optical networks utilize optical cross-connect (OXC) to enable dynamic provisioning and flexible services. As the network traffic keeps increasing rapidly, optical burst switching (OBS) or optical packet switching (OPS) is the key technology to improve network utilization. Most of the optical switches, like MEMS switches, can not cope with the required switching of OBS and OPS that require a switching time of less than 1 μ s. A combination of fast wavelength-tunable transmitter and a passive wavelength router is a good choice as fast optical switch. The switching speed of such optical switch is limited by the switching speed of the tunable lasers. For instance, the HORNET project at Stanford University was aimed at improving the inefficiencies of SONET networks to meet the needs of future optical packets networks [1]. HORNET used a high-speed wavelength-tunable transmitter to enable optical packet switching. Packets with different destination are packed into corresponding wavelength channel directly. Though many manufacturers are developing tunable semiconductor lasers [2], no lasers with <1 μ s tuning speed are available in the market. Moreover, the output wavelengths must meet the DWDM wavelength grid. We first focus on the DBR lasers that can be fast tuned with simple control mechanism. The approach can be applied for other types of tunable lasers. The switching time of the module, including the settling time of DAC and the access time of SRAM-based wavelength map, is less than 100ns. The switching time is currently limited by the electronics and can be improved as faster electronics are available.

2. Wavelength-tunable transmitter module

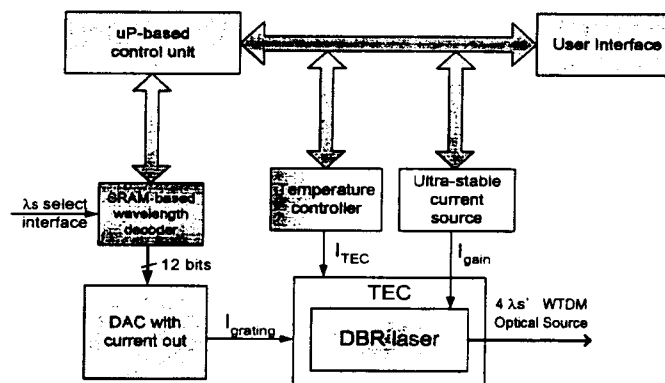


Fig. 1 The schematic diagram of tunable laser module.

The schematic diagram of a wavelength-tunable transmitter module is shown in Fig.1. We use a SRAM as a wavelength map to store the tuning parameters of 4 wavelength channels. It is easy to increase the wavelength channels. If the wavelength numbers of a DBR laser increases to 16 or even 256 channels, the hardware of the module need no modification. The microprocessor-based control unit and user interface are used to download the wavelength map to the SRAM, to set and monitor operation temperature and laser current, and to communicate with other control units. The module also provides a simple self-testing procedure to test the tuning of a DBR laser.

The DBR laser is a conventional 3-section component that consists of the *gain*, *phase* and *grating* electrodes. The *gain* section mainly provides the gain for the laser. By careful design of the device length, wavelength tuning to channels of a given spacing can be achieved by injecting current into the *grating* section. The phase section can be used for fine tuning that allows for accurately setting the initial wavelength. After fabricating into ridge-type lasers, the lasers are cleaved into bars, tested, and packaged into the dual-in-line (DIL) modules with thermoelectric cooler and thermister. The threshold current is 20mA at room temperature and the fiber output power can be larger than 5mW (Fig.2a). By tuning only the *grating* section, the tuning map of a DBR laser is shown in Fig.2b. The output wavelengths of a DBR laser are stair-like with channel spacing close to 100GHz. This can assure that the output wavelength can be tuned to the DWDM wavelength grid at a very fast speed [2]. The side mode suppression ratio (SMSR) is large than 30dB around the channel positions (Fig.2c). The power variation among the channels is only 1.5dB for 4 wavelength channels (Fig. 2d), so no additional power compensation is needed. In Fig. 2e, there is a little red shift at short-wavelength channels due to the imperfect heat sink of the packaged laser module. Fig. 2f shows a 14-pin dual-in-line (DIL) packaged laser with SC/PC connector.

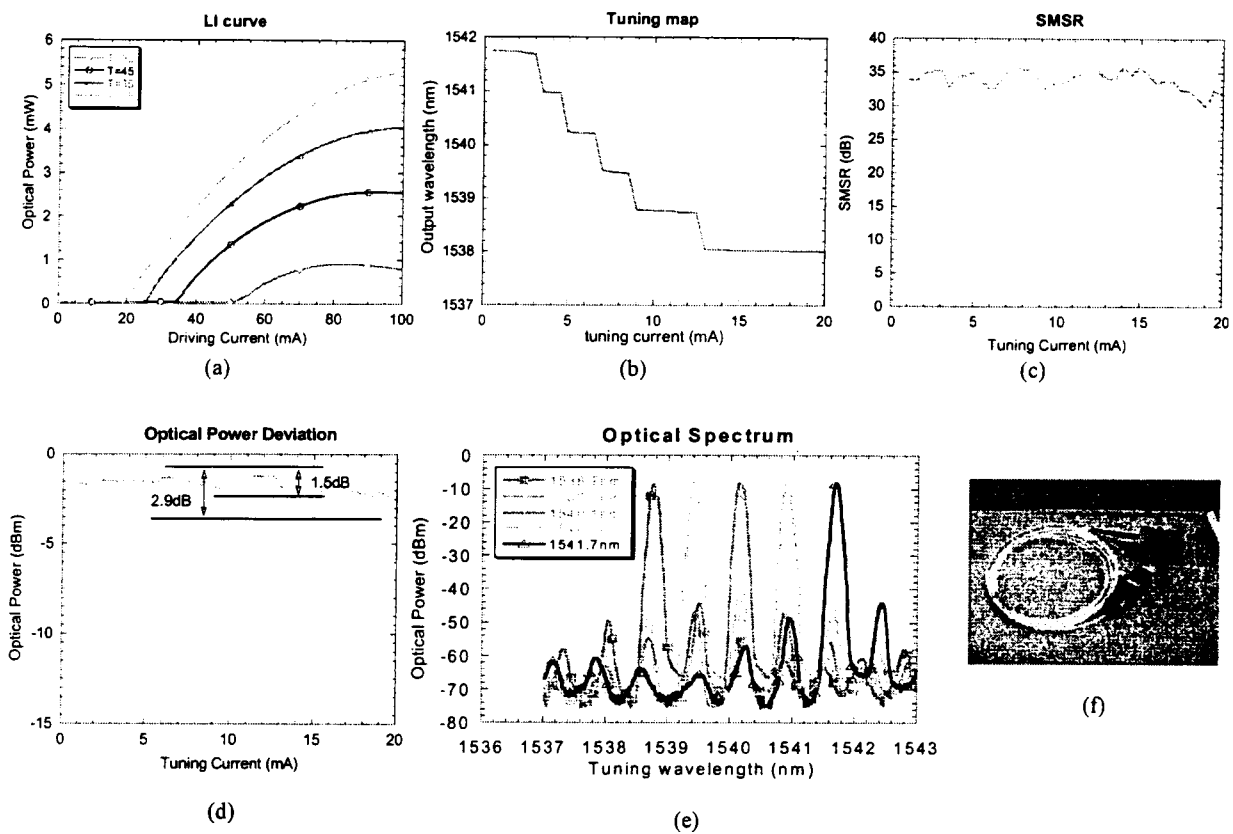


Fig. 2 (a) fiber coupled LI curve, (b) tuning curve, (c) side mode suppression ratio, (d) power deviation, (e) optical spectrum, and (f) the packaged laser module.

The circuit board consists of two parts: driver unit and control unit. In the driver unit, we use a high-speed current-output Digital-to-Analog-Converter (DAC) of 12bit resolution for *grating* current tuning, [3]. The full scale output current can be set to 20mA to satisfy the tuning map of a DBR laser (Fig.2). An ultra-stable current source was used and it can be operated in the constant power mode or constant current mode. It stabilizes the output power of the DBR laser module. The temperature controller can stably control the heat-sink temperature from 15°C to 35°C . We also use the temperature controller for fine-tuning of wavelength. All parts in the driver unit can be controlled internally or by the control unit. In the control unit, we use an EEPROM to store the tuning current of wavelength channels of the DBR laser and other initial parameters of the module. The access time of the SRAM is less than 10ns.

The DBR laser may be subject to the aging-induced wavelength drift or the drift in the operation temperature and other environment conditions. The tuning map of a DBR laser may drift due to the above problems. A wavelength monitoring unit can be included in the control unit to automatically detect the characteristics of DBR laser, like the tuning map and output power, and update the data onto the SRAM in real time though a RS232 serial port.

3. Switching experiments

The setup for measuring the wavelength switching speed is shown in Fig. 3. An optical tunable filter is used as a wavelength discriminator. The rising or falling slope of the tunable filter is tuned to the proximity of the laser output wavelengths. The discriminator gives each wavelength channel a different attenuation and a photo-detector converts the attenuation into different voltage levels. Therefore, the switching characteristics can be clearly displayed on an oscilloscope.

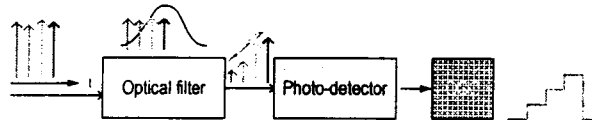


Fig. 3 A simple measurement setup for high-speed wavelength tunable transmitter.

In the experiments, the bias current of the DBR laser is set to 30mA and the output optical power is 1.5mW. The slope of the optical filter is about -10dB per nm of wavelength change. The insertion loss of the filter at center wavelength is about -2.5dB. The measured results are shown in Fig. 4. The DAC outputs a 4-level tuning current where each level corresponds to the required tuning current of a given wavelength. The period of the tuning current is 1 μ s. The trace for the output wavelength clearly indicates that the switching time is less than 0.1 μ s. Since the tuning speed of the tunable laser is typically faster than 10ns, the speed limitation is due to the tuning electronics. The tilt in the output signal for every wavelength channel results from the low-frequency cutoff of the AC-coupled photo-receiver of 1GHz bandwidth. In spite of the signal distortion, we can still observe the rise and fall time for the wavelength switching.

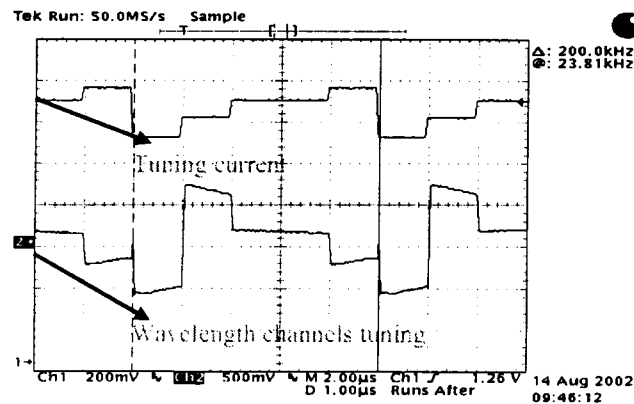


Fig. 4 The tuning characteristic of the high-speed wavelength tunable transmitter.

4. Conclusion

We have demonstrated a low-cost fast λ -tunable transmitter that can be used in OBS or OPS networks. The wavelength can be switched among DWDM channels in <100 ns.

5. Acknowledgment:

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6. References

- [1] K. V. Shrikhande, I. M. White, D. Wonglumsom, S. M. Gemelos, M. S. Rogge, Y. Fukashiro, M. Avenarius, and L. G. Kazovsky, "HORNET: a packet-over-WDM multiple access metropolitan area ring network," *IEEE JOSA in Communication*, vol. 18, no. 10, Oct 2000.
- [2] Franck Delorme, "Widely tunable 1.55- μ m lasers for wavelength division multiplexed optical fiber communications," *IEEE JOQE*. Vol. 34, no. 9, Sept. 1998.
- [3] P. D. Biernacki, M. Y. Frankel, M. E. Gingerich, and P. J. Matthews, "A high-speed mixed digital-analog circuit board for accurate control of wavelength tunable lasers for fiber-optic communications," *IEEE JOLT*., vol. 17, no. 7, July 1999.

附件三

行政院國家科學委員會專題研究計畫成果報告

智慧型寬頻光纖通信網路研究— 子計劃三：智慧型光 CDMA 網路

【第一年】計畫編號：NSC-89-2215-E-032-001
執行期限：88 年 8 月 1 日至 91 年 7 月 31 日

【第二年】計畫編號：NSC-89-2215-E-032-002
執行期限：89 年 8 月 1 日至 90 年 7 月 31 日

【第三年】計畫編號：NSC-90-2215-E-032-003
執行期限：90 年 8 月 1 日至 91 年 7 月 31 日

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摘要

【第一年】

本計畫利用基因演算法找出同時容納量最大的光碼組合，並將它實際應用於光纖分碼多重擷取技術的網路中。此外並設計一個了光纖分碼多重擷取技術網路的通訊協定，根據網路中使用者的需求來找適當的光碼組合以提高網路的效能；並應用模擬軟體來驗證我們的結果。我們分析比較四種相異質數碼系統結合方法:1.水平序列法模型 2.垂直序列法模型 3.隨機排列法模型 4.最佳加入法模型。最後使用最佳加入法模型來增加系統容量。同時參考並分析陣列波導光柵作加-取多工器的性能與研究，在波長分配多工系統上的特性，以光環式共振濾波器來達成可調式加-取多工器的效能，而在光環式共振濾波器的設計中有許多問題，例如：共振長度，耦合係數，如何將波長濾下等等,都正在進行模擬當中，此外再連接光環式共振濾波器與光延遲編碼解碼器並且分析其可行性。

關鍵詞： 光 CDMA，質數碼，分碼，多工器，波導光柵，光環式共振濾波器，分波多工，光延遲編碼解碼器

【第二年】

由於光延遲線的限制和連接容量是產生擁塞現象，且使光電 ATM 交換器形成低交換效率的問題所在，因此具有智慧型的光電 ATM 交換器之高速光骨幹網路能有更快的交換決策，可避免或減低擠塞的狀況。因此，我們提出一智慧型的壓縮決策光電 ATM 交換器架構。此架構包括訊框處理器和光資料壓縮器用來改善交換效率。此智慧型架構中，當有超過二個以上的訊框，其細胞包含相同的目的地，且競爭相同的輸出埠時，失敗者將立即被佇列且壓縮以減低擁塞的程度。因此總交換時間可以減低，同時緩衝器也可以儲存更多

到來的細胞。簡單且有效率的壓縮決策演算法將被提出。壓縮決策花費的時間和飽和性能將會被分析。從模擬的結果得知，此提出的策略藉由佇列延遲可提供更佳的性能相對於不用壓縮決策的光 ATM 交換器。

關鍵詞：非同步傳輸模式(ATM)交換器，壓縮，智慧型光電。

【第三年】

光 CDMA 是利用特殊的光碼將傳輸資料加以編碼後再傳送，如此可容納大量的使用者在同一光纖上同時傳送，故可大量地提升光纖網路的頻寬。然而受限於使用者資料的互相干擾(multi-user interference, MUI)，當同時使用者的數量增加時，錯誤率將隨之增大。本計畫以曼徹斯特編碼的方式，將原本雙極性的 Walsh 碼轉成適合光纖傳送的單極性碼，我們稱之為曼徹斯特編碼式 Walsh 碼 (Manchester Coded Walsh Code, MCWC)。該碼的定干擾量特性讓接收端得以估計出干擾量而加以消除，故可有效地降低錯誤率。此外，針對不同的網路型態，我們提出了三種接收架構：固定多工式、排序式及非排序式接收架構。分析結果顯示，三種接收架構均能大量地提昇網路頻寬，並維持良好的通信品質。尤其是前二種架構，當傳輸量越大時，其通信品質反而越好。

關鍵詞：光纖通信，光分碼多工處理(OCDMA)。

Abstract

【1st year】

We have selected the maximal simultaneous users code set by the genetic algorithm which can be used in optical CDMA network practically. We have also studied the multi-code system in optical CDMA which can provide multimedia and multi-rate services. Some multiplexing techniques of the optical multimedia CDMA system have proposed by using the prime codes with different code lengths. Moreover, we have designed a protocol for the network that can assign the code to users according the users' request. We also try to use the simulation tool to verify the correlation among the codes which are selected by genetic algorithm. Finally, we accomplished a structure using Different Code System Division Multiple Access (CSDMA) techniques with different prime codes to increase the capacity of decoder dramatically. We have used the add-drop multiplexer (ADM) in the WDM system with the arrayed waveguide grating (AWG). Many problems are testing for using the ORR (optical ring resonator) in the WDM system. For example: the relationship between the resonator length, coupling coefficients, and the system transmission rate. Besides, we will propose a new architecture to combine the ORR together with the optical delay line encoder/decoder.

Keywords : multi-code, optical CDMA, multiplexer, prime, waveguide grating, optical ring resonator, wavelength division multiplex, optical delay encoder decoder.

【2nd year】

The congestion due to the limitation of optical delay line queue and link capacity is the low switching efficiency problem in the photonic ATM switch. Therefore, the smart photonic ATM switch for high-speed optical backbone network should has some fast switching strategies so that the congestion can be

avoided or reduced with the smart photonic architecture. In this paper we propose a novel smart photonic ATM switching architecture with compression strategy. This smart architecture includes the frame processor and the optical data compressor to improve the switching efficiency. In the smart architecture, while more than two frames, which consisting of cells with the same destination, content with the same output port, the losers will be queued and compressed right away to reduce the degree of congestion. Therefore the total switching time can be reduced, meanwhile, the scarce buffer is able to store more incoming cells. In this paper, a simple and efficient compression decision algorithm is proposed. The timing of employing compression strategy and the saturated performance of proposed strategy are analyzed. Simulation results show that compared to the photonic ATM switch without compression strategy, the proposed strategy offers a much better performance in terms of queuing delay.

Keywords: Asynchronous Transfer Mode (ATM) switch, compression, smart photonic.

【3rd year】

The Optical code-division multiple access (OCDMA) is one of the techniques to effectively use the large bandwidth of optic fiber. In such systems, the data is encoded by a specific code in the transmitter, and the received signal is decoded by the same code in the receiver. The OCDMA systems allow multiple users to transmit data simultaneously. However, as the number of simultaneous users increased, the bit error rate is increased due to the increased multi-user interference (MUI). In this proposal, a new spreading code called the Manchester coded Walsh code (MCWC) is proposed. In the OCDMA systems using MCWC, the interference contributed by any other user is always a constant and thus the MUI can be easily estimated and cancelled in the receiver. The MCWC encoder and three kinds of MCWC decoders are proposed for the fixed channel networks, scheduled networks, and non-scheduled networks, respectively. The performance

analysis including the APD noise and thermal noise shows that the proposed systems can afford large number of simultaneous users with low bit error rate, especially in the fixed channel networks and scheduled networks.

Keywords: fiber-optic communication, optical CDMA.

一、緣由與目的

【第一年】

在光纖通訊系統的研究上,波長分配多工(Wavelength Division Multiplex, WDM) 技術已被熱烈地討論著; 近年來, 另一項技術是分碼多重擷取(Code Division Multiplex Access, CDMA)技術[1]-[7]無論是在無線或光纖系統中都漸漸地被討論起來;此時兩種技術若相互結合, 將在光纖通信技術更向前一步。目前, 在波長分配多工系統上, 大多是用波導光柵(Waveguide Grating)作多工器和解多工器, 但波導光柵是固定在元件上不可調的, 所以我們使用光環式共振濾波器(Optical Ring Resonator : ORR)[8]-[13]作可調式加-取多工器(Add-Drop Multiplexer : ADM) [14]-[21], 再聯接光編碼器或解碼器, 將波長分配多工系統與分碼多重擷取系統結合起來, 以提高光纖通訊的效能。

在 CDMA 系統中許多是針對提升同時容納量作研究及分析, 但卻沒有人實際將適當的碼組合選擇出來, 這在實際的系統中是很重要的, 因此我們採用一快速演算法能找出適當的碼組合, 並利用模擬軟體驗證其真實性。CDMA 技術因為具有良好之抗干擾性而且可以增加系統容量, 故成為下一代通訊系統之主流, 廣泛運用於無線及光纖通訊, 近年來更深入在多媒體傳輸上之應用, 我們發現在 CDMA 環境中不同長度之編碼可以有條件之共容, 在接收端仍可鑑別部分特定之編碼, 此即為相異質數碼之分碼多工系統及異質碼多工器之設計基礎。將針對在 CDMA 環境中, 不同長度之編碼共同形成相異質數碼系統架構, 可解決解碼容量的不足, 增加同時使用者之數量, 進而在此架構下進行最佳化編碼、異質碼多工器設計、系統容量及可能範圍之應用, 改善了傳統 CDMA 中的缺點。此外研究應用光環式共振濾波器作加-取多工器, 提昇在波域多工技術上的效能[22]-[25], 並結合光延遲解碼器, 增進在光纖分碼多重擷取技術上的性能。

【第二年】

非同步傳輸模式(ATM)是在高速封包交換技術中最有希望的交換技術[26-27]，也是滿足技術上與商業上，逐漸增加的多種在不同成份的通訊和具大寬頻的要求。很多電的交換被侷限在電交換速度，因此，這是非常重要的去研究光電或光交換技術，在未來的高速光骨幹網路。此外，ATM 網路提供多樣的服務品質(QoS)和頻寬在每一連接中[30-31],[33-35]。在此篇研究中，我們提出的智慧型壓縮決策的光電 ATM 交換器架構將討論如下的兩個基本論點：(1)如何決定壓縮決策的花費時間，(2)當所有細胞/訊框還是處於壓縮狀況中的同時，如何避免輸出連結是可用的情況。為了解決此兩問題，我們提出一簡單且有效的壓縮決定演算法(compression decision algorithm, CDA)，是根據細胞的路由資訊，由 CDA 決定是否來執行壓縮決策，因此將可達最佳化的而不會浪費連接容量。

【第三年】

由於光纖的頻寬遠大於目前電子系統所能達到的頻寬，所以如何充分利用光纖的頻寬乃是建立高速光纖網路的一大重要課題。目前除了分波多工(Wavelength-division multiplexing, WDM)技術之外，光分碼多工處理(Optical code-division multiple access, OCDMA) 技術亦是漸獲重視的一個領域[36-44][48]。OCDMA 系統大致可分為同步與非同步兩種系統，其中同步系統雖然需要較為複雜的同步技術，但是它能提供較大的網路規模以及較佳的品質[38]，因此是極具研究價值的一個主題。雖然文獻上已有多種光碼以及相關的架構被提出，但在性能上仍有不小的空間有待提升。因此，本計畫嘗試將傳統的 Walsh code 修改成適合光纖傳輸的 Manchester coded Walsh code(MCWC)以做為新的光碼並提出其對應的架構，期望透過數學分析的方式能證實其具有更優越的性能。

二、 結果與討論

【第一年】

本子計畫目的針對利用質數碼同時容納量的特性設計出智慧型異質碼多工器，並探討其有關之技術。並且針對光環式共振濾波器的特性，探討其有關之技術及延伸。

若用傳統的方法來找出適合共存的碼須耗費許多的資源和時間，以修正式質數碼系統為例，當質數碼為 P 、最大同時容納量為 P_{max} 且 DT 為每次決定所選擇的碼是否能共存所須計算的時間，則整體所須的計算時間為

$$\sum_{K=P}^{P_{max}} C_K^{P^2} \times DT$$

由上式可知，當質數碼變大時所須的時間會呈現指數型式上昇，因此須利用更好的搜尋方式以找尋最大同時容納量的碼。近年來，基因演算法（Genetic algorithm：GA）已廣泛的被應用於解決最佳化的問題，如 multiprocessor scheduling, topology design and bandwidth allocation 和 channel allocation 等問題，也得到很好的結果，因此我們採用基因演算法來降低搜尋所須要的時間，並找出適當的碼。

