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UNIVERSITÉ PARIS-SUD

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DISCIPLINE PHYSIQUE

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par

Melissa ZIEBELL

Transceiver optique en silicium pour les réseaux d'accès

Directeur thèse :

Delphine MARRIS-MORINI

Université Paris-Sud (IEF)

Composition du jury

Rapporteurs :

Examinateurs :

Didier DECOSTER Andrea MELLONI Fréderic BOEUF Beatrice DAGENS Daniel DOLFI Abderrahim RAMDANE Unversité Lille 1 (IEMN) Politecnico di Milano STMicroelectronics Université Paris-Sud (IEF) Alcatel-Thales CNRS/LPN

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To my mother...

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Paris, July 3, 2013

M. Z.

Abstract

Silicon photonics is a research field in full expansion that works towards the integration of photonics and microelectronic components in a single chip. The creation of a broadband optical link that is able to carry high-speed information requires the development of efficient building blocks (laser source, modulator and photodetector) compatible with CMOS technology. Therefore, the work carried out during my Ph.D. focused specifically on silicon optical modulators for high-speed applications from 10 to 40 Gbit/s. The work presented includes design, optimization, fabrication and characterization of the complete device. The modulator is based on the electro-refractive effect obtained by depletion of carriers in PN and PIPIN diodes to obtain a phase change of the optical mode. Intensity modulation is obtained by integration of the active region in a Mach-Zehnder interferometer or a ring resonator. Electrical, optical and radio frequency simulations were conducted on the various elements of the modulator (active region, beam splitters, ...) to design a device for FTTH (Fiber To The Home) applications operating at 1.27 μ m. Additional studies included the design of RF coplanar waveguide electrodes optimized through the development of an electro-optical model that takes into account the propagation of the electrical and optical waves in the active region.

Compatible CMOS processes were proposed, and the necessary fabrication masks were designed considering fabrication tolerances and critical parameters. Finally, various experimental results were obtained on components designed at IEF and fabricated at CEA-LETI. We can specially mention a Mach Zehnder modulator operating at 40 Gbit/s that uses a PIPIN diode to obtain an index variation, and having an extinction ratio of 7.5 dB and losses of only 6 dB. The goal of future silicon modulators' generations is to integrate these devices with the RF driver, and to move towards more complex and efficient modulation formats than the two-level intensity modulation seen so far.

Résumé

La photonique silicium est un domaine de recherche en plein expansion, visant à intégrer la photonique avec la microélectronique. La réalisation d'un lien optique rapide pour transporter une information à haut débit nécessite le développement de blocs de base performants (source laser, modulateur et photodétecteur) et compatibles avec la technologie CMOS.

Dans ce contexte, les travaux menés durant ma thèse ont porté plus spécifiquement sur le modulateur optique en silicium, pour les applications haut débit, de 10 à 40 Gbit/s. Les résultats présentés portent notamment sur la conception, l'optimisation, la fabrication et la caractérisation de ce composant. Le modulateur est basé sur un effet électroréfractif obtenu par déplétion des porteurs dans des structures de type PN et PIPIN pour obtenir une variation de phase de l'onde optique. La modulation d'intensité est obtenue par intégration de la région active dans un interféromètre Mach Zehnder (MZI) ou un résonateur en anneau. Des simulations électriques, optiques et radiofréquences ont été menées sur les différents éléments du composant afin de concevoir un dispositif pour les applications FTTH (Fiber To The Home), fonctionnant à 1.27 μ m. Parmi les points clefs de ces études on peut noter l'optimisation des électrodes RF coplanaires réalisée grâce au développement d'un modèle électro-optique prenant en compte la propagation des ondes électriques et optiques dans la région active. Un procédé technologique compatible avec la filière CMOS a ensuite été mis en place et les masques nécessaires à la fabrication ont pu être dessinés en considérant les tolérances de fabrication et paramètres critiques. Enfin un grand nombre de résultats expérimentaux ont pu être obtenus. On peut notamment retenir un modulateur MZI fonctionnant à 40 Gbit/s utilisant une diode PIPIN et présentant un taux d'extinction de 7.5 dB pour des pertes de seulement 6 dB. Les futures optimisations des modulateurs silicium visent à les intégrer avec l'électronique de commande, et à aller vers des formats de modulation plus complexes mais aussi plus performants que la modulation d'intensité sur 2 niveaux considérée jusqu'à présent.

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Résumé en Français Part I

A Transceiver optique en silicium pour les réseaux d'accès

A.1 Introduction

La photonique silicium est un domaine en pleine expansion depuis quelques années. A terme, l'intégration de l'optique et de l'électronique sur les mêmes circuits doit permettre une augmentation des performances des circuits intégrés microélectroniques. En effet l'utilisation de liaisons optiques permettra de surmonter les limitations actuelles des circuits microélectroniques (puissance dissipée, latence, délai, crosstalk, etc...).

Un lien optique unique peut transporter une grande quantité de données avec peu de dissipation d'énergie. La réalisation de ce lien en silicium facilite grandement l'intégration avec la technologie CMOS de fabrication des circuits intégrés microélectroniques. Les télécommunications optiques bénéficieront également du développement de solution bas coût et haute performance pour les liens optiques rapides. Parmi les applications possibles, le FTTH (pour Fiber to the Home) est une technologie émergente qui est en mesure d'offrir un accès haut débit en apportant des fibres optiques directement au domicile des usagers.

Dans ce contexte, le but principal de mon travail a été de concevoir un modulateur optique en silicium pouvant être intégré dans un émetteur-récepteur silicium selon les exigences fixées par le projet *Transceiver silicium optique intégré à* 10 *Gbit/s pour les réseaux d'accès (SILVER)*, pour des applications FTTH à la longueur d'onde $\lambda = 1,27 \mu$ m. En outre, les résultats de la caractérisation d'un modulateur silicium conçu pour les applications à une longueur d'onde autour de $\lambda = 1,55 \mu$ m sont également présentés et utilisés comme base pour la conception du nouveau modulateur.

Un schéma d'un circuit photonique integré (PIC) est représenté dans la figure A.1. Le

modulateur optique est un des éléments clés d'une chaîne de communication optique. Son rôle est de transformer l'information sous forme électrique en information sous forme optique et il est donc responsable de la conversion correcte des informations transmises.



Figure A.1: Vue schématique d'un lien optique multi-longueur d'onde sur circuit intégré optique / électronique.

Specification	Valeur
Longueur d'onde (λ_0)	$1,27\mu\mathrm{m}$
Rapidité	10 Gbit/s
Structure optique	Inteferomètre Mach Zehnder (MZI) symétrique / assymétrique
Tension	< 5 V
$V_{\pi}L_{\pi}$ (@ 5V)	2 – 3 dB
f _{3dB}	10 GHz
ER	> 10 dB (for 10 Gbit/s)
IL	< 5 dB
Hauteur guide d'onde (H)	440 nm

Table A.1: Cahier des charges du modulateur optique pour les applications à λ = 1,27 μ m.

Le but du projet SILVER est la démonstration d'un circuit intégré photonique en silicium,

bas coût, fonctionnant à 10 Gbit/s. Le cahier des charges du modulateur est reporté dans le tableau A.1. Les points clés fondamentaux de la conception du modulateur incluant la région active, les MMI et les électrodes RF sont présentés ci-dessous, suivi d'une présentation synthétique des principaux résultats obtenus. L'ensemble de ces sujets est traité en détail dans le manuscrit en anglais.

A.2 Conception des modulateurs

Pour les applications 40 Gbit/s requises dans le cadre du projet <u>pH</u>otonics <u>EL</u>ectronics functional <u>Integration on CMOS</u> (HELIOS), un modulateur silicium basé sur la déplétion de porteurs dans une diode PIPIN dans la région active d'un interféromètre Mach Zehnder avait été conçu et optimisé théoriquement avant le début de ma thèse. Cette région active était le résultat d'un compromis entre efficacité, pertes et rapidité.

Pour les application à 1,27 μ m et 10 Gbit/s (cahier des charges du projet SILVER), j'ai tout d'abord mené une étude comparative des performances de la diode PIPIN et d'une diode PN, plus simple à fabriquer d'un point de vue technologique. Cette comparaison est faite pour l'intégration dans un guide d'onde de hauteur (H) 440 nm et de « slab » (h) 100 nm. Ces dimensions ont été imposées par les dimensions des autres éléments formant le transceiver, notamment le laser III-V sur silicium qui devra être intégré à la fin du projet. Les performances du modulateur basé sur la diode PIPIN ou sur la diode PN à la longueur d'onde de 1,27 μ m ont été étudiées en détail.



Figure A.2: Vues schématiques des diodes PIPIN et PN diodes comparées lors de la simulation, h = 100 nm et w = 420 nm.



Figure A.3: Variation d'indice effectif (Δn_{eff}) en fonction de la largeur du guide w d'une diode PN et d'une diode PIPIN (H = 440 nm, h = 100 nm).

Par exemple la variation d'indice effectif obtenue sous une tension de 5 V est reportée figure A.3 en fonction de la largeur du guide d'onde W. On peut constater qu'une plus grande variation d'indice effectif Δn_{eff} est obtenue avec la diode PN. Il est à noter que la largeur du guide doit rester inférieure à 410 nm, sinon le guide devient multimode (région en bleue sur la figure A.3. A partir de ces résultats, les performances de modulation en terme de pertes d'efficacité (VL_{π}) et de pertes d'insertion (IL) ont pu être calculées et sont reportées sur la figure A.4.

Les simulations indiquent que le facteur de mérite VL_{π} diminue pour des guides d'onde étroits, de telle sorte que la valeur la plus faible de VL_{π} (calculée à V= 5 V) a été obtenue pour le guide le plus étroit dans les simulations w = 320 nm. Le choix de cette largeur minimale dans les simulations (320 nm) repose sur l'estimation qu'au-delà, pour des largeurs plus petites, il est probable que les pertes optiques des guides passifs, dues à la diffraction sur les facettes gravées du guide deviennent trop importantes. Il est à remarquer que le VL_{π} se stabilise lorsque w est inférieur à 360 nm, et que diminuer plus encore la largeur des guides ne devrait pas améliorer énormément les performances de modulation de la région active du modulateur. Concernant les pertes optiques dues aux porteurs libres, celles-ci augmentent avec la largeur du guide (figure A.4[b]).



Figure A.4: (a) Efficacité de modulation (VL_{π}) et (b) pertes d'insertion (IL) en fonction de la largeur du guide pour une diode PN et PIPIN.

Les deux types de diode permettent l'obtention de pertes inférieures aux valeurs spécifiées. L'ensemble des résultats de simulation reportés figure A.4 nous ont permis de choisir la diode PN comme région active du modulateur optique du transceiver optique en silicium pour les applications de réseau local (FTTH). Ce choix repose sur le fait que les performances de la diode PN rentrent dans le cahier des charges du projet, tout en étant plus simple à réaliser d'un point de vue de la fabrication des composants.

Concernant les dimensions du guide d'onde, nous avons choisi une largeur de guide de 360 nm. Cette dimension permet un bon compromis entre les différents paramètres (pertes dues aux porteurs libres, VL_{π}) L'étape suivant a été l'optimisation des concentrations de dopants dans les régions p+ et n+. Les performances de la diode PN ont été calculées pour différentes valeurs de concentrations de dopants. Les facteurs de mérite VL_{π} et pertes d'insertion (IL) sont reportés sur la figure A.5.

A partir de ces simulations nous avons choisi une concentration d'ions accepteur dans la région p de la diode ($N_A = 5 \times 10^{17}$ cm⁻³ et une concentration d'ions donneurs dans la région n de la diode de $N_D = 7 \times 10^{17}$ cm⁻³. Ces valeurs correspondent aux régions ombragées des figures A.5 et A.6 et ont été choisies car VL_{π} et IL présentent de bonnes valeurs et relativement peu de variations dues à des variations de concentration de dopants, ce qui rend les performances des modulateurs peu sensibles aux variations technologiques lors de leur fabrication.

De plus, les trous créant une plus grande variation d'indice de réfraction que les électrons, la région de déplétion n'est pas centrée dans le guide d'onde, mais légèrement décalée, pour que le mode optique soit centré sur la région qui est déplétée en trous lors de l'application de la tension inverse sur la diode. Les simulations montrent que la variation d'indice effectif Δn_{eff} maximale est obtenue lorsque le plan de la jonction est décalé du centre du guide de 50 nm vers le côté *n*.



Figure A.5: (a) Efficacité de modulation (VL $_{\pi}$) et (b) pertes d'insertion (IL) en fonction de N_A pour une diode PN, pour différentes valeurs de N_D .



Figure A.6: (a) Efficacité de modulation (VL $_{\pi}$) et (b) pertes d'insertion (IL) en fonction de N_D pour une diode PN diode, pour différentes valeurs de N_A .

La vue schématique finale de la région active du modulateur est reporté figure A.7. La largeur totale de la région active dépend de la conception des électrodes RF, car les régions dopées s'étendent du guide d'onde jusqu'au métal des électrodes. Ainsi en fonction de l'optimisation des électrodes, cette largeur peut varier de quelques μ m jusqu'à environ ~ 50 μ m.

Lors de la conception de ces extensions latérales dopées, entre la région active et le métal, les deux points clefs à prendre en compte sont les pertes optiques et la résistance d'accès du composants R_D .



Figure A.7: Design complet de la région active du modulateur optique pour le transceiver à 1,27 μ m, avec les dimensions et concentrations de dopants optimisées.

La rapidité du modulateur optique dépend de différents facteurs : du temps de réponse intrinsèque de la diode, du temps de réponse de type RC de la diode et du substrat, et de la propagation radio fréquence (RF) du signal électrique le long des electrodes.

Le temps de réponse intrinsèque t_r response correspond au temps de transit des porteurs au travers de la région de déplétion lorsqu'une tension est appliquée sur la diode. Ce temps de réponse intrinsèque a été évalué dans nos composants autour de 1.6 ps. Concernant le schéma équivalent petit signal du modulateur, la région active est équivalente à une capacité C_D en série avec la résistance d'accès de la jonction R_D [1]. Pour maximiser la fréquence de coupure f_c , le produit $R_D \cdot C_D$ doit être minimisé. La résistance R_D est calculée en ajoutant la contribution de chacune des régions entre la région active et les électrodes. Une formule approchée, fonction de la distance D entre les deux électrodes a pu être obtenue par equation A.1 :
$$R_D = \left(5.6 + \frac{2.6 \cdot G}{20}\right) K\Omega \cdot \mu m \tag{A.1}$$

qui correspond à la résistance d'accès de la diode en $K\Omega$ par unité de longueur (en μ m). La capacité équivalente de la région active a également été modélisée. A -2.5 V, $C_D = 0.39 \text{ fF}/\mu\text{m}$. Avec ces valeurs, f_c varie entre 62 GHz et 32 GHz, en fonction des dimensions minimales et maximales des électrodes. Cette fréquence est bien au-delà de la valeur minimale nécessaire pour obtenir une opération à 10 Gbit/s.

De plus, le substrat silicium-sur-isolant (SOI) créée une capacité supplémentaire, en raison de la couche d'oxyde enterrée, formant une constante RC où R est la résistance du substrat non intentionnellement dopé. Sur la figure A.8, le schéma électrique équivalent des différentes régions est superposé à la vue schématique en coupe du composant. L'influence du substrat a été pris en compte lors des simulations des électrodes et de la propagation de l'onde RF le long le région active.



Figure A.8: Vue schématique des constantes de temps RC présentes dans le modulateur du projet SILVER, incluant les constantes de temps RC de la région active et du substrat SOI

Des solutions pour minimiser son influence ont été proposées comme l'utilisation de substrat avec une couche de silice enterrée épaisse, et dont la couche de silicium présente une forte résistivité.

Pour transformer la modulation de phase en modulation d'intensité optique, un interféromètre de Mach Zehnder (MZI) est utilisé. Les coupleurs / combineurs à base de MMI ont été optimisés dans le but de maximiser la transmission à 1,27 μ m. la structure finale est reportée sur la figure A.9. La transmission théorique est de 96 %, ce qui correspond à des pertes d'environ 0.3 dB, probablement lors du couplage entre la région élargie et les guides d'onde de sortie. La figure A.10 illustre la carte de champ à l'intérieur du MMI 1 × 2 optimisé.



Figure A.9: Dimensions du MMI optimisé, choisis comme coupleur et combineur pour le modulateur silicium à $\lambda = 1,27 \ \mu$ m.

Plusieurs géométries de MZI ont été conçues comme illustré sur la figure A.11. Des MZI asymétriques dans une configuration proche des configurations déjà étuidées à 1,55 μ m ont été proposés. De nouvelles structures ont également été proposées, à base de MZI symétrique, en configuration push-pull, et intégrant une région de compensation DC. Le tableau A.2 propose une comparaison entre le cahier des charges et le résultat des simulations concernant le modulateur optique silicium à base de déplétions de porteurs dans une diode PN et fonctionnant à 1,27 μ m.

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Figure A.10: Carte de champ du coupleur MMI 1×2 à $\lambda = 1,27 \mu$ m. Le MMI optimisé présente une transmission de 96.6% ce qui correspond à des pertes de 0.3 dB.



Figure A.11: Vues schématiques des modulateurs à base de MZM (a) symétriques et (b) asymétriques.

	Simulation Results	Project Requirements	
Spec.	PN Diode	SILVER	
λ_0	1,27 $\mu { m m}$	1,27 $\mu { m m}$	
Н	440 nm	440 nm	
w	360 nm	_	
IL	~ 3 dB	< 5 dB	
ER	> 30 dB	>10 dB	
VLπ	1,26 V·cm	$2 - 3 V \cdot cm$	
L_{π}	2,5 mm	_	
f_c	49,73 GHz	_	
R_D (pour $G = 20 \ \mu m$)	8,2 $K\Omega \cdot \mu m$	_	
C_D (à 2,5 V)	$0,39\mathrm{fF}/\mu\mathrm{m}$	-	

Table A.2: Figures de mérite du modulateur : comparaison entre le cahier des charges et le résultat des simulations.

A.3 Fabrication

La première génération de transceiver optique en silicium pour des applications de réseau local, nécessite l'intégration de modulateurs avec les autres éléments du transceiver (laser, détecteur). Le premier masque qui a été réalisé dans ce but prévoir la fabrication des composants individuellement pour tester leurs performances mais aussi des transceiveurs complets. Le réticule a été divisé en trois régions, dédiées au modulateur, aux photodiodes à avalanche (APD) et aux triplexeurs complet. La surface totale du réticule est de 7,3 mm × 7,3 mm , et une vue générale de l'agencement du réticule est reportée sur la figure A.12.

La zone réservées pour le test des modulateurs seul est de 7,3 mm × 4,45 mm, dans laquelle quinze composants de tests ont pu être placés : cinq modulateurs Mach-Zehnder (MZM) asymétrique et dix MZM symétriques ont été conçus, avec différentes longueurs de région active. Trois autres MZM ont été conçus pour être intégrés dans les triplexeurs du réticule, comme illustré sur la figure. A.12.

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La figure A.13 correspond à une vue zoomée de la partie concernant le test des modulateurs seuls. Chaque composant a été numéroté et une description des paramètres de chaque structure est reportée dans le tableau A.3.



Figure A.12: Vue schématique du réticule divisé en trois grandes régions dédiées au test des modulateurs, des photodiode à avalanche et des triplexeurs.

Les étapes principales pour la fabrication des modulateurs sont représentées schématiquement sur les figures A.14 et A.15. Les composants sont fabriqués sur un substrat haute résistivité avec une couche de silice enterrée de 2 μ m d'épaisseur d'épaisseur et une couche de silicium superficiel de 440 nm [2]. Quatre étapes de photolithographie sont nécessaires pour réaliser l'implantation des régions P + +, N + +, P+, and N+ (niveaux 1 à 4), suivi d'un dépôt de masque dur en silice. Une étape de lithographie est ensuite nécessaire pour graver la silice et définir le guide d'onde (niveau 5).Une couche de silice d'épaisseur ~ 700 nm est ensuite déposée dans le but de protéger la région active, suivie qu'un dépôt de métal, lithographie et gravure pour définir les électrodes métalliques (niveaux 6-7).



Figure A.13: Capture d'écran du fichier GDS pour la fabrication des modulateurs. Seule la partie pour le test des modulateurs indépendants est représentée.

I/O	Dispositif	Longueur (mm)	Alignement de la jonction	Interferomètre
1	Guide droit	-	-	-
2	MZI symétrique	0,8	Chevauché	1×2 MMI
3	MZI symétrique	0,8	Aligné	1×2 MMI
4	MZI symétrique	1	Chevauché	1×2 MMI
5	MZI symétrique	0,8	Separé	1×2 MMI
6	MZI symétrique	1	Aligné	1×2 MMI
7	MZI symétrique	1	Separé	1×2 MMI
8	Guide droit	-	-	-
9	MZI symétrique	2,5	Aligné	
10	MZI assymétrique	0,8	Aligné	1×2 MMI
11	MZI symétrique	2	Chevauché	1×2 MMI
12	MZI symétrique	2	Aligné	1×2 MMI
13	Guide droit	-	-	-
14	MZI assymétrique	2	Separé	1×2 MMI
15	MZI symétrique	2	Separé	1×2 MMI
16	Guide droit	-	-	-
17	MZI assymétrique	2	Aligné	1×2 MMI
18	MZI assymétrique	2	Chevauché	1×2 MMI
19	MZI assymétrique	2	Aligné	2×2 MMI
20	MZI assymétrique *	2	Aligné	1×2 MMI
21	MZI assymétrique *	2	Aligné	1×2 MMI
22	MZI assymétrique [*]	2	Aligné	1×2 MMI

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* MZMs utilisé dans les transceivers complets (région triplexeurs du reticule)

Table A.3: Liste des modulateurs inclus dans le réticule, détaillant le type de modulateur et les paramètres spécifiques, ainsi que la position des guides passifs dans le réticule.



Figure A.14: Principales étapes de fabrication des modulateurs (1/2).



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Figure A.15: Principales étapes de fabrication des modulateurs (2/2).

A.4 Électrodes RF

La propagation du signal électrique RF le long de la région active est un point clef pour déterminer les performances globales des modulateurs optiques silicium à haute fréquence. Les trois points à prendre en compte lors de l'étude des électrodes sont d'une part l'atténuation de l'onde électrique le long des électrodes métalliques qui peut diminuer la tension réellement appliqué au bout d'une certaine distance de propagation, d'autre part l'impédance de la ligne chargée par la région active du modulateur, qui si elle n'est pas adaptée à l'impédance de la source et de la charge peut entrainer des réflexions du signal électrique, enfin la différence de vitesse de propagation entre l'onde optique et électrique qui peut diminuer le déphasage subit par l'onde optique au bout d'une certaine distance de propagation.

Ces trois effets peuvent limiter la bande passante de fonctionnement du modulateur. Par conséquent, la conception des électrodes nécessite une modélisation minutieuse, précise et efficace pour atteindre simultanément une grande bande passante, ainsi que des pertes (optique et micro-ondes). Des électrodes à ondes progressives en configuration coplanaire sont utilisées, en configuration masse-signal-masse (GSG pour Ground-Signal-Ground) qui correspond à la configuration des sondes haute fréquence RF disponibles. Pour modéliser le comportement du modulateur, le dispositif complet a été transformé en un circuit équivalent pour simuler les pertes électriques, l'impédance, la vitesse de phase électrique, et la réponse RC.

Les paramètres à optimiser sont la largeur de la ligne de signal (W) et la distance entre la ligne de signal et l'une des électrodes de masse (G). Les études théoriques précédentes pour les modulateurs silicium utilisant des diodes PIPIN [3] avaient permis de définir deux couples de dimensions. Le premier couple permettait d'optimiser l'impédance caractéristique de la ligne chargée par la région active (50 ohms). L'autre couple avait été choisi pour maintenir des pertes électriques faibles, par une augmentation de la largeur de la ligne de signal W. Comme il sera montré plus loin, ces modulateurs fonctionnent à 40 Gbit/s. La caractérisation de ces échantillons a permis de valider le modèle de simulation théorique que j'ai mis en place, en permettant une comparaison entre les performances théoriques et simulées de ces structures.

Le circuit utilisé lors des simulations avec le logiciel Advanced Designed System (ADS) est représenté sur la figure A.16. La région active est divisé en sections de longueur *L_i*. L'élément ligne coplanaire (CPW) présent dans la librairie d'ADS' prend en compte le schéma équivalent *RCLG* des électrodes métalliques sur un substrat donné. La simulation porte sur la bande passante électrique, mais aussi sur un modèle de bande passante électro-optique que j'ai mis en place. Les paramètres importants tels que l'impédance caractéristique et l'indice de propagation électrique sont également modélisés.



Figure A.16: Vue d'une section et du circuit principal utilisé pour les simulations rf du modulateur et l'optimisation des électrodes.

Les paramètres S électriques mesurés expérimentalement et simulés sont présents sur les figures A.17 et figure A.18, pour des dispositifs de longueur de région active L = 0,95 mm et L = 4,7 mm. Une très bonne correspondance est obtenue, notamment pour le dispositif long. Un pic dans la transmission S_{21} est obtenu expérimentalement pour le dispositif court, à environ 15 GHz, qui est peut être causé par le fait que les deux lignes de masses sont physiquement indépendantes. De plus la courbe de réflexion S_{11} montre que la réflexion du signal électrique est inférieure à -10 dB illustrant une bonne adaptation de signal. Suite à cette confirmation de la bonne modélisation de la réponse électrique du modulateur, il a été possible de d'établir un modèle de calcul de la bande passante électro-optique du composant. Le code ADS utilisé pour la simulation peut être trouvé en appendice C.1. Les réponses électro-optiques simulées et mesurées sont comparées sur la figure A.19, pour deux longueurs de région active. Une parfaite correspondance est obtenue.



Figure A.17: Paramètres S du modulateur : simulation et mesure, L = 0,95 mm.



Figure A.18: Paramètres S du modulateur : simulation et mesure, L = 4,7 mm.



Figure A.19: Réponse électro-optique du modulateur : simulation et mesure (a) L = 0,95 mm et (b) L = 4,7 mm.

Le modèle de simulation présenté ci-dessus et illustré sur la figure A.16 a ensuite pu être utilisé pour concevoir les électrodes CPW du modulateur optique conçu pour être integé dans le transceiver pour les applications à $\lambda = 1,27 \ \mu$ m. Une étude comparative de toutes les combinaisons de dimensions G-W, de 5 μ m à 40 μ m a été réalisé pour des composants de longueur L = 1 mm et L = 2 mm. Le couple $G = 20 \ \mu$ m and $W = 20 \ \mu$ m a été choisi car il permet un bon compromis entre les 4 paramètres importants : la bande passante électrique (3 dB de S_{21}), la bande passante électro-optique, l'impédance caractéristique et la variation de vitesse de propagation n_{RF} de l'onde électrique.

Les principales simulations sont reportés sur la figure A.20 pour différentes longueurs (1 mm, 2 mm et 2.5 mm). La transmission électrique (S_{21}) indique des bandes passantes de 10, 7 GHz, 12 GHz, et 17, 2 GHz, respectivement pour les longueurs 1 mm, 2 mm, et 2.5 mm (figure A.20[a]). Le paramètre S_{11} indique une réflexion inférieure à -10 dB pour les trois longueurs (figure A.20[b]). La bande passante électro-optique est estimée à 15, 1 GHz, 17, 7 GHz et 26, 4 GHz respectivement pour les longueurs 2, 5 mm, 2 mm, 1 mm (figure A.20[c]). Pour toutes ces longueurs les bandes passantes sont donc suffisamment grandes pour permettre une modulation à 10 Gbit/s comme requis dans le cahier des charges du projet SILVER.



Figure A.20: Simulation des paramètres S (a) S_{21} , (b) S_{21} et (c) réponse électro-optique du modulateur SILVER avec dimensions des électrodes $G = 20 \ \mu m$ et $W = 20 \ \mu m$, $L = 1 \ mm$.

A.5 Mesures expérimentales

En parallèle à la conception d'un modulateur optique pour les applications réseaux d'accès à 1,27 μ m j'ai caractérisé un grand nombre de composants, fonctionnant à 1,55 μ m, dans différentes configurations. Seuls quelques résultats peuvent être présentés ci-dessous, l'ensemble étant détaillé dans le manuscrit. Des modulateurs optiques basés sur des diodes PIPIN intégrés dans les guides d'onde d'un Mach Zehnder asymétrique, et fonctionnant à 1,55 μ m ont été caractérisés. Les premiers résultats portent sur un modulateur dont la région active est de longueur L = 1.8 mm. Sa caractéristique courant/tension est reportée sur la figure A.21. Un bon comportement électrique redresseur est obtenu, avec un faible courant inverse et une tension d'avalanche autour de -11 V.



Figure A.21: Caractéristique I(V) d'un Mach Zehnder L = 1.8 mm.

La transmission optique normalisée du modulateur, pour une tension inverse allant de 0 à 8 V reportée figure A.22 indique des pertes optiques de -4 dB (niveau maximum de transmission). Un décalage du spectre vers les hautes longueurs d'onde est obtenu lorsque la diode est polarisée en inverse, en raison de l'augmentation de l'indice effectif n_{eff} dû à la déplétion de porteurs dans le bras le plus long de l'interféromètre [4]. Un déphasage inférieur à π est obtenu entre 0 V and -8 V, expliqué par une concentration de dopants réellement obtenue dans le guide inférieure aux valeurs théoriques. Le produit $V_{\pi}L_{\pi}$ a pu être estimé à 6 V·cm.



Figure A.22: (a) Spectre de transmission optique d'un modulateur Mach Zehnder, L = 1,8 mm, pour différentes tensions inverses entre 0 V et -8 V (b) zoom illustrant le point de fonctionnement optimal présentant de faibles pertes et un fort taux d'extinction.

Chapter A. Transceiver optique en silicium pour les réseaux d'accès

Ce modulateur a été caractérisé à 10 Gbit/s. Les diagrammes de l'œil ont été mesurées pour différents points de fonctionnement (figure A.26). Le taux d'extinction et les pertes ont été mesurés pour chaque point de fonctionnement, afin d'évaluer le compromis entre ces deux paramètres. Un taux d'extinction (ER) de 8,1 dB a été obtenu à 10 Gbit/s, avec des pertes optiques de seulement 6 dB.



(d) Diagramme de l'œil à 1564 nm, IL= 6 dB.

Figure A.23: Diagrammes de l'oeil à 10 Gbit/s du modulateur silicium avec une région active de longueur 1,8 mm, illustrant l'augmentation du taux d'extinction lorsqu'on balave les différents points de fonctionnement :(a) à 1561 nm, ER= 1,8 dB, (b) à 1562 nm, ER= 3,6 dB, (c) à 1562, ER= 6, 2 dB, et (d) à 1564 ER= 8.1 dB.

Le taux d'erreur par bit (BER pour Bit Error Rate) du modulateur silicium a été mesuré et compare avec celui d'un modulateur électro-optique commercial en Niobate de Lithium (LiNbO₃) (figure A.24). Un taux d'erreur inférieur à 10^{-9} est obtenu pour le modulateur silicium, avec une pénalité en puissance de l'ordre de 4 dB, probablement due à un taux

d'extinction réduit.



Figure A.24: Comparaison des taux d'erreurs par bit à 10 Gbit/s des modulateurs optique silicium et $LiNbO_3$

Une seconde génération de modulateurs fonctionnant jusqu'à 40 Gbit/s ont été caractérisés. Les meilleurs résultats ont été obtenus avec un composant de longueur 0,95 mm (MZ11) et 4,7 mm (MZ17). Les spectres de transmission optiques pour différentes tensions appliquées sont reportés sur la figure A.25. Les pertes « on-chip » (du guide d'entrée au guide de sortie) du modulateur court (0,95 mm) sont de 2,5 dB (figure A.25[a]) alors que celles du modulateur de 4,7 mm sont de 6 dB (figure A.25[b]). Le produit $V_{\pi}L_{\pi}$ peut être évalué à 3,5 V·cm, qui correspond à une amélioration par rapport aux dispositifs réalisés précédemment, mais qui est toujours plus faible que la valeur théorique de 1,7 V·cm. L'optimisation des dopages qui a été réalisée entre le première et la seconde fabrication est à poursuivre pour atteindre les valeurs optimales théoriques.



Figure A.25: (a) Spectre de transmission optique du modulateur MZ11 (L = 0,95 mm) pour des tensions inverses comprises entre 0 V and -8 V, dans l'encadré : zoom illustrant le taux d'extinction à un point de fonctionnement optimal. (b) Spectre de transmission optique du modulateur MZ11 (L = 4,7 mm) pour des tensions inverses comprises entre 0 V and -8 V, un déphasage égal à π - est obtenu à -8 V

La rapidité de fonctionnement des modulateurs a été illustrée par la mesure de diagrammes de l'œil à 40 Gbit/s. Une séquence pseudo aléatoire à 40 Gbit/s et délivrant une tension pic-pic de 6.5 V_{pp} est utilisée. Pour le modulateur le plus court, le taux d'extinction (ER pour Extinction Ratio) maximum est de 3.2 dB, (figure A.26[a]). Comme la région active est trop courte pour obtenir un déphasage de π , le point de fonctionnement est choisi à une longueur d'onde ne correspondant pas au maximum de transmission, mais légèrement décalé pour optimiser le taux d'extinction. Le choix de ce point de fonctionnement créé des pertes supplémentaires. Des pertes additionnelles de 2 dB ont été tolérées pour obtenir un taux d'extinction de 3.2 dB. Les pertes totales au point de fonctionnement est donc de 4.5 dB.

Pour le modulateur plus long (MZ17), un taux d'extinction de 7,57 dB a été obtenu (figure A.26[b]). Dans ce cas, le point de fonctionnement a été choisi au niveau maximum de la transmission optique, car un déphasage de π peut être obtenu en appliquant la tension inverse. Il n'y a donc pas de pertes optiques additionnelles qui sont donc de 6 dB comme illustré sur la A.25(b). De plus la saturation du signal visible sur le niveau bas du diagramme de l'œil (figure A.26[b]) est dû à une limitation expérimentale, due à l'émission spontanée résiduelle de l'amplificateur optique. Le taux d'extinction de 7.5 dB n'est donc probablement pas limité par le modulateur mais par le set-up expérimental [5]. Ce résultat est cohérent avec le spectre de transmission statique du MZ17, pour lequel un taux d'extinction supérieur à 15 dB est obtenu.



(a) Eye diagram of MZ11, L=0,95 mm.

(b) Eye diagram of MZ17, L=4,7 mm.

Figure A.26: Mesures de diagrammes de l'oeil à 10 Gbit/s du (a) MZ11 (L = 0,95 mm) présentant un ER= 3.2 dB, et (b) MZ17 (L = 4,7 mm) présentant un ER= 7.5 dB.

La réponse électro-optique des modulateurs MZ11 et MZ17 en fonction de la fréquence sont tracés sur la figure A.27 pour des tensions inverses appliquées de -2.5 V. Les bandes passantes électro-optique sont de plus de 40 GHz pour le MZ11, et 22 GHz pour le MZ17.



Figure A.27: Réponse électro-optique petit signal des modulateurs silicium (a) MZ11 (L = 0,95 mm) présentant une bande passante > à 40 GHz et (b) M17 (L = 4,7 mm) présentant une bande passante de ~ 22 GHz.

A.6 Conclusion

Les travaux décrits dans cette thèse couvrent l'étude complète d'un modulateur optique en silicium. Concernant la conception de la région active du modulateur, différents compromis apparaissent entre vitesse, pertes, et efficacité de modulation. Un modulateur a été optimisé en suivant le cahier des charges d'un transceiver pour des applications FTTH (projet ANR SILVER). Des résultats prometteurs sont obtenus, et pourraient conduire aux premiers modulateurs optiques de silicium fonctionnant à la longueur d'onde de 1,27 μ m, et pourraient être utilisés prochainement dans les réseaux FTTH en cours de développement. Un grand nombre de résultats expérimentaux ont également été reportés, à la longueur d'onde de 1,55 μ m, avec des démonstrations de modulateur optique rapide fonctionnant à 40 Gbit/s, présentant un fort taux d'extinction et de faibles pertes (6 dB).

La démonstration d'un modulateur optique silicium isolé n'est qu'une étape vers la démonstration de circuits intégrés photoniques performants. Les perspectives ouvertes par ce travail concernent d'une part la caractérisation des émetteurs-récepteurs silicium conçus, ainsi que la caractérisation de dispositifs complets (laser et modulateur). Les résultats obtenus à l'heure actuelle permettent d'envisager l'utilisation de ces composants dans les interconnexions optiques, pour lesquels une faible consommation électrique, devient un enjeu de taille. Une seconde perspective aux travaux présentés dans cette thèse concerne la diminution de la tension de commande, permettant à la fois de diminuer la consommation électrique mais également d'envisager comme but ultime l'intégration de ces dispositifs optiques avec les circuits électroniques de commande.

Dissertation in English Part II

Silicon Optical Transceiver for Local Network Access

1 Introduction

1.1 Background and Motivation

In the last few years, the continuous growth of data rates in microprocessors for web-based and networking applications and the corresponding larger capacity of data traffic systems have significantly increased the demand placed on contemporary telecom and data infrastructures. Current telecom and datacom networks transport information using light traveling through optical fibers as the communication medium for long-distance telecommunication links. These optical links are usually installed in the backbone and backhaul networks, while electrical wires are used in access networks for the final connectivity leg between providers and customers, due to the high cost of bringing the services to the end users' equipment. Fiber to the Home (FTTH) is an emerging technology that is able to offer customers highspeed access by bringing fiber links directly to their homes. Electronic Integrated Circuits (ICs) continue to improve data traffic systems by integrating more silicon Complementary Metal-Oxide-Semiconductor (CMOS) transistors onto a single chip by reducing feature sizes. However, higher transistor packing density leads to interconnect capacity problems. Power dissipation increases, and interconnect latency becomes an issue due to transmission delay in the long interconnects of ICs. Additionally, attenuation, signal reflection, crosstalk, timing, and dispersion have become increasingly difficult to circumvent with new generations of electronic ICs [6], While High-Performance Computing (HPC) systems have shifted towards partitioning microprocessors into multi-cores, these require an enormous level of complexity that will eventually reach the limits of low power consumption and high bandwidth data communication beyond 10 Gbit/s. There is therefore a need for a new viable replacement that

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meets the continuous high demand on larger capacity data transmission systems.

Optical links have the potential to circumvent the limitations imposed by microelectronic circuits. Photonic and optoelectronic devices based on integrated photonic circuits have grown such that they have started to dominate local network communications as well as long-distance communications and high-bandwidth computer peripherals, but they have also opened up new fields of application [7]. Optical interconnects offer multiple advantages. A single optical link can carry a vast amount of data with little power dissipation. In terms of interconnect capacity, optical links can use light beams, which can touch, cross paths or travel close together without electromagnetic interference, crosstalk, or noise.

Optical links are usually made with III-V semiconductor materials due to their excellent light transmission and absorption properties. However, they do not allow an easily implemented, cost-effective solution. Optical interconnects must achieve high performance at a reasonable cost. This requirement can be achieved using the already advanced infrastructure and CMOS processes of microelectronic devices to fabricate optimized optoelectronic components. Silicon is thus the preferred material to achieve this goal, as it would allow monolithic integration of several photonic and electronic components on a single substrate [8].

Silicon photonics can be used in a wide number of applications such as metro and long-haul telecommunications, datacom in centers and campuses, consumer applications (personal or commercial), high-performance computing, metrology, sensors, medical, and military applications. Figure 1.1 shows a plot of the different markets that can benefit from the use silicon photonics, the data rate needed for each market and their required operation distance [9].

In the last decade, the field of silicon photonics has expanded over the long-distance markets that are in constant need of communication links with enormous bandwidth requirements, such as: long-haul telecommunication applications; most recently, metropolitan telecommunications; and FTTH, which is currently in development in some countries. FTTH is an efficient and promising technology capable of offering a large set of bandwidths, up to several Gb/s, for individuals or small and medium enterprises, where the users can obtain affordable internet connection with data rates of 100 Mbit/s.

Optical fiber communications operate in two wavelength bands around 1.3 μ m and 1.5 μ m. The wavelength band around 1.3 μ m is used because at this wavelength there is no

dispersion and the losses are low. This band was originally used for long-haul telecommunication applications, but it was discontinued mainly because fiber amplifiers at 1.3 μ m are not powerful. However, it is still used for short-distance applications such as FTTH where optical amplifiers are not required. The wavelength bands around 1.5 μ m are the most widely used because in this region the silica losses are the lowest and Erbium-Doped Fiber Amplifiers (EDFAs) offer strong performance[10].



Figure 1.1: Data rate vs. distance of silicon photonics application markets [9].

The new technology of silicon photonics has substantial potential to blend optical technology with low-cost CMOS semiconductor processing. However, there are still many challenges to overcome. Low costs are achieved with high-volume production, but some of the main development problems are the low quantity of integrated product solutions available, as most of the industry has been focused on developing individual silicon photonic components and not integrated transceivers. Additionally, companies and research groups need to create their own Computer-Aide Engineering (CAE)/Computer-Aided Design (CAD) programs, as only a few CAE/CAD design tools begin to be commercially available.

Nevertheless, the development and integration of optical components in a Photonic Integrated Circuit (PIC) are already underway, and the technology will continue to expand in the near future. The latest results have shown silicon photonics' potential as a new technology

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enabling new generations of monolithic optoelectronic devices. Silicon photonics has long been at the research and development level, but recent industrial development has expanded globally. Figure 1.2 shows a forecast of estimated available silicon photonics markets over the next decade. This figure shows high expectations for the industry in the near future.



Figure 1.2: Silicon photonics total available markets [9].

Different building blocks are available to create individual optical components and PICs using CMOS technology, such as fiber coupling interfaces, waveguides, beam splitters, gratings, couplers, polarizers, interferometers, optical modulators, drive circuitries, optical detectors, light sources, and wavelength multiplexers, among others. A schematic diagram of a PIC is depicted in Fig. 1.3. The performance of every components is equally important for the performance of optical links.

One of the key components of optical interconnects is the optical modulator. It is a crucial element in fiber communication systems because it transforms the electrical signal to a light signal and is therefore responsible for the correct conversion of the transmitted information. This process is required for vast bandwidth and for transmitting information such as data or voice through a fiber medium with very low loss at communication bands. The development of the modulator subcomponents, high-efficiency splitters, combiners, and phase modulators are extremely important to satisfy the electrical-to-optical signal conversion requirements. There are many important and critical subcomponents of an optical modulator that guarantee its proper operation. In particular, the design and fabrication of the active region is one of the most crucial components of a silicon optical modulator, as this part is ultimately responsible for the extinction ratio, losses, and modulation efficiency of the device. The Radio Frequency (RF) traveling-wave electrodes are also a sensitive design component, as they govern the interaction between the optical and electrical signals. In addition, they are required to simultaneously achieve low driving voltage, large bandwidth, and low optical and microwave loss. [11].



Figure 1.3: Schematic diagram of a PIC in convergence with a CMOS integrated circuit (electronic-photonic integrated circuit).

1.2 Objective and Dissertation Outline

The main objective of this dissertation is to conceive and develop a high-speed silicon optical modulator to be integrated in a silicon transceiver that will be used for local network applications, specifically FTTH, at a wavelength of 1.27 μ m, as well as to characterize a silicon optical modulator for applications at the telecommunication wavelength band of 1.55 μ m. This work is meant to provide a detailed understanding of the considerations and techniques used for

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the complete design and fabrication of a high-speed silicon optical modulator, in order to introduce a robust device capable of keeping up with the requirements of future high-speed photonics links. This dissertation is divided into five chapters.

- Chapter 2 provides a theoretical review of the mechanisms and theories of optical modulation in silicon by detailing the different physical effects used to obtain modulation in silicon. The chapter also describes the optical and electrical structures used to design the silicon optical modulators of concern for this work, as well as the performance parameters used to evaluate their efficiency. Furthermore, the state of the art of silicon modulators developed in the last two years is presented, using comparison tables to show the different design structures and performance obtained by different research groups around the world. Chapter 2 concludes by introducing the two main projects that grounded this work.
- Chapter 3 presents the complete process followed to design the active region and interferometric structures of the silicon optical modulator for applications at 1.27 μ m.
- Chapter 4 describes the fabrication of the IC mask layout design as well as the process flows followed for the fabrication of both silicon optical modulators: namely, the silicon modulator designed for high-speed applications at the wavelength band around 1.55 μm, and the modulator designed to be integrated on a silicon transceiver used for FTTH applications at wavelength of 1.27 μm.
- Chapter 5 presents a study of the design of RF traveling-wave electrodes used in the Mach-Zehnder modulators (MZMs), including the transmission line theory to create an equivalent lumped circuit of the modulators, and the influence of the modulator's active region and the substrate on the electro-optical response. This chapter also presents a comparison between the experimental and simulated results of the electrical and electro-optical response of the complete device.
- Chapter 6 presents the characterization results of the silicon modulator designed for high-speed applications at 1.55 μ m and ends with a general conclusion of the work presented throughout this dissertation as well as a brief consideration of further study in this field.

2 Silicon Optical Modulation

The dominance of silicon as the semiconductor of choice for electronic applications has led to the investigation of silicon photonic circuits, mainly because of their potential interaction with the already mature CMOS technology [12]. This way, both electronic and photonic circuits can be grown monolithically from the same substrate, reducing overall costs. In view of this outlook, silicon photonics could revolutionize the optical communication industry.

The basic building blocks of an optical interconnect system are light sources, waveguides, detectors, and modulators, the latter being the most essential components of an optical transceiver. Optical modulation is the process of applying information to a beam of light by altering different parameters such as phase, frequency, polarization, or intensity under an applied electrical signal.

Silicon-based materials are the platform of choice for the fabrication of optical modulators because they could be integrated monolithically with the already mature and well-known CMOS technology infrastructure and processes. However, silicon is a frustrating optoelectronic material, as it exhibits weak electro-optic effects in telecommunication wavelengths ranging from 1.3 μ m to 1.55 μ m. This drawback has set off a study of alternate methods to find modulation effects, using different materials and structures.

This chapter reviews the main physical effects, materials, and structures used to obtain high-speed modulation, figures of merit calculated to determine the efficiency of an optical modulator, and a comparison table showing the state of the art of silicon optical modulators designed in the past few years. This review will illustrate the motivations and challenges supporting silicon as the preferred material for high-speed optical modulators using compatible
CMOS processes.

2.1 Physical Effects

Modulators can be classified as either electro-refractive or electro-absorptive. Applying an electric field to a material may change its real and imaginary refractive indices. A change in the real part of the refractive index (Δn) with an applied electric field is known as electro-refraction, whereas a change in the imaginary part of the refractive index ($\Delta \alpha$) is known as electro-absorption [13]. Such changes can be induced by applying external electric fields, injecting free carriers, propagating acoustic waves, or controlling temperature in materials. The following sections will focus on the most common different effects, including the Pockels and Kerr effects, the Franz-Keldysh effect (FKE), the Quantum Confined Stark effect (QCSE), and the plasma dispersion effect, the latter being the main effect of the silicon optical modulators discussed in this dissertation.

2.1.1 Pockels and Kerr Effect

The Pockels and Kerr effect is an electro-optic effect where an electric field applied to a material modifies its refractive indices and thereby its effect on polarized light [14]. The field-dependent change in refractive index is described by

$$\Delta\left(\frac{1}{n_r^2}\right) = rE + sE^2 + \dots$$
(2.1)

where n_r is the refractive index, and *E* is the strength of the electric field. When the first term of Eq. 2.1 is dominant, there is a linear electro-optic effect or Pockels effect, where *r* is the electro-optic Pockels coefficient [15]. The change depends on the direction of the electric field with respect to the crystal axes, for that reason, the Pockels effect occurs only in non-centrosymmetric crystals such as lithium niobate (LiNbO₃), gallium arsenide (GaAs), and indium phosphide (InP). However, the effect disappears completely in bulk silicon due to the inversion symmetric geometry of its crystal. Ongoing research within the last few years has shown that a significant linear electro-optic effect may be induced in silicon by breaking its cubic symmetry with the use of strain, by depositing a straining layer on top of it [16], [17], as

depicted in Fig. 2.1.



Figure 2.1: Strained silicon (a) silicon waveguide fabricated on top of an SOI wafer (b) the same waveguide with a straining layer deposited on top [16].

When the second term of Eq. 2.1 is dominant, the electro-optic effect depends on the square of the applied electric field and is called the Kerr effect, where *s* is the quadratic electro-optic Kerr coefficient. This effect is present in silicon, but theoretical results presented in [18] show that to achieve a considerable refractive index variation Δn_r (in a range between 10^{-4} and 10^{-1}) for a wavelength of 1.3 μ m, the required electric field exceeds the dielectric breakdown electric field for lightly doped silicon.

Modulators using Pockels and Kerr effects reach high speeds over 40 Gbit/s, with low power consumption. However, the materials used for these types of modulators—such as LiNbO₃, LiTaO₃, and KDP crystals—are costly, and cannot be integrated with CMOS technology.

2.1.2 Franz-Keldysh Effect (FKE)

A strong uniform electric field applied to a bulk semiconductor changes the complex dielectric constant of a semiconductor by tilting the energy bands toward the direction of the external field [15]. In general, an electron moves from the valence band to the conduction band by absorbing a photon with energy $hv = E_g$ in the absence of an external field, as shown in Fig. 2.2(a), but if the electron experiences an external field, the wavefunction of holes and electrons could extend to the bandgap between the conduction and valence bands. Electrons in the valence band are excited by photons in the conduction band with lower energy than the bandgap energy E_g that is absorbed at these locations (photo-assisted tunneling). This effect

is shown in Fig. 2.2(b).