圖 1 是一般基因演算法的流程圖，透過一代代複製、交配及突變的過程，我們能找出最好的解，目前我們利用 0 和 1 所組成的字串來代表碼的選擇與否，且交配及突變的過程如圖 2 和圖 3 所示，圖 4 是我們實際所得的結果和其他論文結果比較，我們可發現我們的結果大於其他論文的結果，且利用此方法我們可以直接的得到碼組合。

同時我們採用了一種新式的碼區別多重擷取技術—相異質數碼系統區別多重擷取技術及其設計方法—來提昇整合容量，如圖 5 所示。在光纖波域分工系統中，陣列波導光柵已被頻繁得使用作多工和解多工器，再加上迴路

的原理形成陣列波導光柵的加-降多工器(ADM) 如圖 6 所示。參考陣列波導光柵的技術，我們利用光環式共振濾波器來達成加-降多工器 如圖 7 所示，在光纖通訊系統中的功能，而且還具有可調性等，這是新型的架構。光環式共振濾波器的特性是可在選擇端濾下想要的波形，其它剩餘的波形會在傳輸端濾出，利用此來將波長分工，而在波形上的時間軸與頻率軸與波長軸之間必須相互映射的一個重要問題。以元件的觀點，來考慮光環式共振濾波器中的參數:耦合係數，共振腔長度等及可調範圍，並協調參數之間的影響，形成合適的元件。而濾下的波長也有範圍的限制。

【第二年】

1. 系統架構

此智慧型的壓縮決策光電 ATM 交換器架構，見圖 8 所示。作業流程說明如下：當細胞從輸入連接而來，將進入訊框運算器(frame processor, FP)來偵測此細胞的目的地。有相同目的地輸出的一堆細胞會被貫聯一起成一訊框，如圖 9 所示。FP 會有起始脈波(Fs)和終端脈波(Fe)，分別處於開始端和終止端。為了區別資料信號的起始/終端，這兩個脈波的光功率需大於訊框資料。壓縮率(compression ratio, CA) 定義為壓縮訊框的長度與原訊框長度之比值。如果訊框不能被立即的交換，將會被壓縮和緩衝。每一個被佇列訊框的正確佇列延遲是由競爭控制器(contention controller, CC)來決定是那一個訊框將會被交換至光交換器(optical switch, OS)。同時 CC 會更進一步的偵測是否會跟其他的訊框相碰撞，若有，則失敗者會被光資料壓縮器(optical data compressor, ODC)，先壓縮同時被佇列在光延遲佇列(optical delay queue, ODQ)。否則，訊框將直接至路由交換器(routing switch, RS)。一但訊框通過 OS，RS 將轉換此訊框到正確的輸出埠，而細胞運算器(cell processor, CP)將訊框還原為原細胞並且移除起始/終端脈波。如圖 10 所示。

2. 壓縮決策演算法

Compression Decision Algorithm (CDA)

- 1) Let $j = D_i^k$;
- 2) If $(LAT_j \leq A_i^k)$ {
- 3) If $((S_{i-1}^k + N_{i-1}^k) \leq A_i^k)$ {
- 4) $S_i^k = A_i^k$; /
- 5) $Q_i^k = 0$;
- 6) $LAT_j = S_i^k + O_i^k$;
- 7) } Else {
- 8) $S_i^k = S_{i-1}^k + N_{i-1}^k$;
- 9) $Q_i^k = S_i^k - A_i^k$;
- 10) $LAT_j = S_i^k + N_i^k$;
- 11) }
- 12) } Else {
- 13) If $((S_{i-1}^k + N_{i-1}^k) \leq LAT_j)$ {
- 14) $S_i^k = LAT_j$;
- 15) $Q_i^k = S_i^k - A_i^k$;
- 16) $LAT_j = S_i^k + N_i^k$;
- 17) } Else {
- 18) $S_i^k = S_{i-1}^k + N_{i-1}^k$;
- 19) $Q_i^k = S_i^k - A_i^k$;
- 20) $LAT_j = S_i^k + N_i^k$;
- 21) }
- 22) }

3. 模擬結果

圖 11 說明在所提出的決策下，在不同壓縮比率和網路負荷時，當交換器的大小($N=2$)，訊框的大小($L=50$)時的平均佇列延遲(average queueing delay, AQD)。CR=1.0 為傳統的交換方式(即沒有壓縮)，由此圖知 AQD 的大小取決於 CR 的大小。而且在沒有壓縮情況下，只要交通負荷為 0.55，AQD 就趨近無窮大。如果所有的輸入資料將壓縮比率都為 60%，系統的產量會有明顯的改進。

圖 12 說明在不同的交換器大小(N)和不同的網路負荷(Λ)下，AQD 和 CR 的關係。直覺的，越大的交換器越容易發生嚴重的目的地競爭。

圖 13 證明 CR=0.8 在不同的交換器大小，不同的網路負荷，和不同的訊框大小($L=10, 50, 100$ 細胞)下的 AQD。很清楚地，如果大的訊框被佇列著，將影響往後到達的訊框也被佇列，因此 AQD 明顯的增加。當網路負荷飽和時，我們可以得到幾乎一致的 AQD，這是因為在此情況下幾乎所有的訊框都被佇列著，而不管是有多大的訊框。然而不同的大小訊框，在低網路負荷下，AQD 明顯地會各不相同，小的訊框會有小的 AQD。

【第三年】

1. MCWC 之特性

基本的 Manchester Coded Walsh Code (MCWC)乃是將 Walsh code 以 Manchester code 的方式呈現，亦即以'01'取代原 Walsh code 的 chip '0'，以'10'取代原 Walsh code 的 chip '1'。若將基本的 MCWC 中的每個 chip 重複 2^L 次，還可得到新的 MCWC，其中長度為 $2K$ 的 MCWC 共有 $(2K-1)$ 個，它們的特性如下：

- (1) 每個碼的權重(weight)為 K 。
- (2) 相同碼的內積等於 K ，反相碼的內積等於 0。
- (3) 不同碼的內積等於 $K/2$ 。
- (4) 由於 MCWC 採互補式編碼，亦即 bit '0' 和 bit '1' 使用的是反相碼，所以 N 個同時使用者的總權重等於 NK ，每個使用者受到其餘 $(N-1)$ 個使用者的總干擾量為 $K(N-1)/2$ ，因此很容易由接收信號中預測出干擾量。

2.系統架構

2.1 MCWC 編碼器

MCWC 編碼器的架構如圖 14 所示，採互補式編碼，當 data bit $b=1$ 時，由下路編出 MCWC 碼，當 data bit $b=0$ 時，由上路編出反相 MCWC 碼。因此不論所欲傳送的位元是 0 或 1，其輸出碼的權重均能保持固定，並且根據特性 (3)，其對其他通道的干擾量亦能保持固定。若是所欲傳送資料的目的地必須是可調的，則可使用圖 15 所示的可調式 MCWC 編碼器，圖中的 K 個光開關之動作為：將 MCWC 視為 K 個曼徹斯特碎元(chip)，每一碎元對應一個光開關，若為 '10' 則光開關為 '交叉'，若為 '01' 則光開關為 '平行'。例如，當 MCWC=01011010 時，由上而下的 4 個光開關應被調至 "平行、平行、交叉、交叉"。

2.2 MCWC 解碼器

本計畫針對不同的網路應用，提出了下列三種 MCWC 解碼器架構。

第一型：固定多工式

固定多工式的系統如圖 16 所示，其通道數目為固定，而且其發射接收配對為固定，因此各通道所受到的干擾量亦為固定。此種系統的解碼器架構如圖 17 所示，系統中使用較簡單的固定式光解碼器及光限制器 [39-40][42][44-45]，而且最後之電子決策電路(threshold circuit)的臨界值亦可用固定值，是三種 MCWC 解碼器中最簡單的一種架構。

第二型：排序網路式

另一種網路架構如圖 18 所示，它是一種星型分佈網路，是屬於廣播-選取型(broadcast-and-select) 的網路。在這種網路中，最有效率的方式是使用排序方式[46]，在資料傳送前，預先排出下一時段(time-slot)有哪些使用者使用哪些通道，如此不僅可避免資料發生衝撞，也可事先得知下一時段將受到的干擾量。這種排序網路的解碼器架構如圖 19 所示，由於在資料傳輸前已由排序程序獲得所欲接收的碼及同時使用量 N ，所以其光解碼器及光限制器以及電子決策電路必須是可調式，而且必須在每一資料接收時段之前由控制器調整好。圖中的可調式相關器如圖 20 所示，至於可調式光限制器，由於現在仍無此類成品，我們設計了一種如圖 21 所示的量化型的可調式光限制器架構，該圖可調至 15 種臨界值，其特性示於圖 22 中。

第三型：非排序網路式

非排序網路式解碼器架構如圖 23 所示，由於未經排序，事先無法得知干擾量，所以該接收器必須自行估測同時使用量 N ，因此它比排序網路式解碼器多了一調路徑以估測同時使用量 N ，其估計原理乃是根據 MCWC 碼的特性

(4)。控制器根據估計結果，再調整光限制器以及電子決策電路。

3. 分析結果

在分析上述的 MCWC 光 CDMA 系統時，我們除了考慮 MCWC 的 MUI 之外，還將 APD 雜訊及熱雜訊一併列入考慮[47-18]，分析結果分述如下。

3.1 第一/二型接收器

本型系統之性能如圖 24 所示，圖中並畫出了不含光限制器的修正式質數碼(Modified prime code)系統、雙光限制器的修正式質數碼系統[40]、以及完全差距碼(Perfect difference code, PDC)系統[43]的性能以做為比較。不同於其他的光 CDMA 系統，當同時使用量 N 越大時，MCWC 系統的錯誤率反而越低，這是因為光限制器的輸入功率有兩種可能：當 $b=0$ 時為純干擾量，當 $b=1$ 時為干擾量加上信號量，經過光限制器後，輸出的功率變成：當 $b=0$ 時為 0，當 $b=1$ 時為干擾量加上信號量，所以當 N 越大時， $b=0$ 和 $b=1$ 的功率差距越大，造成錯誤率隨之降低。由圖中之數值可看出，當 N 夠大時，MCWC 系統只需較低的接收功率即能達到遠低於其它系統的錯誤率。

3.2 第三型接收器

本型系統之性能如圖 25 所示，圖中仍與雙光限制器的修正式質數碼系統[41]、以及 PDC 系統做比較。當 N 較小時，錯誤率仍隨著 N 之增大而下降，猶如第一/二型那樣。但是當 N 不小時，由於第三型接收器在估計同時使用量 N 時所出現的誤差，會造成錯誤率隨著 N 之增大而上升。若和 PDC 系統比較，雖然原始的第三型 MCWC 系統性能略差於 PDC 系統，但若在圖 23 的控制器

中加入一個簡單的自移式平均器(moving average)，將連續 W 位元所估測得的 N 值加以平均，如此將可顯著地降低估測 N 值的錯誤方差，進而降低錯誤率。由圖 25 可看出，只需加入 $W=2$ 的自移式平均器，MCWC 系統的性能即可超過 PDC 系統。

三、 已完成之具體成果

【第一年】

論文

1. Yang-Han Lee, Rainfield Y. Yen, Kuo-Ting Lin, Rong-Hou Wu, Chung-Shuo Yang, "Design of Optical Multi-layer Prime Code using Optimal Design Algorithm," *Journal of Optical Communications*, vol.21 pp.2-9, Feb 2000.
2. M. H. Chen and Y. H. Lee, "A Genetic Algorithm for the Central Code Controller of the FO-CDMA Network", will be published in the *Microwave and Optical Technology Letters* at October 5, 1999
3. Y. H. Lee, S. T. Sheu, M. H. Chen, "Application of Genetic Algorithm for Optimal Simultaneous Code Set Selection in Optical CDMA", IASTED International Conference on Modelling and Simulation (MS'99), Philadelphia, PA (Cherry Hill, New Jersey) USA, May 5-8, 1999
4. Yang-Han Lee, Meng-Hong Chen and Jen-Hung Huang, "Maximal Simultaneous Users Code Set Selection in FO-CDMA using Genetic Algorithm", 1999 International Conference on Computational Electromagnetics and Its Applications (ICCEA'99), Beijing, China, Nov. 1-4, 1999
5. Yang-Han Lee, Rong-Hou Wu, Chorng-Hann Niou and Meng-Hong Chen "Performance analysis of synchronous different codes system division multiple access techniques with prime code in optical spread spectrum communication"(Accept) ONDM'99, Paris, February 8-9, 1999
6. Yang-Han Lee and Chorng-Hann Niou "Novel technique to increase simultaneous capacity of optical multi-media CDMA system," submitted to *Electronics Letters* (11/21/1998)
7. Yang-Han Lee and Chorng-Hann Niou "Smart optical multiplexer for optical

multi-media CDMA,” submitted to Electronics Letters (11/21/1998)

專利

- 1.“高解碼容量的階層式質數碼之解碼器之延遲邏輯架構解碼器”(完成申請)
- 2.發明人：李揚漢, 王智源. 發明創作名稱：“光電碼域交換器”. (發明第 138641 號：中華民國).
3. 發明人: 牛崇翰, 李揚漢, 陳孟宏, 王智源, 魏國純 發明創作名稱：“智慧型相異質數碼多工傳輸系統”(發明第 136989 號：中華民國)

【第二年】

- (1) Shiann-Tsong Sheu, Yang-Han Lee, and Chih-Chiang Wu “ A Smart Photonic ATM Switch Architecture with Compression Strategy”, *Journal of Lightwave Technology*, Vol. 19, No. 1, Jan. 2001.

【第三年】

論文

1. L. L. Jau, Y. H. Lee, ”Synchronous Optical CDMA system using Manchester Coded” Revised and Resubmitted to IEEE J. Lightwave Technology.
2. R. H. Wu, Y. H. Lee, ”Study of Optical Multiplexing Techniques with Different Codes System in Optical Multimedia CDMA Network,” Accepted by J. Opt. Comm.

3. R. H. Wu, Y. H. Lee, "Performance Analysis of Serial Attachment Modified Prime Sequence Code with Multi-layer /Parallel Decoder Architecture," J. Opt. Comm., vol.22, pp.184-190, Dec. 2001.
4. Y. H. Lee, et. al, "Optical Code-Division Switching Network Based on Simultaneous Capacity Prime Code," Microwave and Optical Technology Letters, vol.31, No.5, pp.336-340, Dec. 5, 2001.
5. R. H. Wu, Y. H. Lee, "Design of a Photonic Code Division Switch Based on Simultaneous Capacity Prime Code," Microwave and Optical Technology Letters, March 2001.

四、 參考文獻

- [1] C. L. Lu, et. al., "CORD — a WDMA optical network: subcarrier-based signaling and control scheme," *IEEE Photon Technol. Lett.*, vol. 7, no. 5, pp. 555-557, May 1995.
- [2] L. Tancevski, et. al., "Hybrid wave-length hopping/time spreading scheme for use in massive optical network with increased security," *IEEE J. Lightwave Technol.*, vol. 12, no. 6, pp.2636-2647, Dec. 1996.
- [3] L. Tancevski, I. Andonvic, M. Tur and J. Budin, "Hybrid wavelength hopping/time spreading code division multiple access systems," *IEE Proc.-Optoelectron.*, vol. 143, no. 3, pp. 161-166, June 1996.
- [4] G. C. Yang and W. C. Kwong, "Performance comparison of multiwavelength CDMA and WDMA + CDMA for fiber-optic networks," *IEEE Trans. Commun.*, vol. 45, no. 11, pp. 1426-1434, Nov. 1997.
- [5] J. M. Simmons, et. al., "Optical Crossconnects of Reduced Complexity for WDM Networks with Bidirectional Symmetry," *Photonics Technol. Letters*, vol. 10, no. 6, pp. 819-821, 1998.
- [6] M. Ibsen, et. al., "Sinc-Sampled Fiber Bragg Gratings for Identical Multiple Wavelength Operation," *Photonics Technol. Letters*, vol. 10, no. 6, pp. 842-844, 1998.
- [7] J. L. Jackel, et. al., "Acousto-optic Tunable Filters (AOTF's) for Multiwavelength Optical Cross- Connects: Crosstalk Considerations," *J. Lightwave Technol.*, vol. 14, no. 6, pp. 1056-1066, 1996.
- [8] G. C. Yang, "Some New Families of Optical Orthogonal Codes for Code-Division Multiple-Access Fiber-Optic Network," *IEE Proc. Commun.*, vol. 142, no. 6, pp. 363-368, Dec. 1995.
- [9] W. C. Kwong, et. al., "2ⁿ prime-sequence codes and coding architecture for optical code-division multiple access," *IEEE Trans. Commun.*, vol. 44, no. 9, pp. 1152-1162, Sep. 1996.
- [10] J. G. Zhang, "Shortened prime code and their cost-effective encoders for use in all optical CDMA networks," *IEICE Trans. Commun.*, vol. E79-B, no. 2, pp. 198-201, Feb. 1996.

- [11] C. S. Hsu, et. al., "Performance analysis of unslotted fiber optic code division multiple access(CDMA) packet network," *IEEE Trans. Commun.*, vol. 45, no. 8, pp. 978-987, Aug. 1997.
- [12] C. S. Hsu, et. al., "Performance analysis of slotted fiber optic code division multiple access(CDMA) packet network," *IEEE Trans. Commun.*, vol. 45, no. 7, pp. 819-828, Jul. 1997.
- [13] C. L. Ho, et. al., "Performance Analysis of CDMA Optical Communication Systems with Photodiodes," *J. Lightwave Technol.*, vol. 12, no. 6, pp. 1062-1072, 1994.
- [14] S. Tao, et. al., "Design of Self-Routing Frequency Division Multiplex (SF-FDM) Network Using an Optical Ring Filter with or without Gain as a Router," *IEEE J. Lightwave Technol.*, vol. 13, no. 11, pp. 2168-2182, 1995.
- [15] K. Takiguchi, et. al., "Bias of an Optical Passive Ring-Resonator Gyro Caused by the Misalignment of the Polarization Axis in the Polarization-Maintaining Fiber Resonator," *J. Lightwave Technol.*, vol. 10, no. 4, pp. 514-522, 1992.
- [16] Y. H. Ja, "Optical Fiber Filter Comprising a Single-Coupler Fiber Ring (or Loop) and a Double-Coupler Fiber Mirror," *J. Lightwave Technol.*, vol. 9, no. 8, pp. 964-974, 1991.
- [17] F. Sanchez, "Matrix Algebra for All-Fiber Optical Resonators," *J. Lightwave Technol.*, vol. 9, no. 3, pp. 838-844, 1991.
- [18] K. Oda, et. al., "A Wide-FSR Waveguide Double-Ring Resonator for Optical FDM Transmission Systems," *J. Lightwave Technol.*, vol. 9, no. 6, pp. 728-736, 1991.
- [19] N. Q. Ngo, et. al., "Novel Realization of Monotonic Butterworth type Lowpass, Highpass, and Bandpass optical Filters Using Phase-Modulated Fiber-Optics Interferometers and Ring Resonators," *J. Lightwave Technol.*, vol. 12, no. 5, pp. 827-841, 1994.
- [20] B. E. Little, et. al., "Wavelength Switching and Routing Using Absorption and Resonance," *Photonics Technol. Letters*, vol. 10, no. 6, pp. 816-818, 1998.
- [21] D. A. Smith, et. al., "Evolution of the Acousto-optic Wavelength Routing Switch" *J. Lightwave Technol.*, vol. 14, no. 6, pp. 1005-1019, 1996.

- [22] Y. Tachikawa, et. al., "Arrayed- Waveguide Grating Add-Drop Multiplexer with Loop-Back Optical Paths," *Electronics Letters*, vol. 29, no. 24, pp. 2133-2134, 25th November 1993.
- [23] L. H. Spiekman, et.al., "Design and Realization of Polarization Independent Phased Array Wavelength Demultiplexers using Different Array Orders for TE and TM," *J. Lightwave Technol.*, vol. 14, no. 6, pp. 991-995, 1996.
- [24] H. H. Yaffe, et. al., "Resonator Couplers Acting as Add-Drop Filters Made with Silica-on-Silicon Waveguide Technology," *J. Lightwave Technol.*, vol. 12, no. 6, pp. 1010-1022, 1994.
- [25] Y. Tachikawa, et. al., "Arrayed- Waveguide Grating Multiplexer with Loop-Back Optical Paths and Its Applications," *J. Lightwave Technol.*, vol. 14, no. 6, pp. 977-984, 1996.
- [26] CCITT, "B-ISDN ATM Cell Transfer Performance", Draft Recommendation I35B, *IEEE Communications Magazine*, vol. 30, no. 4, April 1992.
- [27] J-Y. L and Boudex, "The Asynchronous Transfer Mode : a Tutorial", *Computer Networks and ISDN Systems*, Vol. 24, pp. 279-309, 1992.
- [28] W. T. Chen, H. J. Liu and Y. T. Tsai, "High-Throughput Cell Scheduling for Broadband Switching Systems", *IEEE Journal of Selected Areas in Communications*, vol. 9, no. 9, pp. 1510-1523, December 1991.
- [29] Y. Choi, et. al., "Performance Study of the Multi-Cell Transfer Scheme in the Optical Backbone Network," *IEICE Trans. Commun.*, vol. E78-B, no. 5, pp. 729-735, 1995.
- [30] Y. Choi, et. al., "Photonic ATM Switch Architecture for WDM Optical Networks", *IEICE Trans. Commun.*, vol. E78-B, no. 9, pp. 1333-1335, 1995.
- [31] J.W., Lockwood, et. al., "Scalable Optoelectronic ATM Networks: The IPOINTE Fully Functional Testbed", *Journal of Lightwave Technology*, vol. 13, no. 6, pp. 1093-1103, 1995.
- [32] M. Prycker, "Asynchronous Transfer Mode : Solution for Broadband ISDN", Ellis Horwood, 1993.
- [33] Y. Shimazu, et. al., "Ultrafast Photonic ATM Switch with Optical Output Buffers", *Journal of Lightwave Technology*, vol. 10, no. 2, pp. 265-272, 1992.
- [34] W. D. Zhong, et. al. "Terahertz : A Modular and Expandable Terabit/Second

- Hierachically Miltiplexing Photonic ATM Switch Architecture”, *Journal of Lightwave Technology*, vol.12, no. 7, pp. 1307-1315, 1992.
- [35] W. D. Zhong, et. al., ”Optical Bit-Interleave Multiplexing ATM Switch”, *Electronic Letter*, vol. 31, no. 12, pp. 995-997, 1995.
- [36] P. R. Prucnal, M. A. Santoro, and T. R. Fan, “Spread spectrum fiber-optic local area network using optical processing,” *J. Lightwave Technol.*, vol. LT-4, pp. 547-554, May 1986.
- [37] J.A. Salehi, “Code division multiple-access techniques in optical fiber networks—Part I: Fundamental principles,” *IEEE Trans. Commun.*, vol. 37, pp. 824-833, Aug. 1989.
- [38] W. C. Kwong, P. A. Perrier, and P. R. Prucnal, “Performance comparison of asynchronous and synchronous code-division multiple-access techniques for fiber-optic local area networks,” *IEEE Trans. Commun.*, vol. 39, pp. 1625-1634, Nov. 1991.
- [39] J. H. Wu and J. Wu, “Synchronous fiber-optic CDMA using hard-limiter and BCH codes,” *J. Lightwave Technol.*, vol. 13, pp. 1169-1176, June 1995.
- [40] T. Ohtsuki, K. Sato, I. Sasase, and S. Mori, “Direct-detection optical synchronous CDMA systems with double optical hard-limiters using modified prime sequence codes,” *J. Select. Areas Commun.*, vol. 14, pp. 1879-1887, Dec. 1996.
- [41] H. M. H. Shalaby, “Synchronous fiber-optical CDMA systems with interference estimators,” *J. Lightwave Technol.*, vol. 17, pp. 2268-2275, Nov. 1999.
- [42] C. L. Lin and J. Wu, “A synchronous fiber-optic CDMA system using adaptive optical hardlimiter,” *J. Lightwave Technol.*, vol. 16, pp. 1393-1403, Aug. 1998.
- [43] C. S. Weng and J. Wu, “Perfect difference codes for synchronous fiber-optic CDMA communication systems,” *J. Lightwave Technol.*, vol. 19, pp. 186-194, Feb. 2001.
- [44] J. J. Chen and G. C. Yang, “CDMA fiber-optic systems with optical hard limiters,” *J. Lightwave Technol.*, vol. 19, pp. 950-958, July 2001.
- [45] L. Brzozowski and E. H. Sargent, “All-optical analog-to-digital converters,

hardlimiters, and logic gates,” *J. Lightwave Technol.*, vol. 19, pp. 114-119, Jan. 2001.