Figure 2.2: (a) Equilibrium state and (b) Franz-Keldysh effect and tilting of energy bands in a uniform electric field [15].

The FKE can be used for two effects: the electro-absorption and the electro-refraction. If the energy of the incident light is near the bandgap E_g , the absorption spectrum shifts to longer wavelengths when a uniform electric field is applied. However, if the incident energy is smaller than the bandgap energy E_g , then the reactive index change is more dominant causing phase modulation. Electro-absorption modulators are designed with III-V direct bandgap semiconductors because the absorption coefficient changes very strongly in the vicinity of the band edge. In general, indirect semiconductors such as silicon and germanium result in a very weak absorption near the band edge, making them poor candidates for strong optical modulation.

The effect of the FKE was measured experimentally in silicon by [18] by calculating the change in the refractive index using the Kramers-Kroning relation on the measured electro-absorption spectra. The results showed that the greatest change of refractive index in silicon occurred at a wavelength of 1.07 μ m; the effect diminishes significantly at the telecommunications wavelengths of 1.3 μ m and 1.55 μ m [19]. Because of the FKE electro-absorption and

electro-refraction silicon modulators are not suitable for high-speed applications. However, germanium is often used as quasi-direct bandgap material due to its unique band structures [20], [21].

2.1.3 Quantum-Confined Stark Effect (QCSE)

A quantum well is a semiconductor structure of small bandgap energy sandwiched between a material of larger bandgap energy, and is typically no wider than 20 nm. A quantum well at equilibrium consists of symmetrically distributed electron and hole wave functions inside a well coupled together by coulomb forces to form excitons [22] as illustrated in Fig. 2.3(a).



Figure 2.3: Schematic diagram of the quantum-well energy bands and wave functions (a) at equilibrium state and (b) under a strong electric field [15].

If the exciton absorption is considered, the transition in optical absorption occurs at a photon energy lower than the bandgap energy [15]. Therefore, photon energy ($\hbar\omega$) emitted or absorbed in quantum well structures is given by

$$\hbar\omega = E_g + E_{el} - E_{hl} - E_{ex} \tag{2.2}$$

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where E_g is the bandgap energy of the well material, and E_{el} , E_{hl} , and E_{ex} are the electron sub-band energy, the hole sub-band energy, and the binding energy of the exciton, respectively.

A QCSE occurs when an electric field is applied perpendicular to the quantum well. The electron sub-band energy E_{el} decreases while the hole sub-band energy E_{hl} increases, reducing the energy difference between the two. At the same time, the binding energy E_{ex} also decreases. Thus, the electric field pushes the wavefunctions of an electron and a hole towards the well wall on opposite sides of the quantum wel, as depicted in Fig. 2.3(b), as the bandgap energy decreases and the band-edge is shifted towards longer wavelengths.

The QCSE is conventionally used in III-V semiconductor materials such as InGaAs and InAlAs because they induce larger refractive index change quantum well structures than in silicon-based materials [15]. However, the discovery of a strong QCSE in Ge/SiGe quantum wells in [22] has potentially opened the door for the development of reliable high-speed Ge/ SiGe quantum well devices [23].

2.1.4 Plasma Dispersion Effect

The plasma dispersion effect (also called the free carrier dispersion effect) depends on the free carrier concentration of a semiconductor material to change its optical absorption and refractive index coefficients. If free carriers are added to an undoped semiconductor material, the absorption coefficient increases, while the refractive index decreases. The opposite effect occurs if free carriers are removed from a doped semiconductor material.

Changes in the absorption ($\Delta \alpha$) of a semiconductor can be theoretically described by the Drude-Lorentz Eq. 2.3 [19]

$$\Delta \alpha = \frac{e^3 \lambda_0^2}{4\pi^2 c^3 \varepsilon_0 n} \left(\frac{\Delta N_e}{\mu_e m_e^{*2}} + \frac{\Delta N_h}{\mu_h m_h^{*2}} \right)$$
(2.3)

and the equation for the refractive index variation (Δn) change is given by Eq. 2.4

$$\Delta n = \frac{-e^2 \lambda_0^2}{8\pi^2 c^2 \varepsilon_0 n} \left(\frac{\Delta N_e}{m_e^*} + \frac{\Delta N_h}{m_h^*} \right)$$
(2.4)

where N_e , N_h , m_e^{*2} , m_h^{*2} , μ_e , μ_h are the free electron and hole densities, the effective masses of an electron and a hole, and the mobilities of electrons and holes, respectively. Soref and Bennett [18] used experimental results on the absorption spectrum of doped silicon to calculate the change in the refractive index implementing Kramers-Kroning relations, whose results were used to determine the relationship between carrier concentration variation and refractive index at wavelengths 1.3 μ m as seen in Fig. 2.4(a), and 1.55 μ m, as seen in Fig. 2.4(b). The plots presented in Fig. 2.4 exhibit a larger refractive index change due to free holes for $\Delta N < 2 \times 10^{19} \text{ cm}^{-3}$, which means that holes are more efficient in changing the refractive index of silicon. The experimental expressions to determine the refractive index and absorption coefficient changes in silicon due to carrier concentration variation are at $\lambda = 1.3 \ \mu$ m

$$\Delta n = \Delta N_e + \Delta N_h = -6.2 \times 10^{-22} \Delta N_e - 6.0 \times 10^{-18} (\Delta N_h)^{0.8}$$
(2.5)

$$\Delta \alpha = \Delta \alpha_e + \Delta \alpha_h = 6.0 \times 10^{-18} \Delta N_e + 4.0 \times 10^{-18} \Delta N_h \tag{2.6}$$

and at $\lambda = 1.55 \,\mu\text{m}$

$$\Delta n = \Delta N_e + \Delta N_h = -8.8 \times 10^{-22} \Delta N_e - 8.5 \times 10^{-18} (\Delta N_h)^{0.8}$$
(2.7)

$$\Delta \alpha = \Delta \alpha_e + \Delta \alpha_h = 8.5 \times 10^{-18} \Delta N_e + 6.0 \times 10^{-18} \Delta N_h \tag{2.8}$$

where $\Delta \alpha$ is expressed in cm⁻¹, and N_e and N_h are expressed in cm⁻³. Soref's equations Eq. 2.5 and Eq. 2.7 can be related to the Drude-Lorentz Eq. 2.4 by taking the ratio between the change of refractive indexes for given doping concentrations $\Delta n(1.55)/\Delta n(1.3)$. The result is approximately equal to the square ratio between both wavelengths ~ $(1.55/1.3)^2$ as shown in

Eq. 2.4. Soref's equations also show that the plasma dispersion effect allows for important changes of refractive index variation in silicon at the telecommunication wavelengths.



Figure 2.4: Refractive index change in Si with doping concentration (a) at $\lambda = 1.3 \ \mu m$ (b) at $\lambda = 1.55 \ \mu m$, [18].

In silicon modulators, the plasma dispersion effect is obtained with accumulation, injection or depletion of carriers to change the refractive index and absorption coefficient of silicon. High-speed operation in silicon is mostly obtained with MZMs or Ring Resonator (RR) modulators based on the carrier dispersion effects in PN or PIN diodes. The phase change is accomplished by injecting or depleting the carriers of a PN or PIN junction (phase shifter) that is embedded in both arms of a Mach-Zehnder Interferometer (MZI), or in the circular (ring) waveguide of a RR. Alternatively, a Metal-Oxide-Semiconductor (MOS) capacitor can be used as phase shifter [24] using carrier accumulation to obtain carrier dispersion in silicon. Nevertheless, carrier depletion is the most widely used and advanced approach of the three methods to design high-speed devices due to superior performance in terms of modulation efficiency, speed, and losses [25], [26], [27], [5]. The silicon optical modulators presented in this dissertation are based on the free-carrier dispersion effect. Phase shifters embedded in the arms of MZIs and RRs were used to change the refractive index of silicon using carrier depletion.

2.1.5 Additional Effects

Alternative modulation effects known to achieve modulation of silicon include the thermooptic and acousto-optic effects. In the *thermo-optic effect*, the refractive index of silicon changes with increasing temperature. The thermo-optic coefficient of silicon is given by $dn/dT = 1.86 \times 10^{-4} K^{-1}$, which leads to a refractive index change that is approximately 2 to 3 order larger than other effects. However, the thermo-optic effect is a slow mechanism, and controlling the temperature rise is challenging [19], [28].

In the acousto-optic effect, the change of refractive index occurs when an acoustic wave propagates in a medium, creating a perturbation of the refractive index such that light scatters at the traveling periodic refractive index grating created by the acoustic wave [14]. However, to produce a perceptible change of refractive index, the acoustic-optic coefficient of the material must be very high. Additionally, high-intensity optical fields may induce a change in the refractive index of a material by nonlinear effects [29].

2.2 Waveguides

The silicon photonic waveguide is the main basic element required to build a silicon optical interconnect system. The entire silicon photonics platform relies on a comprehensive understanding of such structures. This section explores some main waveguide topics that are relevant for a clear understanding of the devices designed in this dissertation.

2.2.1 Silicon-On-Insulator Substrate

Silicon photonic circuits require an optical waveguide system with features capable of adapting to passive and dynamic photonic devices with monolithic electronic-photonic integration. Consequently, waveguides need to be fabricated in silicon substrates using robust and compatible fabrication processes (CMOS), and materials [30]. To meet such requirements, silicon photonic circuits use SOI substrates for waveguide fabrication. SOI substrates are ideal surfaces on which to design waveguides where light propagation is primarily confined to the high refractive index medium. An SOI substrate consists of a silicon dioxide (SiO₂) layer (also called [Buried Oxide (BOX)]) sandwiched between the silicon substrate and a crystalline silicon layer on top. Ideally, the BOX layer is thick enough to prevent light leakage into the silicon substrate. The thicknesses of the BOX and the crystalline silicon layers are usually 1-2 μ m and 200 – 400 nm, respectively. There are five types of waveguides as shown in Fig. 2.5: namely, plane waveguide, strip waveguide, rib (or ridge) waveguide, strip-loaded waveguide, and buried strip waveguide [31].



Figure 2.5: Types of dielectric WGs (a) Thin-film waveguide, (b) strip waveguide, (c) rib (or ridge) waveguide, (d) strip-loaded waveguide, and (e) buried strip waveguide.

From all the five waveguide configurations shown in Fig. 2.5, the strip-loaded waveguide shown in Fig. 2.5(d) is the most complex structure to fabricate. The raised and buried strip waveguides shown in Fig. 2.5(b) and Fig. 2.5(e), respectively, are not suitable for high-speed applications, as they require electrode placement near the active region, which contributes to considerably to losses of the complete device, as it will be studied throughout this dissertation. The thin-film (or slab) waveguide shown in Fig. 2.5(a) does not offer acceptable optical confinement for design of a high-speed optical modulator. Finally, the rib waveguide shown in Fig. 2.5(c) offers both a simple structure to fabricate as well as an optimal configuration to achieve maximum optical confinement for the envisioned application. All the devices discussed in this dissertation, are based on rib waveguides.

2.2.2 Optical Confinement

The Filling Factor (FF) of a region *R* is defined as the fraction of the mode power flux confined and guided over the region *R* (usually the waveguide core or a section in the middle of the core) of the waveguide to the total power flux, as illustrated in Fig. 2.6. The FF (Γ_{FF}) is given by



Figure 2.6: Filling factor of a rib waveguide.

$$\Gamma_{FF} = \frac{\int\limits_{0}^{R} P_Z(s) \, ds}{\int\limits_{-\infty}^{\infty} P_Z(s) \, ds}$$
(2.9)

where P_Z is the power flux in the direction of light propagation (*z* direction), and where the power is the Poynting vector $P_Z(x, y) = (E \times H^*) \cdot z$. The FF is an important parameter because it provides a modal analysis of the transmitted light inside a waveguide core. The calculation of

the FF is used to find the waveguide performance by measuring the amount of light confined in a specific region *R*.

Additionally, the FF depends on the waveguide dimensions, such as rib height H, slab height h, waveguide with w, as well as on the wavelength, where, according to modal theory, shorter wavelengths are more confined inside a waveguide than longer wavelengths. The calculation of the FF provides an insight on the waveguide design, as it requires the careful optimization of the FF to ensure maximum light confinement in the desired region.

2.3 Optical Structures

Silicon optical modulators can be designed using a great variety of optical devices, effects, and materials according to the targeted application. The silicon optical modulators presented in this dissertation use the free-carrier dispersion variation effect to change the refractive index of silicon. This change in refractive index causes a phase shift of the guided mode, which must be converted to intensity variation using a combination of interferometric structures so that the optical modulation can be measured.

The two optical structures most widely used to design silicon optical modulators are the RRs and the MZIs. This section will present the operation principle and characteristics of both the RR and the MZI, but it will also include the operation principle and design parameters of Multimode Interferometer (MMI) couplers. These devices are the main focus of this section as they are an important part of the design of the optical modulators studied in this dissertation.

2.3.1 Ring Resonator

A Operation Principle

A ring resonator consists of a circular-shaped waveguide that is placed near a straight waveguide. The phase shifter is embedded in the circular waveguide. Light traveling along the straight waveguide is coupled in the ring and circulates many times around the closed path. The light beam is coupled back to the straight waveguide, as depicted in Fig. 2.7. An incoming light beam with wavelength λ_0 propagates along the straight waveguide of the resonator. When the light reaches the coupler, a small power fraction is coupled into the resonator, while the rest of the light continues on its path along the straight waveguide. The coupled light interferes constructively to build up a large intensity within the ring, while at the same time the transmitted light at the output waveguide decreases. The constructive interference in the ring is given by

$$n_{\rm eff}L = m\lambda_0, \quad \text{for} \quad m \in N$$

$$\tag{2.10}$$

where *L* is the circumference of the ring $2\pi r$, and n_{eff} is the effective refractive index of the guided mode in the ring. By tuning the effective refractive index of the ring waveguide, the resonance wavelength is modified, which induces a strong modulation of the transmitted signal.



Figure 2.7: Schematic diagram of the coupling principle of a ring resonator.

Light circulation within the ring can be described as [32]

$$a_2 = \tau \exp(i\phi)b2 \tag{2.11}$$

where ϕ is the round-trip phase shift, and τ is the transmission factor. By solving a set of equations that describe the relationship between the coupling of light in and out of the

resonator, the intensity transmission is given by [32]

$$T = \left|\frac{a_2}{a_1}\right|^2 = \frac{(\alpha - |t|)^2}{(1 - \alpha |t|)^2}$$
(2.12)

where α is the loss factor, and t the transmission coefficient. Equation 2.12 shows that when the internal losses in the ring are equal to the coupling losses, $\alpha = |t|$, the total transmission drops to zero. At this condition, the ring is critically coupled due to perfect destructive interference at the output waveguide. If $\alpha = < |t|$, then the ring resonator is under-coupled, and for $\alpha = > |t|$ the ring resonator is over-coupled.

The carrier dispersion effect is used in RR modulators using a PN or PIN junction embedded in the ring waveguide, where the refractive index of the ring is modulated by applying an electrical signal to the ring [33], as illustrated in Fig. 2.8. When bias is applied to the phase shifter, the optical mode propagating in the ring varies, creating a shift of the resonance wavelength. In the figure a shift of the transmission spectrum is seen when voltages vary from 0 V to 4 V



Figure 2.8: (a) Schematic diagram of a RR silicon modulator based on the carrier dispersion effect [33] and (b) Spectra of the RR modulated optical transmission with applied voltage.

In general, RR are versatile elements with small footprints that are widely used in applications ranging from filters to sensors and biosensors, and for telecommunications. The operation principle of RR has been used to create resonators with different shapes and designs, including racetracks, disk, and Manhattan configurations; RR can also be cascaded, and they can use vertical or lateral coupling with the straight waveguide. The resonant wavelength of a RR can also be tuned using other physical effects described earlier in this chapter to change the effective refractive index of the ring.

B Figures of Merit

The spectral performance of RRs can be measured in terms of Free Spectral Range (FSR), Full-Width at Half Maximum (FWHM), finesse and Quality Factor (Q). The FSR is the wavelength difference $\Delta\lambda$ between consecutive resonance dips, as shown in Fig. 2.9, and is given by [34], [35]



Figure 2.9: Transmission spectra of a ring resonator showing the FSR, and the FWHM.

$$FSR = \frac{\lambda_0^2}{n_g L}$$

(2.13)

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where n_g is the group refractive index, and L is the perimeter of the ring. It can be seen in Eq. 2.13 that the FSR is inversely proportional to the size of the RR, thus, to achieve a large FSR, the ring must be small. The FWHM measures the sharpness of the resonance, and is defined as the width of the 3-dB resonance peak. The finesse of the ring is the ratio of the FSR and the FWHM of the resonance for a specific resonance wavelength; and it is given by

$$F = \frac{FSR}{FWHM}$$
(2.14)

Finally, the Q factor is the ability of a cavity to confine the field. It measures the sharpness of the transmission peak and is defined as

$$Q = \frac{\lambda_0}{FWHM} \tag{2.15}$$

The advantage of using RR for silicon optical modulators is the compactness of this structure (typical dimensions are in the order of a few dozens of microns) because it could be integrated with current microelectronic circuits. However, the disadvantages of these structures lie in their high sensitivity to small temperature variations and fabrication tolerances, as well as their short spectral bandwidth.

2.3.2 Mach-Zehnder Interferometer

A Physical Principle

A MZI consists of input and output waveguides, a beam splitter, a beam combiner, and phase shifters (at least one), as seen in Fig. 2.10. An incident optical signal splits into two beams at the beam splitter. The two optical waves travel across the phase shifters that are embedded in the arms of the MZI. Modulation of the output signal occurs by applying a voltage to one or both arms of the MZI.



Figure 2.10: Schematic diagram of a MZI.

A phase difference occurs between the two optical waves when the signals interfere, producing an intensity modulation of the output beam. The intensity varies when the phase changes from 0 to π .

The phase variation caused by refractive index variation at the output waveguide of the MZI is given by

$$\Delta\phi = \frac{2\pi\Delta n_{\rm eff}L}{\lambda_0} \tag{2.16}$$

where Δn_{eff} is the effective refractive change difference between both MZI arms, and *L* is the phase shifter length. The normalized output intensity can be written as [36]

$$I = I_0 \cos^2\left(\frac{\pi}{\lambda} \Delta n_{\rm eff}L\right) \tag{2.17}$$

where I_0 is the amplitude of the output intensity. In the case of an asymmetric MZI the arm length difference needs to be considered. Then, the normalized output intensity becomes

$$I = I_0 \cos^2 \left[\frac{\pi}{\lambda} \left(\Delta n_{\rm eff} L + n \Delta L \right) \right]$$
(2.18)

where ΔL is the arm length difference.

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The MZI is one of the most widely used interferometers to obtain modulation in silicon due to its advantages over other interferometers, including operation over a great range of wavelengths, as well as immunity to temperature and process fluctuations.

B Figures of Merit

The modulation performance of a silicon optical modulator can be determined from the response characteristics of the strength of interaction between the optical and electrical signals. These parameters are commonly known as Figures of Merit (FoM). One of the most crucial FoM when using MZMs is the modulation efficiency ($V_{\pi}L_{\pi}$), where V_{π} and L_{π} are the applied voltage and phase shifter length required to obtain a phase shift difference between two optical waves of π . This FoM is used to determine the length of the phase shifter.

$$L_{\pi} = \frac{\lambda}{2\Delta n_{\rm eff}(V_{\pi})} \tag{2.19}$$

where Δn_{eff} is the change of the effective refractive index at V_{π} . Typical V_{π} and L_{π} values are in the order of a couple volts and mm, respectively.

The Insertion Loss (IL) is the ratio of the optical transmission launched at the input of the MZM I_{IN} to the optical transmission measured at the output I_{ON} (Eq. 2.20), and the Extinction Ratio (ER) is the efficiency at which the transmitted optical power is modulated, as shown in Fig. 2.11. It is defined as the ratio of the transmitted power in the I_{ON} state to the transmitted power in the I_{OFF} state (Eq. 2.21). There are two different ERs to consider, the static (DC) ER, which is measured with a sole DC electrical signal applied to the modulator, and the dynamic ER, which is measured when an RF signal is added. The static ER is measured from the DC transmission spectra of the modulator, while the dynamic ER is evaluated between the high and low levels of an eye diagram measurement [37].



Figure 2.11: IL and ER.

$$IL = 10\log_{10}\left(\frac{I_{ON}}{I_{IN}}\right) \tag{2.20}$$

$$ER = 10\log_{10}\left(\frac{I_{ON}}{I_{OFF}}\right) \tag{2.21}$$

The optical transmission of an asymmetric MZI as a function of the wavelength is depicted in Fig. 2.12. The figure shows the transmission shift with applied voltage as well as the the measurements of the IL and the ER (in DC). Only the optical transmission of an asymmetric MZM is shown here because the experimental results of all the MZMs presented in this dissertation were designed with asymmetric arms of $\Delta L = 40 \ \mu$ m.

This ΔL was calculated from the experimental setup to obtain a FSR that would allow visualization of at least three periods of the optical transmission on the computer screen. The transmission spectra shift of a symmetric MZI is determined by the difference in the transmission level with applied voltage, and it does not depend on the wavelength.

A possible drawback of using a MZI for silicon photonic applications is their size when compared to the considerably smaller RRs. However, the choice of interferometric device relies on the application, as size might not be a real concern for many applications where temperature stability or larger spectral bandwidth is paramount.



Figure 2.12: Transmission spectra of a MZI.

The light launched at the input waveguide of a MZIs used for silicon optical modulators must be equally divided between its arms, and properly combined so that the maximum amount of light can exit the structure. Consequently, the design of the waveguide beam splitter and combiner of a MZI must be cautiously selected and optimized because these can be extremely sensitive to waveguide size and application wavelength. Various waveguide splitters have been used for this task, including y-junctions, star couplers, or MMI couplers, among others, according to the application, device size, or wavelength for which the MZI has been designed. MMI couplers and star-coupler-based beam splitters [38] were used to design the MZMs presented in this dissertation. Nevertheless, the MMI coupler is of particular interest, as it was the interferometric structure used for the design of the MZMs presented in Chapter 3. Its principle of operation is presented in the next.

2.3.3 Multimode Interference (MMI) Coupler

An MMI coupler is a photonic structure that can be used in applications involving light coupling, splitting, and combining, as well as in switches and wavelength multiplexing/demultiplexing. Light is sent into and recovered from a central waveguide region using single-mode waveguides placed at the beginning and end of the MMI. The central waveguide structure employs properties among the propagation constants of multimode waveguides, usually more than three modes (Figure 2.13(a)), to obtain self-imaging at particular distances along the propagation direction of the guide. An $N \times M$ MMI coupler consists of N input waveguides and M output waveguides. Figure 2.13(b) shows the schematic diagram of the top view of a 1×2 MMI coupler of length L_{MMI} , width W_{MMI} , input waveguides W_{IN} , output waveguides W_{OUT} , and distance between the center of the output waveguides D.



Figure 2.13: (a) Lateral field profiles of the first 7 guided modes of a multimode wavguide. Re-digitized from [39]. (b)Schematic diagram of the 1×2 MMI couplers with length L_{MMI} , width W_{MMI} , input waveguides W_{IN} , output waveguides W_{OUT} , and distance between the center of the output waveguides D.

There are several approaches for the operation analysis, one of which includes the Multimode Propagation Analysis (MPA) that will be used to analyze the MMIs designed in this

work. The MPA offers easy understanding into the mechanisms of multimode interference. Using the self-imaging theory given in [39], the propagation analysis is often illustrated for a two-dimensional (*xz*-plane) approximation of the multimode waveguide as it usually has a significant ratio of width to height. The behavior of all modes in the vertical direction is considered equal due to the vertical confinement [19]. Using the guided-mode propagation analysis and the general interference theory developed in [39], the beat length between the two lowest-order modes is given by

$$L_{\pi} \approx \frac{4n_r W_{MMI}^2}{3\lambda_0} \tag{2.22}$$

where λ_0 is the free-space wavelength, n_r is the ridge refractive index (core), and W_{MMI} is the width of the multimode waveguide. The propagation constants have an approximately quadratic dependance on the mode number v, and the spacing between the propagation constants of the fundamental and higher-order modes can be expressed by [39] as

$$\beta_0 - \beta_v = \frac{v \left(v + 2\right) \pi}{3L_{\pi}} \tag{2.23}$$

An input field $\Psi(x, 0)$ at z = 0 and completely contained within W_{MMI} does not excite unguided modes. The input field can produce a profile at a distance z. Therefore $\Psi(x, 0)$ can be expressed as the superposition of all the guided mode fields

$$\Psi(x,L) = \sum_{\nu=0}^{m-1} c_{\nu} \psi_{\nu}(x) \exp\left[j \frac{\nu(\nu+2)\pi}{3L_{\pi}}L\right]$$
(2.24)

where c_v is the field excitation coefficient. Properties of the mode phase factor determine the type of images formed. It can be seen from Eq. 2.24 that a self-image (at the same location in the lateral direction) of the input field can be completely reproduced when

$$\exp\left[j\frac{\nu(\nu+2)\pi}{3L_{\pi}}L\right] = 1 \quad \text{or} \quad (-1)^{\nu}$$
(2.25)

These conditions are fulfilled when

$$L = p\left(3L_{\pi}\right) \tag{2.26}$$

where *p* is an integer. When *p* is even, the image will be direct, and when *p* is odd, the image formed is a mirrored self-image (at the opposite location in the lateral direction) with respect to x = 0.

If the first condition of Eq. 2.25 occurs (the exponent term equals to 1), then the phase changes of all the guided modes differ by integers multiples of 2π , and therefore, all the modes interfere with the same relative phases as in z = 0, producing a direct replica of the input mode. When the exponent term of Eq. 2.25 follows the second condition $(-1)^{\nu}$ occurs, the phase changes are alternatively even and odd multiples of π . In this case, the even modes will be in phase and the odd mode in anti-phase. The self-imaging principle produces three different types of self-imaging [40].

- A 1 × N MMI coupler has waveguide symmetry because it is fed by a single input waveguide located in the middle of the multimode waveguide at *d* = 0 (symmetric interference). Therefore, symmetric guided modes are excited, and self-imaging occurs at a distance *L* = 3*L*_π/4, and the *N*-fold image is found at *L* = 3*L*_π/4N as shown in Fig. 2.14(a).
- For a 2 × *N* MMI coupler, the coefficients of the 2nd, 5th, 8th, etc. modes are zero when the guided modes are excited by an input field that is placed off the waveguide center at approximately $d = \pm W_{MMI}/6$ (restricted self-imaging). The self-image, mirrored image, and *N*-fold images are found at $2L_{\pi}$, L_{π} , and L_{π}/N respectively.

• In an $M \times N$ MMI coupler, the coefficients of modes are non-zero when the MMI coupler is fed by one of M input ports anywhere within the range $-W_{MMI}/2 < d < W_{MMI}/2$ (general self-imaging) . From Eq. 2.25, the self-images or mirrored images (at the same location in the lateral direction) are found at distances $p(L_{\pi})$. When p is an even integer the self-images are found, and when p is an odd integer, the mirrored images are found as depicted in Fig. 2.14(b). The N-fold images are located at $L = p(3L_{\pi})/N$, for $p \ge 0$ and $N \ge 1$.

MMI couplers were chosen to split and combine the light in the MZIs-based silicon optical modulators designed for local network applications at $\lambda = 1.27 \ \mu$ m. The principle design of this structure for our application will be developed in Chapter 3.



Figure 2.14: (a) Symmetric interference showing access waveguide modes at z = 0, self-image at $3L_{\pi}/4$, 2-fold image at $3L_{\pi}/8$ and a 3-fold image at $3L_{\pi}/12$. (b) General self-imaging showing access waveguide modes at z = 0, self-image at $2(3L_{\pi})$, mirrored-image at $3L_{\pi}$, and 2-fold image at $3L_{\pi}/2$ [19].

2.3.4 Other Devices

Additional optical structures of concern in this dissertation are star coupler-based beam splitters and grating couplers. Star coupler-based beam splitters work by diffracting incoming light in a slab waveguide whose width is designed to avoid light from getting to the edges of the waveguide. The light couples into two output waveguides that are specifically placed at an angle according to the wave front, to collect the maximum amount of light [38]. The device is optimized by choosing the appropriate dimensions of the waveguide length *L* and the inclination angle α of the output waveguides. The structure is illustrated in Fig. 2.15(a).

A grating coupler is a waveguide with periodic variation (in the order of a few hundreds of nanometers) of the rib waveguide's height, as in Fig. 2.15(b). It is often used to overcome coupling problems due to mode mismatch between a sub-micrometer waveguide at the chip edge and an external single-mode fiber. An incident plane wave propagating at an angle θ_i with respect to the *z* axis is has a phase-matched interaction with the *m*th guided wave when there is a grating of period Λ that satisfies the phase matching condition [41]



Figure 2.15: Additional optical devices used to design silicon optical modulators (a) star coupler-based beam splitter, and (b) Grating coupler.

$$\kappa_i \sin(\theta_i) + p \frac{2\pi}{\Lambda} = \beta' \tag{2.27}$$

where $\kappa_i = 2\pi/\lambda$ is the module of the incident wave vector, *p* is the diffraction order, $\beta' = (2\pi/\lambda)n_{\text{eff}}$ is the real part of the propagation constant, and n_{eff} is the effective refractive index of the guided mode. The grating will diffract energy from the incident wave into the guided wave.

2.4 Electrical Structure

One of the main electrical structures used to design silicon optical modulators based on the plasma dispersion effect is the PN junction, as it offers the simplest configuration of all the possible structures used for this purpose. The PN junction is able to provide the largest effective index variation Δn_{eff} ; however, it also delivers the largest losses among all the electrical structures, caused by doping concentration in the modulator's active region. To overcome this problem, other types of configurations such as the PIN and PIPIN diodes are commonly used. The PIN diode offers fewer losses due to the presence of an intrinsic region (less doping concentration in the active region), at the expense of decreasing n_{eff} by a great amount. In contrast, the PIPIN diode consists of a p-doped region embedded in the intrinsic region of a PIN diode with the purpose of increasing n_{eff} , while maintaining the most reduced losses possible. The design of these different structures shows that a compromise between n_{eff} and insertion loss must always be made when choosing the modulator's electrical structure. This section describes the operation principle and configurations inside a waveguide of a PN junction; these apply to both the PIN and PIPIN diodes.

2.4.1 The PN Junction

All the modulators studied in this dissertation rely on carrier depletion in a PN junction—or variants PIN and PIPIN junctions—to change the refractive index of silicon. Silicon can be doped with donor atoms (N_D), usually obtained from phosphorous or arsenic, and acceptor atoms (N_A), usually obtained from boron. If the concentration of N_D in silicon is larger than the concentration of N_A then the silicon material is *n*-type (larger concentration of electrons *n*). Silicon is *p*-type when the concentration of N_A exceeds the concentration of N_D (larger concentration of holes *p*), respectively. The theoretical electron-hole concentration in an undoped semiconductor material (intrinsic doping concentration) is $n_i = 10 \times {}^{10} \text{ cm}^{-3}$ at room temperature, though in our devices the residual electron-hole concentration was experimentally calculated to be ~ $n_i = 10 \times {}^{14} \text{ cm}^{-3}$. A PN junction is created when a *p*-type and an *n*-type regions are in close contact creating a potential difference at their boundary. A bilateral carrier diffusion leaves an imbalance of charges that generates a space-charged region on each side of the junction. This region is called the depletion region, as illustrated in Fig. 2.16.

The built-in potential V_{bi} across the depletion region prevents the penetration of electrons into the *p*-side and holes into the *n*-side. Under zero bias V_{bi} is determined by



Figure 2.16: PN junction showing (a) the space charge zone and (b) the energy band diagram at thermal equilibrium.

$$V_{bi} = \frac{kT}{q} \ln\left(\frac{N_D N_A}{n_i^2}\right) \tag{2.28}$$

where kT/q is the thermal voltage $V_T \approx 26 \ mV$ at room temperature. When the PN diode is reverse biased, the electric field at the junction increases where the width of the depletion region is given by

$$x_d = \sqrt{\frac{2\varepsilon}{q} \left(\frac{1}{N_A} + \frac{1}{N_D}\right) + (V_{bi} + |V_a|)} \tag{2.29}$$

where *q* is the elementary electric charge, and *V*_a is the applied reverse voltage, ε is the electrical permittivity, which is the product between silicon relative permittivity $\varepsilon_r = 11.68$, and vacuum permittivity $\varepsilon_0 \approx 8.854 \times 10^{-12}$ F/m. When *x*_d increases, holes and electrons are pulled further away form the junction creating a depletion region capacitance *C*_D. The small signal *C*_D is given by

$$C_D = \frac{\varepsilon A}{x_d} \tag{2.30}$$

where *A* is the surface area of the parallel plate capacitor. There are three main mechanisms that govern the speed of optical modulators, the time response of the PN junction, the carrier drift velocity in the depletion region (drift time response), and the propagation of the RF signal on the device.

• The time response τ of the PN junction is determined from the equivalent circuit of the diode, which includes the depletion region capacitance C_D and the diode's series resistance R_D . The RC response of the PN junction usually dominates the speed of optical modulators based on the carrier dispersion effect. The time response of the PN diode is given by $\tau = R_D C_D$, ergo, $f_{RC3dB} = 1/(2\pi\tau)$. For high-speed operation τ

needs to be as small as possible to allow for a large bandwidth, which means R_D and C_D must be small. The conditions to reduce R_D will be explained in the next chapter. However, according to Eq. 2.30, C_D can be reduced by decreasing the junction's depletion region, but at the same time, the depletion region must be sufficiently large to obtain a substantial refractive index variation of silicon. Consequently, a trade-off between speed response and modulation efficiency is necessary.

• The drift time t_r response is the transit time of carriers through the depletion region when a bias is applied to the diode, and is calculated from the distance carriers travel to cross the depletion region and the drift velocity v_d . The carrier-drift bandwidth is determined by [42]

$$f_{3dB} = \frac{2.4}{2\pi t_r}$$
(2.31)

where $t_r \equiv x_d / v_d$, and x_d is the depletion region calculated in Eq. 2.29. It can be seen from Eq. 2.31, that f_{3dB} increases with a narrower depletion region, or a higher carrier drift velocity.

• The propagation of the electrical RF wave on the phase shifter gives rise to signal attenuation in the traveling-wave electrodes, impedance mismatch, and different propagation speeds of the optical and electrical signals. Due to its complexity, this topic is studied in detail in Chapter 5.

2.4.2 Diode Configurations Inside the Waveguide

Lately, the fastest operation of silicon modulators has been obtained with carrier depletion of PN, PIN and PIPIN junctions. The search for a structure that provides the best modulation efficiency has led researchers to design junctions in different and creative configurations inside the waveguide. The most common configurations are lateral, vertical, and interleaved, as depicted in Fig. 2.17. If the PN junction is in the lateral (Fig. 2.17[a]) or vertical (Fig. 2.17[b]) configurations, then junction is parallel to the direction of the light propagation, whereas in the interleaved configuration (Fig. 2.17[c]) periodic consecutive depletion regions lie perpen-

Chapter 2. Silicon Optical Modulation

dicular to the direction of light propagation. PN junctions designed parallel to the direction of light propagation must be optimized so as to provide the maximum overlap between the depletion region and the optical field. The perpendicular configuration of interleaved junctions makes the light beam overlap the maximum depletion regions across the waveguide without the need for accurate alignment.



Figure 2.17: PN junction configuration in the waveguide (a) lateral, (b) vertical, and (c) interleaved.

High-speed operation of up to 10 Gbit/s has been demonstrated on devices based on vertical [25], [43], [26], [44], lateral [4], [5], [45], [46], [47], [48], and interleaved [49], [50], [51], [52], [53], [54] configurations. However, comparing the vertical and horizontal PN diodes, modulators designed with a lateral junction offer fewer losses than modulators designed with vertical junctions. This is because one of the doped regions covers the entire area of the

waveguide, as seen in Fig. 2.17(b), which means that one electrode must be in contact with the rib. Usually, the metal is placed far away from the waveguide, but this requires a highly doped region to reduce the access resistance, which also adds losses to the active region.

The link between dopant concentration, resistance, and losses will be explained in depth in Chapter 3. The lateral PN junction is thus the preferred configuration for achieving high-speed modulation in silicon, although the interleaved junction is becoming more popular. The optical modulators analyzed in this dissertation are based on lateral PN and PIPIN diodes, as well as interleaved PN junction configurations.

2.4.3 The PIPIN Diode

A PIPIN diode is a variation of a PN diode. It consists of a *p*-doped slit embedded in the intrinsic region of a PIN diode as depicted in Fig. 2.18. The purpose of this design is to create an additional intrinsic region to reduce the optical loss in the rib waveguide. Furthermore, by separating the p-doped region (p+ and p regions) a central p-doped slit is created between both intrinsic regions. Having two different p-doped regions is advantageous, as the doping concentration in the center slit can be reduced in order to have only the doping concentration region that can be effectively depleted when under bias. This design decreases the losses of the active region even further, while the p+ region can be maintained at a higher doping level to decrease the access resistance [1]. The center slit was chosen to be p-doped because holes are more efficient in changing the refractive index of silicon for doping concentrations below 10^{19} cm⁻³, as was seen in Section 2.1.4. The depletion region of the PIPIN diode occurs in the PIN region, as illustrated in Fig. 2.18.

Both intrinsic regions have an influence on the -3 dB cutoff frequency of the device; x_{pin} determines the capacitance of the junction, while x_{pip} increases the access resistance of the diode. Nevertheless, it has been determined that the PIPIN configuration can reach the same frequency values as a PN diode, as the capacitance is reduced [1]. The PIPIN diode is therefore an innovative design that is able to achieve a high cutoff frequency while decreasing optical losses in the active region. This diode was used as the active region of most of the optical modulators characterized and presented in Chapter 6.



Figure 2.18: Schematic diagram of the cross-sectional view of a PIPIN diode. The diode consists of a *p*-doped region embedded in the intrinsic region of a PIN diode.

2.5 State of the Art

The development of silicon optical modulators has consistently increased in the last decade since the appearance of the first high-speed working device at 1 Gbit/s developed by *Intel* in 2004 [24], which was based on carrier accumulation of a MOS capacitor embedded in an asymmetric MZI. Since then, ongoing research has continued with successful results. All the silicon modulators developed until now have exhibited a wide range of designs in terms of materials, dimensions, junction configurations, optical structures (e.g., interferometers, resonators, and couplers), traveling-wave RF electrodes, driving devices, etc. Tables 2.1 and 2.2 summarize the latest and most relevant results on silicon optical modulators to date. The tables include diode type, optical structure, and performance (FoM) comparison. Only all-silicon modulators based on carrier depletion were considered. The highlighted columns are the silicon optical modulators designed and characterized by the silicon photonics group *Institut d'Électronique Fondamentale (IEF)* (one of the research units of *Université Paris-Sud* –UPS), and fabricated at *CEA-LETI*.

Looking at the FoM comparison shown in both tables, we can see that each modulator has strong and weak parameters. The tables show strong parameters in *blue*, and weak parameters in *red*. The first silicon optical modulator working at 40 Gbit/s (column 1) was designed in 2007 using a vertical PN junction embedded in a MZI. This modulator was the fastest device at that time, although the ER was only 1 dB and the $V_{\pi}L_{\pi}$ was 4 V·cm. Such FoMs are not efficient for current modulators. However, this modulator helped to set the bar for future generations to come.

Modulators operating between 10 - 20 Gbit/s include RRs- and MZIs-based devices (columns 6, 7, 10, 16, and 17). In this group of modulators, driving voltages in the range of 0.5 - 1 V_{pp} are already used on *IMEC*'s (columns 16-17) and *MIT/Sandia*'s (column 10) modulators, although they also delivered largest losses or lowest ERs at high-speed operation. In the 20 - 30 Gbit/s range (columns 11, 13, and 14), *U.Washington*'s devices (column 13) delivered a large ER of 10 dB, in comparison with 8.5 dB for the *Bell-Labs* modulator (column 11). However, *U.Washington*'s modulator delivered an IL of 15 dB at the operating point. *Bell-Labs* did not report an IL at the operating point, but the reported on-chip loss was 9 dB, which is also large. While the *Bell-Labs* modulator had a smaller V_πL_π than *U.Washington*'s, the latter used a driving voltage of only 1 V_{pp} (vs. 3.5 V_{pp} for the *Bell-Labs* modulator). For speeds over 40 Gbit/s, *Surrey*'s (columns 2-4) and *IMEC*'s (column 15) modulators delivered the largest ERs (10 dB for *Surrey*'s modulator and 11.3 dB for *IMEC*'s), at the expense of producing the highest ILs, which was between 8 - 25 dB at the operating point for *Surrey*'s modulator (the IL at the operating point of *IMEC*'s modulator was not reported).

IEF's modulators (columns 6-9) did not exceed all other modulators in terms of ER or driving voltage, but they delivered comparative FoMs. These modulators succeeded in delivering the lowest IL of all the modulators at the operating point, and among the least on-chip losses. *IEF*'s active region design based on a PIPIN diode can be attributed the low-loss response of the modulator, as it will be explained and shown throughout this document.

In general, the tables show that the overall performance of silicon optical modulators based on the free-carrier dispersion effect are mostly limited by the losses, and that a compromise must always be made to obtain a reasonable equilibrium between modulation efficiency and losses. For that reason, creative PN junction types, orientations, shapes, and dimensions are constantly being designed so that the challenges and limitations of this type of modulator can be reduced with every new modulator design.

	[1]	[2]	[3]	[4]	[5]	[6]	[7]	[8]	[9]	[10]
Year	2007[43]	2011[55]	2011[55]	2011[45]	2012[27]	2011 [56]	2011 [49]	2012 [5]	2012 [5]	2011 [44]
Institution	Intel	Surrey	Surrey	Surrey	Surrey	UPS	UPS	UPS	UPS	MIT/Sandia
Speed	40 Gbit/s	40 Gbit/s	40 Gbit/s	40 Gbit/s	50 Gbit/s	10 Gbit/s	10 Gbit/s	40 Gbit/s	40 Gbit/s	12.5 Gbit/s
Wavelength	1.55 <i>µ</i> m	$1.55\mu{ m m}$	$1.55\mu{ m m}$	1.55 μm	1.55 <i>µ</i> m	$1.55\mu{ m m}$	$1.55\mu{ m m}$	$1.55\mu{ m m}$	$1.55\mu{ m m}$	1.55 <i>µ</i> m
Mechanism	Depletion	Depletion	Depletion	Depletion	Depletion	Depletion	Depletion	Depletion	Depletion	Depletion
Electrical	Vertical	Lateral	Lateral	Wrapped	Lateral	Lateral	Interleaved	Lateral	Lateral	Vertical
Structure	PN	PN	PN	PN	PN	PIPIN	PN	PIPIN	PIPIN	PN
Optical	Asymmetric	Asymmetric	Asymmetric	Asymmetric	Asymmetric	$r = 50 \ \mu \mathrm{m}$	$r = 50 \ \mu m$	Asymmetric	Asymmetric	$2r = 3.5 \mu\mathrm{m}$
Structure	MZI	MZI	MZI	MZI	MZI	RR	RR	MZI	MZI	Microdisk
Length	-	1 mm	3.5 mm	1.35 mm	2 mm	252 <i>µ</i> m	252 µm	0.95 mm	4.7 mm	0.25 mm
$V_{\pi}L$	<4 V·cm	2.7 V·cm	2.7 V·cm	14 V·cm	2.8 V⋅cm	3 V∙cm	2.5 V⋅cm	3.5 V⋅cm	3.5 V⋅cm	-
Capacitance	_	200 pF/m [*]	_	410 fF/m	_	-	54 fF	0.22 fF/µm	0.21 fF/μm	_
f _{3dB}	~ 30 GHz	$47\mathrm{GHz}^*$	-	> 40 GHz	> 40 GHz	-	-	> 40 Ghz	20 Ghz	$20\mathrm{GHz}^*$
ER	1 dB	7.5 dB	10 dB	6.5 dB	2.2 dB	5.9 dB	4 dB	3.2 dB	6.6 dB	3.2 dB
IL	< 4 dB	8 dB	15 dB	25 dB	~ 5.6 dB	-	-	4.5 dB	6 dB	-
On-Chip IL	-	5 dB	15 dB	15 dB	4.1 dB	~ 6 dB	~ 6 dB	2.5 dB	6 dB	< 1 dB
Power	16pJ/bit	2/5.3 pJ/bit	2/5.3 pJ/bit	4.5 pJ/bit	4.4 pJ/bit	_	287 fJ/bit	2 pJ/bit	10 pJ/bit	3 fJ/bit
RF Voltage	6 V _{pp}	$4/6.5 \mathrm{V}_{pp}$	$4/6.5 \mathrm{V}_{pp}$	6 V <i>pp</i>	$6.5 \mathrm{V}_{pp}$	$\sim 5 V_{pp}$	$\sim 5 V_{pp}$	$\sim 6.5 \mathrm{V}_{pp}$	$\sim 6.5 \mathrm{V}_{pp}$	$\sim 1 \mathrm{V}_{pp}$

* Theoretical/predicted value

Table 2.1: State of the art of silicon optical modulators (1/2).

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	[11]	[12]	[13]	[14]	[15]	[16]	[17]	[18]	[19]	[20]
Year	2012[48]	2012[48]	2011[57]	2012 [58]	2012[58]	2012[58]	2012[58]	2012 [59]	2013 [47]	2013 [47]
Institution	Bell Labs	Bell Labs	U. Washington	IMEC	IMEC	IMEC	IMEC	PETRA ¹	IS ²	IS
Speed	30 Gbit/s	50 Gbit/s	25 Gbit/s	< 25 Gbit/s	40 Gbit/s	10 Gbit/s	10 Gbit/s	50 Gbit/s	50 Gbit/s	60 Gbit/s
Wavelength	$1.55\mu{ m m}$	$1.55\mu{ m m}$	$1.55\mu{ m m}$	1.55 μm	1.55 <i>µ</i> m	$1.55\mu{ m m}$	$1.55\mu{ m m}$	1.55 <i>µ</i> m	1.55 <i>µ</i> m	1.55 µm
Mechanism	Depletion	Depletion	Depletion	Depletion	Depletion	Depletion	Depletion	Depletion	Depletion	Depletion
Electrical	Lateral	Lateral	Lateral	Interleaved	Lateral	Interleaved	Lateral	Depletion	Lateral	Lateral
Structure	PN	PN	PN	PN	PN	PN	PN	PN	PN	PN
Optical	Asymmetric	Asymmetric	Traveling-Wave	Asymmetric	Asymmetric	$r = 40 \ \mu m$	$r = 40 \ \mu m$	Asymmetric	Asymmetric	Asymmetric
Structure	Push-pull MZI	Push-pull MZI	MZI	MZI	MZI	RR	RR	MZI	MZI	MZI
Length	6 mm	2 mm	3 mm	3 mm	3 mm	4.7 mm	2 mm	0.25 mm	0.75 mm	0.75 mm
$V_{\pi}L$	1.86 V·cm	2.4 V⋅cm	2.7 V·cm	1.12 V·cm	1.55 V⋅cm	3.5 V⋅cm	2.4 V⋅cm	1.3 V⋅cm	2.05 V·cm	2.05 V⋅cm
Capacitance	$0.3 - 0.4 {\rm fF}/\mu{ m m}$	$0.3 - 0.4 {\rm fF}/\mu{ m m}$	0.12 nF/mm	1.59 nF/mm	0.44 nF/mm	_	-	_	330 fJ/mm	330 fJ/mm
f _{3dB}	20 GHz	_	15 GHz	-	15 Ghz	3.8 Ghz	11 Ghz	_	27.7 GHz	27.7 GHz
ER	8.5 dB	5.2 dB	10 dB	-	11.3 dB	9.7 dB (DC)	9.1 dB (DC)	4.3 dB	7.5 dB	3.6 dB
IL	_	_	15 dB	-	_	4.8 dB	7.9 dB	_	6.5 dB	3.5 dB
On-Chip IL	9 dB	4.1 dB	6.2 ±0.5 dB	5.4 dB	2.6 dB	_	_	4 dB	1.9 dB	1.9 dB
Power	_	_	400 fJ/bit	_	_	_	_	_	~ 4.5 mW dB	~ 9.5 mW dB
RF Voltage	$3.5 \mathrm{V}_{pp}$	$3.5 \mathrm{V}_{pp}$	$1 \mathrm{V}_{pp}$	$5 V_{pp}$	5 V _{pp}	0.5 V _{pp}	0.5 V _{pp}	4.3 V _{pp}	$\sim 6.5 \mathrm{V}_{pp}$	$\sim 6.5 \mathrm{V}_{pp}$

¹ Photonics Electronics Technology Research Association, Japan

² Institute of Semiconductors, China

Table 2.2: State of the art of silicon optical modulators (2/2).

2.6 Projects

This dissertation focuses on two silicon optical modulators, each specifically designed to fulfill the requirements stipulated by the pHotonics ELectronics functional Integration on CMOS (HELIOS) and SILicon based integrated optical transceiVER for local access networks (SILVER) (SILVER) projects. Each project had a different set of specifications according to the application. The first silicon optical modulators had been designed for HELIOS before my arrival at *IEF*. My involvement in this project consisted in the characterization of the devices that were fabricated at *CEA-LETI*: namely, batches T852B and U532P, for which results are presented in Chapter 6. At the same time, I was fully involved in the conception and design of the silicon optical modulators required for SILVER, whose complete design process is presented in Chapters 3-5. Further details of each project are presented next.