- [46] E. Modiano and R. Barry, “A novel medium access control protocol for WDM-based LAN’s and access networks using a master/slave scheduler,” *J. Lightwave Technol.*, vol. 18, pp. 461-468, Apr. 2000.
- [47] J. B. Abshire, “Performance of OOK and lower-order PPM modulations in optical communications when using APD-based receivers,” *IEEE Trans. Commun.*, vol. 32, pp. 1140-1143, Oct. 1984.
- [48] H. M. Kwon, “Optical orthogonal code-division multiple-access system—part I: APD noise and thermal noise,” *IEEE Trans. Commun.*, vol. 42, pp.2470-2479, July. 1994.

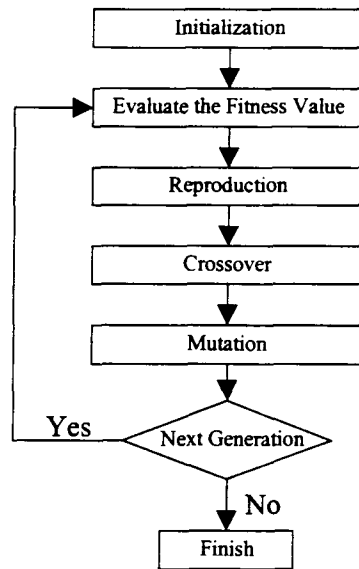


圖 1 基因演算法

<u>Before Crossover</u>		<u>After Crossover</u>
String1: <u>1</u> 1 <u>0</u> <u>0</u> <u>0</u> <u>0</u> <u>1</u> <u>0</u> <u>0</u>	$\Rightarrow$	<u>0</u> 1 <u>0</u> <u>0</u> <u>0</u> <u>0</u> <u>1</u> <u>0</u> <u>0</u>
Mask: <u>1</u> 0 <u>1</u> <u>1</u> <u>0</u> <u>1</u> <u>1</u> <u>0</u> <u>1</u>		<u>1</u> 0 <u>1</u> <u>1</u> <u>0</u> <u>1</u> <u>1</u> <u>0</u> <u>1</u>
String2: <u>0</u> 1 <u>0</u> <u>0</u> <u>0</u> <u>0</u> <u>1</u> <u>1</u> <u>0</u>		<u>1</u> 1 <u>0</u> <u>0</u> <u>0</u> <u>0</u> <u>1</u> <u>1</u> <u>0</u>

圖 2 交配

<u>Before Mutation</u>		<u>After Mutation</u>
String1: 11000 <u>0</u> 100	$\Rightarrow$	11000 <u>1</u> 100

圖 3 突變

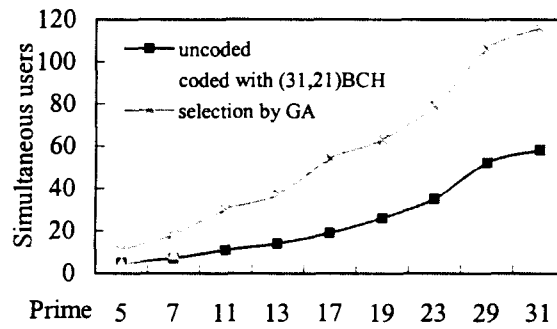


圖 4 使用基因演算法所找出的最大同時容納量和其他論文的比較

Different Optical Code Division Multiple Access System (DOCDMAS)

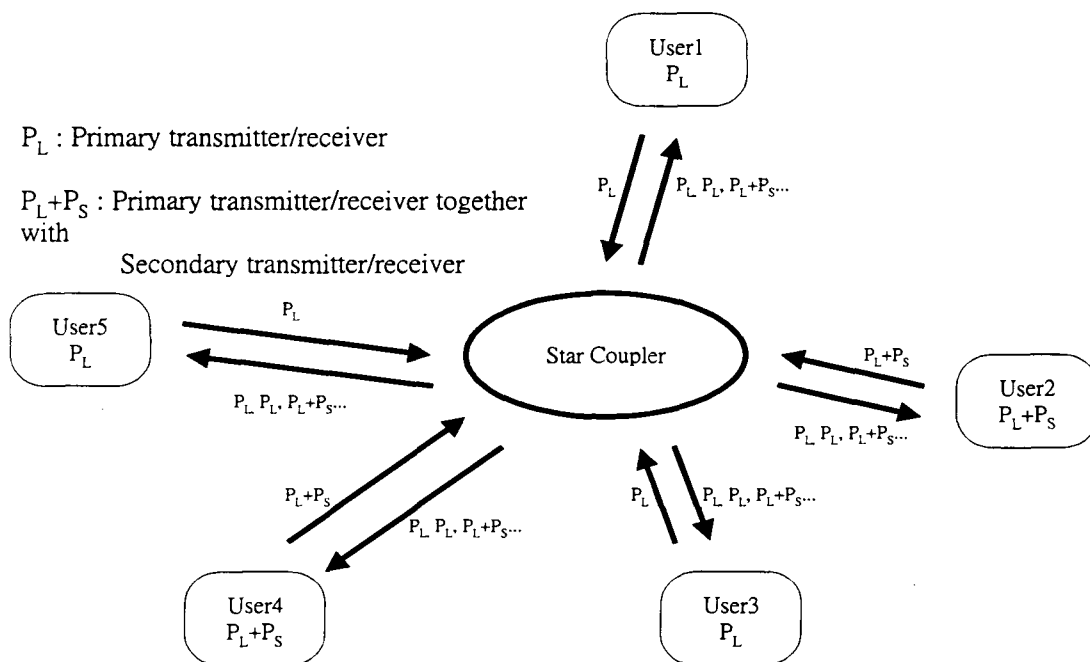


圖5 異質碼系示意圖

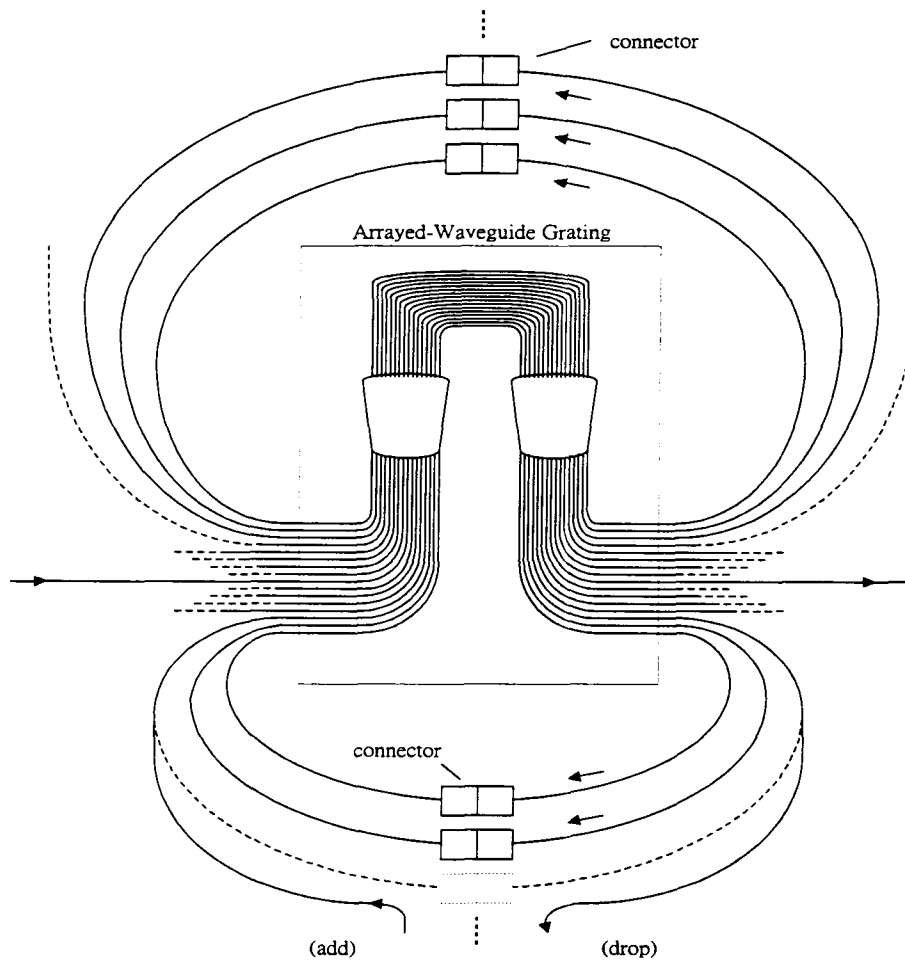


圖 6. 16*16 陣列波導光柵之 15 個通道的加-取多工器

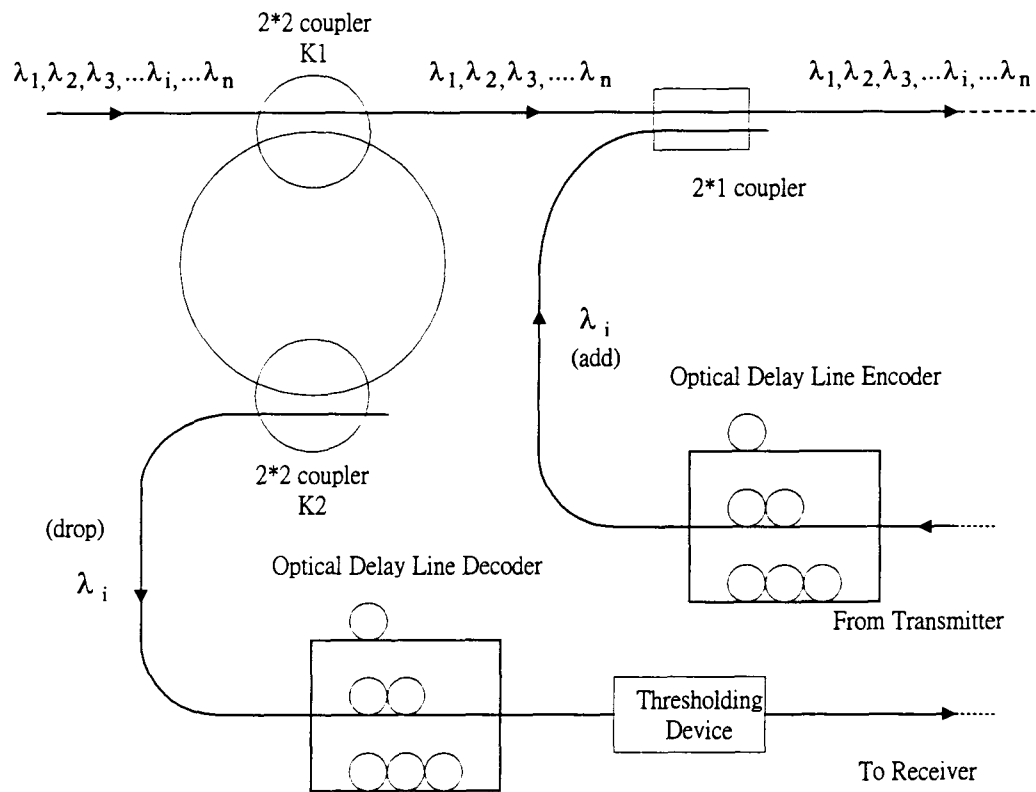


圖 7. 光環式共振加-取多工器與光延遲線編碼/解碼器

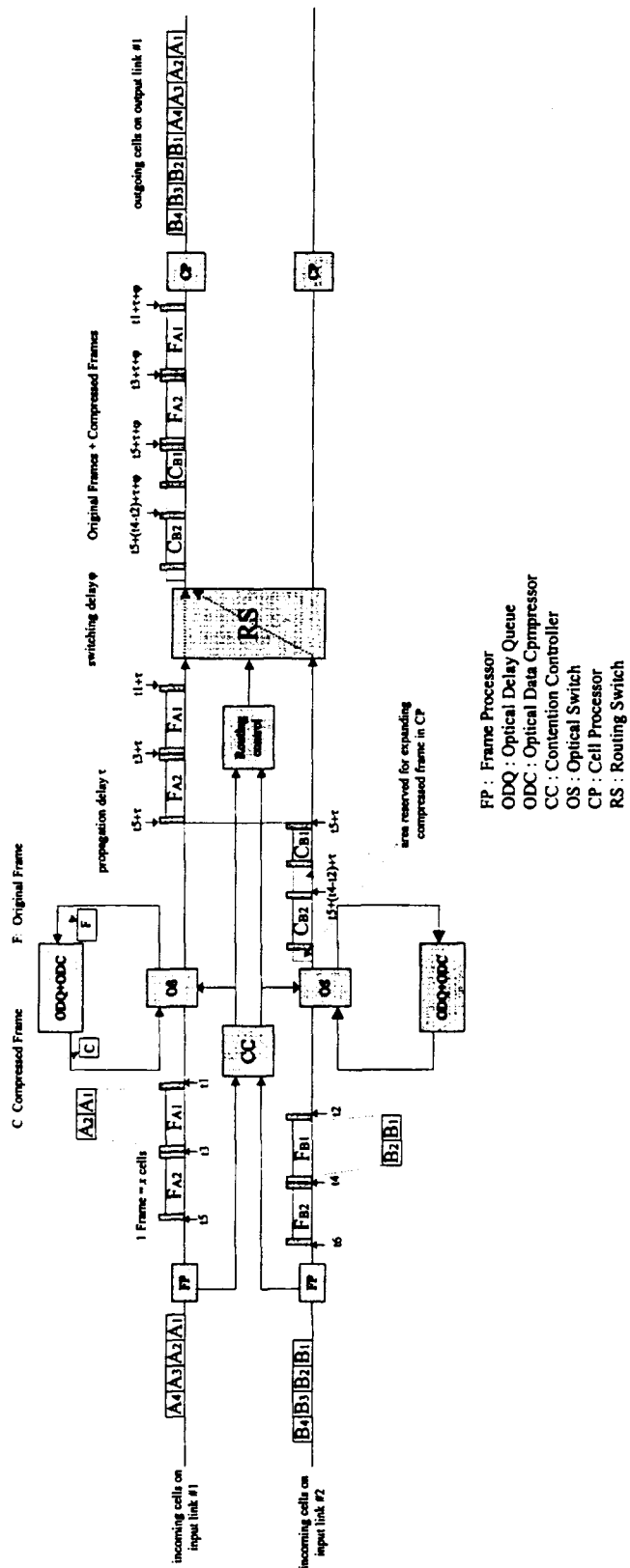


圖 8 智慧型的壓縮決策光電 ATM 交換器之架構

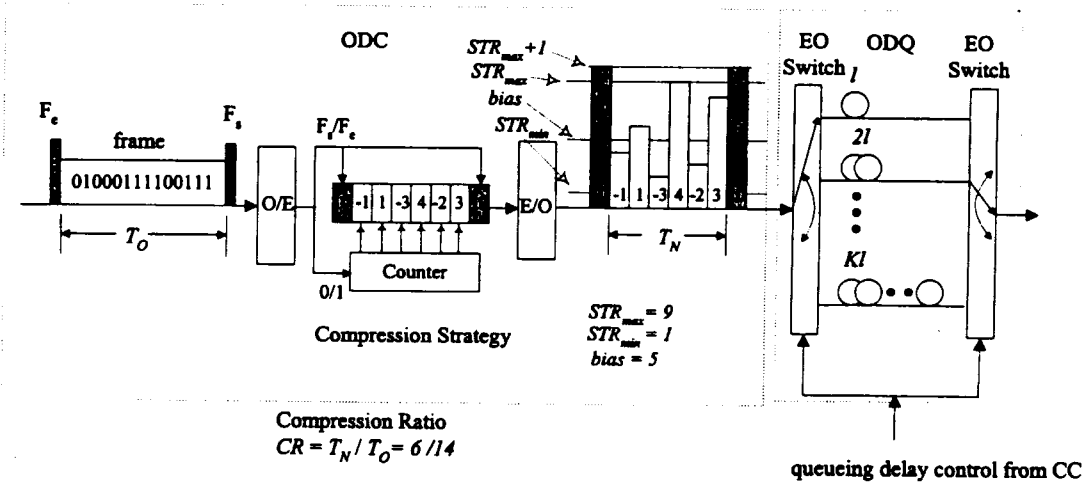


圖 9 智慧型光電 ATM 交換器之 ODQ+ODC 架構

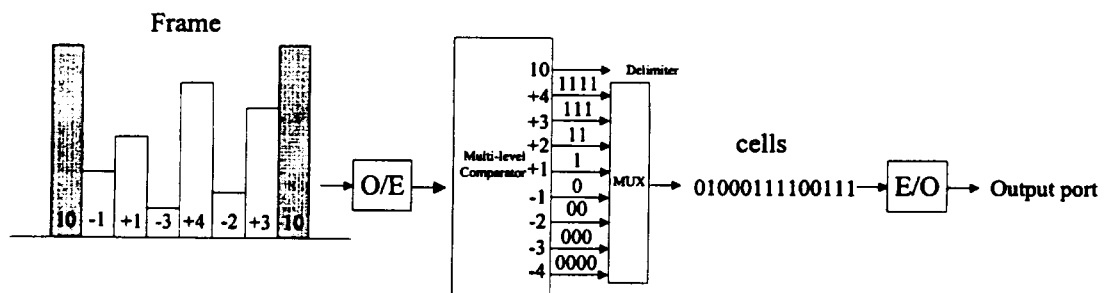


圖 10. 細胞運算器之架構

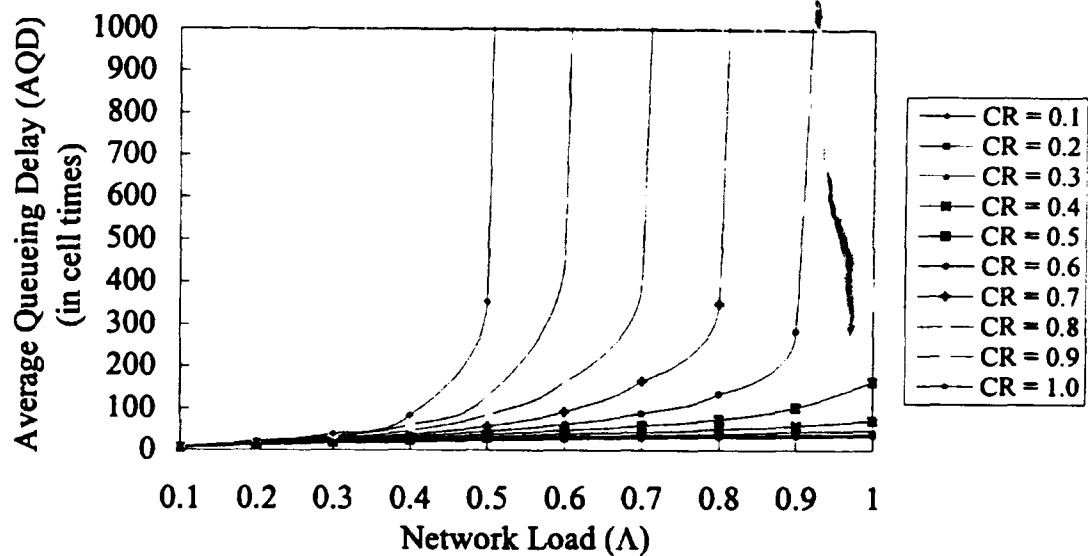


圖 11. 在不同壓縮比率(CR)和網路負荷(Λ)時的平均佇列延遲(AQD)

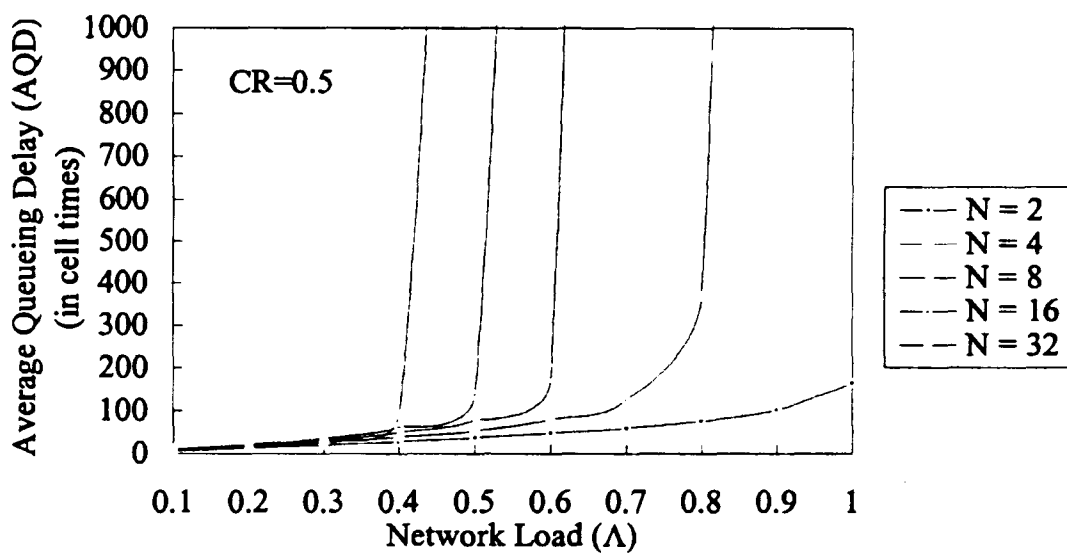


圖 12. 在不同的交換器大小(N)和不同的網路負荷(Λ)下的平均佇列延遲(AQD)

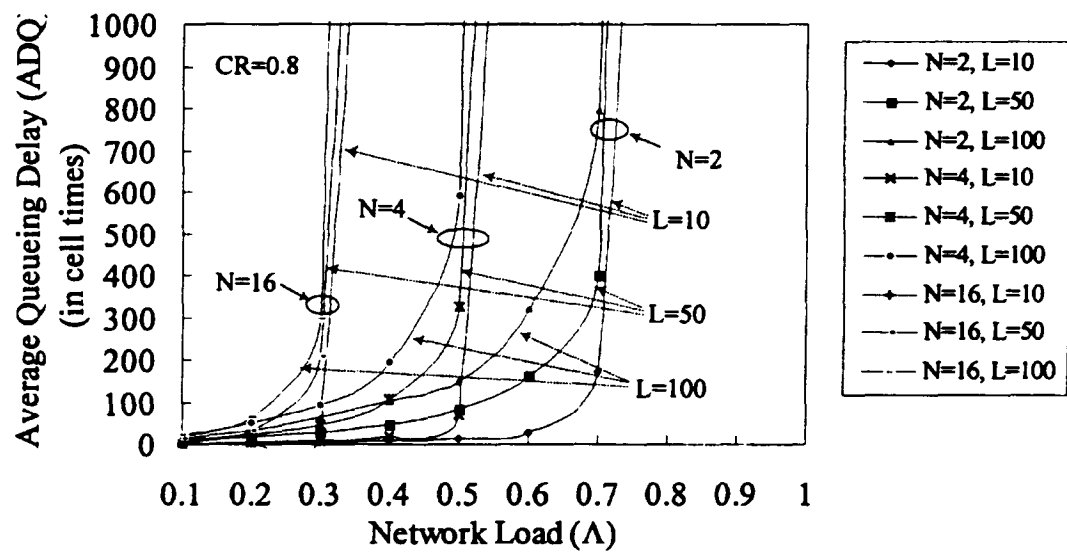


圖 13.不同的交換器大小(N)和不同的訊框大小(L)之 AQD 的比較

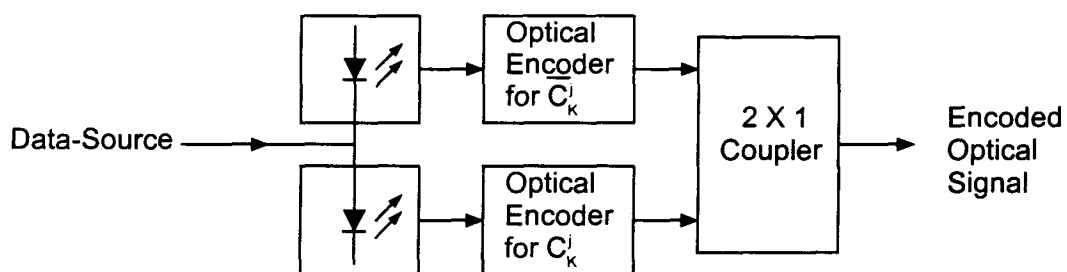


圖 14. MCWC 編碼器架構

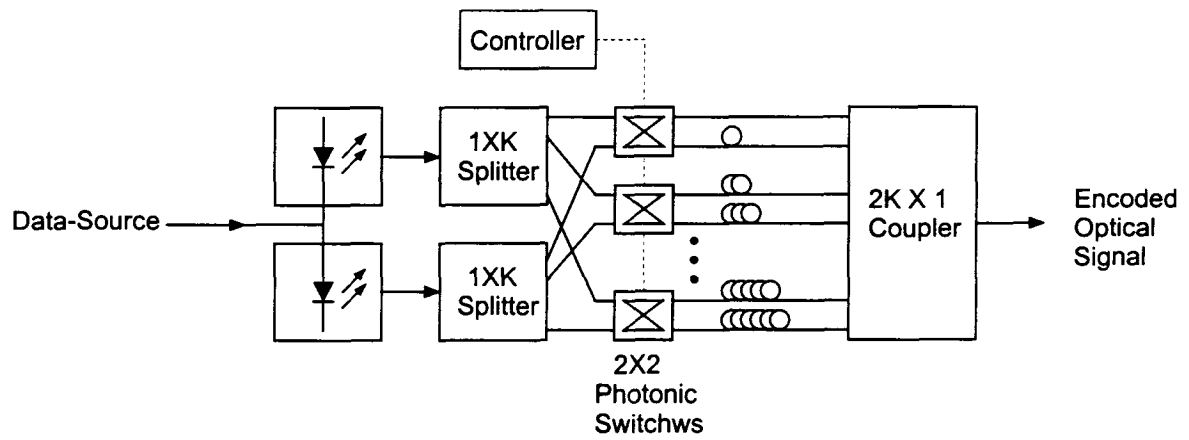


圖 15. 可調式 MCWC 編碼器架構

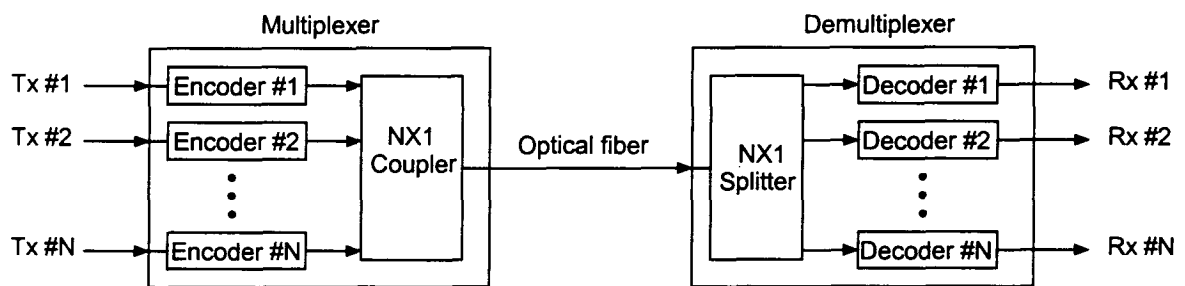


圖 16. 固定多工式系統架構

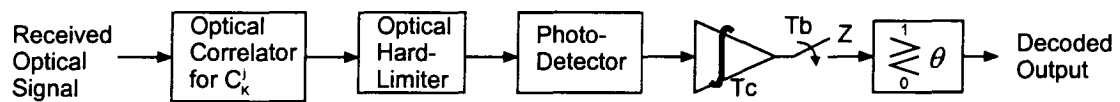


圖 17. 固定多工式解碼器架構

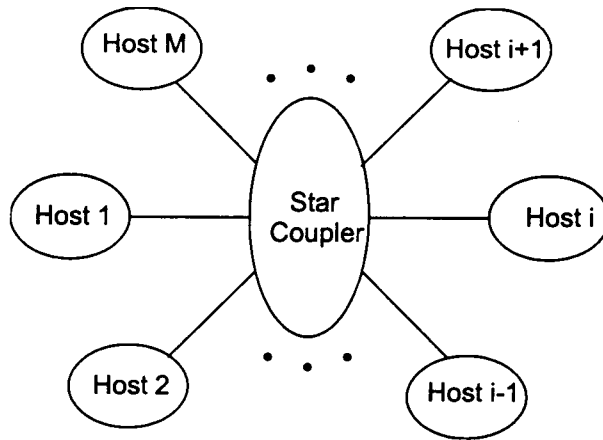


圖 18. 星狀廣播-選取型網路

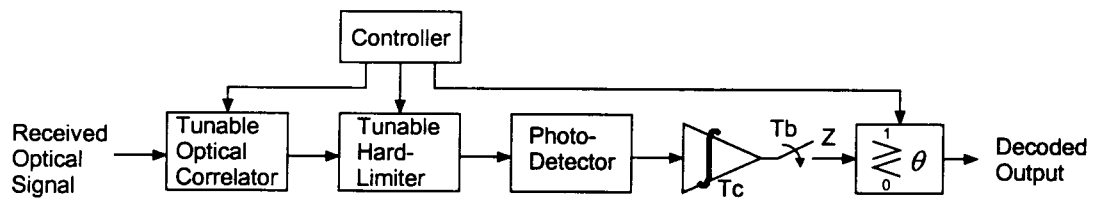


圖 19. 排序網路式解碼器架構

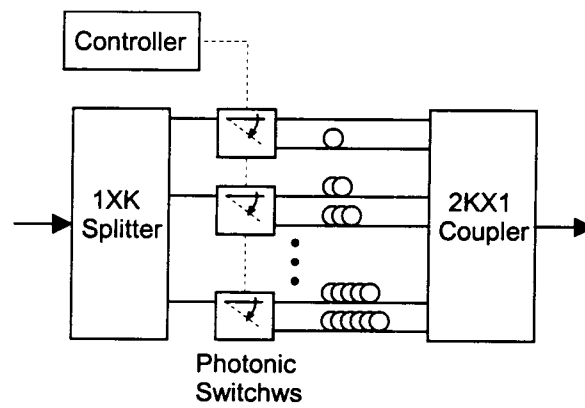


圖 20. 可調式相關器

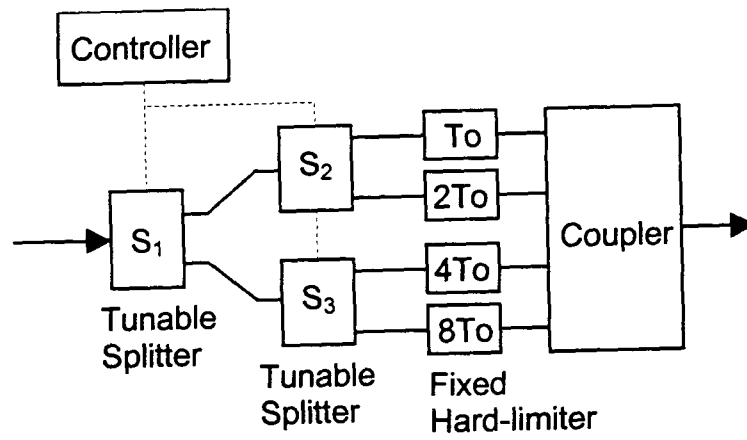


圖 21. 具 15 臨界值之量化式可調型光限制器

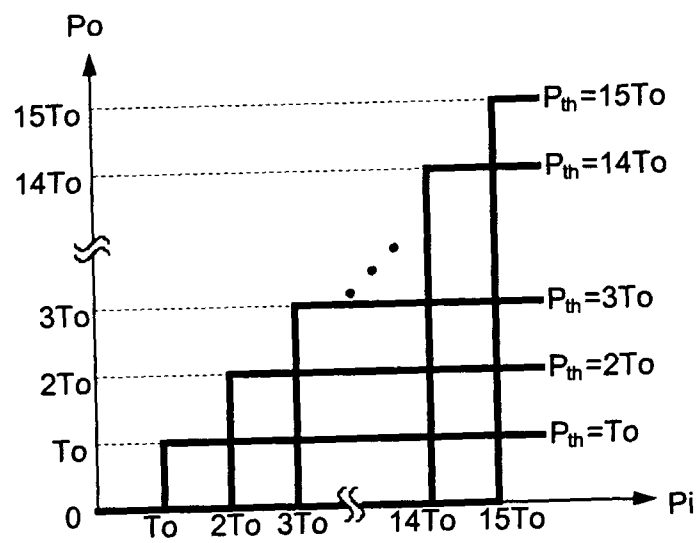


圖 22. 具 15 臨界值之量化式可調型光限制器的特性曲線

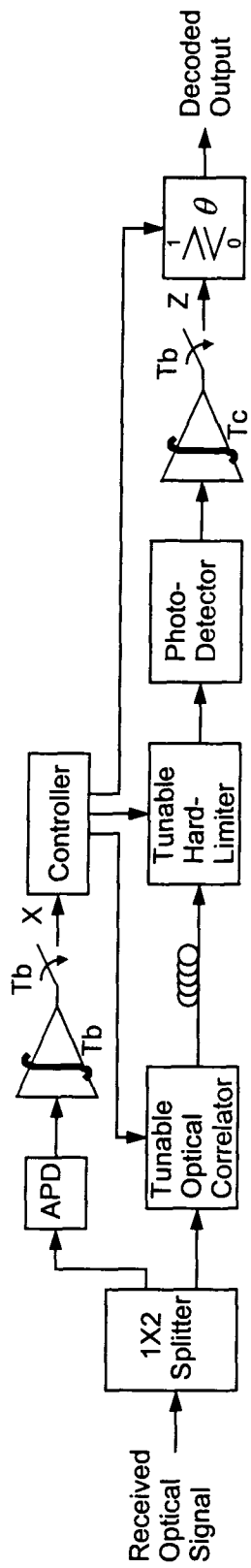


圖 23. 非排序網路式解碼器架構

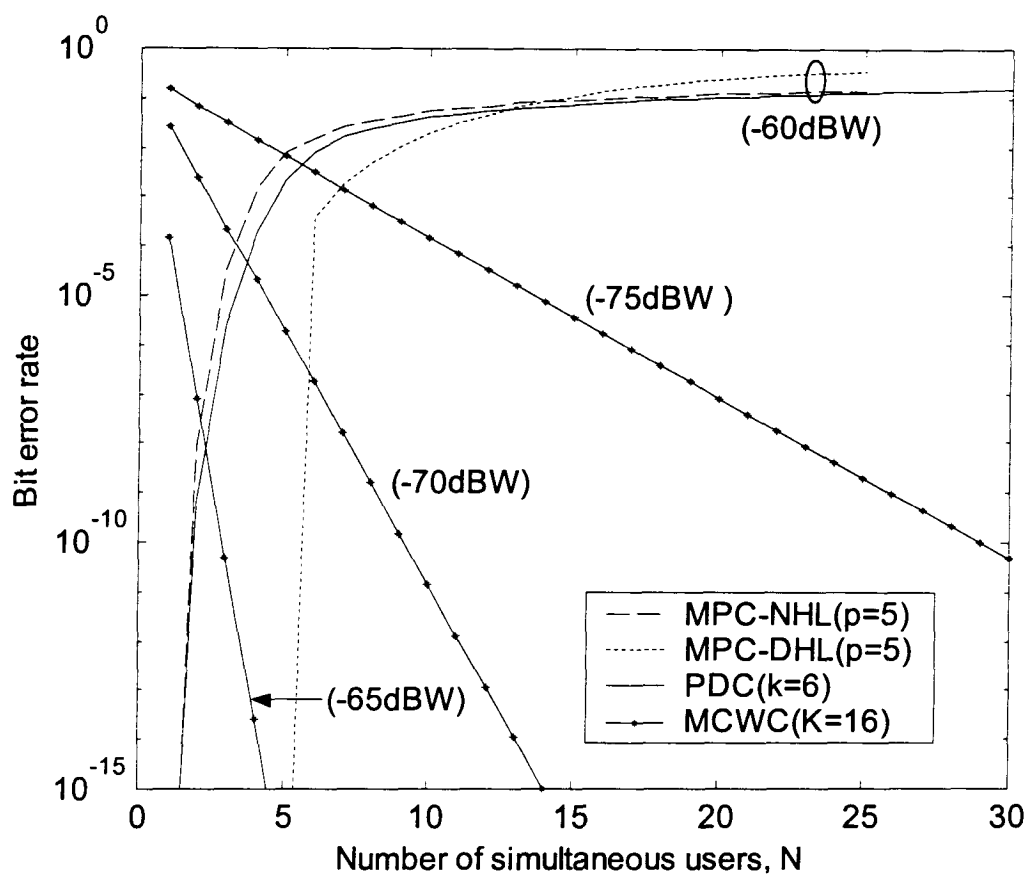


圖 24. 第一/二型接收器之性能

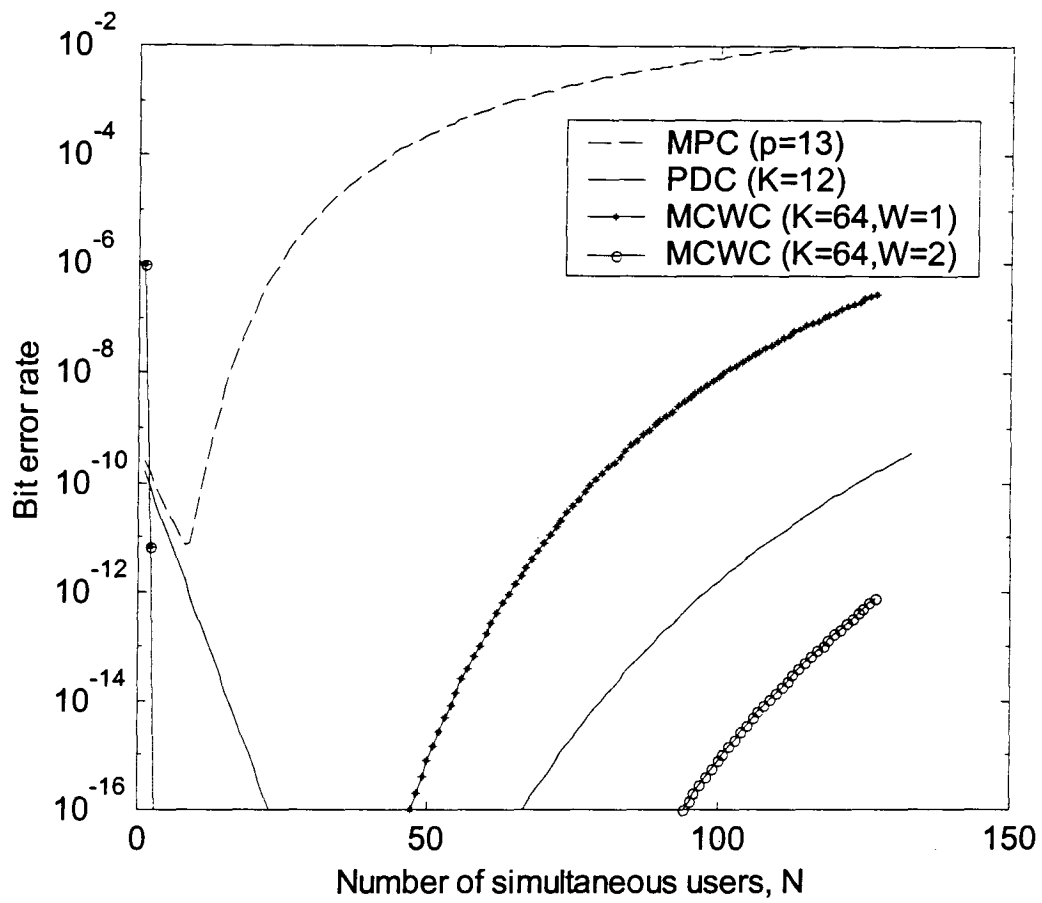


圖 25. 第三型接收器之性能

附件四

行政院國家科學委員會專題研究計畫成果報告

光柵模組在光通信系統之應用

計畫編號：NSC 90-2215-E-002-007

執行期限：90 年 8 月 1 日至 91 年 7 月 31 日

主持人：王倫 國立台灣大學光電光程學研究所教授

一、中文摘要

我們利用光束傳播法來分析去年所執行的光纖耦合架構。此架構乃是利用在光纖透鏡之後埋入長週期光纖光柵，以達到光源與光纖之間光耦合的目的。我們利用光束傳播法來分析光纖的核蕊對光纖透鏡的影響以及分析兩種典型的光纖，即色散偏移光纖與單模步階式光纖對此架構的影響。利用光束傳播法，我們可以決定出在不同的光源光腰下有最佳耦合效率的透鏡半徑與工作距離。另外，我們也比較兩種分析方法，即 ABCD 法與光束傳播法，可發現光束傳播法與實驗結果比較吻合。此外，我們提出了在二元長週期光纖光柵中非線性開關的理論模型。與過去利用非線性模場耦合理論分析在正弦長週期光纖光柵的非線性開關相較起來，此一新的模型清楚說明了非線性相位調制如何改變光柵的相位匹配條件，而且能顯示出週期個數對穿透率的影響。利用這個模型，我們探討入射脈衝波形對非線性開關的影響，證實了在某特定強度的理想方波入射時會有完全的非線性耦合。同時我們進一步發現在長週期光纖光柵中完全耦合的條件，而這些條件廣義地適用在線性與非線性耦合的情形。再將此模型應用在類週期性長週期光纖光柵中，我們在理論上預測了非線性開關存在於此種光柵，並且在某特定強度的理想方波入射時亦有完全的非線性耦合。

二、英文摘要

The beam-propagation method (BPM) is employed to analyze our unique coupling scheme carried out last year, which combines a lensed fiber and a long-period fiber grating (LPFG). The influences of a core within the fiber lens are investigated. As for the fiber dependence of our coupling scheme, two typical types of fibers are studied, namely dispersion-shifted and step-index fibers. With the BPM, the optimal coupling efficiencies for various source waists with corresponding lens radii and working distances are determined. We also compare the results obtained by ABCD method, and find that BPM has better agreement with experimental results. We also propose a theoretical model of nonlinear switching in binary long period fiber gratings (LPFG's). In contrast to the previous model of nonlinear switching in sinusoidal LPFG's based on nonlinear coupled-mode theory, this model illustrates how nonlinear phase modulations change the phase-matching condition, and also exhibits the influences of the number of periods on transmission. Based on this binary model, we investigate the dependence of pulse shapes on nonlinear switching performance, and then demonstrate that the complete switching can be achieved by use of an ideal square pulse with a given intensity. We further find out the general conditions for either linear or nonlinear complete switching in LPFG's. Moreover, by applying the binary model to studying the nonlinear switching performances in quasi-periodic LPFG's, we predict the existence of nonlinear switching in these LPFG's as well as the achievement of their complete switching when an ideal square pulse is used.

三、研究結果

The BPM has been applied to the analysis of beam evolution through the fiber lens in the coupling scheme reported previously (see Fig. 1) [1-3]. Since the effects of lens boundary and core cannot be properly accounted for in the *ABCD* method, the BPM is considered a better approach for the analysis. We consider two kinds of waist sizes $1\mu\text{m}$ for the small-waist source and $3.25\mu\text{m}$ for the large-waist source (see Fig. 2). From the BPM simulation, because of the large divergence angle, the incident beam from a small-waist source is truncated by the lens boundary, and partial energy is reflected from the interface. The reflected field then causes the oscillation of the coupling efficiency over the working distance. However, for the large-waist source, this effect is not significant. In addition, the existence of a fiber core has no significant effect on the coupling efficiency for the large- and the small-waist sources. Note for the large-waist source, efficiencies can reach as much as $\sim 83\%$ and 84% for the cored and coreless lenses, respectively. Also by our comparing the results from the BPM, the *ABCD* method is found quite applicable to such analysis if the source has a large waist. The calculated coupling efficiencies versus working distances are compared with the experiment results, and an overall agreement is obtained. Fig. 3 shows the coupling efficiencies versus the working distance for fiber-to-fiber coupling and LD-to-fiber coupling scheme. Fig. 4 shows the fiber dependence of our coupling scheme, the coupling constant $\kappa_{\text{co-cl}}$ of a DSF is larger than that of an SI-SMF, which explains why the transmission loss of the first cladding mode of an SI-SMF, is small. By studying the coupling behavior of two kinds of fibers, it was found that the longitudinal tolerance for a DSF is also larger. In Fig. 5, it shows the dependence of optimal coupling efficiencies on source beam waists. We can find that BPM is more consistent with the measured efficiencies and that the larger beam waist would lead to a better optimal coupling efficiency. Finally, Fig. 6 shows the optimal conditions versus the beam waist. Shown in Fig. 7(a) is the input-output relationship for two LPFG's, each with a total cell number of 120 ($Lg = 69.5\text{ mm}$). At this length the transmission curves of the two gratings are barely distinguishable in the linear limit and in higher-intensity regions where the pulse detuned from the resonance. Little deviation is found the power $\sim 1 P_c$ at which nonlinear phase modulations bring the pulse into resonance. This deviation can be attributed to the difference in interplay between linear couplings and nonlinear phase modulations of the two gratings.