2.6.1 HELIOS

HELIOS was a 4-year European project (2008 to 2012) that reunited nineteen partners among major European CMOS photonics and electronics institutions with the aim of developing the building blocks for the demonstration of complex silicon photonics functions addressed to industrial needs. The *Institut d'Électronique Fondamentale* made part of a collaboration team whose task was the design of thick (slab height over 220 nm), thin (slab height below 220 nm), and slow-wave modulators to reach the performance targets up to 40 Gbit/s [60]. In particular, *IEF*'s task in the project was to design thick silicon optical modulators for highspeeds of 10 Gbit/s and 40 Gbit/s at a wavelength of 1.55 μ m. Table 2.3 summarizes the required specifications of the HELIOS project.

Specification	Measurement				
Wavelength (λ_0)	1.55 μm				
Speed	10 Gbit/s and 40 Gbit/s				
Optical	Symmetric / asymmetric				
Structure	MZI				
Voltage	< 5 V				
$V_{\pi}L_{\pi}$ (@ 5V)	2.5 V⋅cm				
f_{3dB}	10 GHz (for 10 Gbit/s)				
	35 GHz (for 40 Gbit/s)				
ER	> 15 dB (DC)				
	6 – 8 dB (for 10 Gbit/s)				
	2 dB (for 40 Gbit/s)				
IL	< 5 dB				
WG Height (H)	400 nm				

Table 2.3: Specifications of an optical modulator for applications at 1.55 μ m.

2.6.2 ANR SILVER

FTTH is fast-growing method of providing high-speed access to consumers and businesses. Since its deployment a few years back, FTTH has become an efficient solution for the continuous demand for high-speed access. The upgrade to FTTH could be completed more quickly if inexpensive, and high-speed performant Optical network unit (ONU) transceivers could be integrated in end-to-end high-bit-rate networks, which could lead to a network upgrade [61]. The SILVER project was conceived with the aim of demonstrating a low-cost photonic integration approach on silicon for 10 Gbit/s ONU for passive optical networks Passive Optical Network (PON). The project is a collaboration among four industrial and academic institutions, each with a different task, but equally involved in the common final task of integrating the silicon transceiver. SILVER is a three-year project that kicked off in 2010. The partners include coordinator *Alcatel Thales III-V Lab*, in charge of the design of a III-V/Si laser, *CEA-LETI* in charge of designing the wavelength multiplexers and tapers, as well
as the complete integration of the the ONU transceivers, *France Telecom R&D*, in charge of specifications and and system validations, and *L'institut d'Électronique Fondamentale* who is in charge of the Si/Ge Avalanche Photodiode (APD) and the silicon modulator. Table 2.4 shows the specifications of an optical modulator for FTTH applications at $\lambda = 1.27 \ \mu$ m.

Specification	Measurement
Wavelength (λ_0)	1.27 μm
Operation Speed	10 Gbit/s
Optical	Symmetric / asymmetric
Structure	MZI
Voltage	< 5 V
$V_{\pi}L_{\pi}$ (@ 5V)	2-3 V·cm
f_{3dB}	10 GHz
ER	> 10 dB (for 10 Gbit/s)
IL	< 5 dB
WG Height (H)	440 nm

Table 2.4: Specifications of an optical modulator for FTTH applications at 1.27 μ m.

Conclusion

This chapter has presented a literature review of all the theoretical concepts relevant for this dissertation. The first part of the chapter was dedicated to the analysis of the main physical effects that make modulation in silicon possible. These effects show the potential of fabricating opto-electronics devices with current CMOS processes—most importantly, those that can be integrated with microelectronic chips. The plasma dispersion effect is of particular interest, as it was the effect used for the development of the silicon optical modulators presented in the next chapters.

General waveguide aspects were also presented, followed by the operation principle of the main optical and electrical structures, and their respective FoM needed to design performant silicon optical modulators. Furthermore, two thorough state of the art tables showed—with the purpose of visualizing in a parallel fashion— the latest results obtained on devices designed with carrier depletion as the mechanism of choice to obtain modulation in silicon. Finally, the two projects for which this work was possible were presented, including envisioned applications for the silicon optical modulator design specifications. The following chapter will explain the design process of these devices.

3 Modulator Design

The silicon optical modulators developed at the *Institut d'Électronique Fondamentale (IEF)* had always been conceived for applications using a wavelength of 1.55 μ m. Three main electrical structures (PN and PIPIN diodes, and interleaved PN junctions) were usually optimized and integrated on Mach-Zehnder interferometers (MZIs) and ring resonators (RRs). Due to the performance requirements set by the HELIOS project such as a high-speed operation of 40 Gbit/s, and losses below 5 dB, the modulator selected was based on carrier depletion of a PIPIN diode embedded in the active region of a MZI. All the performance requirements were reached with this structure as a result of its strategic design, in which the maximum refractive index was obtained while limiting the losses. My Ph.D. work required development of a high-speed optical modulator to be integrated in a silicon transceiver for local access networks (at a wavelength of 1.27 μ m) according to the requirements of the SILVER project. Table 3.1 lists the modulator specifications for both applications.

Because the PIPIN worked effectively for the HELIOS project, this device was selected as a starting point for the design of the SILVER modulator. However, the entire design of an optical modulator for a silicon transceiver working at 1.27 μ m required the simulation of the various parts of a modulator, including passive elements, active region, interferometric structure as well as the designing of the IC layout—Graphic Data System File (GDS) file—to create masks for the different fabrication levels. This chapter presents a complete explanation of the design and integration techniques that were developed during this work, except for design of the RF electrodes whose theory and design will be explained in Chapter 5

Specifications	HELIOS	SILVER
Wavelength (λ_0)	$1.55\mu\mathrm{m}$	1.27 µm
Operation Speed	40 Gbit/s	10 Gbit/s
Voltage Range	$0-5 \mathrm{V}$	$0-5 \mathrm{V}$
Length	< 5 mm	_
ER	6–8 dB (at 10 Gbit/s)	> 10 dB
IL	< 5 dB	< 5 dB
WG Height (H)	400 nm	440 nm

Table 3.1: Comparison of requirements between an optical modulator for applications at 1.55 μ m and an optical modulator for applications at 1.27 μ m.

The first section of the chapter explains the process of the complete design of the modulator's active region, followed by section describing the design of the interferometric structures: namely, the MZI using MMI couplers. The last section is focused on the design of the IC layout for fabrication.

3.1 Phase Shifter Optimization

The different operation wavelengths between the already-optimized modulators (1.55 μ m) and the silicon transceiver (1.27 μ m), made the main structure of the existing modulators incompatible with the new application. This incompatibility is caused by the effect of the waveguide on the mode confinement and the modulation efficiency. Tehrefore, a new active region had to be fully optimized. The first task in the optimization process was to simulate the active region of the previously designed PIPIN diode for a wavelength 1.27 μ m. This was accomplished using the in-house optical simulation software *SolverFDM* (developed by Gilles Rasigade during his Ph.D. work [3], [62]). The *SolverFDM* simulation software is based on a finite-difference method created to simulate electro-optical devices as it has a direct interface with ISE DESSIS TCAD 9.0 Integrated Systems Engineering (ISE), a software used to perform the electrical simulations.

SolverFDM allows the designer to draw the cross-section of the electro-optical device by choosing its dimensions and materials as well as doping types and concentrations. The con-

ceptualization of the main structure is followed by the generation of multiple files that define the designed geometry, including its dimensions, materials, doping type and concentrations, and electrical contact type and placement. These are the files are used for the ISE electrical simulation. The electro-optical simulations are conducted by applying Soref's equations (Eq. 2.5 - Eq. 2.8) to the electrically simulated files once they are imported back to *SolverFDM*.

The next simulation step consists in solving Maxwell's equations in *SolverFDM* using a finite difference frequency domain (FDFD) method. This step provides paired calculated values of the effective refractive index and effective absorption coefficient of each optical mode, as a function of the applied voltage. The results can be used in analytical models to obtain the optical transmission of MZIs and RRs as a function of the voltage, and to calculate the main parameters that determine the efficiency of an optical modulator, such as the insertion loss (IL), and L_{π} required for a V_{π} = 5 V (maximum allowed DC voltage range) [3].

The rest of this chapter will describe the optimization process used to design the active region of the modulator needed for a silicon transceiver working at 1.27 μ m. The main parameters considered in the design of the active region included the waveguide dimensions, doping concentrations, and electrical structure type and configuration, all of which will be discussed individually in the next sections.

3.1.1 Waveguide Dimensions

The schematic diagram of a cross-sectional view of the PIPIN diode that meets the requirements of applications at 1.55 μ m is shown in Fig. 3.1 [3]. The figure shows all the dimensions and the doping concentrations of the active region. This structure had been designed before I began my work. This PIPIN diode was mainly developed with the aim of reaching a tradeoff between device speed, efficiency, and loss, as the silicon modulator was required to reach operation speeds of up to 40 Gbit/s. The design of an unintentionally doped section in the active region limited the amount of doping concentration in the waveguide, as it is known that dopants are a direct source of optical losses [1], [4], [49] and [5]. However, the PIPIN diode is a sensitive device to fabricate because it requires extremely precise alignment to define the *p*-doped region in the center of the waveguide.

The optical modulator required for integration with a silicon transceiver for applications at a 1.27 μ m wavelength had to reach a maximum operation speed of 10 Gbit/s, as listed in

Chapter 3. Modulator Design

Table 3.1. The operation speed of this modulator was well below that of the previous one, and because no stronger requirement in optical loss was needed it was possible to explore the fabrication of a simpler device. The PN diode potentially met these requirements, as it can be created with fewer fabrication levels than the PIPIN diode.



Figure 3.1: Dimensions and doping concentrations of a PIPIN diode optimized for applications at 1.55 μ m [3].

A change in the application wavelength induces a variation of the optical mode profile inside the rib waveguide, but it was important to analyze the change using an equivalent structure. The design of the PN diode began by choosing a rib waveguide with the same dimensions, doping concentrations, and locations as the previously optimized PIPIN diode. A schematic diagram comparing the PIPIN diode and the first design of the PN diode is depicted in Fig. 3.2. The PN junction was designed 50 nm away from the center of the waveguide towards the *n*-doped region so that there was a maximum overlap between the depletion region and the optical mode. The *p*-doped region was designed to be wider than the *n*-doped region, as it is known that the *p* dopants are more efficient in obtaining refractive index variations in silicon [1]. It can be seen in Fig. 3.2(a) (represented by a green line) that the PIN junction location of the PIPIN diode corresponds to the same junction location as the PN junction seen in Fig. 3.2(b). The figure also shows that the doping concentrations of the PN junction were chosen to match the *p* + and *n* + doping concentrations of the PIPIN diode, which were $p_{+} = 8 \times 10^{17}$ cm⁻³, and $n_{+} = 1 \times 10^{18}$ cm⁻³. A weaker concentration of the *p*-doped region guarantees that the PN junction space charge zone expands mainly over that region.



Figure 3.2: PIPIN and PN diodes used for simulation for h = 220 nm.

To reduce fabrication complexity, compatible waveguide and slab heights were established among the necessary photonic structures to be integrated in the silicon transceiver proposed by the SILVER project. The dimensions were selected by the partners involved in the SILVER project. The waveguide (*H*) and slab (*h*) heights of the PIPIN and PN diodes were designed to be 440 nm and 220 nm, respectively, these values were imposed by the laser dimensions of the silicon transceiver.

In the first trial, an electro-optical simulation of both diodes was performed for different waveguide widths from 320 nm to 440 nm. The narrowest waveguide dimension of the range was chosen because the losses inside the rib waveguide increase due to sidewall roughness when the waveguide width is reduced. The widest waveguide value was chosen, as it was the maximum dimension expected for the PIPIN diode designed at 1.55 μ m. The results of the effective refractive index Δn_{eff} change as a function of the waveguide width of PIPIN and PN diodes designed with a waveguide height of 440 nm, and a slab height of 220 nm (as determined by the SILVER project) at a wavelength $\lambda = 1.27 \,\mu$ m are shown in Fig. 3.3. The variation of Δn_{eff} is shown at 5 V, the maximum allowed DC voltage. The change of Δn_{eff} with increasing waveguide width w, is the result of changes in the overlap between the optical mode and the depleted region inside the waveguide. The results show that the PN junction offers larger effective refractive index variation than the PIPIN diode. The simulated change in refractive index of both structures was used to calculate the modulation efficiency $VL_{\pi} = V \cdot \lambda/(2\Delta n_{\text{eff}})$, at V = 5 V, and the losses due to the active region 10log₁₀($e^{-\alpha L}$), at $\alpha = 0$ V.



Figure 3.3: Change of the effective refractive index (Δn_{eff}) vs. waveguide width (w) of a PN diode and a PIPIN diode of slab height 220 nm at 5 V.

The results are illustrated in Fig. 3.4. When analyzing the VL_{π} plotted in Fig. 3.4(a), it can be seen that the modulation efficiency is superior in the PN junction, as the VL_{π} value is lower than the one obtained with the PIPIN diode. A small value of VL_{π} also leads to the conclusion that shorter phase shifters will be able to achieve high-speed modulation, because all the VL_{π} values were always calculated at V = 5 V. However, when looking at the results of the active region losses plotted in Fig. 3.4(b), we soon realized that the PN diode supplies higher losses. The smallest loss is 5.42 dB, which is seen at 320 nm waveguide width. This loss value corresponds to the active region alone, as it does include the additional loss contribution from the waveguide roughness and the beam splitter/combiner. This value exceeds the maximum loss requirement of the silicon transceiver and is therefore not suitable for such applications. The losses obtained with the PIPIN diode fell within the maximum loss requirements; however, it was important to look for a simpler device to fabricate. With the chosen waveguide dimensions, only the PIPIN diode fulfilled the project requirements, but all the other collaborators preferred a simpler structure.



Figure 3.4: (a) Modulation efficiency (VL_{π}) and (b) (IL) vs. waveguide width (w) of a PN diode and a PIPIN diode of slab height 220 nm at 5 V

Chapter 3. Modulator Design

Looking at the optical confinement inside the waveguide (shown in Fig. 3.5), it is clearly noticeable that a considerable amount of light is spread out in the slab. To see the exact amount of light confined in the depletion region, the FF was simulated choosing a rectangular region centered in the middle of the waveguide (green rectangle in Fig. 3.5). This region was chosen to have the same height as the rib waveguide and a width of 100 nm, and was selected because the width corresponds to roughly the maximum width of the depleted region regardless of the electrical structure. The same rectangular region was used for all the FF simulations presented in this dissertation. The simulation produced a FF as low as 27%.





The FF increases proportionally with frequency, which means that from the relationship $v = c/\lambda$, where v is the normalized frequency, the FF is inversely proportional to the wavelength, demonstrating that for a given waveguide, we can expect a larger FF for a 1.27 μ m wavelength than for a 1.55 μ m wavelength. Additionally, a study of the FF described in [1] showed that the change in the effective refractive index increases when the ratio (r) between the slab and the waveguide heights (r = h/H) decreases. A decrease of r will produce an increase of the mode confinement, which is obtained from the computation of the FF. It was also shown that the effective refractive index changes proportional to the FF. Consequently, it can be said that waveguide design required the careful optimization of the FF to ensure maximum light confinement in the desired region. This optimization was carried out on a waveguide of height 440 nm and width 420 nm, where the slab height (h) was modified to increase light confinement in the waveguide. The simulation was carried out using the same rectangular area—centered in the middle of the waveguide—as described above. The results

are shown in Fig. 3.6.



Figure 3.6: FF of a 420 nm wide waveguide of height H = 440 and slab height h = 220 nm

The graph shows that the light confinement started at 27 % for the maximum slab height of h = 220 nm. The FF increased as the slab height was reduced, and at h = 100 nm, the FF reached at value of 33 %. The simulation process of the FF was repeated for waveguide widths from 300 nm to 500 nm. The results are shown in Fig. 3.7. The plot also shows that by decreasing the slab width (w), the FF increased by a significant amount. For h = 100 nm, the increase of the FF seen as the rib waveguide becomes narrower starts to level off for waveguide widths below 340 nm, which was an important observation to consider in the optimization process of the waveguide width, as it shows that narrower waveguides these dimensions did not necessarily deliver better optical confinement.

The figure also shows the region in which the optical mode confinement becomes multimode according to waveguide dimensions w and h. The analysis of the FF revealed that a 220 nm slab height leads to considerably poor light confinement at 1.27 μ m. A picture of the simulation results of light confinement inside four waveguides with slab heights h = 100 nm, h = 140 nm, h = 180 nm, and h = 220 nm, is shown in Fig. 3.8.



Figure 3.7: FF comparison among waveguides of various widths (*w*) and slab heights (*h*), w = 420 nm.



Figure 3.8: Light confinement in waveguides of various slab heights (*h*).

After analyzing the results obtain in Fig. 3.7 and Fig. 3.8, it was determined that the most appropriate slab height was 100 nm, the same as that used on the modulator designed for applications at 1.55 μ m, which delivered reasonable light confinement. Changing the slab height of the PN diode meant additional design and fabrication steps for the integration of

the transceiver. However, the change represented a real opportunity to reach the modulation performance needed for high-speed applications at a 1.27 μ m wavelength.

Modulation performance of the PN and PIPIN diodes was calculated for the new rib waveguide at 1.27 μ m. Except for the slab height, all the other dimensions, as well as the doping concentrations of the PN and PIPIN diodes, remained the same, as illustrated in Fig. 3.9. The simulations of the updated structures were resumed at the starting point to find the new Δn_{eff} . The results presented in Fig. 3.10 show a noticeable increase in the Δn_{eff} of both diodes when compared to the results presented in Fig. 3.3. The largest Δn_{eff} occurred at w = 320 nm in both diodes, which is the lowest w used in the simulations to avoid additional losses due to sidewall roughness caused by narrowing the waveguide width. The Δn_{eff} value of the PN diode increased from 2×10^{-4} at h = 220 nm to 2.7×10^{-4} at h = 100 nm, while in the PIPIN diode Δn_{eff} increased from 1.4×10^{-4} at h = 220 nm to 2.4×10^{-4} at h = 220 nm.



Figure 3.9: PIPIN and PN diodes used for simulation for h = 100 nm and w = 420 nm.

The PN diode produced the highest Δn_{eff} . However, it was seen that a decrease of the waveguide slab height caused the emergence of additional modes, which is an important issue that needs consideration in the design process. Multimode propagation of the light beam started to occur when $w \ge 410$ nm, as seen in the shaded region of Fig. 3.10, which meant that the rib waveguide width had to be narrower than 410 nm.

Using the results obtained for the change of the effective refractive index, FoMs VL_{π} and IL were calculated and plotted, as illustrated in Fig. 3.11. The results of the modulation efficiency seen in 3.11(a) improved when compared to the results obtained in 3.4(a) for both diodes. The plot shows that VL_{π} decreases for narrower waveguides, such that the mini-

mum simulated VL_{π} (calculated at V = 5 V) occurred for the narrowest simulated waveguide w = 320 nm, whose values were 1.13 V·cm and 1.31 V·cm for the PN and PIPIN diodes, respectively. However, the figure shows the figure shows that the progression of the VL_{π} value towards narrower waveguides starts to even out, which means that narrower waveguides will not necessarily produce an improved VL_{π} .



Figure 3.10: Change of the effective refractive index ($\Delta n_{\rm eff}$) vs. waveguide width of a PN diode and a PIPIN diode of slab height 100 nm.

It is also notable that the losses at L_{π} and 0 V decreased considerably when the results shown in Fig. 3.11(b) are compared to those obtained in Fig. 3.4(b). This simulation delivered maximum losses at w = 400 nm, of 4.1 dB and 1.6 dB for the PN and PIPIN diodes, respectively. It is abundantly clear from this plot that the PIPIN diode delivers considerably fewer losses than the PN diode. The results of Fig. 3.11(b) also show that the losses increase proportionally with wider waveguides. However, as the results obtained for VL_{π} , it can be seen that the losses start to become slightly constant for waveguide widths below 360 nm.



Figure 3.11: (a) Modulation efficiency (VL_{π}) and (b) IL vs. waveguide width of a PN diode and a PIPIN diode of slab height 100 nm.

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At this point, the simulation results obtained in Fig. 3.11 were sufficient for us to choose the PN diode as the main structure in designing the optical modulator of a silicon transceiver used for local network access applications. The PN diode was chosen over the PIPIN diode, as we were able to obtain satisfactory modulation efficiency as well as insertion loss within the project's requirements. However, what makes the PN diode stand out over the PIPIN diode is the fact that it is easier to fabricate than the PIPIN diode, as will be seen in the fabrication chapter. Moreover, we chose 360 nm as the waveguide width of the active region, as this dimension shows moderate losses and adequate VL_{π} for the optical modulator.

These dimensions show a fair compromise among all parameters, including loss modulation efficiency, sidewall-roughness-induced losses, and ease of fabrication. A schematic diagram showing the chosen dimensions of the active region for the optical modulator is shown in Fig. 3.12. These dimensions were H = 440 nm, h = 100 nm, and w = 360 nm. The green line indicates the placement of the PN junction within the rib waveguide, which up to this point had not been optimized.



Figure 3.12: PN diode dimensions chosen to design the optical modulator of a silicon transceiver at $\lambda_0 = 1.27 \ \mu$ m. The dimensions are $H = 440 \ \text{nm}$, $h = 100 \ \text{nm}$, and $w = 360 \ \text{nm}$. The green line indicated a provisional placement of the diode's PN junction.

In summary, we were able to find waveguide dimensions capable of achieving highspeed modulation, while staying within the specifications of a silicon transceiver working at a 1.27 μ m wavelength. Nevertheless, additional improvements were needed to fully optimize the active region of the complete optical modulator: in particular, doping concentration and junction location within the rib waveguide.

3.1.2 Doping Concentration

Upon selection of diode type and dimensions, the next step was optimization of the dopant concentrations of the p+ and n+ regions, as all the simulations made to this point had used fixed doping concentrations. The PN diode was simulated for different doping concentrations of both dopant types. Plots showing the calculated VL_{π} and IL as a function of N_D (n doping concentration) for various N_A values are shown in Fig. 3.13, and the VL_{π} and IL as a function of N_A (p doping concentration) for various N_D values, are show in Fig. 3.14.

One type of dopant was first selected and kept constant to simulate the change in the effective refractive index, modulation efficiency, and losses while varying the concentration of the other type of dopant. The same simulations were performed by inverting the constant and varying type of dopant.

All doping concentrations were varied from 1×10^{17} cm⁻³ to 1×10^{18} cm⁻³. The results of the modulation efficiency as a function of N_D (*n*-type dopants) from Fig. 3.13(a) show that the lowest N_A dopant concentration, 1×10^{17} cm⁻³, was not large enough to obtain a reasonable VL_{π} . In general, low doping concentrations lead to a small Δn_{eff} , which will increase the VL_{π} . A large VL_{π} represents a long device (because L_{π} increases), which also causes higher propagation losses, as seen in Fig. 3.13(b), where N_D doping concentrations below 3×10^{17} cm⁻³ produced an increase of the VL_{π} , therefore increasing the IL. In the case of the lowest N_A doping concentration, the loss trend is different because the Δn_{eff} extremely weak. This doping concentration provides a Δn_{eff} between 1.11×10^{-4} cm⁻³ and 1.53×10^{-4} cm⁻³, and an absorption coefficient (α) at 0 V, between ~ 0.4 and 2.8 cm⁻¹. These simulated results show that the *p*-doping concentration is so weak as to produce a negligible waveguide absorption which is directly related to the losses. Doping concentrations of above 3×10^{17} cm⁻³, produced a relatively constant VL_{π} between 1 and 1.5 V·cm for N_D concentrations above 3×10^{17} cm⁻³. In this range, the losses increased due to the direct proportionality between dopant concentration and absorption.

The analysis made for N_D can be also applied for N_A . Figure 3.14(a) shows similar curve trends than Fig. 3.13(a); that is, N_A doping concentrations below 3×10^{17} cm⁻³ supplied a weak modulation efficiency, leading to the increase of losses in the same doping concentration range as seen in Fig. 3.13(b). However, a careful comparison of the results seen in Fig. 3.13(b) and Fig. 3.14(b) show a small difference between the loss curves of both doping concentrations.



Figure 3.13: (a) Modulation efficiency (VL_{π}) and (b)IL vs. N_D of a PN diode for N_A doping concentration variations from 1×10^{17} cm⁻³ to 1×10^{18} cm⁻³.



Figure 3.14: (a) Modulation efficiency (VL_{π}) and (b)IL vs. N_A of a PN diode for N_D doping concentration variations from 1×10^{17} cm⁻³ to 1×10^{18} cm⁻³.