For a sinusoidal LPFG, both processes distributed along the grating, whereas, for a binary LPFG, couplings take place only at heterointerfaces between regions 0 and 1. However, when fewer number of cells are compared, as for the case shown in Fig. 7(b), where $N = 15$ ($Lg = 8.7\text{ mm}$), a larger deviation at the critical power is observed, while two gratings remain similar in the linear limit and higher-intensity regions. This result implies that, as the number of periods decreases, the nonlinear coupling behaviors that are due to these two different mechanisms near the critical power are more distinct. However, at the intensity region where the pulse is detuned, these two gratings are almost the same.

Fig. 8 depicts the relative transmission curve for an almost ideal square pulse (FWHM, 75 ps), where the relative transmission is defined as the ratio of the integrated output power to the initial power, and the input peak power is normalized to P_c . The input pulse's profile is shown schematically in the inset. The resonance cladding mode is chosen to be LP_{04} , as the same as used for the simulation in Fig. 7. The central

wavelength of the pulse is set to $1.55\ \mu\text{m}$, which corresponds to off-resonance operation. Grating strength $\Delta n_{UV} = 2.51 \times 10^{-4}$ and the length of the LPFG are chosen to produce complete linear coupling at $\lambda_{\text{res}} = 1.541\ \mu\text{m}$. It can be seen that the best switching ratio for the square pulse is more than 95% at the input peak power of $\sim 1.52 P_c$.

Fig. 9 shows the evolution of normalized transmitted power along a uniform LPFG with respect to the almost complete switching input peak power $1.52 P_c$ for the square pulse. For a 50-ps square input pulse after propagating through a Fibonacci LPFG, the relative transmission curve is as shown in Fig. 10. Except for a little degraded linear transmission that is due to coupling to several weak cladding modes at this wavelength, the existence of intensity-dependent switching is clearly seen for the Fibonacci LPFG, just as for a uniform LPFG. And complete switching is almost achieved at the input peak power of $P_{\text{in}} = 1.15 P_c$. To understand the local coupling behavior in the Fibonacci LPFG, we choose the input peak power $P_{\text{in}} = 1.15 P_c$ as an example because it has the best switching ratio. Fig. 11 shows the evolution of the normalized transmitted power. It can be found that, except for a little fluctuation along the length of the Fibonacci LPFG, the evolution curve is similar to that shown in Fig. 9.

四、計畫結果自評

We have completed 3/4 of the year's project and published some results in the renowned journals. The uncompleted item has moved to the recently approved project and been pursued accordingly. Thanks to NSC's continual supports, we feel confident that our research group could establish her reputation in the world in the near future.

五、參考文獻

- [1] W. T. Chen and L. A. Wang, "Laser-to-fiber coupling scheme by utilizing a lensed fiber integrated with a long-period fiber grating," *IEEE Photon. Technol. Lett.*, Vol.12, pp. 501–503, 2000.
- [2] W. T. Chen and L. A. Wang, "Optical coupling method utilizing a lensed fiber integrated with a long-period fiber grating," *Appl. Opt.*, Vol.39, pp. 4490–4500, 2000.
- [3] Wen-Yueh Su, Gia-Wei Chern and L. A. Wang, "Analysis of Cladding-Mode Couplings for a Lensed Fiber Integrated with a Long-Period Fiber Grating by Use of the Beam-Propagation Method," *Applied Optics*, Vol. 41, No. 31, pp.6576–6584, 2002.
- [4] Gia Wei Chern, Jui Fen Chang, and Lon A. Wang, "Modeling of nonlinear pulse propagation in periodic and quasi-periodic binary long-period fiber gratings," *J. Opt. Soc. Am. B*, Vol. 19, no. 7, pp. 1497–1508, 2002.
- [5] B. J. Eggleton, R. E. Slusher, C. M. de Sterke, P. A. Krug, and J. E. Sipe, "Bragg grating solitons," *Phys. Rev. Lett.* **76**, pp. 1627–1730 (1996).
- [6] S. Trillo, S. Wabnitz, N. Finlayson, W. C. Banyai, C. T. Seaton, G. I. Stegeman, and R. H. Stolen, "Picosecond nonlinear polarization switching with a fiber filter," *Appl. Phys. Lett.* **53**, pp. 837–839 (1988).
- [7] G. W. Chern and L. A. Wang, "Transfer-matrix method based on perturbation expansion for periodic and quasiperiodic binary long-period gratings," *J. Opt. Soc. Am. A* **16**, pp. 2675–2689 (1999).
- [8] G. P. Agrawal, *Nonlinear Fiber Optics* (Academic, Boston, Mass., 1996).

六、圖表

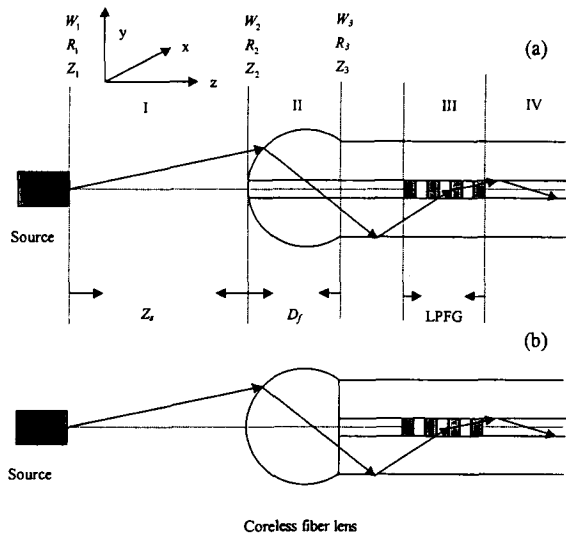


Fig. 1. Schematic diagrams of the coupling scheme studied in the text for (a) a cored fiber lens and (b) a coreless fiber lens integrated with an LPFG.

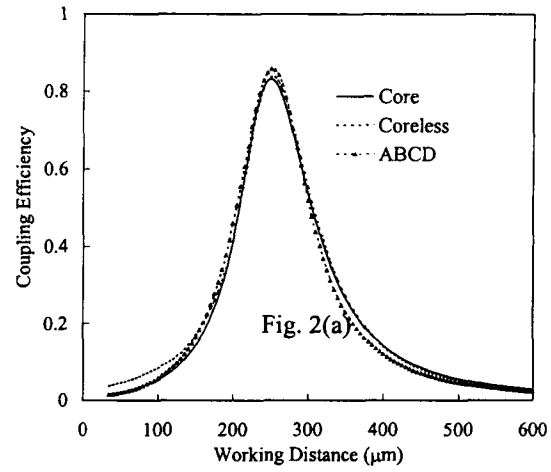


Fig. 2. Calculated coupling efficiencies versus the working distance by use of the BPM (cored and coreless lens) and by use of the *ABCD* method for couplings from the source(a) with an initial waist of $3.25\mu\text{m}$

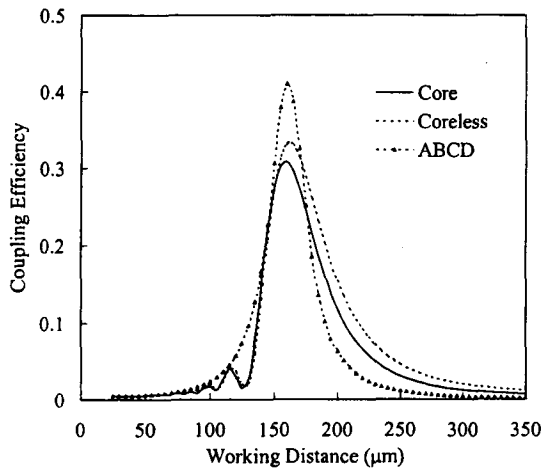


Fig. 2(b)

Fig. 2 (b) with an initial waist of $1\mu\text{m}$

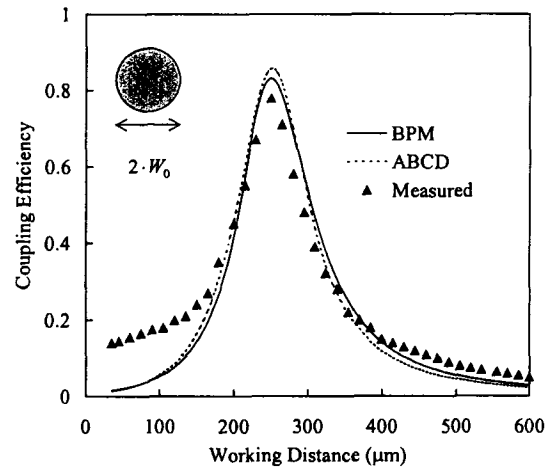


Fig. 3(a)

Fig. 3. Measured coupling efficiencies versus the working distance for (a) fiber-to-fiber coupling, where W_0 ($3.25\mu\text{m}$) with a lens radius of $120\mu\text{m}$ and

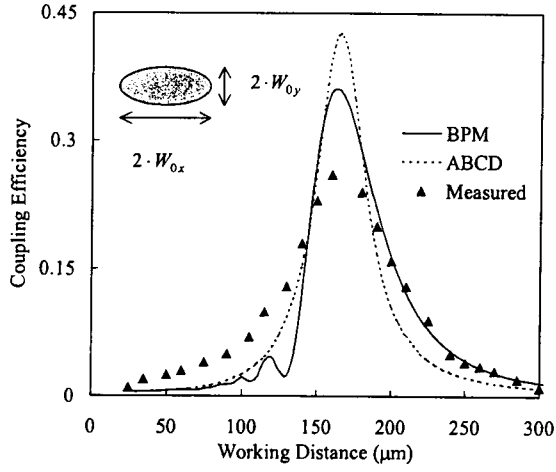


Fig. 3(b)

Fig. 3(b) LD-to-fiber coupling, where $W_{0x} = 1.78 \mu\text{m}$ and $W_{0y} = 0.71 \mu\text{m}$ with a lens radius of $75 \mu\text{m}$. The results are also compared with the calculated ones by use of the BPM and by use of the ABCD method

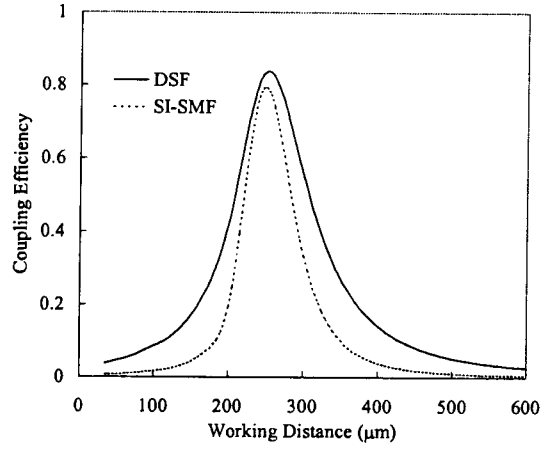


Fig. 4. Comparison of calculated coupling efficiencies of a DSF and an SI-SMF as a function of the working distance for couplings from a source with a $3.25 \mu\text{m}$ waist. The inset shows the field distribution for the first cladding modes of a DSF and an SI-SMF.

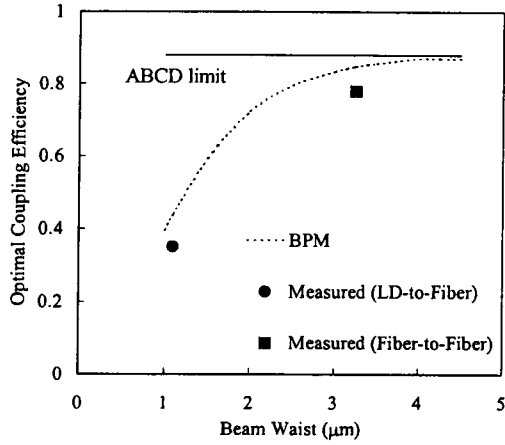


Fig. 5. Calculated and measured optimal coupling efficiency versus beam waist.

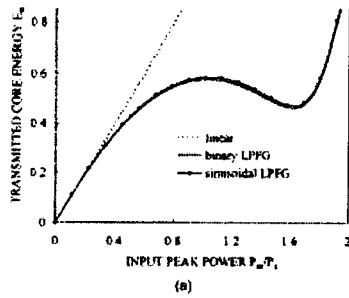


Fig. 7. Core energy transmitted through a binary and a uniform sinusoidal LPFG as a function of normalized input peak power. The number of unit cells is (a) 120

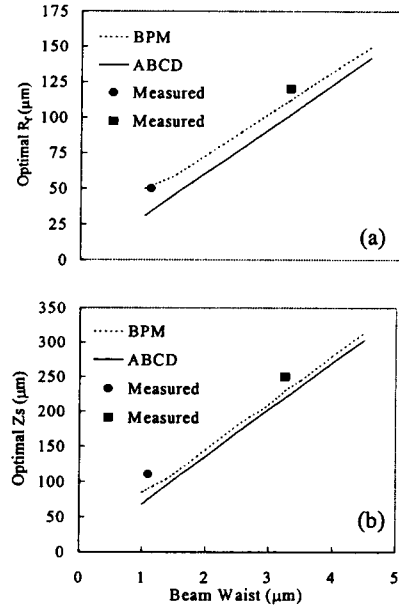
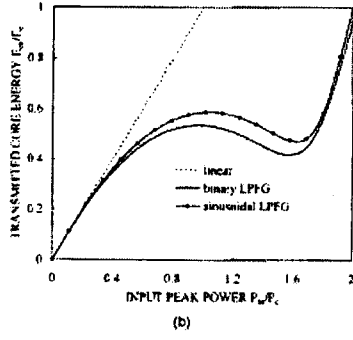


Fig. 6. Calculated and measured optimal conditions of (a) the lens radius R_f and (b) the working distance Z_s versus the beam waist.



b) 15 in each grating. E_c is defined as the total energy of the pulse with input peak power P_c .

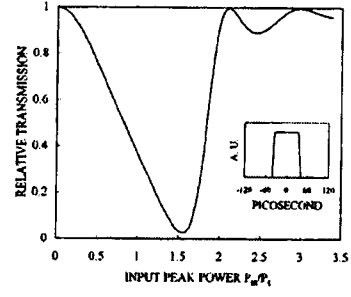


Fig. 8. Fraction of the output power emerging from the core plotted versus normalized input peak power for a 75-ps square pulse as input. Inset, normalized input pulse profile.

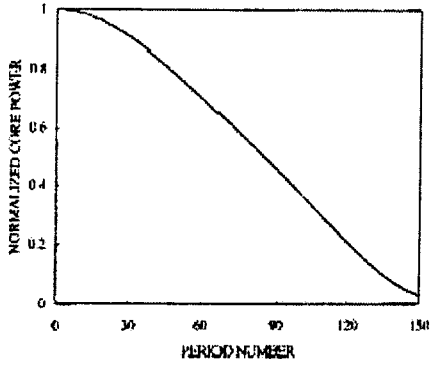


Fig. 9. Evolution of the normalized transmitted core power through a uniform LPFG with respect to input peak power $P_{in} = 1.52 P_c$ of the square pulse.

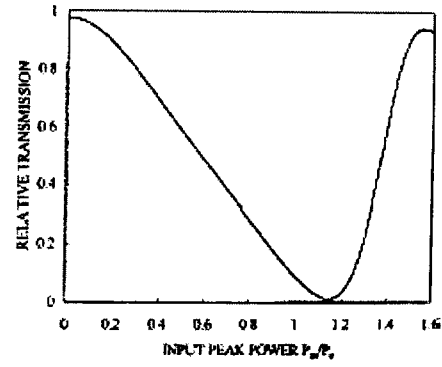


Fig. 10. Relative transmission of a square pulse in a Fibonacci LPFG as a function of normalized input peak power. Note that almost complete switching occurs at $1.15 P_c$.

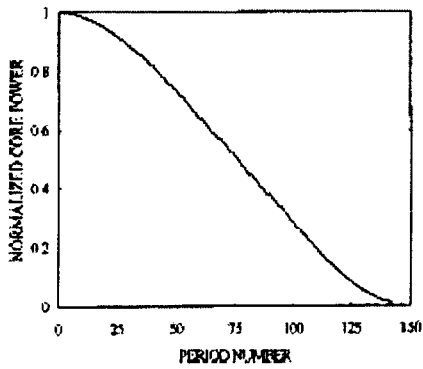


Fig. 11. Evolution of the normalized transmitted core power through a Fibonacci LPFG with respect to input peak power $P_{in} = 1.15 P_c$ of a square pulse.

附件五

行政院國家科學委員會專題研究計畫結案報告
智慧型寬頻光通信網路監測技術之研究
Study of Monitoring Techniques in Smart Broadband Optical Networks

本子計畫第一年計畫編號：NSC 89-2215-E-155-002

執行期限：88 年 8 月 1 日至 89 年 7 月 30 日

本子計畫第二年計畫編號：NSC 89-2215-E-155-011

執行期限：89 年 8 月 1 日至 90 年 7 月 31 日

本子計畫第三年計畫編號：NSC 90-2215-E-003-001

執行期限：90 年 8 月 1 日至 91 年 7 月 31 日

主持人：曹士林 副教授 國立師範大學 光電研究所

一.中文摘要

寬頻光纖網路發展日新月異，特別是採用分波多工(WDMA)，分碼多工(CDMA)或次載波多工(SCMA)方式可提供高容量資訊於一條光纖中傳輸。可是一旦光纖中斷，或在一條傳輸線路徑上有光電元件受外在物理量(如溫度，壓力)之變化，而使系統傳輸失效，將對高速資訊網路通信產生極大的威脅，進而影響到日常生活。因此，近年來對高可靠度的高容量資訊網路有極為迫切的需求。

在本子計畫第一年中，我們嘗試設計及分析一種使用光纖布拉格光柵之類神經型光監測網路。除此之外，我們亦嘗試設計及分析一種新型絕緣層上矽晶(SOI)波導麥克生干涉儀作為光纖網路溫度監測之使用。藉由比較光纖布拉格光柵感測器(FBGs)與絕緣層上矽晶(SOI)波導布拉格光柵感測器，我們發現絕緣層上矽晶(SOI)波導麥克生干涉儀對於溫度感測之靈敏度，將遠高於傳統的光纖布拉格光柵感測器。除此之外，波導麥克生干涉儀之FWHM亦遠小於光纖布拉格光柵感測器及絕緣層上矽晶波導布拉格光柵感測器。吾人亦嘗試將所設計之感測器應用於三種最常見之分波多工感測網路；匯流排形、星形及樹形拓撲結構。藉由適當的佈置波導麥克生干涉儀於星形分波多工感測網路，高容量光纖網路之可靠度問題，將可獲得

解決。

在本子計畫第二年，著重於智慧型分碼多工網路監測訊號與次載波多工通信訊號同時傳輸技術之研究，將光纖雷射應用於智慧型分碼多工通信訊號及次載波多工訊號傳輸，分析光纖雷射載著智慧型分碼多工通信網路監測訊號並應用於次載波多工系統中之可行性。為了能有效載入檢測訊號，必須製作雙埠環型光纖雷射，利用其雙埠同時調變訊號傳輸特性，以同時載入網路偵測 CDMA 訊號。在此計畫中，我們自行研發完成一寬頻光源並供一種使用光纖布拉格光柵之類神經型光監測網路使用。吾人所設計之寬頻光源可達 100nm 頻寬，而雙埠環型光纖雷射也已開發完成，並申請專利，其特性也作完量測與分析，以提供智慧型寬頻通信網路監測使用。

在本子計畫第三年，著重於提出一有效的路由自動切換方式，在此我們設計出一具波長選擇性的元件，並研究智慧型光纖分波多工網路之保護網路，以提高智慧型光纖分波多工網路的可靠度。此波長選擇性的元件是一積體化的可調式光分叉復用器，利用結合布拉格光柵及多模干涉區所組成之光分插復用器元件，我們在布拉格光柵兩側摻雜硼和磷離子，再利用電壓調變布拉格光柵中雜質的濃度變化，利用雜質的變化來改變折射率，藉此從波導中取出特定信號，以達到可控制特定波長信號的路由路徑，接

著，我們將此可調式光分插復用器應用於高密度分波多工的保護網路系統中，如此才可充分利用有限的波長資源。

關鍵詞：溫度感測器、類神經網路、麥克生干涉儀、分波多工、絕緣層上矽晶、波導、光網路、光纖網路、光感測器、光網路監測、光纖雷射、分碼多工、次載波多工、光分插復用器、保護網路

Abstract

Broadband fiber network attracts lot of attentions recently. Especially using WDMA, CDMA, SCMA techniques can increase capacity. Nowadays highly reliable optical fiber networks are strongly required because optical fiber cables have been widely introduced and high capacity signals are transmitted through one fiber. This means that serious problems will happen when fiber fail. This project will develop monitoring techniques for smart broadband optical fiber networks.

In the 1st year, we demonstrate a novel high-resolution neural temperature sensor using two fiber Bragg gratings (FBG's) and a modular artificial neural network in this report. Because of the fast computational ability of the modular artificial neural network and sensitivity of FBG's, the sensor can make high-resolution temperature and frequency measurements in real time. Further, a new silicon-on-insulator (SOI) waveguide Michelson interferometer sensor is also designed and analyzed in this report. The authors compare an optical SOI waveguide Michelson interferometer sensor, a silicon-on-insulator Bragg waveguide grating sensor, and a fiber Bragg grating sensor for temperature sensing. The sensing accuracy of SOI waveguide Michelson interferometer sensor is higher than the fiber Bragg grating sensor. Moreover, the full width of half

maximum (FWHM) of the frequency responses of pass-band of our proposed SOI Michelson interferometers can be designed much narrower than fiber and SOI Bragg waveguide grating sensors for sensing resolution enhancement. Three popular WDM sensor networks are studied with using the SOI waveguide Michelson interferometer sensor. Therefore, the environmental information can be monitored by our designed high-capacity and accuracy enhanced WDM networks with SOI waveguide Michelson interferometer sensors.

In the 2nd year, we focus on the development of smart SCMA communication network with CDMA monitoring technique. Applying our designed fiber lasers in smart CDMA/SCMA network, the simultaneous transmission characteristics and noise are analyzed. We developed five two-port fiber lasers for the CDMA/SCMA fiber-optic monitoring networks. The flexibility of SCMA/CDMA transmission system has been demonstrated from simulation results.

In the 3rd year, we focus on the new technology of router. Here, we design an integrated device, which could select the filtering wavelengths and increase the reliability of optical network. The tunable optical add-drop multiplexer which combines Bragg-grating structure and multimode interference structure. We dope the Boron and Phosphorous ions beside Bragg-grating structure and add voltages to change the carrier concentration distribution. By changing carrier concentration distribution, the index will change. We can drop the specific wavelength from waveguide. We could apply such a device into optical network to save the finite channel wavelength numbers.