The results show that *p*-type dopants (N_A) for doping concentrations above 3×10^{17} cm⁻³, have an influence on the losses (Fig. 3.13[b]), while the *n*-type dopants (N_D) do not (Fig. 3.14[b]). The loss curves seen in Fig. 3.13(b), show that higher N_A dopant concentrations provide fewer losses, which contradict the fact that the increase of doping concentrations is responsible for the losses' increase. However, in this case, a higher N_A concentration leads to less losses because the influence of the significant Δn_{eff} rise—and thus a decrease of L_{π} , with increasing N_A —prevails over the absorption influence.

After analyzing the doping concentration simulation results, we decided to choose $N_A = 5 \times 10^{17} \text{ cm}^{-3}$ and $N_D = 7 \times 10^{17} \text{ cm}^{-3}$, whose values can be seen in the shaded regions of Fig.3.13 and Fig. 3.14, respectively. These values were chosen because VL_{π} and IL provided a fairly constant and superior values in these concentration ranges, while allowing for a margin of error during fabrication. A schematic view of the PN diode designed until this point is shown in Fig. 3.15.



Figure 3.15: PN diode design the optical modulator of a silicon transceiver at 1.27 μ m, including the chosen doping concentrations $N_A = 5 \times 10^{17}$ cm⁻³ and $N_D = 7 \times 10^{17}$ cm⁻³.

3.1.3 Junction Location

The influence of the different doping concentrations in the PN diode has by now been repeatedly explained and illustrated. That is, the *p*-dopants have a greater influence on the effective refractive index variation than the *n*-dopants' counterpart. For that reason, the depletion region was not expected to be located in the center of the rib waveguide. The last design parameter needed for a full optimization of the PN diode involved finding the exact location of the PN junction within the rib waveguide. The location was found by simulating Δn_{eff} from -5to 0 V as a function of the PN junction's location using the doping concentrations previously chosen, $N_A = 5 \times 10^{17}$ cm⁻³ and $N_D = 7 \times 10^{17}$ cm⁻³. The simulation results are depicted in Fig. 3.16. The junction position was assumed to be at zero in the middle of the rib waveguide, and from that point on, it was shifted every 30 nm to perform the simulation. The resulting curve shows that the maximum Δn_{eff} occurred when the junction position was shifted 50 nm to the right of the middle of the waveguide (towards the *n*-doped region).



Figure 3.16: (a)Simulation of the Δn_{eff} with respect to the junction location inside the rib waveguide. (b) Schematic diagram of the rib waveguide (w = 360 nm) showing the optimized location of the PN junction (50 nm off center towards the *n*-doped region).

Screen captures of the finalized PN diode design simulation can be seen in Fig.3.17. Figure 3.17(a) shows a complete view of the waveguide and how the dopants get depleted with increasing reverse bias, while Fig. 3.17(b) shows the carrier concentration in the *x* direction, considering the *y* direction in the middle of the rib waveguide.



Figure 3.17: Carrier depletion of a PN junction with increasing reverse voltage. (a) Map of carriers inside the waveguide. (b) Cut view: carrier concentration along the *x*-direction.

Overall, the figure shows dopant behavior under applied reverse voltage. It is noticeable in the figure that the p dopants are more responsive to the applied reverse bias than the ndopants; for example, at -5 V, the p-doped region exhibits a larger depleted region than the n-doped region, due to a lower dopant concentration. An updated version of the PN diode design is illustrated in Fig. 3.18 showing the dimensions of the p and n regions within the waveguide.



Figure 3.18: PN diode design showing the junction location in the rib waveguide.

The waveguide shown in Fig. 3.18 corresponds to only ~ 1.4 μ m of the entire active region width, which was the most critical part to design. In fact, the actual active region width relies on the traveling-wave electrodes' design, as it extends from the rib waveguide to the metal electrodes. The total of the active region could be estimated between a couple of μ m and up to ~ 50 μ m for our modulators and devices designed with similar dimensions. Two important considerations to keep in mind when designing the lateral extensions of the active region are: device losses and resistance R_D .

This chapter has discussed the influence of the optical mode in the waveguide, but some electrical effects also need consideration in the design of the active region. Because electrical effects and design considerations will be fully studied in Chapter 5, only those relevant to the design of the active region will be briefly discussed in this chapter. It has been repeatedly seen that high doping concentrations have a direct influence on optical loss, but it will be seen that the device resistance also has a significant influence on the electrical bandwidth.

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Figure 3.19: PN diode design an optical modulator of a silicon transceiver at 1.27 μ m showing optimized dimensions and doping concentrations.

The device resistance must be small to obtain a larger electrical bandwidth. Resistance can be reduced if the cross-sectional area of the measured material is increased, or if the material is heavily doped. An additional advantage of reducing silicon resistivity is that the material behaves more like a conductor under these conditions and can thus create a proper ohmic contact—and therefore a smooth transition—with the metal of the traveling-wave electrodes.

The remaining part of the active region was then designed to fulfill the above premises. The complete design is shown in Fig. 3.19. A 300-nm-wide section at each side of the waveguide that extends from the outermost edge of the waveguide slab (beyond 500 nm) was designed with a doping concentration of 1×10^{19} cm⁻³ (p + + and n + +). The height was kept at 100 nm to match the slab height. Two additional sections were added at each side of the waveguide slab, but they were designed with the same height as the rib waveguide (440 nm) to increase the cross-sectional area and reduce the device resistance. Additionally, the doping concentration was kept high to create ohmic contacts with the metal electrodes while reducing electrical resistivity, contributing to the reduction of the device resistance. The width of these sections depends on the traveling-wave electrodes. In other words, to create the rib waveguide, the silicon layer is partially etched in a region that covers 800 nm from the edge of the rib waveguide, as shown in Fig. 3.19.

The complete active region was simulated, again taking into consideration all the optimized parameters. A summary of the PN diode specifications is listed in Table 3.2. The table shows a comparison between the optimization results and the specifications of the

optical modulator needed to integrate with a silicon transceiver at 1.27 μ m (SILVER project). It can be seen that the optimized structure delivered a promising modulation efficiency V·L_{π}, as its simulated value is well below the minimum required for this FoM. At the same time, the small simulated V·L_{π} value established the possibility of designing short MZMs that would contribute to reducing the footprint of the silicon transceiver.

	Simulation Results	Project Requirements
Spec.	PN Diode	SILVER
λ_0	$1.27~\mu{ m m}$	$1.27\mu\mathrm{m}$
Voltage	5 V	< 5 dB
IL	< 5 dB	< 5 dB
VL_{π}	1.26 V·cm	2 - 3 V·cm
L_{π}	2.5 mm	_
Н	440 nm	440 nm
w	360 nm	-

Table 3.2: Simulated design parameters and FoM summary obtained for the optical modulator needed for applications at 1.27 μ m (1/2).

3.1.4 Modulator Speed

As mentioned in Chapter 2, the speed of an optical modulator depends on the RF signal propagation of the signal, the time free carries take to be driven out of the junction (intrinsic response), and the RC response time of the diode and the substrate; all three factors need careful consideration to guarantee high-speed operation. The influence of the electrical wave propagation through the phase shifter requires a separate optimization. This topic is explained in detail in Chapter 5.

The intrinsic time response of a PN diode can be calculated using Eq. 2.31. $f_{3dB} = 2.4/2\pi t_r$ [42], where the transit time $t_r = w/v_d$, and v_d is the drift velocity. A silicon modulator based on carrier depletion may work at high-speed operation when the electric field of a depleted PN junction reaches a value such as 10⁵ V/cm so that carriers can travel close to the

saturation velocity $v_s = 10^7$ cm/s in silicon [63]. Therefore, t_r can be approximated with the saturation velocity and the waveguide width w = 360 nm. Replacing all the values in the Eq. 2.31, the intrinsic time response of our device is ≈ 1.06 ps

The RC response time refers to the RC time constant of the active region, where R is the access resistance to the junction and C is the capacitance [1]. Additionally, the SOI substrate adds additional capacitance due to the sandwiched silicon oxide layer between the silicon layers and a resistance from the unintentionally doped silicon substrate. Fig. 3.20 shows a cross-sectional view of the device showing a superposed equivalent lumped circuit to represent the resistances and capacitances in the substrate.





The influence of the substrate's RC constant is a key parameter for the RF traveling-wave electrodes' design. Further explanation of this influence is provided in Chapter 5. However, the influence of the RC constant of the PN diode $R_D C_D$ is analyzed in this section. In general, the electrical cutoff frequency f_c , due to the $R_D C_D$ constant of the optical modulator, can be calculated as

$$f_c = \frac{1}{2\pi (R_D C_D)} \tag{3.1}$$

Equation 3.1 shows that a large f_c is obtained provided that $R_D C_D$ is minimized. The total resistance of the active region R_D is calculated using the electrical resistance Eq. 3.2

$$R_i = \rho_i \frac{l_i}{A_i} \tag{3.2}$$

where R_i is the resistance, ρ_i is the electrical resistivity, l_i corresponds to the width of the region analyzed, and A_i the cross-sectional area, calculated from the product between the height of the region H, and the length of the phase shifter L, as shown in the axonometric view of the PN diode from Fig. 3.21. All these parameters are different for each section of the active region.



Figure 3.21: Method to calculate the resistance of the active region of an optical modulator, where each section of the active region with doping concentration, doping type, or geometry changes needs to be calculated individually.

The value of the electrical resistivity, or specific electrical resistance ρ , depends on the material, and in our case, it depends on doping type and doping concentration. The theory and equations for calculating the resistivity with impurity density for boron- and phosphorous-doped silicon were obtained from [42] and [64]. Because our PN diode is built with different

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dimensions, dopant types, and doping concentrations, the total resistance of the active region R_D was calculated based on the relationship between doping concentration and resistivity (shown in Appendix A) by dividing the PN diode in regions sharing the same geometry, doping type, and doping concentration, as showed in the cross-sectional view of the PN diode of presented in Fig. 3.22. The total device resistance was calculated by adding the resistance of each region R_i . The figure illustrates that the widths of regions R_1 and R_8 depend on the gap between the signal and ground lines of the Coplanar Waveguide (CPW) electrodes (G), as the outermost heavily doped regions extend from ~ 800 nm from the rib waveguide to the metal electrodes, while regions R_2 to R_7 account for the same width regardless CPW electrodes' geometry. The geometries and resistivity values used for each section of the active region, as well as the calculated resistance when $G = 20 \ \mu$ m, are listed in Table 3.3



Figure 3.22: Schematic diagram of the process needed to calculate R_D of the optical modulator based on a PN diode, where each region of the active region with different geometry, dopant type, or dopant concentration requires an individual calculation.

Using these assumptions, an approximation to calculate the R_D was evaluated in Eq. 3.3 as

$$R_D = \left(5.6 + \frac{2.6 \cdot G}{20}\right) K\Omega \cdot \mu m \tag{3.3}$$

Section	Α (μm ²)	<i>l</i> (µm)	L (μ m)	ρ (Ω·cm)	Resistance ($\Omega \cdot \mu m$)
R_1	1.1×10^{-3}	3	2,500	0.0089	605
R_2	0.25×10^{-3}	0.30	2,500	0.0089	270
R_3	0.25×10^{-3}	0.50	2,500	0.0671	3,350
R_4	1.1×10^{-3}	0.23	2,500	0.0671	350
R_5	1.1×10^{-3}	0.13	2,500	0.0280	80
R_6	0.25×10^{-3}	0.50	2,500	0.0280	1,400
R_7	0.25×10^{-3}	0.30	2,500	0.0058	170
R_8	1.1×10^{-3}	G-5	2,500	0.0058	1,975

Table 3.3: Calculation of the total device resistance R_D method by adding the resistance of each section determined by different geometry or doping concentration.

which corresponds to the diode's resistance in $K\Omega$ per unit length (in μ m). The equation was approximated knowing that the resistance of the active region, which is composed of the sum of the individual resistance of sections R_2 to R_7 , is defined to be 5.6 $K\Omega \cdot \mu$ m. The resistance of the heavily-doped regions is composed of regions R_1 and R_8 which is calculated depending on the G dimension of the CPW electrodes, thus the G term in Eq. 3.3; e.g., if $L = L_{\pi} = 2.5$ mm, the resistance of the heavily-doped regions ($R_1 + R_8$) when $G = 20 \ \mu$ m is ~ 2.6 $K\Omega$, and is why the second term of Eq. 3.3 is normalized with respect to 20.

Furthermore, the capacitance of the active region was simulated using Eq. 2.29 and 2.30 obtained from electrostatic analysis in section 2.4.1, using the optimized N_A and N_D doping concentrations. The results of the calculated capacitance (C_D) per unit length (μ m) as a function of the reverse voltage are shown in Fig. 3.23. At -2.5 V, $C_D = 0.39 \text{ fF}/\mu\text{m}$. Using the calculated values of R_D and C_D we can determine the electrical cutoff frequency f_c of the entire active region of the optical modulator. This frequency varies from $f_c \approx 62 \text{ Ghz}$ for $G = 5 \mu$ m, to $f_c \approx 32 \text{ Ghz}$ for $G = 50 \mu$ m. This frequency range is well above the 10 Gbit/s operation requirement of the silicon transceiver.

The calculation of the resistance, capacitance, and electrical bandwidth contribution from the device completed the optimization process of the modulator's active region (phase shifter). Once the active region had been properly designed, the next step was to optimize the interferometric structures needed to change the phase modulation obtained in the active region to intensity modulation, so that it could be detected by the measurement equipment. The simulation of the interferometer structures needed to design the silicon optical modulator is described in the next section.



Figure 3.23: Calculation of the capacitance vs. reverse voltage of the PN junction of the SILVER modulator.

3.2 Interferometric Structures

The main interferometric structure selected for the silicon modulator designed to be part of a silicon transceiver used for applications at a 1.27 μ m wavelength was the MZI. This type of interferometer offers the temperature stability that a RR cannot offer, and which is a imperative feature for the expected application. Additionally, the structure selected as beam splitter and combiner is the MMI coupler due to its tolerance for fabrication inaccuracies and low excess loss at the output [54]. A complete analysis of the optimization carried out on these structures to fulfill the requirements of the SILVER project is provided below.

3.2.1 Multi-mode Interference (MMI) Couplers

To split and combine the light signal to the MZI, we designed 1×2 MMI and 2×2 MMI couplers. These interferometric devices were chosen because they are robust and have a high fabrication tolerance. As mentioned in Chapter 2, section 2.3.3, MMI couplers have a central waveguide of length L_{MMI} , and width W_{MMI} , as well as access waveguides of width W_{IN} and output waveguides of width W_{OUT} , that are separated by a distance between the center of the output waveguides D as shown in Fig. 3.24 for a 1 × 2 and a 2 × 2 MMI couplers.



Figure 3.24: MMI design dimensions chosen for the silicon modulator for applications at $\lambda = 1.27 \ \mu m$ (a) $1 \times 2 \ MMI$ and (b) $2 \times 2 \ MMI$.

The simulation and optimization of these structures at 1.27 μ m for our particular application was carried out using *FIMMWAVE* and *FIMMPROP Tools* by PhotonDesign. We started the optimization process of the 1 × 2 MMI design by choosing 1 μ m as the width dimension of the access (W_{IN}) and output (W_{OUT}) waveguides. This dimension was preferred over the 360 nm designed waveguide width to reduce refraction caused by a narrower waveguide. When choosing the main waveguide width W_{MMI} it is important to note that the multimode portion of the MMI needs to support more than three optical modes for proper interference, and that the distance between the two output waveguides must be sufficiently large to ensure easy fabrication of the trench between rib waveguides.

For the last consideration, the trench distance was set to 2 μ m, and because $W_{OUT} = 1$ μ m, the distance between the centers of the output waveguides was $D = 3 \mu$ m; also, the output waveguides were placed approximately $W_{MMI}/4$ with respect to the MMI center, using general self-imaging theory [40], [39]. Therefore, the width of the multimode section $W_{MMI} = 6 \mu$ m. The design of the 1 × 2 MMI is shown in Fig. 3.24(a).

An initial simulation was performed to determine if W_{MMI} was wide enough to support more than three optical modes. The 1 × 2 MMI was designed and simulated using *FIMMWAVE* and *FIMMPROP*, where it was found that the designed waveguide supported eighteen modes as shown in Fig. 3.25



Figure 3.25: Highest-order mode of a 1 × 2 MMI with multimode waveguide width W_{MMI} = 6 μ m.

To calculate multimode waveguide length L_{MMI} , the beat length (L_{π}) of the two lowestorder modes must be estimated first. Using Eq. 2.22, $L_{\pi} \approx 4n_r W_{MMI}^2/3\lambda_0$, where $\lambda_0 = 1.27 \,\mu$ m, $n_r = 3.497$ (ridge index), and $W_{MMI} = 6 \,\mu$ m, $L_{\pi} \approx 132.2 \,\mu$ m. For a 1 × 2 MMI, the first two-fold can be calculated using the equation for symmetric interference $L_{MMI} \approx 3L_{\pi}/8$, which is $L_{MMI} \approx 49.6 \,\mu$ m. Light propagation through the 1 × 2 MMI was then simulated starting with the calculated L_{MMI} , this value was varied until the proper L_{MMI} was found. Figure 3.26 shows the results of the light output (in percentage) as a function of L_{MMI} , where the curves of the outputs 1 and 2 overlapped as they yielded nearly the same output power. The total output is also seen in the figure. The maximum optical output was found at $L_{MMI} = 46.5 \,\mu$ m, and it delivered 96 % of the input power, which corresponds to a \sim 0.3 dB losses that are most likely caused by light coupling from the central waveguide to the output waveguides. Figure 3.27 shows the simulation results of the 1 × 2 MMI where the optical transmission output is at its maximum.



Figure 3.26: Light output percentage of the 1×2 MMI coupler designed for SILVER. The optimized MMI provides a 96.6% optical transmission, which corresponds to 0.3 dB of losses.



Figure 3.27: Optimization results of the 1×2 MMI coupler designed for SILVER. The optimized MMI provides a 96.6% optical transmission, which corresponds to 0.3 dB of losses.

The same procedure was followed to calculate the dimensions of the 2 × 2 MMI coupler. The widths of the access and output waveguides were chosen to be $W_{IN} = W_{OUT} = 1 \ \mu m$, and the distance between the centers of the two access and output waveguides was D = 2.4 μ m. This dimension was decreased from the chosen $D = 3 \mu$ m in the 1 × 2 MMI to limit the value of L_{MMI} to less than 100 μ m. In the case of 2 × 2 MMI couplers, the input and output waveguides are placed at 1/6 with respect to the MMI center using general restricted-imaging theory [40], [39]. Therefore, D represents one third of the width of the multimode waveguide, and thus, $W_{MMI} = 7.2 \mu$ m. Because this waveguide width is larger than the one designed for the 1 × 2 MMI coupler (which was 6 μ m), it was certain that the width of the 1 × 2 MMI coupler supported more than three modes. Using all the known dimensions, the beat length of the lowest-order mode was approximated $L_{MMI} \approx 190.33 \mu$ m. In the case of a 2 × 2 MMI coupler the 2-fold image was found at $L_{MMI} = L_{\pi}/2$, which yielded $L_{MMI} \approx 81.5 \mu$ m. The 2 × 2 MMI coupler the 2-fold image was found at $L_{MMI} = L_{\pi}/2$, which yielded $L_{MMI} \approx 81.5 \mu$ m. The 2 × 2 MMI coupler the 2-fold image of the most favorable length for maximum transmission output was found at $L_{MMI} = 90.5 \mu$ m, supplying 0.928 of the input power, which corresponds to a 0.64 dB of loss. The final design of the 2 × 2 MMI coupler is shown in Fig. 3.24(b).

A plot of the total output optical transmission as a function of the multimode waveguide length is shown in Fig. 3.28, and a screen shot of the 2 × 2 MMI coupler simulation can be seen in Fig. 3.29. Finally, Table 3.4 summarizes the design dimensions and simulation results of the designed MMI couplers.



Figure 3.28: Light output percentage of the 1×2 MMI coupler designed for SILVER. The optimized MMI provides a 92.8% optical transmission, which corresponds to 0.64 dB of losses.



Figure 3.29: Optimization results of the 1×2 MMI coupler designed for SILVER. The optimized MMI provides a 92.8% optical transmission, which corresponds to 0.64 dB of losses.

Parameter	1×2 MMI	$2 \times 2 \text{ MMI}$
L _{MMI}	46.5 <i>µ</i> m	90.5 μ m
W_{MMI}	$6\mu{ m m}$	7.2 μm
W_{IN}	$1\mu{ m m}$	$1\mu{ m m}$
W_{OUT}	$1\mu{ m m}$	$1\mu{ m m}$
D	$3\mu{ m m}$	$2.4 \mu \mathrm{m}$
Output Power	96.6 %	92.8 %
IL	0.3 dB	0.6 dB
λ	$1.27~\mu{ m m}$	$1.27~\mu{ m m}$

Table 3.4: Dimensions and simulation results of 1 × 2 and 2 × 2 MMI couplers at λ = 1.27 μ m.

3.2.2 Mach-Zehnder Interferometer (MZI)

Different geometries of MZIs were considered to design the interferometers needed to measure the optical transmission of the designed silicon optical modulators, as depicted in Fig. 3.30:


Figure 3.30: Schematic diagrams of the asymmetric and symmetric MZM designed for the SILVER project.

- The asymmetric MZI has proven to be an useful device to characterize silicon optical modulators because precise Δn_{eff} can be calculated from the optical transmission shift measured at different voltages.
- The symmetric MZI in push-pull configuration designed with a DC compensation region on each arm was also considered to balance the voltage and to obtain the largest bandwidth possible at high-speed operation (maximum phase shift of the MZI). The design of this type of MZI has rarely been demonstrated in other silicon optical modulators before now.

Static optical transmission simulations (applied DC voltage only) were performed by adding a loss of ~ 0.3 dB from passive waveguides (< 0.5 dB/mm), 0.6 dB from both 1 × 2 MMI couplers (beam splitter and combiner), and 1.3 dB from both 2 × 2 MMI couplers. The length difference between the arms of the asymmetric MZI was designed to be 40 μ m; this length was calculated so that the FSR of the optical transmission was small enough to allow for a measurement of at least three transmission periods. The simulation was completed using the power transmission equation of an asymmetric MZI as follows

$$I_{\rm MZI} = \frac{1}{4} \left[\exp\left(\frac{-\alpha L_1}{2} - j\frac{2\pi}{\lambda}n_{\rm eff}L_1\right) + \exp\left(\frac{-(\alpha + \Delta\alpha)L_1}{2} - j\frac{2\pi}{\lambda}\left(n_{\rm eff} + \Delta n_{\rm eff}\right)L_1\right) \right]^2 \quad (3.4)$$

where L_1 and L_2 correspond to the length of each of the modulator's arms (labeled as arm 1 and arm 2, respectively), and $\Delta \alpha$ and Δn_{eff} correspond to the absorption and effective refractive index change for an applied voltage between 0 and -5 V. The transmission of the MZI was used to calculate the entire transmission of the silicon optical modulator using the following equation:

$$I_{\text{total}}(\text{dB}) = I_{\text{MZI}} - 2IL_{\text{MMI}} - IL_{\text{prop}}$$
(3.5)

where $I_{\rm MMI}$ is the loss of each coupler and $I_{\rm prop}$ is the propagation loss due to waveguide roughness; all parameters are expressed in dB. Figure 3.31 shows the simulation results of an asymmetric MZM's optical transmission for reverse voltages from 0 to -5 V. The results provided an expected red shift of the transmission with increasing reverse bias. The simulated device produced an overall IL of approximately 3 dB and an ER of over 40 dB. These results fall within the required specifications of a silicon transceiver for applications at 1.27 μ m (SILVER project). Table 3.5 shows an updated comparison of the simulated results, where the value of the IL has been modified to include the losses from the MZI, the MMI couplers, and the passive waveguides. Additionally, the cutoff frequency range obtained from the *RC*_D calculation earlier in this chapter, as well as the ER obtained in the simulation of the MZI has been added for a complete comparison of the figures of merit.



Figure 3.31: Simulated DC transmission of SILVER's asymmetric MZI for reverse voltage from 0 to 5V. The simulated transmission shows that the simulated IL and ER fall within SILVER's requirements.

	Simulation Results	Project Requirements
Spec.	PN Diode	SILVER
λ_0	$1.27~\mu{ m m}$	$1.27\mu{ m m}$
Н	440 nm	440 nm
w	360 nm	_
IL	~ 3 dB	< 5 dB
ER	> 30 dB	> 10 dB
VL_{π}	1.26 V·cm	2 - 3 V·cm
L_{π}	2.5 mm	_
f_c	49.73 GHz	_
R_D (for $G = 20 \ \mu m$)	8.2 $K\Omega \cdot \mu m$	_
C_D (at 2.5 V)	$0.39~{ m fF}/\mu{ m m}$	_

Table 3.5: Final FoM comparison between the simulated optical modulator design for applications at $\lambda = 1.27 \ \mu m$ and SILVER requirements.

Conclusion

The design of silicon optical modulators is a challenging task that requires the consideration of several parameters, all interacting in different ways. Designing an all-silicon modulator requires that compromises be made among multiple parameters to allow for the best possible results. It was shown throughout this chapter that a small change in one parameter could change the response of the entire device. For that reason, orderly and careful consideration was followed to complete this complex task.

As shown throughout this chapter, the optimization of the PN diode was carried out with the purpose of finding the most appropriate active region to be used in an optical modulator that will eventually become part of a silicon transceiver for local network applications. In view of the SILVER project's required specifications for the silicon transceiver's design, the performance of various parameters of the PIPIN and PN diodes were compared. Overall, the PN diode proved to be the most appropriate device for the task, as it fulfills all the project requirements, even if the IL are higher than those of the PIPIN diode. The greatest advantage of selecting a PN diode is that it is a much simpler device to fabricate.

The optimization results of the active region based on a PN diode produced an optical performance that fulfills the requirements of SILVER project in terms of ER and IL (as presented in Table 3.5). Even a promising $V \cdot L_{\pi} = 1.26$ V·cm is foreseen for the PN diode design, guaranteeing high-speed operation of a short device.

The next challenge rests on the design of the RF traveling-wave electrodes, which are responsible for losses due to the different propagation speed of the optical and electrical signals, attenuation, and signal reflection caused by impedance mismatch. These topics will be discussed in Chapter 5 in the context of the entire design of the silicon optical modulators. Because the RF traveling-wave electrodes were designed using simulations and characterization results from the HELIOS devices, the processes and techniques used to fabricate the modulators are presented in advance in Chapter 4 to allow for a better understanding of the final design process.

4 Fabrication

Based on the designed parameters studied in the two previous chapters, complete Match-Zehnder modulators (MZMs) were assembled in GDS layout files at IEF for fabrication at CEA-LETI with compatible CMOS processes. CEA-LETI, one of the partner institutions of the SILVER project. This chapter presents a description of the process behind the fabrication of silicon modulators, beginning with the design of the IC mask layout and the process flow followed for completion. The first section of this chapter presents a description of the process followed to design and fabricate silicon optical modulators for applications at $\lambda = 1.55 \ \mu m$. Although these modulators had been previously designed, it is important to summarize essential aspects of their fabrication process, as they will be of great importance in Chapter 6. This section will include a summary of relevant modulator design considerations needed to understand the characterization results, followed by presentation of the processes that were utilized to fabricate the silicon modulators. The second section studies the different GDS layout parameters used to draw the modulators for integration with the silicon transceiver at $\lambda = 1.27 \ \mu m$ (SILVER modulators) in the IC mask layout, as well as the process flow and fabrication procedures designed for the fabrication of the optical MZMs with phase shifters based on PN diodes, for applications at $\lambda = 1.27 \ \mu m$.

4.1 High-Speed Optical Modulators at $\lambda = 1.55 \,\mu m$

4.1.1 Mask Design

The characterization results of the silicon optical modulators presented in this dissertation came from two different batches: namely, T582B and U532P. Batch T582B produced the first silicon optical modulators for high-speed operation up to 10 Gbit/s, while batch U532P delivered modulators working at a high-speed operation of up to 40 Gbit/s. A summary of the main components fabricated in T582B and U532P are presented in this chapter as needed for a thorough understanding of the characterization results presented in Chapter 6.

Batch T582B contained optical modulators using on ring resonators (RRs) and Mach-Zehnder interferometers (MZIs) for applications at $\lambda = 1.55 \,\mu$ m. Figure 4.1 shows a general view of the fabricated chip. Each chip contained over one hundred modulators with different parameters, such as length, type of interferometric structure, type of active region and CPW electrodes. The chip was divided into two regions, the MZMs region and the RRs region, and a total of 9 wafers (P01 to P09) were fabricated, each of which was fabricated with specified different doping concentrations. T582B was fabricated using standard lithography processes to create the *p*-slit in the middle of the rib waveguide, as described in Section 4.1.2.A. This batch produced silicon optical modulators with high-speed operation up to 10 Gbit/s designed with RRs based on PIPIN diodes and interleaved junctions, as well as with MZMs based on PIPIN diodes.

Overall, the measured modulation efficiency from batch T582B as well as the weak Signal-to-Noise Ratio (SNR) made the measurements at higher operation speeds more challenging. Additionally, the standard fabrication process of this batch increased fabrication variations in the alignment of the *p*-doped slit in the center of the waveguide that could have contributed to the moderate results of the modulation efficiency. The characterization results of the optical modulators from T582B, and the analysis due to its fabrication process, will be presented in the next chapter.

The latest generation of MZMs were fabricated on batch U532P. The fabrication process of batch U532P had a few differences in comparison with batch T582B; in particular, this batch was fabricated with the self-alignment process, which defines the active region in the center of the waveguide, as will be seen in Section 4.1.2.B. It is important to mention that

the fabrication of both batches was already in progress at the time of my arrival to *IEF*, but I carried out the characterization process of all the devices presented in this document. A summarized description of the devices of concern for this dissertation is presented next.



Figure 4.1: IC mask layout used for the fabrication of batch T582B.

A Mach-Zehnder Interferometers

The MZM side of chips T582B and U352P had several MZMs designed with PIPIN diodes in their active regions, as shown in Fig. 3.1. Table 4.1 summarizes the different parameters of the MZMs of both batches. The longer modulators were all designed with CPW electrode dimensions $W = 30 \ \mu$ m, and $G = 8.5 \ \mu$ m, while the shorter modulators were designed with CPW electrode dimensions $W = 5 \ \mu$ m, and $G = 25 \ \mu$ m. The CPW electrodes were created

High-Speed Optical Modulators at λ = 1.55 μ m

to provide a device-characteristic impedance of around 50 Ω or to deliver the fewest losses. Additionally, the width of the waveguides was targeted at 420 nm ±20 nm because there is an oxidation process to smooth out the flanks of the waveguide right after its patterning. This oxidation process reduces the waveguide width by ~ 10 nm the waveguide width.

Label	Active Region	Length (mm)	Electrodes		Batch	
			W (µm)	G (µm)	T582B	U532P
MZ01	PIPIN diode	6.7	30	8.5	1	
MZ03	PIPIN diode	4	30	8.5	1	
MZ04	PIPIN diode	1.8	10	25	1	
MZ06	PIPIN diode	4	30	8.5	1	1
MZ07	PIPIN diode	0.95	5	25	1	✓
MZ09	PIPIN diode	0.95	5	25	1	1
MZ11	PIPIN diode	0.95	5	25	1	1
MZ12	PIPIN diode	0.95	5	25	1	1
MZ14	PIPIN diode	2.9	30	8.5	1	1
MZ17	PIPIN diode	4.7	30	8.5	1	✓
MZ19	PIPIN diode	6	30	8.5	1	1

Table 4.1: List of MZMs based on a PIPIN diode fabricated in batches T582B and U352P.

B Ring Resonators

A varied group of devices was fabricated in the RR section of the chips of batch T582B. The RRs—based on a PIPIN diode and interleaved PN diodes—are of particular interest for the presentation of characterization results in Chapter 6. Table 4.2 lists the characteristics of the RRs of interest for this dissertation. The RRs based on a PIPIN diode were designed with a 50 μ m radious, and a 200 nm gap distance between the ring and the straight waveguide. These values were chosen from the results of a careful optimization of the structures to obtain a maximum ER at critical coupling [3]. The center of the rings was *n*-doped, while a squared exterior was strategically designed to be able to fit the RF electrode probes. An important

characteristic of RR01 and RR11 is that the PIPIN diode does not extend over the coupling region between the ring and the straight waveguide. For this reason, this region does not contribute to the change of refractive index. The length of the active region was therefore 292 μ m. Figure 4.2 shows a schematic diagram of the RRs design, with a zoomed-in view of the details inside the structure.



Figure 4.2: RR based on a PIPIN diode of batch T582B.

Label	Device	Active Region Length (mm)	Gap (nm)	Radius (µm)
RR01	PIPIN diode	292	200	50
RR11	PIPIN diode	292	200	50
RR64	Interleaved PN diodes	292	200	50

Table 4.2: Partial list of RRs fabricated in batch T582B.

The RRs based on interleaved junctions were designed with the same dimensions as the RR described above. Because phase modulation efficiency depends on doping concentration as well as on the length of p and n regions, the space charge extensions were analytically

described as a function of the applied reverse bias [62], assuming the space charge regions were empty. Using this analysis, the designed dimensions and doping concentration values were chosen to establish a large space charge extension variation when the modulator underwent reverse bias from 0 to 10 V. The lengths of the *p* and *n* regions were chosen at 400 nm and 300 nm, respectively, and their doping concentrations at 4×10^{17} cm⁻³ and 1×10^{17} cm⁻³, respectively [49]. Unintentionally doped regions were designed immediately next to both sides of the rib waveguide with the intention of enhancing carrier depletion inside the rib. Figure 4.3 illustrates a zoomed-in view of the design details of the interleaved junctions inside the RR.



Figure 4.3: RR based on interleaved diodes of batch T582B.

Interleaved diodes present a large fabrication process advantage in comparison with other structures because the position of the junction inside the waveguide is considerably more tolerant to shifts due to mask alignment. This lithography tolerance can create a random shift of the junction position, but even a PN junction shift of 100 nm in the waveguide direction may only induce an influence on the phase shifter's modulation efficiency.

4.1.2 Process Flow

The fabrication of silicon optical modulators takes place at *CEA-LETI. IEF* always works in close collaboration with *CEA-LETI* to come up with the best possible process for the fabrication of the silicon devices. The main roll of *IEF* is to draw the IC layouts that are utilized to fabricate the masks required for each technological fabrication level.

Two processes are described in this section to fabricate silicon modulators for applications at $\lambda = 1.55 \ \mu$ m. They were created to find the most accurate way of aligning the *p*-slit region inside the rib waveguide of the modulators whose active regions were based on PIPIN diodes. Complete fabrication of the silicon modulators required a total of ~ 50 technological steps [65]. However, in this document only the most significant fabrication levels will be described to provide a general visualization of the technology as well as to illustrate critical fabrication processes that ultimately influence the overall response of the devices.

A Standard Process

Batch T582B was fabricated using standard fabrication processes. Schematic diagrams of the fabrication steps are illustrated in Fig. 4.4 to Fig. 4.6. In the figures, the lithography levels are shown in azure blue boxes located between the diagram and the process description.

The process started with the preparation of an 8-inch, high-resistivity SOI *SOITEC* substrate (resistivity > 1 K Ω ·cm) [66] with a silicon dioxide—BOX—layer thickness of 2 μ m, and a silicon overlayer of 400 nm. The thickness of the silicon layer was initially measured, and and a passivation layer of 10 nm was deposited on top of the substrate by thermal oxidation. A silica hard mask with a thickness of 240 nm was deposited using a Hight Temperature Oxide (HTO) technique. The first lithography level was used to pattern the rib waveguide, using a careful dimensional control to create the side openings with accurate dimensions. The hard mask was etched down to the silicon layer (240 nm) and the photoresist was stripped off (level 1). This step is one of the most critical, as it patterns the width of the waveguide by etching ~ 285 nm of the exposed silicon layer to create the rib waveguide. After the etching was completed, a slab of ~ 110 nm was created. The 240 nm of the silica hard mask were etched off completely. A new protection layer via thermal oxidation was deposited in the superficial silicon layer, and a new lithography step (level 2) was taken to create the opening region of the central slit for doping implantation. This is also a critical step, as it required

accurate alignment in the middle of the rib waveguide. A central slit of ~ 100 nm was created and boron implantation was carried out over a depth of 390 nm. The photoresist was then stripped off and additional four lithography steps were carried out to create each of the doping regions of the active region P + +, N + +, P +, and N + (levels 3 to 6). It is important to note that the implantation of the N + region was carried out using a directional plasma method, as the width of this region was only 50 nm. All four lithography steps required precise alignment to ensure proper doping implantation in the designed regions.

At this point, a thermal annealing was carried out on the substrate at 1050° C for 30 seconds to activate the dopants in the waveguide. A protection silicon dioxide layer of ~ 700 nm was deposited on the substrate with High Density Plasma (HDP). A new lithography step was performed to pattern the vias of the modulators (level 7), followed by the complete etching of the silicon dioxide layer. The remaining photoresist was stripped off, and a silicide layer of at least 400 nm was deposited onto the substrate to create an ohmic contact between the silicon layer and the metal electrodes. This process was followed by the deposition of metal combination stack Ti/TiN/AlCu/Ti/TiN to create electrodes. A final lithography step was taken to pattern the area of electrodes and remove the material from the rest of the substrate (level 8), followed by the stripping of the remaining photoresist material in the substrate.

Process FlowHELIOS: Standard-Aligned				
	Process	Lithography Level	Description	
			SOI Substrate Si layer thickness: 400 nm BOX thickness: 2 μm	
			Deposition: SiO ₂ hard mask HTO of thickness 240 nm	
			Lithography and SiO₂ etching Hard mask etching	
			Stripping	
			Silica etching: 285 nm waveguide patterning	
			Etching: complete hard mask stripping	
		2	Lithography and doping implantation P region (boron): 3×10 ¹⁷ cm ⁻³	
Si	SiO ₂ Metal Resist	P++ P+	P N+ N++	

Figure 4.4: Process flow HELIOS: standard alignment (1/3).



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Figure 4.5: Process flow HELIOS: standard alignment (2/3).



Figure 4.6: Process flow HELIOS: standard alignment (3/3).

B Auto-Aligned Process

Batch U532P was fabricated using an auto-aligned fabrication process to accurately define the *p*-slit in the middle of the rib waveguide. Schematic diagrams of the main part of this process are shown in Fig. 4.7 and Fig. 4.8. The preparation of the SOI substrate mached the previous process. The particularity of the auto-aligned process is that the first lithography mask was used to pattern the openings (in the hard mask) of the waveguide and center slot simultaneously to obtain perfect alignment between the two. The process was followed by lithography steps to create the rib (level 2) and to implant the *p*-dopants in the center slit (level 3) using the same silica hard mask to define both regions. Two additional lithography steps were implemented for the doping implantation of P+ and N+ regions (levels 4 to 5) while taking advantage of the hard mask to define the boundary limits of both regions in the rib waveguide. In this process, directional plasma implantation of the N+ region was also used. The hard mask was etched off after the fabrication of the central region was competed. Two additional lithography steps were carried out for the doping implantation of the P + +and N + + regions (levels 6 to 7). From this point forward, the same fabrication steps of the previous process were carried through completion of the optical modulators (steps 4 of Fig.4.5 to step 8 of Fig. 4.6).

Process FlowHELIOS: Auto-Aligned					
	Process	Process Lithograpy Deso Level			
			SOI Substrate Si layer thickness: 400 nm BOX thickness: 2 μm		
			Deposition: SiO ₂ hard mask HTO of thickness 240 nm		
			Lithography: auto-alignment hard mask etching. Simultaneous patterning of waveguide and center slot openings		
			Stripping		
		2	Lithography: slot protecction		
			Silica Etching: waveguide patterning Critical: dimensional control of waveguide etching		
			Stripping		
		3	Lithography: doping implantation P region (boron): 3×10 ¹⁷ cm ⁻³		
Si	SiO ₂ Metal Resist	P++ P+	P N+ N++		

Figure 4.7: Process flow HELIOS: auto-alignment (1/2).



High-Speed Optical Modulators at $\lambda = 1.55 \ \mu m$

Figure 4.8: Process flow HELIOS: auto-alignment (2/2).

4.1.3 Characterization of Fabricated Wafers

Once the fabrication of the devices was completed, there were nine wafers with optical modulators to characterize on both batches. Various methods were followed to dice the wafers. Some wafers were entirely diced at *IEF*, while others were partially diced at *CEA-LETI* with optical quality, and completely separated at *IEF* upon arrival. Usually, wafer halves, quarters, or in some cases specific pieces were sent to *IEF* for characterization. The handling of the wafers after dicing is completed is one of the biggest concerns as chip-to-chip rubbing causes chip facets to crumble, affecting the transmission measurements such as the characterization of the modulators' required butt-coupling of the light into the chip.

In general, every 8-inch wafer had 21 lines and 21 columns filled with chips, which is a total of 441 chips of each design (RRs and MZMs. A matrix-like system was used to label the chips as depicted in Fig. 4.9: e.g., *L02C11* designated the chip located in the second row (*ligne* in French) *L02*, and the eleventh column *C11*.



Figure 4.9: Chip labeling system used for all the fabricated wafers.

The microscope pictures presented in Fig. 4.10 show a section of each of the two different chips fabricated in batch T852B.





After the wafers had been properly diced, the first step in the characterization process was to test the optical DC transmission response of modulators from chips located in different places within the wafer. This method is used to locate an area where the modulators produced an acceptable response in terms of losses, transmission shape, and—naturally—modulation. This was a challenging and lengthy process, as the optical responses were particularly different from chip to chip. To understand such results, the active regions of a few modulators where the optical transmission did not meet the expected results were inspected using a Scanning Electron Microscope (SEM). The results are shown in Fig. 4.11.

The micrographs show the active region of a MZM from each batch. The micrograph from Fig. 4.11a(a) corresponds to the active region of a MZM from wafer P05 of batch T582B where the rib waveguide, as well as the thick silicon dioxide protection layer, are clearly seen. On top of the silica layer are some leftover silicide materials located exactly above the areas were the silicon layer was etched off to form the rib waveguide. The leftover silicide is the result of not being able to completely etch off the metal layer from the surface. The fact that these silicides are so close to the area where modulation occurs is the reason why the optical transmission in batch T582B modulators had multiple issues, including substantial losses due to metal material in close proximity to the modulation region. It is worth noting that the leftover silicides were not present in all the fabricated wafers, as we were able to characterize modulators from one of the wafers (P01) of this batch that worked at a high-speed operation of 10 Gbit/s [49], [4], [56].



Figure 4.11: SEM micrographs showing fabrication defects in the active region of a MZM of (a) batch T582B and (b) batch U352.

The same procedure was followed for the optical modulators tested from batch U352P. The cross-sectional view of a typical waveguide is shown in the micrograph depicted in Fig. 4.11b.No leftover silicides were found in any of the wafers of this batch; however, it can be seen that the active region does not have a sharp profile, but a rounded, wet-etch type profile. Contrary to the previous leftover silicides, the shape of the modulators of batch U352P was found in all of the wafers.

Upon measuring the dimensions of some devices from chips located in different places across the same row, we found that the width of the active region changes from column to column. This was a known *CEA-LETI* fabrication technique to produce devices with an interesting range of waveguide width dimensions that was accomplished by changing the lithography doses for each column. Micrographs of the waveguide width measurements can be seen in Fig. 4.12. The four micrographs were taken from the left half of wafer P04 by choosing chips over a range that spanned from the left edge of the wafer (P05-L05C02) to the rightmost column available (P05-L05C11). The figures show how waveguide width decreases from left to right (increasing column number), going from ~ 420 \pm 10 nm down to ~ 321 \pm 10 nm corresponding to the wafer center (according to the SEM measurement at the top of the rib) within a range of nine columns. It was later found that the modulators that worked at high speed were found in columns toward the middle of the wafer (columns C09 to C11) where the waveguides had the narrowest widths. Optical transmission response on devices

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characterized near the edge of the wafers was measured unsuccessfully, due to the fact that at this location the waveguide width gets close to the limit for the input light beam to propagate in single-mode.



(c) Chip P04-L05C07





CEA-LETI also inspected some of the modulators, especially parts that required critical dimension control, such as waveguide widths and start coupler-based beam splitters. SEM micrographs of this inspection can be seen in Fig. 4.13. Figure 4.13(a) shows the top view of a waveguide profile with a 394 nm width measurement, validating that the waveguide dimensions obtained are suitable for single-mode propagation of the guided beam. Figure 4.13(b) shows a star coupler-based beam splitter, depicting all the corners and right angles that make the structure challenging to fabricate. It is noticeable in the figure that all the

designed right angles in reality become smooth rounded corners, which is an important design consideration when dealing with structures that require the intersection of segments with right angles or smaller ones.



(a) Waveguide width

(b) Star coupler

Figure 4.13: SEM micrographs showing the dimensions of the active regions of different MZMs located in different places throughout wafer U352P. Picture courtesy of *CEA-LETI*.

4.2 High-Speed Optical Modulators at $\lambda = 1.27 \ \mu m$

4.2.1 Mask Design

Designing the layout of the silicon transceiver required all of the collaborators on the SILVER project to coordinate the proper integration of all the different devices into a single transceiver. The first generation of silicon transceivers for local network applications included the fabrication of single modulators, as well as complete transceivers, in the same die. The reticle area was divided into three regions among the most crucial devices of the transceiver—APDs, triplexers, and modulators—where half of the space was assigned to the modulators. The total reticle area was established to be 7.3 mm \times 7.3 mm, and the area assigned to the modulators was 7.3 mm \times 4.45 mm.

The GDS file was drawn in the CAD and layout design editor *L-Edit* by *Tanner EDA*. Figure 4.14 shows a general view of the complete reticle layout design, including APDs, triplexers,

and modulators. All the modulators drawn in the layout file are different. Each modulator was chosen with different parameters to cover all possible outcomes in the characterization process, as well as common fabrication issues, such as dimensional control, alignment, and doping concentration.



Figure 4.14: GDS schematic of SILVER's designed reticle including all three main parts of the silicon transceiver design: modulators, APDs, and triplexers.

The reticle shown in Fig. 4.14 has a total of fifteen MZMs distributed in a vertical configuration and parallel to each other. The active region of all the MZM is based on the PN diodes, whose design was presented in Chapter 3. The GDS layout for SILVER included a combination of asymmetric MZIs, symmetric MZIs, asymmetric MZIs integrated with other

components (complete transceivers). Additionally, three straight waveguides were designed to calculate the losses in the modulators due to passive waveguide. Due to the setup of the characterization bench at *IEF*, the electrical pads and the waveguides were designed orthogonally to each other. In our case, the electrical pads were designed in the same direction as the arms of the MZMs, which means that the access input and output waveguides were designed with a 90° angle turn at both ends. The different characteristics of the MZMs' components and passive devices designed in the IC layout are described next.

Doping implantation is a particularly sensitive step in the silicon optical modulator fabrication process, as the outcome of the implantation ultimately determines the operation of the device. The results obtained in HELIOS helped determine areas for improvement in future generations of silicon modulators. It was verified in [5] that theoretical doping implantation is difficult to achieve in the fabrication process. The implantation optimization process depends on many factors, such as lithography tolerances, energy, dose, and implantation time, as well as dopant migration during thermal annealing. The design of the optical modulator for a silicon transceiver at 1.27 μ m eliminates a few fabrication steps by using a PN diode. Additionally, to include possible alignment outcomes of the doped implantation regions, three design configurations of the diode's PN junction were proposed, as illustrated in Fig. 4.15. In addition to the obvious PN junction design shown in Fig. 4.15(a), where the dopant implantation at the junction is exactly aligned, it was also suggested that the dopant implantation be made by leaving a 50 nm gap at the junction between the *p*- and the *n*-doped regions, as shown in Fig. 4.15(b), and by overlapping both doping regions by 50 nm, as shown in Fig. 4.15(c). The implementation of these three proposed junction designs will allow us to account for unmanageable fabrication end results during the doping implantation step.

A Mach-Zehnder Modulators

The modulator part of the reticle has in total fifteen MZMs. There are five asymmetric MZM with an arm difference ΔL of 40 μ m, of which four have phase shifters lengths of L = 2 mm, and one has a length of L = 0.8 mm. These lengths were chosen after considering results obtained from the previously designed modulators (based on a PIPIN diode), in which high-speed results at 40 Gbit/s were obtained on a modulator with a phase shifters length of 0.95 mm



Figure 4.15: Schematic diagram of three different designs of the PN diode's junction configuration. (a) Both doping regions are designed at the junction of the PN diode. (b) The doping concentration leave a 50nm gap at the junction. (c) Both doping concentrations overlap 50 nm at the junction.

[5]. In the new design, it was decided that a 2-mm-long modulator could provide improved results over the 0.95-mm-long device because, as seen in Chapter 3, the VL_{π} of the modulator designed for applications at a 1.27 μ m wavelength was simulated to be 1.26 V·cm, therefore ensuring maximum modulation efficiency with low losses. Moreover, a length of 0.8 mm was chosen to test high-speed results on a shorter device with reduced ER.

Each of the 2-mm-long asymmetric MZMs has different parameters. There is one modulator for each of the PN junctions alignment configurations shown in Fig. 4.15, and the fourth modulator was designed with an aligned PN junction, and a 2 × 2 MMI coupler. The short asymmetric MZM was also designed with an aligned PN junction. Figure 4.16 shows one of the asymmetric MZMs drawn in the ICs layout reticle.