Keywords : temperature sensor, neural network, Michelson interferometer, wavelength-division

multiplexing (WDM), silicon-on-insulator(SOI), waveguide, optical networks, optical fiber network, optical sensor, fiber laser, code division multiplexing access (CDMA), sub-carrier multiplexing access (SCMA)

二. 緣由與目的

由於多媒體資訊時代的來臨，各式網路系統逐步的整合，並以高容量的光纖網路作為媒介，在本國已有光纖網路應用於有線電視及電信傳輸方面，特別是分波多工網路[1~2]，分碼多工網路[3~4]及混合光纖/電纜網路[5~6]均在國內外快速發展中。為了有效的將不同的光纖網路做整合，本整合型計畫希望能針對智慧型寬頻光纖通信網路作長期而有系統研究。由於寬頻光纖網路發展快速，目前一條光纖可傳資料量非常大，一但光纖中斷，或元件損壞，將對信號傳輸及網路服務造成非常大的影響，因此，高可靠度的光纖網路實有迫切的需要，目前在國內外已有一些針對雷射元件或網路結構進行的研究[7~8]。

本子計畫為整合型計畫之子計畫六，目的是研究智慧型寬頻光纖通信網路的監測技術。本子計畫第一年的目的是研究光纖光柵及波導光柵感測元件之製作及特性分析，以提供寬頻光纖網路中作為監測元件使用，並利用模糊-類神經網路提升運算速度，研發出快速運算，而又不怕雷射光頻率漂移的模糊-類神經型光監測系統。本子計畫第二年的執行進度為進行雙埠環型雷射的研究並應用於傳輸智慧型分碼多工通信訊號及次載波多工網路偵測進行分析。在國外，針對智慧型分碼多工通信訊號及次載波多工訊號同時傳輸之研究甚少，因此本計畫研究開發新型雙埠雷射五種，並分析同時傳輸分碼多工訊號與次載波訊號之可行性。本子計畫第三年的執行內容為進行智慧型雙向光纖監測網路整合

分波多工路由之實驗，並設計有效之偵錯及波長重置監測網路。

三、結果與討論

在本子計畫第一年，藉由適當佈置光感測器，可提供網路偵錯、容錯等功能。在本年度計畫中，我們提出以光纖布拉格光柵為溫度感測器，並利用模糊-類神經之演算法，以達到高效率、高精度之感測技術。並利用光功率計與計頻器量測光經由光纖布拉格光柵後，反射光頻譜分布情形。經由適當改變外在溫度，我們發現：光頻譜將隨溫度變化而產生飄移情形。並藉由分析比較，我們發現其監測精度可達到 0.005℃。

在本年度計畫中，我們亦嘗試設計分析新型 SOI 積體光波導感測器，利用 SOI 之低成本、高速率之 CMOS 積體電路製程特色，未來可達到光電元件積體化整合之效果。我們使用 SOI 為基材所設計之波導麥克生干涉儀結構圖。經由分散矩陣法 (scattering matrix method) [9]，我們可以得知此波導感測元件之輸入與輸出光場間關係，如下所示：

$$E_{r1} = \frac{re^{-i\beta l_1}}{(1-K)} [Ke^{-2i\beta l_3} + (1-K)e^{-2i\beta l_4}] E_{in} \quad (1)$$

$$E_{r2} = \left[\frac{i\sqrt{K}}{\sqrt{1-K} + \sqrt{1-K}} r e^{-2i\beta l_3} + e^{-2i\beta l_4} e^{-i\beta(l_1+l_2)} \right] E_{in} \quad (2)$$

$$E_{r3} = [i\sqrt{K}(1-r)^2 e^{-i\beta(l_1+l_4)}] E_{in} \quad (3)$$

$$E_{r4} = [\sqrt{1-K}(1-r)e^{-i\beta(l_1+l_4)}] E_{in} \quad (4)$$

其中， E_{in} 為輸入光場， K 為 SOI 波導耦合器之耦合係數， β 為傳播常數， l_i 為耦合器之各臂長， r 為 SOI 波導布拉格光柵之光反射率。並利用耦合模態方程式 (couple mode equation) [10] 及光譜隨溫度飄移之原理 [11] 達到溫度

感測之目的。利用麥克生干涉儀、SOI 波導布拉格光柵與光纖布拉格光柵於外在溫度變化為 10°C 之光譜飄移作比較。由此可知，我們所設計的麥克生干涉儀，對於溫度的靈敏度將 20 倍高於傳統之光纖布拉格光柵感測器。且其 FWHM 亦將遠小於單一 SOI 波導布拉格光柵與光纖布拉格光柵感測器。

由於 SOI 麥克生干涉儀具有高溫靈敏度及低 FWHM 特性，我們嘗試將其應用至分波多工光通訊網路中，並經由分析其於階梯型、星型及樹型等拓撲結構下之訊雜比(SNR)，了解其於網路監測之功能。而星型拓撲結構在分波多工光通訊網路中，可容納較多數目的 SOI 麥克生干涉儀。我們亦針對感測範圍(光纖佈放長度)做分析比較。

經由上述分析，我們發現星型拓撲結構在分波多工光通訊網路中，將可容納 82 個高溫靈敏度之 SOI 麥克生干涉儀，且其感測範圍將廣達 220 公里以上。其優越之網路監測特性，並可作為未來智慧型光通訊網路之參考。

在本子計畫第二年，我們利用自行研製之寬頻光源，可當作待測光纖網路光源並結合第一年成果利用光纖布拉格光柵，供網路之監測端作發射光源之研究使用。此一寬頻光源可應用於以光纖光柵或波導光柵做監測網路之用。

為了研製 CDMA/SCMA 同時載入訊號之雷射，首先我們在此寬頻放大器的兩端分別接上一個 2x2 光電調變器與一個極化控制器，而形成兩個主動光纖環的第一種光纖雷射。此架構有四個主要的調變參數，包含了半導體光放大器的輸入電流、兩個 2x2 光電調變器的調變電壓及摻鐳光纖放大器的輸入功率。我們調整半導體光放大器的輸入電流和摻鐳光纖放大器

的輸入功率，並分別觀察其對系統輸出功率的影響。

最後，我們發現當 2x2 光電調變器的調變電壓 V_1 等於 2.65 伏特、 V_2 等於 1.5 伏特時，其光纖雷射有最佳的功率輸出。

此外，第二種光纖雷射系統使用了一組 $1.3\mu\text{m}$ 半導體光放大器作為主要的增益物質，而一端由中心波長 1319nm 的光柵所組成，另一端則由一個 2x2 光電調變器及極化控制器組成一個主動光纖環，其信號傳輸方式形成一非對稱共振腔。

然後，我們調變半導體光放大器的輸入電流由 70 至 250 毫安培、並且將 2x2 光電調變器的調變電壓由 0 調整至 3.8 伏特，隨著極化控制器的調變，我們可依數據得到半導體光放大器的輸入電流、2x2 光電調變器的調變電壓與系統雷射輸出功率值。

由半導體光放大器與系統雷射輸出功率關係圖，我們將系統的第二輸出端接至光譜儀，並調變半導體光放大器的輸入電流，每隔 5 毫安培記錄一次，可發現當電流超過 105 毫安培時，將產生多模態雷射現象。

再者，我們提出第三種新型光纖雷射系統，其主要組成的元件包含了 1x2 和 2x2 光電調變器及光循環器各一組，其中 1x2 和 2x2 光電調變器的頻寬分別可達到 8GHz 和 1.25GHz。在系統中，我們仍利用了半導體光放大器當作增益物質，再使用一組極化控制器控制信號在光纖中傳輸時的偏極態，以達到穩定且最佳的輸出結果。

為了了解此光纖雷射系統的特性，我們調變系統中半導體光放大器的輸入電流由 50 至 120 毫安培，可發

現在 50 毫安培以上將生成單模態雷射，而超過 120 毫安培會轉成多模態雷射。

經由一連串的實驗過程，我們調整半導體光放大器的輸入電流至 116 毫安培、1x2 光電調變器的調變電壓為 2.2 伏特、2x2 光電調變器的調變電壓為 1.0 伏特，在此狀況下將可得到最理想的系統雷射輸出值。

第四種新型光纖雷射使用了一對耦合係數為 50:50 的 3dB 耦合器組成兩個被動光纖環，再採用一個 2x2 光電調變器來連結兩光纖環，以調變信號之耦合率，並提供一個交流調變信號的輸入端和兩個系統輸出測試端，除此，我們使用了一個相同的半導體光放大器做為系統重要的增益介質。

首先，我們調變半導體光放大器的輸入電流並仔細觀察光譜儀的波形，可發現系統的觸發電流為 25 毫安培，隨著電流增加，輸出功率也不斷提升，直至輸入電流大於 70 毫安培時，系統的輸出波形開始產生不穩定的現象，其半導體光放大器電流調變值與系統測試端輸出功率關係圖。

由以上的實驗，我們可知將半導體光放大器的輸入電流固定在 70 毫安培，可產生最穩定且最大輸出功率，然後再調變 2x2 光電調變器的調變電壓由 0 至 3.4 伏特，可由實驗數據得知當半導體光放大器的輸入電流為 70 毫安培、光電調變器的調變電壓為 2.7 伏特，將可生成系統最佳的雷射輸出。

第五種光纖雷射系統與第四種類似，其系統依然使用了一對 3dB 耦合器、一個 2x2 光電調變器及重要的半導體光放大器增益介質作為主要的系統元件，不同的是兩耦合器與其它元

件的連結方式有所不同，而在兩耦合器所組成的兩被動光纖環中，我們各使用了一個極化控制器，以調變系統輸出之穩定性。

接著，我們調變半導體光放大器的輸入電流由 35 至 195 毫安培，可發現當電流為 195 毫安培時有最佳的雷射輸出值，然後再調變 2x2 光電調變器的調變電壓由 0 至 3.6 伏特，並隨著兩組極化控制器的相位調變，可得到電壓調變值與系統輸出功率的數據曲線圖。

最後，經由上述的實驗過程所示，我們固定半導體光放大器的輸入電壓在 195 毫安培、2x2 光電調變器的調變電壓為 0.3 伏特，可得到此系統最佳的雷射輸出，其最高輸出雷射功率可達 -13.59dBm。

此外，我們將光纖雷射應用於 SCMA 與 CDMA 同時在光纖網路中傳輸，除此之外，本計畫所採 CDMA 多工通道有四十個，利用光纖傳輸雷射功率在 0.02 毫瓦至 0.1 毫瓦，在改變次載波多工訊號調變指數下 CDMA 信號功率變化曲線，並可藉由次載波多工訊號二次拍差比與三次拍差比在第二頻道中，對於不同之光纖雷射功率與不同次載波多工訊號調變指數下之曲線關係圖，我們將 CDMA 信號與二次拍差比與三次拍差比作比較，可知 CDMA 信號在次載波多工訊號中可視為雜訊項，最後次載波多工訊號的特性亦即其訊雜比可得知，在 CDMA 在傳輸系統中放置二十個通道時，其訊雜比可達到三十五分貝，若是利用光濾波器進一步濾除光纖雷射的光雜訊，最後次載波光通訊的信號品質相信可提升，此一光通訊系統可於傳送 SCMA 光通訊中傳送 CDMA 之網路監測訊號。

在本子計畫第三年，我們利用自行設計的積體化的可調式光分叉復用

器來進行智慧型雙向光纖監測網路之系統誤碼率之模擬。

在光纖通信網路中，為了確保網路的安全性及通信品質，保護網路的存在是必要的，系統利用光電開關來控制路由，當有兩條路由中有一條發生斷線時，如何切換保護網路，經路由重置以確保網路的存活。圖(一)為運用可調式光分叉復用器於光纖保護網路的示意圖及以兩個光分叉復用器所組合而成的節點，圖(二)為當節點一發生故障時，保護網路運作的方式。傳統的可調式光分插器，必須用解多工器將所有波長分別獨立，再透過電路控制選擇要下載的波長，如果用我們所設計的可調式光分插器來取代解多工器，則不須將個別波長分別獨立，同時我們所設計的可調式光分叉復用器可結合半導體製程技術大量製造。圖三為積體化可調式光分叉復用器的示意圖，主要由布拉格光柵、多模干涉元件和電極所組合而成的，此元件共有四個腳位，一個輸入埠(port 1)，一個輸出埠(port 3)，一個取出埠(port 2)和上傳埠(port 4)，基本的操作原理為兩不同波長信號 λ_1 、 λ_2 由輸入埠輸入，透過電壓控制訊號，可任意由取出埠中擷取兩波長 λ_1 、 λ_2 之一，並可再上傳埠中上傳另一訊號，最後兩信號 λ_1 、 λ_2 可由輸出埠輸出，如要處理更多信號，可經由串聯不同週期之元件來達成。由圖一中可知布拉格光柵的總長度(N_g)和電極的長度(L_e)同為 $7800\mu\text{m}$ ，多模干涉元件和四腳位的寬度分別為 $12\mu\text{m}$ 、 $4\mu\text{m}$ 。操作電壓控制為($V_p=0\text{V}$, $V_n=0\text{V}$)和($V_p=0.6\text{V}$, $V_n=-0.6\text{V}$)兩種模式，當操作電壓控制為($V_p=0\text{V}$, $V_n=0\text{V}$)，可擷取 λ_1 信號，相對的當操作電壓控制為($V_p=0.6\text{V}$, $V_n=-0.6\text{V}$)，可擷取 λ_2 信號。

圖(四)為積體化可調式光分叉復用器輸入埠和取出埠之剖面圖，由此圖中，我們可看出此元件是一SOI(Silicon-On-Insulator)結構所組

成，最下層為一矽晶圓層，其次為二氧化矽層和最上層的矽晶圓層，在上層矽晶圓層下有摻雜一層雜質層，其依序的高度為 H_{base} 、 H_i 、 H_p 和 H_l 為 $2\mu\text{m}$ 、 $0.4\mu\text{m}$ 、 $0.2\mu\text{m}$ 和 $2.8\mu\text{m}$ 。

圖(五)為積體化可調式光分叉復用器之剖面圖及載子分布圖，在此元件中，摻雜硼和磷以當 n-type 和 p-type，濃度高達 10^{18} 和 $2 \times 10^{17} 1/\text{cm}^3$ ，深度同為 $0.5\mu\text{m}$ 。我們在布拉格光柵兩側摻雜硼和磷離子，上面再鍍上一鋁金屬薄膜，再利用電壓驅動雜硼和磷離子向布拉格光柵中移動，如此布拉格光柵中就產生了雜質的濃度變化，我們利用雜質的變化來改變折射率，藉此從波導中取出特定信號，以達到可控制特定波長信號的路由路徑。圖(六)為時間對折射率變化示意圖，我們利用程式可模擬在一定時間內的，折射率變化，由此圖，我們可知當在時間在 $20 \times 10^{-9}/\text{sec}$ 時，載子已注入完成，及此元件的切換速度可控制在 $20 \times 10^{-9}/\text{sec}$ 。圖七為電壓對折射率變化示意圖，由此圖我們可知當電壓越大時，相對的折射率也變化的越大，在我們此元件中。我們操縱在電壓為正負 0.6V 。

圖(八)加電壓與未加電壓時，由於在布拉格光柵、多模干涉元件區域的折射率產生變化，所以可在取出埠(port 2)擷取出不同之波長。圖(九)為模擬在不加電壓下，在取出埠(port 2)和輸出埠(port 3)所得信號強度比對波長關係圖，由此圖，我們可看出當在取出埠(port 2)取出信號時，信號穿透至輸出埠(port 3)，而產生損耗的量非常的少。圖(十)為模擬在加電壓正負 0.6 伏特下，在取出埠(port 2)和輸出埠(port 3)所得信號強度比對波長關係圖。

圖(十一)為運用可調式光分叉復用器於點對點光纖通訊系統之示意圖，我們將之前所設計的可調式光分叉復用器納入圖(十一)的結構中，由波

長 $\lambda_1=1549.5\text{nm}$ 和 $\lambda_2=1550.4\text{nm}$ 的雷射打入多工器中，經過摻鉍光纖放大器增強信號後，從我們設計的元件輸入埠輸入，當我們不加電壓訊號時，我們可從取出埠擷取出 $\lambda_1=1549.5\text{nm}$ 的信號，同時我們可以從上傳埠上傳一新的且波長同為 1549.5nm 的信號，最後經由解多工器，圖(十二)為當 $\lambda_1=1549.5\text{nm}$ ，在取出埠(port 2)所模擬得之眼圖，圖(十三)為當 $\lambda_1=1549.5\text{nm}$ 且取出 λ_1 後，再上傳 λ_1 ，在輸出埠(port 3)所模擬得之眼圖，由圖(十二)和圖(十三)我們可從眼圖中可知其具有良好的通訊品質。

四、計畫成果自評

本子計畫於第一年已建立光纖光柵及波導光柵之數學理論模型，並模擬不同結構參數之光纖感測器頻率響應，設計不需要用光頻譜分析儀，也能偵測物理量之光柵信號檢測技術，同時結合模糊-類神經型光監測系統，實驗測量感測系統解析度，克服光譜漂移對系統的影響，並開發新型 SOI 波導麥克生干涉儀，取代傳統之光纖布拉格光柵感測器應用至分波多工光通訊監測網路中。

第二年中完成開發寬頻光源以與第一年計畫成果整合成寬頻光網路監測系統，並另開發五種新型雙埠光纖雷射，並可同時載入 CDMA 與 SCMA 多通道信號供網路偵測使用，可應用在智慧型次載波多工光通信網路各節點。

在第三年中完成研究一有效分波多工路電元件之設計，可以用電壓控制路由，使得在系統偵測有錯時可以藉由保護網路，重置路由，使得系統繼續存活。此研究三年內共發表之期刊論文十二篇[12-23]，會議論文發表一百篇[24-123]，專利申請十五項[124-138]，成果豐碩。

五、參考文獻

- [1] Chan, C.K. "Demonstration of an add-drop network node with time slot access for high-speed WDMA dual bus/ring packet networks", *Optical Fiber Communication, Technical Digest Series 48261*, pp. 62-64, 1998.
- [2] Tong, F. "32-Channel tunable receiver module for wavelength-division multiple-access networks", *IEEE Photonics Technology Letters*, vol. 9, pp. 1523-1525, 1997.
- [3] Ohtsuki, Tomoaki, "Channel interference cancellation using electrooptic switch and optical hardlimiters for direct-detection optical CDMA systems", *Journal of Lightwave Technology*, vol. 16, pp. 520-526, 1998.
- [4] Shalaby, "Chip-level detection in optical code division multiple access", *Journal of Lightwave Technology*, vol. 16, pp. 1077-1087, 1998.
- [5] Tzerefos, P., "Standards for high speed digital communications over cable TV networks", *IEE Conference Publication*, vol. 451, pp. 224-229, 1998.
- [6] Angelopoulos, J.D., "TCP/IP over ATM based HFC systems", *IEE Conference Publication*, vol. 451, pp. 6-10, 1998.
- [7] Livermore, Eric, "Architecture and control of an adaptive high-capacity flat network", *IEEE Communications Magazine*, vol. 36, pp. 106-112, 1998.
- [8] Kulkarni, P.T., "Impact of link failures on the performance of multihop lightwave networks", *Computer Communications*, vol. 21, pp. 179-185, 1998.

- [9] F. Scanzhez, "Matrix algebra for all-fiber optical resonators", *IEEE J. of Lightwave Technology*, vol. 9, pp. 838-844, 1991.
- [10] A. Yariv and P. Yeh, "Optical waves in crystals", *New York: Wiley*, 1984, Chapter 6.
- [11] M. G. Xu, L. Reekie, Y. T. Chow, and J. P. Dakin, "Optical in fiber grating high pressure sensor", *Electron. Lett.*, vol. 29, pp. 398-399, 1993.
- [12] Shyh-Lin Tsao, Jiang-Hung Tien and Chun-Wei Tsai, "Simulations on An SOI Grating-Based Optical Add/Drop Multiplexer" Accepted in *Journal of Selected Topics In Quantum Electronics*, Sep., 2002
- [13] Jingshown Wu, Shyh-Lin Tsao and Chun-Hua Ho, "Forward pumping PDFA Modelling and Application on the CATV Systems", To be published in *Journal of optical communications*, Oct, 2003.
- [14] Shyh-Lin Tsao and Chun-Yi Lu, "BPM simulation and comparison of 1×2 Directional Waveguide Coupling and Y-Junction Coupling Silicon-on-Insulator Optical Couplers", To be published in *Fiber and Integrated Optics*, Oct., 2002.
- [15] Shyh-Lin Tsao and Wen-Ming Cheng, "Simplified Formula of Bending Loss for Optical Fiber Sensors", *Fiber and Integrated Optics*, Vol.21, pp323-331, 2002.
- [16] Shyh-Lin Tsao, Huang-Chen Guo and Yi-Jr Chen, "Design of A 2x2 MMI MZI SOI Electro-Optic Switch Covering C-band and L-band", To be published in *Microwave and Optical Technology Letters*, 2002.
- [17] Shyh-Lin Tsao, Hao-Chih Yu and Yi-Chih Lin, "Analysis of an optical satellite communication system with stabilized microwave reference note transmission", *Microwave and Optical Technology Letters* vol. 33, no.3, pp.149-154, May 5, 2002.
- [18] Shyh-Lin Tsao and Peng-Chun Peng, "Design and Simulation of a 2×2N SOI Optical Power Splitter", *Microwave and Optical Technology Letters*, vol.32,no.4, pp.307-310, Feb., 2002.
- [19] Shyh-Lin Tsao, Wei-Liang Chen and Shien-Cheng Chiou "An Active IIR Filter using Coupled Optical waveguide Rings", *Journal of optical communication* vol. 23, no.2, pp52-55, April, 2002.
- [20] Shyh-Lin Tsao and Peng-Chun Peng, "An SOI Michelson Interferometer Sensor with Waveguide Bragg Reflective Gratings for Temperature Monitoring", *Microwave and Optical Technology Letters*, vol.30,no.5, pp.321-322, September,

- 2001.
- [21] Jyh-Ping Lee, Shyh-Lin Tsao, Hen-Wei Tsao, "Broadband hybrid analog/digital CATV trunk network", IEEE Transactions on Broadcasting, Vol.45, No.3, pp.339-347, September, 1999.
 - [22] Yang-Han Lee, Jyh-Yuan Wang and Shyh-Lin Tsao, "Design of packet-switched network using optical signal processing techniques", Microwave and Optical Technology Letters, October, Vol.23, No.2, 104-109, 1999.
 - [23] Shyh-Lin Tsao, Jingshown Wu and Bih-Chyun Yeh, "High resolution neural temperature sensor using fiber Bragg gratings", IEEE Journal of Quantum Electronics, November, vol. 35, No.11, 1590-1596, 1999.
 - [24] Shyh-Lin Tsao, Shu-Tsung Kuo, Tang-Yuan Chang, "A fiber ring laser with fourteen output", Technical Digest of 1998 Optoelectronics and Communications Conference(OECC'98), pp.256-257, 1998.
 - [25] Shyh-Lin Tsao, Shu-Tsung Kuo, Tang-Yuan Chang, "Study of multiple output fiber ring laser", Proceedings of The 16th (1998) International Telecommunication Symposium (ITS'98), Vol. I, September 15-17, Taipei, pp. 85-89, 1998.
 - [26] Shun-Chen Yang and Shyh-Lin Tsao, "Analysis and demonstration of adjustable correlator", Technical Digest of International Conference on Application of Photonic Technology(ICAPT'98), SPIE vol. 3491, pp. 128-132, 1998.
 - [27] Shyh-Lin Tsao and Bi-Chung Yeh, "Extremely high resolution fiber temperature sensor", Technical Digest of International Conference on Application of Photonic Technology(ICAPT'98), SPIE vol. 3491, pp. 873-877, 1998.
 - [28] Shyh-Lin Tsao, Shu-Tsung Kuo, Tang-Yuan Chang, "Characteristic measurement of multiple output laser with arrayed-waveguide grating", Technical Digest of International Conference on Application of Photonic Technology(ICAPT'98), pp.145-150, 1998.
 - [29] Shyh-Lin Tsao, "Characteristics study of a fiber-optic storage loop using MQW semiconductor optical amplifier", Asia Pacific Symposium on Optoelectronics, Proceedings of SPIE, vol. 3420, pp. 318-327, 1998.

- [30] Shyh-Lin Tsao, Hao-Chih Yu, "Performance analysis of a satellite optical heterodyne communication system", 1998 National Symposium on Telecommunications (NSTC'98), December 18-19, Taipei, pp.119-123, 1998.
- [31] Shyh-Lin Tsao, Jingshown Wu and Bih-Chyun Yeh, "A fiber Bragg grating temperature sensor with artificial neural networks", Conference Proceedings of 13th international Conference on Optical Fiber Sensors, Kyongju, Korea, pp. 438-441, April 12-16, 1999.
- [32] Shyh-Lin Tsao, Jingshown Wu and Shun-Chen Yang, "Design of optical fiber CDMA systems with dynamic encoder/decoders", The fourth Symposium on Computer and Communication Technology, pp. 373-376, Taouan, Taiwan, October 7-8, 1998.
- [33] Shyh-Lin Tsao and Hao-Chih Yu, "Vibration analysis of an optical satellite coherent communication system", The fourth Symposium on Computer and Communication Technology, pp.63-68, Taouan, Taiwan, October 7-8, 1998.
- [34] Shyh-Lin Tsao, Hao-Chih Yu, "Design of a satellite optical heterodyne communication system with two local oscillator lasers", Proceeding of 1998 International Photonics Conference (IPC'98), pp. 346-348, Taipei, Taiwan, December 15-18, 1998.
- [35] San Liang Lee, Shyh-Lin Tsao, Chun-Liang Yang, Yann-Liang Chen, Chih-Hui Chiang, "Characterization of transparent properties in semiconductor laser amplifier", Proceeding of 1998 International Photonics Conference (IPC'98), pp.523-525, Taipei, Taiwan, December 15-18, 1998.
- [36] Jhy-Ping Lee, Hen-Wei Tsao, Shyh-Lin Tsao, Henry Y. Lin, Kuang-Yi Cheng and Y. Hsu, "Wideband analog and digital CATV distribution network", Proceeding of 1998 International Photonics Conference (IPC'98), pp. 760-762, 1998.
- [37] Jingshown Wu, Tsun-Hua Ho, Wei-Chou Wu, Shyh-Lin Tsao, Herry Lin, Steven Chen, Holner Lin, and Michael Chin, "Experimental PDFA and its application on the CATV system", Proceeding of 1999 IEEE/LEOS Summer optical Meetings, San Diego, U.S.A., July 26-30, pp.17-18, 1998.
- [38] Shyh-Lin Tsao and Sheen-Je Lee, "Simulation of Optical SOI Directional Coupler with Polymer

- Cover", to be published in Proceeding SPIE's 44th Annual Meeting, The Symposium on Optical Science, Engineering, and Instrumentation, July 18-23, Denver, Colorado, U.S.A., pp.322-332, 1999.
- [39] Shyh-Lin Tsao, Sheen-Je Lee, Peng-Chun Ping and Tien-Sheng Chao, "Simulation of optical Silicon on insulator directional coupler", Proceeding of The Sixth Symposium on Nano Device Technology, May 12-13, Hsin-Chu, Taiwan, pp.169-172,1999.
- [40] Shyh-Lin Tsao, Shin-Ge Lee and Ming-Chun Chen and Tien-Sheng Chao, " Maximum Optical Power Output Analysis of Single Mode SOI Waveguides", Proceeding of The Sixth Symposium on Nano Device Technology, May 12-13, Hsin-Chu, Taiwan, pp166-168, 1999.
- [41] Shyh-Lin Tsao, Hao-Chih Yu and Yi-Chih Lin, "Analysis of a new high-speed coherent optical satellite communication system", Technical Digest of The Pacific rim Conference on Lasers and Electro-Optics (CLEO/Pacific Rim'99), Aug.30-Sept.3, Seoul, Korea, Vol. 2, pp.720-721, 1999.
- [42] Shyh-Lin Tsao and Shu-Tsung Kuo, " A new multiple output two-ring fiber laser", Technical Digest of The Pacific rim Conference on Lasers and Electro-Optics (CLEO/Pacific Rim'99), Aug.30-Sept. 3, Seoul, Korea, Vol. 2, pp. 517-518, 1999.
- [43] Shyh-Lin Tsao, Sheen-Je Lee, Peng-Chun Ping and Tien-Sheng Chao, "simulation of optical Silicon on insulator directional coupler", Proceeding of The Sixth Symposium on Nano Device Technology, May 12-13, Hsin-Chu, Taiwan, pp.169-172,1999.
- [44] Shyh-Lin Tsao, Hao-Chih Yu and Yi-Chih Lin, "Performance evaluation of a new optical heterodyne FSK satellite communication system", Proceeding of 1999 International Microwave and Optoelectronics Conference (IMOC'99) CD-ROM,Ri de Janeiro, Brazil, pp.463-466, August 9-12,1999.
- [45] Shyh-Lin Tsao and Thi-Chi Liou, "A new two-way opto-electronic probe of optical component analyzers", Proceeding of the Eighth National Conference on Science and Technology of National Defence, TouYuan, Taiwan, Nov. 20, B43 in CD-ROM, pp.1-10, 1999.
- [46] Shyh-Lin Tsao and Yi-Chih Lin,

- "Image transmission through an experimental wireless communication system", Proceeding of 1999 12th IPPR Conference on Computer Vision, Graphics and Image Processing (GVGIP'1999), Taipei, Taiwan, Aug.22-24, pp.134-137, 1999.
- [47] Shyh-Lin Tsao, Yi-Chih Lin and Hao-Chih Yu, "Analysis of a new coherent optical satellite communication system", Proceedings of The 2nd(1999) Radio Science Symposium, Chung-Li, Dec. 20-21, Taiwan, pp.77-84, 1999.
- [48] Shyh-Lin Tsao and Thi-Chi Liou, "Design and implementation of a wideband microwave matching circuit of circulator for communications", Proceedings of The 2nd(1999) Radio Science Symposium, Chung-Li, Dec. 20-21, Taiwan, pp.71-76, 1999.
- [49] Shyh-Lin Tsao and Shin-Ge Lee, "Reflective spectra analysis for Silicon-on-Insulator Bragg waveguide grating temperature sensor", Proceedings of The 2nd(1999) Radio Science Symposium, Chung-Li, Dec. 20-21, Taiwan, pp. 386-390, 1999.
- [50] Yi-Fan Tsai and Shyh-Lin Tsao, " Design of high-pass optical digital filters(HPODFs) for enhancing speech recognition in wireless optical communication", Proceeding of Optics and photonics Taiwan'99 (OPT'99) ,Dec.16-17, Chung-Li, Taiwan, pp.10630-1066,1999.
- [51] Jian-Shuen Fang and Shyh-Lin Tsao, "A two-dimensional optical position sensor with dual photodetectors", Proceeding of Optics and photonics Taiwan'99 (OPT'99) ,Dec.16-17, Chung-Li, Taiwan, pp.1343-1346, 1999.
- [52] Shyh-Lin Tsao and Shien-Cheng Chiou, "A fiber laser with two active fiber loop mirrors", Technical Digest of The Fifth Optoelectronics and Communications Conference (OECC2000), July 11-14, Chiba, Japan, pp.332-333, 2000.
- [53] Shyh-Lin Tsao and Peng-Chun Peng, "Design of A Two-Dimensional 1×32 SOI Optical Waveguide Coupler", Technical Digest of The Fifth Optoelectronics and Communications Conference(OECC2000), July 11-14, Chiba, Japan, pp.278-279, 2000.
- [54] Shyh-Lin Tsao and Huang-Chen Guo, "Simulation of 1x4 MMI

- Coupler with Eight Square Pads”, Technical Digest of The Fifth Optoelectronics and Communications Conference(OECC2000), July 11-14, Chiba, Japan, pp.282-283, 2000.
- [55] Shyh-Lin Tsao and Shin-Ge Lee, “Characteristics Analysis of SOI Waveguide Michelson Interferometers for Developing Biomedical Fiber Temperature Sensing Head”, Proceeding of SPIE Optical Sensing, Imaging and Manipulation for Biological and Biomedical Applications Conference, International optoelectronics Symposium on Photonics Taiwan, July 26-28, Taipei, Taiwan, Vol. 4082, pp.278-285, 2000.
- [56] Shyh-Lin Tsao and Yu-Long Shen, “Experimental Study of the temperature effect on the polarization state of MQW semiconductor optical amplifier”, Proceeding of SPIE Optoelectronic Materials and Devices II Conference, International optoelectronics Symposium on Photonics Taiwan, July 26-28, Taipei, Taiwan, Vol.4078, pp.414-421, 2000.
- [57] Shyh-Lin Tsao and Der-Yu Chen, “Comparison of Two 1550nm Ultra Narrow-Band Optic Infinite Impulse Response Filters for High-Speed Optical Signal Processing”, Proceeding of SPIE Optical Storage and Optical Information Processing Conference, International optoelectronics Symposium on Photonics Taiwan, July 26-28, Taipei, Taiwan, Vol. 4081, pp.210-219, 2000.
- [58] Shyh-Lin Tsao and Huang-Chen Guo, “Analysis of MMI Characteristics with Periodic Air Pad on SOI Waveguide-Based Optical Interconnects”, Proceeding of SPIE Optoelectronic Materials and Devices II Conference, International optoelectronics Symposium on Photonics Taiwan, July 26-28, Taipei, Taiwan, Vol.4078, pp.397-405, 2000.
- [59] Shyh-Lin Tsao and Peng-Chun Peng, “Design of Two-Dimensional 1×16 and 1×32 Array Waveguide Optical Power Splitters”, Proceeding of SPIE Optoelectronic Materials and Devices II Conference, International optoelectronics Symposium on Photonics Taiwan, July 26-28, Taipei, Taiwan, Vol. 4078, pp.373-382, 2000.
- [60] Shyh-Lin Tsao and Ming-Chun Chen, “Analysis of the Optimal Light Launching Position of SOI Tapered Waveguides”, Proceeding of SPIE Optoelectronic Materials

- and Devices II Conference, International optoelectronics Symposium on Photonics Taiwan, July 26-28, Taipei, Taiwan, Vol. 4078, pp.406-413, 2000.
- [61] Shyh-Lin Tsao and Shien-Cheng Chiou, "Implementation of A 1280nm-1380nm Light Source for Reflective Spectrum Measurement of Biomedical Fiber Sensors ", Proceeding of SPIE Optical Sensing, Imaging and Manipulation for Biological and Biomedical Applications Conference, International optoelectronics Symposium on Photonics Taiwan, July 26-28, Taipei, Taiwan, Vol.4082, pp.286-297, 2000.
- [62] Shyh-Lin Tsao and Wen-Ming Cheng, "Tightly Bending Loss Measurement for 1.3 μm and 1.5 μm broadband wavelength Division Multiplying Fiber Sensor Systems", Proceeding of SPIE Optical Sensing, Imaging and Manipulation for Biological and Biomedical Applications Conference, International optoelectronics Symposium on Photonics Taiwan, July 26-28, Taipei, Taiwan, Vol. 4082, pp.265-274, 2000.
- [63] Shyh-Lin Tsao and Thi-Chi Liou, "Design and Implementation of A New Two-Way Opto-electronic Probe for Optical Information Processing Components Analysis", Proceeding of SPIE Optical Storage and Optical Information Processing Conference, International optoelectronics Symposium on Photonics Taiwan, July 26-28, Taipei, Taiwan, Vol. 4081, pp.242-249, 2000.
- [64] Shyh-Lin Tsao, Peng-Chun Peng, Shin-Ge Lee and Tien-Sheng Chao, "A Single Mode SOI Optical Directional Coupler with S-Bend Waveguide I/O Ports", Proceeding of The Seventh Symposium on Nano Device Technology, May 24-25, Hsin-Chu, Taiwan, pp.56-59, 2000.
- [65] Shyh-Lin Tsao and Peng-Chun Peng, "Design and Simulation of 1 $\times$ 4 and 1 $\times$ 8 Silicon-on-Insulator Multimode Interference Optical Power Splitters ", Proceeding of The Seventh Symposium on Nano Device Technology, May 24-25, Hsin-Chu, Taiwan, pp. 259-262, 2000.
- [66] Shyh-Lin Tsao, Shin-Ge Lee, Peng-Chun Peng, and Tien-Sheng Chao, "Temperature Response of An Optical Silicon-on-Insulator Channel Waveguide Grating", Proceeding of The Seventh Symposium on Nano Device Technology, May 24-25, Hsin-Chu, Taiwan, pp.255-258, 2000.

- [67] Shyh-Lin Tsao, Ling-Fan Tsao, Herman Rao and Kang Hsin Lu, "Experimental study of Compressed Image Transmission over The Fast-Ethernet Optical fiber Communication System", Proceeding of 2000 13th IPPR Conference on Computer Vision, Graphics and Image Processing (GVGIP'1999), Taipei, Taiwan, Aug.21-23, pp.316-320, 2000.
- [68] Shyh-Lin Tsao and Ling-Fan Tsao, "Experimental Study of Image Transmission Through A VHF over Fast-Ethernet Fiber-Optical Communication System", Proceeding of 2000 Taiwan Area Network Conference (TANET2000), pp.918-923, October 19-21, Tainan, Taiwan, 2000.
- [69] Shyh-Lin Tsao and Shien-Cheng Chiou, "A Novel Fiber Laser with symmetric Resonant Cavity for Optical Communications", Proceeding of The 6th Asia-Pacific Conference on Communication (APCC2000), Vol. I, pp. 215-217, Seoul, Korea, Oct. 30-Nov. 2, 2000.
- [70] Shyh-Lin Tsao and Shien-Cheng Chiou, "A Novel Wavelength Tunable Fiber Laser with Non-Symmetric Resonant Cavity", Proceeding of 2000 IEEE Lasers and Electro-Optics Society 2000 Annual Meeting(LEOS2000), Vol. 2, pp.397-398, Nov. 13-16, Rio Grande, Puerto Rico, 2000.
- [71] Shyh-Lin Tsao, Peng-Chun Peng and Shin-Ge Lee, "A Novel MMI-MI SOI Temperature Sensor", Proceeding of 2000 IEEE Lasers and Electro-Optics Society 2000 Annual Meeting(LEOS2000), Vol.2, pp.464-465, Nov. 13-16, Rio Grande, Puerto Rico, 2000.
- [72] Shyh-Lin Tsao, Shien-Cheng Chiou and Yuan-Ze Lee, "A Fiber Ring Laser with A Pair of Passive Fiber Loop Mirrors and A 2x2 Electro-Optical E/O Switch", Proceeding of 2nd International Photonic Conference (IPC2000), pp. 610-612, December 12-15, Hsinchu, Taiwan, 2000.
- [73] Shyh-Lin Tsao, Shin-Ge Lee and Peng-Chun Peng, "Performance Analysis of WDM Temperature Monitoring Networks using Silicon-on-Insulator Waveguide Michelson Interferometers", Proceeding of 2nd International Photonic Conference (IPC2000), pp. 699-701, December 12-15, Hsinchu, Taiwan, 2000.
- [74] Shyh-Lin Tsao, Shien-Cheng Chiou and Ta-Chun Lin, "A Fiber Laser with An Active Fiber Loop Mirror and A Narrow-Bandwidth Fiber

- Grating Mirror”, Proceeding of 2nd International Photonic Conference (IPC2000), pp. 386-388, December 12-15, Hsinchu, Taiwan, 2000.
- [75] Shyh-Lin Tsao, Shien-Cheng Chiou and Che-Yuan Hsu, “Implementation of An Optical Circulator Included Active Fiber Loop Mirror for A Fiber Laser”, Proceeding of 2nd International Photonic Conference (IPC2000), pp. 691-693, December 12-15, Hsinchu, Taiwan, 2000.
- [76] Shyh-Lin Tsao and Huang-Chen Guo and Yi-Jr Chen, “Design and Simulation of A 1x2 MMI SOI Waveguides with 0.78/1.55mm WDM Coupler”, Proceeding of 2nd International Photonic Conference (IPC2000), pp. 598-600, December 12-15, Hsinchu, Taiwan, 2000.
- [77] Shyh-Lin Tsao, Shin-Ge Lee, Peng-Chun Peng and Hsin-Chun Huang, “Measurement of Optical Waveguides on Silicon-on-Insulator Wafer”, Proceeding of 2nd International Photonic Conference (IPC2000), pp.541-543,December 12-15, Hsinchu, Taiwan, 2000.
- [78] Shyh-Lin Tsao, Peng-Chun Peng and Hsin-Chun Huang, “Design of Vertical Coupler Mach-Zehnder Interference Multiplexer Using Silicon-on-Insulator Wafer”, Proceeding of 2nd International Photonic Conference (IPC2000), pp. 613-615, December 12-15, Hsinchu, Taiwan, 2000.
- [79] Shyh-Lin Tsao and Wen-Ming Cheng, “An Optical Wavelength to Power Converter Using Fiber Spiral Bending Technique”, Proceeding of 2nd International Photonic Conference (IPC2000), pp. 619-621, December 12-15, Hsinchu, Taiwan, 2000.
- [80] Shyh-Lin Tsao and Ming-Chun Chen, “BPM Simulation on The Optimal Input Light Spot-Size and Launching Position of SOI Tapered Waveguides”, Proceeding of 2nd International Photonic Conference (IPC2000), pp. 607-609, December 12-15, Hsinchu, Taiwan, 2000.
- [81] Shyh-Lin Tsao and Shu-Tsung Kuo and Ching-Sheu Wang, “A Hybrid CATV and ADSL Broadcasting Optical Fiber Communication System”, Proceeding of 2000 National Symposium on Telecommunications (NSTC'2000), pp. 2-253-2-258, December 15-16, Chung-Li, 2000.
- [82] Shyh-Lin Tsao and Huang-Chen Guo, “Wavelength Response Simulation of A 2x2 MMI SOI Coupler for WDM Optical

- Communications”, Proceeding of 2000 National Symposium on Telecommunications (NSTC'2000), pp. 2-259-2-262, December 15-16, Chung-Li, 2000.
- [83] Shyh-Lin Tsao, Peng-Chun Peng, Hsin-Chun Huang, and Shun-De LinZhong, “Design of A 2x2 MMI-MZI Elcetro-Optic Switch Based on SOI Stubstrate with S-bend Waveguide”, Proceeding of 2000 International Electron Devices and Materials Symposium (EDMS2000), pp. 296-299, Chung-Li, Taiwan, Dec.20-21, 2000.
- [84] Shyh-Lin Tsao, Ming-Chun Chen and Hsin-Chun Huang, “Study of the Exponentially and the Linearly Double-Tapered Waveguides Using SOI Technology”, Proceeding of The Seventh Symposium on Nano Device Technology, April. 24-25, Hsin-Chu, pp.216-219, 2001.
- [85] Shyh-lin Tsao, Huang-Chen Guo, Yi-Jr Jou, “Design of a 1x2 1.3um/1.55um SOI MMI wavelength division multiplexer”, Proceeding of The Seventh Symposium on Nano Device Technology, April. 24-25, Hsin-Chu, pp.209-211,2001.
- [86] Shyh-Lin Tsao, Peng-Chun Peng , Chun-Yi Lu, Li-Wu Huang and Tien-Sheng Chao, “Mode analysis of silicon-on-insulator optical rib waveguide”, Proceeding of The Seventh Symposium on Nano Device Technology, April. 24-25, Hsin-Chu, pp.56-59,2001.
- [87] Shyh-Lin Tsao, Shun-De LinZhong, Zhao-Hong Zeng and Tien-Sheng Chao, “Insertion loss analysis of 1x4 SOI MMI optical power splitter”, Proceeding of The Seventh Symposium on Nano Device Technology, April. 24-25, Hsin-Chu, pp.212-215,2001.
- [88] Shyh-Lin Tsao, Huang-Chen Guo and Yi-Jr Jou, “Simulation of a 2x2 MMI SOI WDM coupler”, Proceeding of The 4th Pacific Rim Conference on Lasers and Electro-Optics (CLEO/Pacific Rim 2001), July 15-19, Chiba, Japan, Vol. II, pp. II-210-II-210, 2001.
- [89] Shyh-Lin Tsao, Shien-Cheng Chiou and Yuan-Ze Lee, “Study of A Coupled Fiber Ring Laser”, Proceedings of Opto-Electronic and optical communication/Integrated optics and Optical Communication (OECC/IOOC 2001) Conference, pp.423-424, July 1-5, Sydney, Australia, 2001.
- [90] Shyh-Lin Tsao , Shien-Cheng Chiou and Ta-Chun Lin, “Wireless Safeguard Signal Transmission

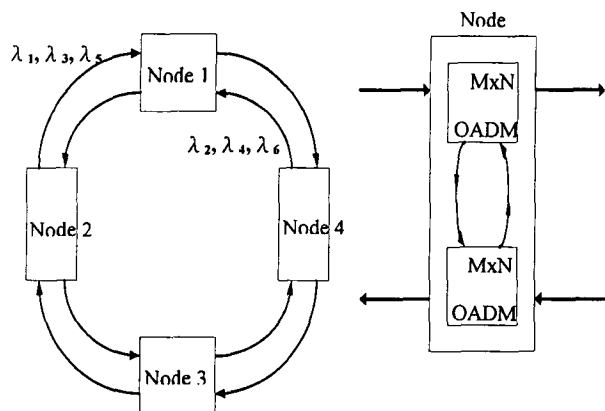
- Through A Fiber Laser”, Proceedings of Opto-Electronic and optical communication/Integrated optics and Optical Communication(OECC/IOOC 2001) Conference, pp.128-129, July 1-5, Sydney, Australia, 2001.
- [91] Shyh-Lin Tsao , Shien-Cheng Chiou and Che-Yuan Hsu, “Application of an optical circulator included fiber laser in CATV system”, Proceedings of Opto-Electronic and optical communication/Integrated optics and Optical Communication (OECC/IOOC 2001) Conference, pp. 433-434, July 1-5, Sydney, Australia, 2001.
- [92] Shyh-Lin Tsao, Zhang-Hong Tine and Yi-Jr Jou, “BPM Simulation on A Vertical Coupling Optical Switch Based on SOI Structure”, Proceedings of Opto-Electronic and optical communication/Integrated optics and Optical Communication (OECC/IOOC 2001) Conference, pp. 404-405, July 1-5, Sydney, Australia, 2001.
- [93] Shyh-Lin Tsao and Zhang-Hong Tine, “Design and Simulation of A Digital Optoelectronic Switch on Silicon-on-Insulator Wafer”, Proc. of The 46th SPIE’s Annual Meeting, ‘Design, Manufacturing, and Testing of Planar Optical Waveguide Decices’, Vol. 4439, California, U.S.A., 29 July-3 August, pp.1-9,2001.
- [94] Shyh-Lin Tsao and Yu Min Lu, “Design and comparison of 8x8 optical switches with adaptive wavelength routing algorithm”, to be appeared in Proc. of The 46th SPIE’s Annual Meeting, California, U.S.A., 29 July-3 August,2001.
- [95] Shyh-Lin Tsao, Shien-Cheng Chiou and Ta-Chun Lin, “ An active fiber-optic FIR filter included IIR optical filter”, to be appeared in Proc. of The 46th SPIE’s Annual Meeting, California, U.S.A., 29 July-3 August,2001.
- [96] Shyh-Lin Tsao, Ling-Fan Tsao and Yu-Ching Ni, “ Medical image transmission via Giga-Ethernet optical free space communication system with SOA”, to be appeared in Proc. of The 46th SPIE’s Annual Meeting, California, U.S.A., 29 July-3 August,2001.
- [97] Shyh-Lin Tsao and Ling-Fan Tsao, “Application of DSP Technique on Medical Image Transmission Though Gigabit Ethernet Optical-Fiber Communication System”, to be appeared in Proc. of The 46th SPIE’s Annual Meeting, California, U.S.A., 29 July-3 August,2001.

- [98] Shyh-Lin Tsao, Tai-Chi Liou and Yu-Chia Hsu, "Two-Way Optical Signal Processing Element S-Parameter Measurement", to be appeared in Proc. of The 46th SPIE's Annual Meeting, California, U.S.A., 29 July-3 August, 2001.
- [99] Shyh-Lin Tsao, Huang-Chen Guo, and Yi-Jr Chen, "An SOI X-Crossing Optical Switch", to be appeared in Proc. of The 46th SPIE's Annual Meeting, California, U.S.A., 29 July-3 August, 2001.
- [100] Shyh-Lin Tsao, Ming-Chun Chen and Hsin-Chun Huang, "Study of Linearly Tapered Waveguides on SOI for Optical Interconnect Waveguides to Fibers Coupling", to be appeared in Proc. of The 46th SPIE's Annual Meeting, California, U.S.A., 29 July-3 August, 2001.
- [101] Shyh-Lin Tsao, Hsin-Chun Huang and Peng-Chun Peng, "Analysis of A Mach-Zehnder Interferometer Switch With Vertical Coupling Filter Waveguide Structure", to be appeared in Proc. of The 46th SPIE's Annual Meeting, California, U.S.A., 29 July-3 August, 2001.
- [102] Shyh-Lin Tsao, Huang-Chen Guo and Yi-Jr Jou, "A Novel Wavelength Switch with 2 X 2 MMI SOI Photonic Crystal Inside", to be appeared in Proc. of The 46th SPIE's Annual Meeting, California, U.S.A., 29 July-3 August, 2001.
- [103] Shyh-Lin Tsao, Shien-Cheng Chiou and Che-Yuan Hsu, "Design of A Fiber-Optical Integrator for Optical Computing", to be appeared in Proc. of The 46th SPIE's Annual Meeting, California, U.S.A., 29 July-3 August, 2001.
- [104] Shyh-Lin Tsao, Shien-Cheng Chiou and Che-Yuan Hsu, "Implementation of A High Bandwidth Fiber Ring Laser for Subcarrier Microwave Communications" Proceeding of Asia-Pacific Microwave Conference (AMPC2001), Dec.3-6, Taipei, Vol. I, pp. 342-345, 2001.
- [105] Shyh-Lin Tsao, Shien-Cheng Chiou and Ta-Chun Lin, "Microwave Frequency Response Measurement of A Novel Fiber Laser", Proceeding of Asia-Pacific Microwave Conference (AMPC2001), Dec.3-6, Vol. I, pp. 346-349, Taipei, 2001.
- [106] Shyh-Lin Tsao, Hsin-Chun Huang and Peng-Chun Peng, "Design of A 1x4 MMI-MZI Electro-Optic Switch Using SOI Technology", Proc. of Electronics Devices and

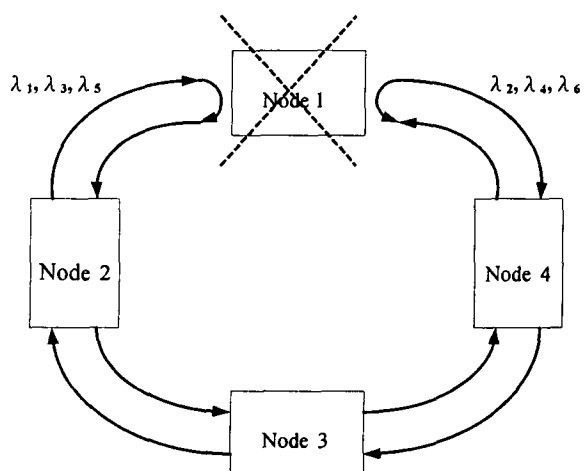
- Materials Symposia/Taiwan'01, Kaohsiung, Taiwan, Dec. 13-14, pp.455-458, 2001.
- [107] Hsin-Chun Huang, Shyh-Lin Tsao, Ming-Chun Chen, Jauh-Jung Yang, Heng-Chun Huang, "Analysis of SU-8 Polymeric Optical Waveguide for 1x4 Array Waveguide Grating", Proc. of Electronics Devices and Materials Symposia/Taiwan'01, Kaohsiung, Taiwan, Dec. 13-14, pp.679-682, 2001.
- [108] Shyh-Lin Tsao, Peng-Chun Peng and Chun-Yi Lu, "Design and Simulation of A 2x 16 Silicon-on-Insulator Multimode Interference Optical Power Splitter", Proc. of Optics and Photonics Taiwan'01, Kaohsiung, Taiwan, Dec. 14-15, Vol.I, pp.1031-1033, 2001.
- [109] Zhen-Yu Jiang and Shyh-Lin Tsao, "Design of A New Fiber-Optic Filter Based on IIR Filter Module"
- [110] Shyh-Lin Tsao, Ta-Chun Lin and Chung-Cheng Lee, "Study of An Internal Modulated Fiber Laser' Proc. of Optics and Photonics Taiwan'01, Kaohsiung, Taiwan, Dec. 14-15, Vol.I, pp.1118-1121, 2001.
- [111] Huang-Cuang Lin, Shyh-Lin Tsao and Shu-Tsung Kuo, " Application Fiber Ring Structures with Optical Semiconductor Amplifier in A Fiber-Optic CATV Network", Proc. of Optics and Photonics Taiwan'01, Kaohsiung, Taiwan, Dec. 14-15, Vol. I, pp. 608-610, 2001.
- [112] Che-Yuan Hsu, Shyh-Lin Tsao and Shien-Cheng Chiou, " Application of a two-outputs fiber laser in CATV transmission system", Proc. of Optics and Photonics Taiwan'01, Kaohsiung, Taiwan, Dec. 14-15, Vol. I, pp.1064-1066, 2001.
- [113] Shyh-Lin Tsao, Che-Yuan Hsu and Huang-Ping Chen, "Design of a fiber optical differentiator for optical computing", Proc. of Optics and Photonics Taiwan'01, Kaohsiung, Taiwan, Dec. 14-15, Vol. I, pp. 1149-1151, 2001.
- [114] Shyh-Lin Tsao, Huang-Chen Guo, Yi-Jr Jou, and Yi-Jr Chen, "A 1x2 1.3mm/1.55mm MMI SOI WDM Switch Device", Proc. of Optics and Photonics Taiwan'01, Kaohsiung, Taiwan, Dec. 14-15, Vol. I, pp. 920-922, 2001.
- [115] Shyh-Lin Tsao, Huang-Chen Guo, Yi-Jr Chen, and Yi-Jr Jou, "Design of A Coupler Switch with SOI Waveguides", Proc. of Optics and Photonics Taiwan'01, Kaohsiung,

- Taiwan, Dec. 14-15, Vol.I, pp. 923-925, 2001. Proc. of Optics and Photonics Taiwan'01, Kaohsiung, Taiwan, Dec. 14-15, Vol. I, pp.1051-1053,2001.
- [116] Hsin-Chun Huang, Shyh-Lin Tsao, Ming-Chun Chen, Jauh-Jung Yang, and Heng-Chun Huang, "Integrated Optical SU-8 Polymeric With Array Waveguide Grating", Proc. of Optics and Photonics Taiwan'01, Kaohsiung, Taiwan, Dec. 14-15, Vol. I, pp.942-944, 2001.
- [117] Shyh-Lin Tsao, Yu-Long Shen, Huei-Ju Tsai and Yu-Chia Hsu, "Implementation of a Two-Way Optical Polarization Measurement Analyzer", Proc. of Optics and Photonics Taiwan'01, Kaohsiung, Taiwan, Dec. 14-15, Vol. I, pp.1034-1037, 2001.
- [118] Ta-Chun Lin, Shyh-Lin Tsao and Shien-Cheng Chiou, "Analysis of a CDMA/CATV optical fiber communication system using fiber laser", Proc. of Optics and Photonics Taiwan'01, Kaohsiung, Taiwan, Dec. 14-15, Vol. I, pp.1067-1070, 2001.
- [119] Shyh-Lin Tsao, Huang-Ping Chen and Huang-ang Lin, "Transmission of CATV signals through a coupled fiber ring laser", Submitted to 7th Opto-Electronic and optical communication and Optical Communication (OECC 2002) Conference, Yokohama, Japan, July 8-12, 2002/1/31.
- [120] Shyh-Lin Tsao, Ta-Chun Lin and Chung-Cheng Lee, "Characteristic measurement of a novel fiber laser includes a phase modulator", Submitted to 7th Opto-Electronic and optical communication and Optical Communication (OECC 2002) Conference, Yokohama, Japan, July 8-12, 2002/1/31.
- [121] Shyh-Lin Tsoa, Hsin-Chun Huang, and Ji-Chung Chen, "Fabrication and Sumulation of SU-8 Polymeric Optical Waveguide for Array Waveguide Grating", Proc. of Asia SPM4 & Taipei Symposiun on Nanotech., Aug. 12-15, pp.106, 2002.
- [122] Shyh-Lin Tsao, Yu-Wei Tan, Tze-Yu Lin, and Hsin-Chun Huang, "Study of Optical Switch Based on Mach-Zehnder Interferometer with 100nm Thickness Vertical Coupling on SOI Wafer", Proc. of Asia SPM4 & Taipei Symposium on Nanotech., Aug. 12-15, pp.107, 2002.
- [123] Shyh-Lin Tsao, Hsin-Chun Huang, and Po-Ling Yeh, "Analysis of Optical Vortex Spinning in the

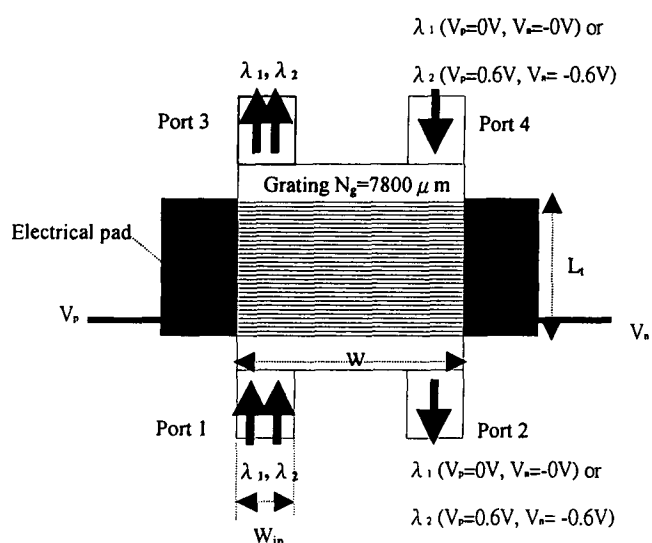
- Photonic Crystals on SOI Wafer", Proc. of Asia SPM4 & Taipei Symposium on Nanotech., Aug. 12-15, pp.108,2002
- [124] 中華明國專利,“偵測溫度及光源頻率裝置及其應用”
- [125] 美國專利,“Optical frequency and Temperature Sensor System”
- [126] 美國專利,“Highly Accurate Temperature Sensor Using Two Fiber Bragg Gratings”
- [127] 中華明國專利,“多輸出光纖環型雷射”
- [128] 中華明國專利,“光通信系統”
- [129] 中華明國專利,“偵測溫度及光源頻率裝置及其應用—追加專利”
- [130] 中華明國專利,“一種可調式光學編碼器與解碼器”
- [131] 美國專利,“Tunable Fiber Optical Encoder/Decoder”
- [132] 美國專利,“Optical Communication System”
- [133] 中華明國專利,“一種雙向光學元件網路分析儀探頭”
- [134] 中華明國專利,“以有機材料覆蓋之 Si/SiO₂/Si 結構積體光學波導耦合器”
- [135] 中華明國專利,“新型絕緣層上矽晶 1 對 32 光分歧器”
- [136] 中華明國專利,“衛星光通信系統之同調光接收器”
- [137] 中華明國專利,“一種微帶傳輸線分頻多工電路佈局”
- [138] 中華明國專利,“絕緣層上矽晶波導光學麥克森型溫度感測器”
- [139] 中華明國專利,“光纖捲繞式波長-功率轉換器”
- [140] 中華明國專利,“一種使用於 1.3 μ m 波段之寬頻光放大器及非對稱共振腔式波長可調光纖雷射”
- [141] 中華明國專利,“絕緣層上矽晶 1 對 2 之 1.3 μ m/1.55 μ m 光波長多工/解多工器”



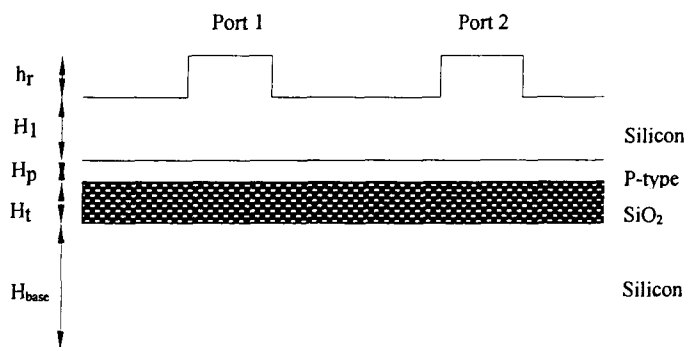
圖(一) 運用可調式光分叉復用器於光纖保護網路的示意圖



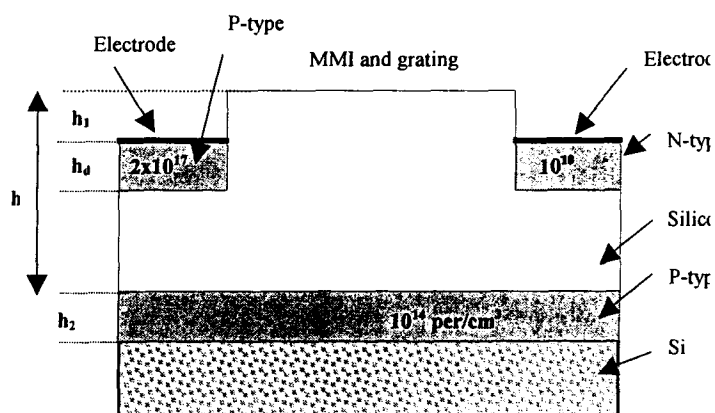
圖(二) 當節點一發生故障時，保護網路運作的方式



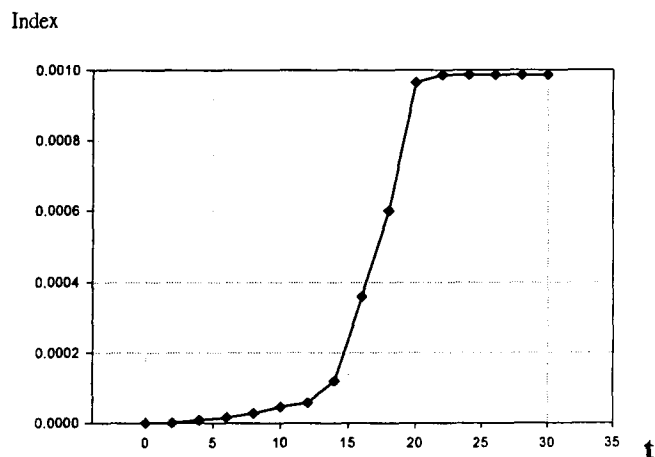
圖(三) 積體化可調式光分叉復用器的示意圖



圖(四) 積體化可調式光分叉復用器輸入埠和取出埠之剖面圖

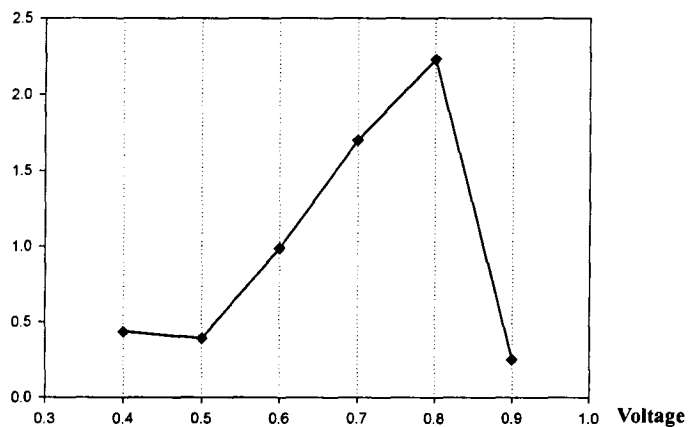


圖(五) 積體化可調式光分叉復用器之剖面圖及載子分布圖



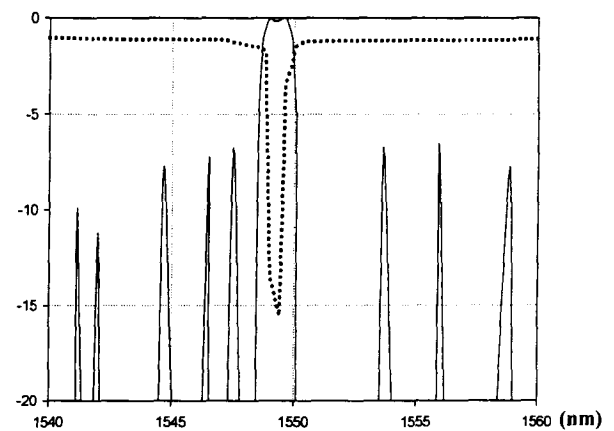
圖(六) 時間對折射率變化示意圖

Index changes



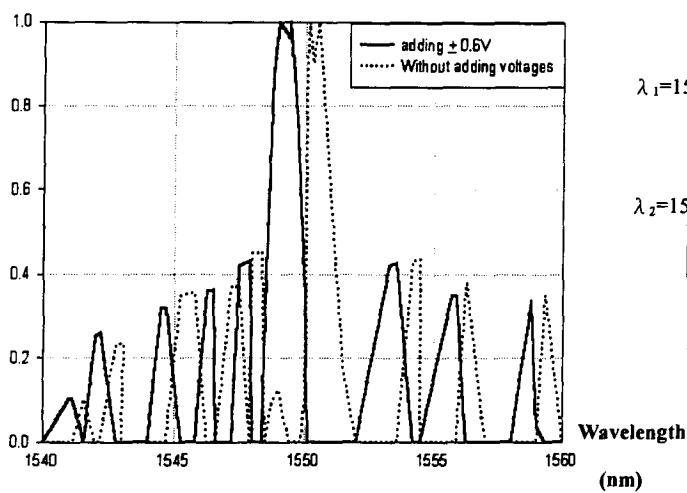
圖(七) 電壓對折射率變化示意圖

Intensity power

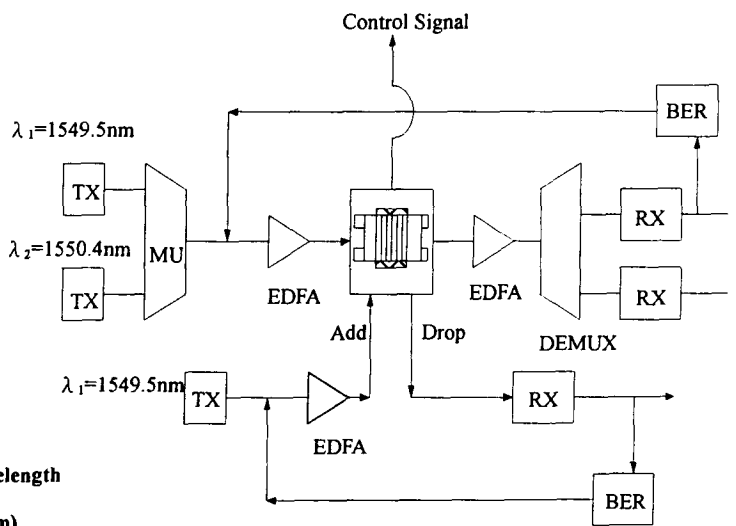


圖(十) 模擬在加電壓下，在取出阜(port 2)和輸出阜(port 3)所得信號強度比對波長關係圖

Intensity

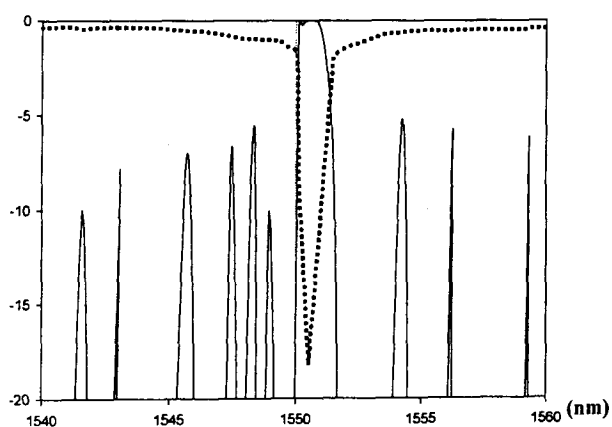


圖(八) 加電壓與未加電壓時，取出阜(port 2)濾出不同之波長

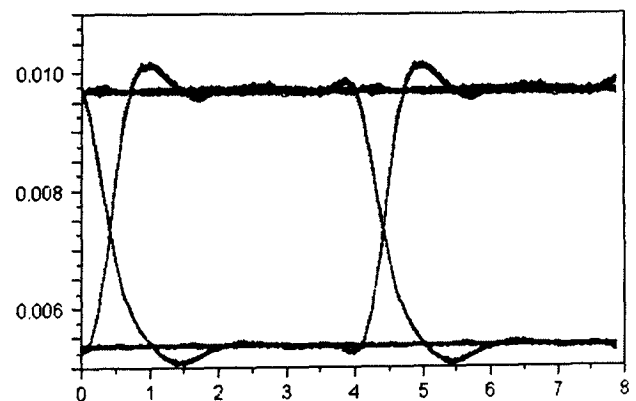


圖(十一) 運用可調式光分叉復用器於點對點光纖系統之示圖

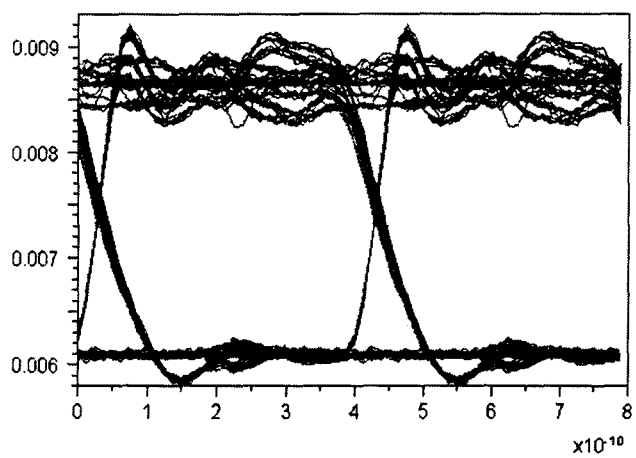
Intensity power



圖(九) 模擬在不加電壓下，在取出阜(port 2)和輸出阜(port 3)所得信號強度比對波長關係圖



圖(十二) 當 $\lambda_1=1549.5\text{nm}$ ，在取出阜(port 2)所模擬得之眼圖



圖(十三) 當 $\lambda_1=1549.5\text{nm}$ 且取出 λ_1 後，再上傳 λ_1 ，在輸出埠(port 3)所模擬得之眼圖