Figure 4.16: IC layout view of (a) an asymmetric MZM showing complete details of the (b) location of the active region, (c) pn junction, and (d) junction alignment.

Figure 4.16 shows one of the asymmetric MZMs drawn in the IC's reticle. Figure 4.16(a) shows a top view of a complete device and inset views of the details between the CPW electrodes, such as the waveguide's exact location between the signal and ground electrodes (Fig. 4.16[b]), doping concentrations (Fig. 4.16[c]), and PN junction configurations (Fig. 4.16[d]). The reticle also includes the first designed symmetric MZMs with DC push-pull compensation zones at 1.27 μ m. There are ten symmetric MZMs with phase shifter lengths of 0.8 mm, 1 mm,

2 mm, and 2.5 mm. There is one modulator with a 2.5 mm phase shifter length and aligned PN junction, and three modulators for every other length, each having one of the PN junction alignments. Although the electrode configuration of the symmetric MZMs is different, the location of the waveguide inside the signal and ground electrodes was designed with the same dimensions as the asymmetric MZMs CPW. Thus, when zooming in to a symmetric MZI, the active region looks exactly the same as in Fig. 4.16(b), Fig. 4.16(c), and Fig. 4.16(d). Three additional MZMs were designed in the triplexers' part of the reticle, as seen in Fig. 4.14. These modulators were specifically designed for and integrated with the other components, to form a complete transceiver, which has been integrated into the first generation of SILVER chips. An asymmetric 2-mm-long MZM was the modulator design chosen to be integrated with the other components.

B CPW Electrodes

The signal and gap of the CPW electrodes were designed with the dimensions found in Chapter 5, $W = G = 20 \ \mu$ m. Additionally, the ground electrode was designed with a width of 100 μ m, as determined by a previous study on the ground electrode width influence. Because of the geometry of the CPW RF probes available in the characterization bench, most specifically the pitch (overall distance G-S-G electrodes), tapered pads were added to the CPW electrodes. These pads consist of two main regions measuring 100 μ m in length. One region is used to taper up the signal electrode from 20 μ m up to 40 μ m, and taper down the ground electrodes by keeping a 20 μ m gap between the signal and ground electrodes in the transition area. The second region provides sufficient room for the probes to be placed comfortably. SILVER's IC mask layout has two different electrode pad designs.

In the modulator's reticle, the CPW electrodes of all the stand-alone modulators were drawn parallel to the MZMs' arms to fulfill the *IEF*'s characterization bench, as illustrated in Fig. 4.17(a). In the triplexer's section, the CPW electrodes of the modulators integrated in the transceivers were drawn perpendicular to the MZMs' arms. This second design was created to fulfill the setup of the characterization bench at *III-V Lab*, as they are in charge of the complete transceiver's characterization. The 90° electrode turn was designed by using concentric quarter-circles, while maintaining the electrode-gap width dimensions in the turns. The pad regions were designed with a length of 50 μ m, as shown in Fig. 4.17(b). An

additional traveling-wave electrode was designed next to the CPW electrodes and surrounding the unbiased arm of the MZMs, with the purpose of applying a DC voltage to the unbiased arm of the MZMs to balance the phase shift and increase the modulation efficiency of the modulators.



Figure 4.17: Types of pads designed to access the CPW electrodes of a MZM: (a) parallel to the MZM's arms, (b) perpendicular to the MZM's arms.

C Passive Optical Devices

All the dimensions of the passive components were drawn 40 nm larger than the designed dimensions; e.g., the waveguides were drawn with a 400 nm width by adding 20 nm on each side of the optimized 360 nm waveguide, because the fabrication process includes an oxidation step (HTO passivation) to reduce the roughness of the waveguide flanks that produces a reduction of the overall waveguide width. To couple the maximum amount of light to the sample, all input and output waveguides were designed with a width of 3 μ m. From that width, progressive tapers were used over a length of 300 μ m to gradually reach the designed waveguide length of 400 nm. Grating couplers were added at the input and output waveguides. They were designed with the dimensions seen in Fig. 4.18. The grating couplers were designed and placed in the reticle so as to allow to them to be cut off, if needed, without affecting other devices in the chip (e.g., if we want to characterize the modulators over a large spectral range,

as the response of grating couplers specifically depends on a specific wavelength—limited spectral range).



Figure 4.18: (a) Bragg grating's designed dimensions for the input and output waveguides of the MZMs with an (b) inset showing the dimensions of the waveguide height variation period.

The MMI couplers were drawn using the dimensions calculated in Section 3.2.1. Since the input and output waveguides of the MMI couplers were designed with a width of $W_{IN} = W_{OUT} = 1 \ \mu$ m, transition tapers were drawn over a length of 10 μ m to gradually move from or to a waveguide width of 400 nm.

The vertical placement of the modulators was another constraint faced, as waveguide turns took up a considerable amount of space in the mask. To make light turn smoothly, ample 90° turns were preferred and were designed with quarter-circles. The turns of the MZMs' arms were designed with circumference arcs of 30° and a radius of 190 μ m, as illustrated in Fig. 4.19. A radius of ~ 200 μ m was used for the longer arm of the asymmetric MZM. These dimensions were chosen so that both arms of the MZM ended up at the exact distance as the waveguide separation *D* of the MMI couplers.



Figure 4.19: Schematic diagram of the waveguide turns designed to draw the arms of the optical modulators.

D Wafer Variants

Doping concentration is a sensitive fabrication process, as it depends on multiple parameters to achieve the targeted concentration. Furthermore, precise measurement of the final concentrations in the device is almost impossible. Because silicon modulators depend greatly on doping concentration, multiple wafers will be fabricated for the optical modulators designed to be integrated in the silicon transceiver for local network applications. Table 4.3 lists the different wafers and doping concentrations proposed, where *PXX* denotes the wafer labeling system. Wafer P01 is planned to have the silicon optical modulators fabricated with the optimized p+ and n+ doping concentrations. The other six pairs of doping concentrations were chosen to have higher or lower doping concentration levels of both dopants on the same wafer, or a combination of high and low dopant concentration levels for each type of impurity.

Finally, Fig. 4.20 shows a screen capture of the finalized reticule modulator region of the ICs mask layout. Each modulator has been labeled, and a summarized description of the location and design parameters of every modulator is shown in Table 4.4.

High-Speed Optical Modulators at λ = 1.27 μ m

Doping Concentration			
Wafer	P+	N+	
P01	$5 \times 10^{17} \text{ cm}^{-3}$	$7 \times 10^{17} \text{ cm}^{-3}$	
P02	$5 \times 10^{17} \text{ cm}^{-3}$	$1 \times 10^{18} \mathrm{~cm}^{-3}$	
P03	$5 \times 10^{17} \text{ cm}^{-3}$	$4 \times 10^{17} \mathrm{~cm}^{-3}$	
P04	$1 \times 10^{18} \mathrm{~cm}^{-3}$	$7 \times 10^{17} \mathrm{~cm}^{-3}$	
P05	$7 \times 10^{17} \text{ cm}^{-3}$	$7 \times 10^{17} \text{ cm}^{-3}$	
P06	$1 \times 10^{18} \mathrm{~cm}^{-3}$	$1 \times 10^{18} \mathrm{~cm}^{-3}$	
P07	$4 \times 10^{17} \mathrm{~cm}^{-3}$	$4 \times 10^{17} \mathrm{~cm}^{-3}$	

Table 4.3: Projected wafer batches and doping concentrations of the optical modulator designed for local access network applications.



Figure 4.20: GDS screen capture of the complete design of SILVER's modulators. Only the part of the reticle with stand-alone modulators is shown.

I/O	Device	Length (mm)	Junction Alignment	Interferometer
1	Straight Waveguide	-	-	-
2	Symmetric MZI	0.8	Overlapped	1×2 MMI
3	Symmetric MZI	0.8	Aligned	$1 \times 2 \text{ MMI}$
4	Symmetric MZI	1	Overlapped	1×2 MMI
5	Symmetric MZI	0.8	Separated	1×2 MMI
6	Symmetric MZI	1	Aligned	$1 \times 2 \text{ MMI}$
7	Symmetric MZI	1	Separated	1×2 MMI
8	Straight Waveguide	-	-	-
9	Symmetric MZI	2.5	Aligned	
10	Asymmetric MZI	0.8	Aligned	1×2 MMI
11	Symmetric MZI	2	Overlapped	$1 \times 2 \text{ MMI}$
12	Symmetric MZI	2	Aligned	$1 \times 2 \text{ MMI}$
13	Straight Waveguide	-	-	-
14	Asymmetric MZI	2	Separated	$1 \times 2 \text{ MMI}$
15	Symmetric MZI	2	Separated	$1 \times 2 \text{ MMI}$
16	Straight Waveguide	-	-	-
17	Asymmetric MZI	2	Aligned	$1 \times 2 \text{ MMI}$
18	Asymmetric MZI	2	Overlapped	$1 \times 2 \text{ MMI}$
19	Asymmetric MZI	2	Aligned	2×2 MMI
20	Asymmetric MZI *	2	Aligned	$1 \times 2 \text{ MMI}$
21	Asymmetric MZI *	2	Aligned	$1 \times 2 \text{ MMI}$
22	Asymmetric MZI *	2	Aligned	$1 \times 2 \text{ MMI}$

 $^{\ast}\,$ MZMs integrated in the transceivers drawn in the triplexers region of the reticle

Table 4.4: List of MZMs included in the SILVER reticle. The list includes type and specific parameters of each modulator as well as the location of the straight waveguides drawn in the IC mask layout.

4.2.2 Process Flow

The process flow created for the SILVER devices was based on the processes described in Section 4.1.2, differing in the order that the lithography levels are to be completed. This process flow implements the doping implantation before the etching of the rib waveguide (levels 1-4) in order to implant the silicon layer when it still has the same thickness everywhere. This is necessary because the rib waveguides of the silicon modulators designed with PN diodes have the same doping concentrations in regions of different heights, as shown in Fig. 4.21. The PIPIN diode shown in Fig. 4.21(a) was mostly designed with regions that had the same doping type, concentration and height, except for the N+ section that had a 50-nm-wide region in the waveguide rib that was implanted using a directional plasma method. The PN diode, on the other side, shares the same doping type and concentration in waveguide regions with different heights, as depicted in Fig. 4.21(b). For that reason, the dopants need to be uniformly distributed in all the regions that make up the rib waveguide.



Figure 4.21: Doping implantation reasoning for choosing of the order of the fabrication steps of a (a) PIPIN diode, and (b) a PN diode.

Schematic diagrams of the main steps projected for the fabrication of the modulators designed for local network access are depicted in Figs. 4.22 and 4.23. These devices are set to be fabricated in a high-resistivity SOI substrate with a 440-nm-thick silicon overlayer and a 2- μ m-thick BOX [2]. Four consecutive photolithography steps are needed for the doping implantation of regions *P* + +, *N* + +, *P*+, and *N*+ (levels 1-4), followed by the deposition of a silica hard mask. A new lithography step is needed to etch the silica mask and pattern the rib waveguide (level 5). Upon completion of this process, the photoresist will be stripped off so that the silicon is etched in the openings patterned by the hard mask. This is a critical step, as it needs dimensional control to ensure that the width of the waveguide is fabricated within the designed dimensions. At this point, the hard mask can be etched off completely,
and a passivation HTO layer may be deposited on the surface. A silicon dioxide layer with a thickness of \sim 700 nm will be deposited to protect the active region and avoid unwanted oxidation. The lithography steps to create the metal electrodes (levels 6-7) will follow the same process as for previous generations of silicon optical modulators.



Figure 4.22: Process flow for SILVER's modulators (1/2).

Process Flow SILVER				
Process	Lithography Level	Description		
	5	Lithography and hard mask etching Pattering of waveguide openings		
		Stripping		
		Silicon etching: waveguide patterining Critical: dimensional control of waveguide etching		
		Hard mask etching		
		SiO_2 Deposition : PECVD 700 nm Protects the active region, avoids unwanted oxidation		
		Stripping		
	6	Lithography and SiO₂ etching Via patterning		
		Stripping		
		Metal deposition		
	7	Lithography and metal etching Electrode patterning		
		Stripping		
Si SiO ₂ Metal	Resist P++	P+ P N+ N++		

Figure 4.23: Process flow for SILVER's modulators (2/2).

Conclusion

This chapter has delivered essential insights on the relevance of the fabrication process, where no trivial fabrication steps exist and every step is highly dependent on the previous one. The fabrication processes and the results obtained on the devices designed for applications at 1.55 μ m were presented in order to elucidate the most critical fabrication concerns. In particular, the importance of choosing appropriate waveguide rib dimensions and envisioning fabrication tolerances that may otherwise cause the modulator not to work—especially if the waveguide width chosen is overly close to the limit of multimode propagation—was emphasized.

Accurate device fabrication can also be achieved using creative methods and known techniques to vary certain parameters. This was the case of the waveguide width variation with chip horizontal position within the wafer, where different lithography doses were utilized to achieve these results. Two different process flows were presented to fabricate the same device. While both processes produced high-speed silicon modulators, the advantages and disadvantages of choosing one process over the other will become clearer in the next chapter. Nevertheless, it was important to test both processes to improve the technology and learn from the results obtained.

The design considerations and results seen in T582B and U532P were used as guidelines for drawing the IC mask layout for the modulators designed for applications at $\lambda = 1.27 \ \mu m$. The advantage of these modulators is that the active region was designed with a simpler structure (PN diode versus PIPIN diode), reducing the number of fabrication steps that require close attention to critical dimensions. The process is more straightforward, yet not trivial, as it consists of a complete new set of dimensions, optical structures and integrated devices. However, the successful fabrication of the previous generations of silicon modulators provided new insights on different fabrication problems that may not have been considered for the fabrication of future generations of silicon optical modulators.

5 RF Electrodes

RF signal propagation is a key element in determining the overall performance of high-speed silicon optical modulators. An applied electrical signal is needed to drive the modulator, but this signal needs to co-propagate in such a way that it does not limit the device's speed of operation. The problem with applying an RF signal to a MZM, is that these devices are usually a few millimeters long, and when the frequency increases, the electrical wave attenuates, causing a decrease in the ER. In addition, the co-propagation speed mismatch between the electrical and optical signals is an inherent effect that is more visible in longer devices because the speed difference between both waves becomes larger with length. Impedance mismatch is the third concern when applying an RF signal to both, MZMs and RR modulators, as it causes signal reflections that contribute to the ER decrease of the device. Consequently, the design of the driving electrodes needs careful, efficient and accurate modeling techniques to simultaneously achieve large bandwidth as well as low optical and microwave loss. In the case of RR modulators, signal propagation is not an issue of concern because their sizes are usually only a few dozens of micrometers.

The traveling-wave electrodes of MZMs presented in this dissertation were designed with CPW electrodes in Ground-Signal-Ground (GSG) configuration to adjust to the high-speed RF probes available in the characterization bench. The complete device was transformed into a transmission line equivalent circuit to simulate the total loss, impedance, phase velocity, and RC response. This chapter presents the analysis of the CPW electrode design used for silicon optical modulators based on PIPIN and PN diodes embedded in the arms of MZMs. The theory of the RF traveling-wave electrodes, and the influence of the electrical signal propagation

Chapter 5. RF Electrodes

on modulation speed, electrical, and optical bandwidths will be explained, followed by a description of the RF electrode model chosen for the optical modulators. A comparison between the simulation and experimental results of the CPW electrodes of two silicon optical modulators designed for applications at 1.55 μ m (HELIOS project) are presented to validate the effectiveness of the optimization model used to design CPW electrodes. Finally, the expected results of the CPW electrodes designed for the optical modulator required for applications at 1.27 μ m (SILVER project), and the explanation for the chosen design are analyzed.

5.1 Transmission Line Fundamentals

5.1.1 Basic Theory

Various planar transmission line configurations include stripline, slot line, microstrip, and coplanar waveguide, with the latter two topologies being the most common used for high-speed applications. The microstrip configuration consists of a strip conductor on the surface of a substrate (dielectric or semiconducting) with a conduction ground plane on the bottom of the substrate. This configuration requires a challenging characterization setup because the electrical signal cannot be applied on the same plane. The CPW configuration has the signal and the ground conductors on the surface of the substrate. The electrode configurations along with their respective electric field distribution are depicted in Fig. 5.1. Usually, the center conductor carries the signal, while the other two conductors are grounded (GSG configuration). There is also a two-electrode configuration in signal-ground configuration (SG). Although both of these configurations allow for easy access to all the connections in the same layer, CPW electrodes in GSG configuration provide lower propagation loss than other transmission line configurations. Additionally, this configuration was compatible with the experimental equipment in the characterization room. For that reason, they were selected as the traveling-wave configuration of choice to drive the available silicon optical modulators.

In general, a CPW transmission line is represented with an equivalent circuit model that has the four distributed line lumped parameters *R*, *L*, *C*, and *G*, where *R* is the loss from the transmission line conductors (DC and the high-frequency loss due to the skin-effect), the *L* is the inductance (the energy stored in the magnetic fields that link the conductors), *G* is the conductance due to dielectric loss in the media surrounding the conductors, and *C* is the distributed capacitance between the transmission line conductors.



Figure 5.1: CPW transmission line configuration and TEM mode E-field distributions.

All four components are frequency-dependent. The equivalent circuit model is depicted in Fig. 5.2. The equivalent circuit can be divided into two major parts: the impedance $Z(\omega)$ and the admittance $Y(\omega)$, so that the characteristic device impedance Z_D and the propagation constant γ can be calculated as follows

$$Z_D = \sqrt{\frac{Z(\omega)}{Y(\omega)}}$$
(5.1)

$$\gamma = \sqrt{Z(\omega)Y(\omega)} \tag{5.2}$$

where $Z(\omega)$ and $Y(\omega)$ are calculated from the series lumped elements and the shunt elements of the transmission line equivalent, respectively. The series and the shunt parameters are combined into series impedance $Z(\omega) = R(\omega) + j\omega L(\omega)$ and admittance $Y(\omega) = G(\omega) + j\omega C(\omega)$, respectively, so that the transmission line (device) characteristic impedance Z_D and propagation constant γ can be calculated from

$$Z_D = \sqrt{\frac{R(\omega) + j\omega L(\omega)}{G(\omega) + j\omega C(\omega)}}$$
(5.4)

$$\gamma = \alpha + j\beta = \sqrt{\left[R(\omega) + j\omega L(\omega)\right] \left[G(\omega) + j\omega C(\omega)\right]}$$
(5.5)

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The traveling-wave electrodes influence the frequency response of silicon optical modulators, mainly due to three aspects: impedance mismatch caused by wave reflections effects on the transmission line introduced by the difference between the device and load impedances, propagation speed difference between the electrical and optical waves, and electrical signal attenuation.



Figure 5.2: Transmission line lumped parameter equivalent circuit

5.1.2 Scattering Parameters

In electrical transmission line systems, the propagation of RF energy through a multi-port network is described as a black box by means of a scattering matrix that contains the scattering parameters (S-parameters) of the system [67]. When the RF signal is incident on one port, a fraction of that signal gets reflected to that port, another fraction is scattered and transmitted to other ports, and the rest is attenuated. A schematic representation of the input and output configuration of the transmitted and reflected signals of a 2-port network is presented in Fig. 5.3. The S-parameters are important in microwave design because they are easier to measure and work with than any other types of parameters

The S-matrix for an *N*-port contains N^2 coefficients (S-parameters), each one representing a possible input-output path. The S-parameters are ratios that describe the incident to the reflected and transmitted waves at the ports. They are defined for a given frequency and system impedance, vary as a function of frequency, have a magnitude and a phase, and can be represented in decibels as they are voltage ratios of the waves [67].



Figure 5.3: Schematic diagram of the two-port model black box used to calculate the S-parameters of the transmission line equivalent of a silicon optical modulator.

The numbering convention is always out/in, and for that reason, S-parameters with the same port numbers are always reflection measurements, while S-parameters with different port numbers are always transmission measurements. The four S-parameters of a two-port network are defined by

$$S_{11} = \frac{b_1}{a_1}$$
, for $a_2 = 0$ (Input reflection coefficient) (5.6)

$$S_{12} = \frac{b_1}{a_2}$$
, for $a_1 = 0$ (Reverse transmission coefficient) (5.7)

$$S_{21} = \frac{b_2}{a_1}$$
, for $a_2 = 0$ (Forward transmission coefficient) (5.8)

$$S_{22} = \frac{b_2}{a_2}$$
, for $a_1 = 0$ (Output reflection coefficient) (5.9)

where S_{11} is the forward reflection (input match), S_{12} is the reverse reflection (output match), S_{21} is the forward transmission (gain/loss) and S_{22} is the reverse transmission (leakage or isolation).

The equivalent transmission line of an optical modulator was simulated using *Agilent EEsof EDA* the electronic design automation software system Advanced Designed System (ADS), and S-parameters S_{11} and S_{21} were extracted from the simulation. The result of S_{11}

Chapter 5. RF Electrodes

was important to calculate the device impedance Z_D and to determine if the signal reflection fell below -10 dB, which is an acceptable value to avoid the reduction of the RF power that could be coupled into the modulator. The S-parameter S_{21} was used to determine signal loss, and change in the refractive index due to propagation of the RF wave. These values were used to simulate the optical bandwidth of the silicon modulator so that we could determine if high-speed operation was possible, as presented in the next section.

5.2 CPW Electrode Design

The CPW electrodes are designed by converting the silicon optical modulator into several lumped equivalent distributed circuits throughout the transmission line length. This equivalent was created using all the device parameters and dimensions to reproduce the transmission line equivalent such that the impedance Z_D and the electrical propagation constant γ can be related to the design variables of the device. By reproducing the transmission line model, the electro-optical frequency response of the silicon optical modulator can be calculated and optimized accordingly. This section focuses on the different variables that influence the frequency response of the CPW electrodes and all the considerations taken for an accurate design of the traveling-wave electrodes that drive the silicon optical modulators.

5.2.1 Equivalent Circuit Model

The equivalent circuit model of the silicon optical modulator was created, as shown in Fig. 5.4.Table 5.1 shows the lumped elements of the complete device's equivalent circuit. All elements are given in units of length.

The main parameters needed to simulate and optimize CPW electrodes are the the width of the signal line Signal Line Width (W) and the gap between the signal line and that of the ground lines Gap Width (G) (Fig. 5.1). By changing the dimensions of W and G, all the element values of the circuit shown in Fig. 5.4 also change: i.e., a wider G increases R_D as the highly doped regions need to be wider to cover the entire area between the signal and ground lines. Also, a change in G changes all the resistances and capacitance from the substrate, making the optimization process an extremely challenging task to accomplish.



Figure 5.4: (a) Schematic view of CPW electrodes fabricated on a SOI substrate with the overlapped lumped elements of a silicon optical modulator's equivalent circuit. (b) Equivalent circuit model of the complete device.

The lumped model of silicon modulators was initially validated in [3] using the R_DC_D of the designed PIPIN diode for the HELIOS project. The comparison between the simulated and experimental results was carried out on 1-mm-long devices whose measured electro-optical bandwidth reached up to 15 GHz. The model was validated using the design automation software ADS. The equivalent circuit model shown in Fig. 5.4 was carefully reduced to three main parts: a sole element representing the *RLCG* circuit of the CPW electrodes, as shown in Fig. 5.5(a), an *RC* node (R_D and C_D) to represent the influence of the active region, and an *RC* node (R_{Si} and C_{Si}) to represent the influence of the substrate. The simplified circuit is depicted in Fig. 5.5(b).

Circuit Element	Comment
R	Access resistance from the signal line conductor
L	Inductance from the signal line conductor
C_x	Lateral capacitance between signal and ground lines
G_{x}	Power dissipation admittance between signal and ground lines
C_{SSi}	Capacitance between the signal line and the substrate layer
C_{Si}	Substrate capacitance
G_{Si}	Substrate admittance
R_D	Device resistance
C_D	Device capacitance

Table 5.1: Lumped elements of the equivalent circuit of a silicon modulator (per unit length)

The ADS simplified model was deemed successful, as it provided accurate results in comparison with experimental measurements as presented in [3]. However, the model presented in [3] was only tested on devices of length L = 1 mm, and for high-speeds up to 15 GHz. When the ADS model was used to simulate longer devices operating at higher speeds larger than 20 GHz, the model did not provide such accurate results because it was later found that changes in certain key parameters were not considered, such as the relative permittivity of silicon (ε_s), whose value should be evaluated from the interaction between the substrate's Si/SiO₂ layers, as will be later seen in this chapter. The model was thus carefully optimized to provide more accurate results for devices of any length. Additionally, the model was used to calculate optical bandwidth within ADS, using the S-parameters simulated from the device's transmission line. The optimized ADS model as well as the calculated electro-optical response were validated with the experimental results obtained from the silicon optical modulators measured at a high-speed operation of 40 Gbit/s, and it was later used to design the CPW electrodes of the optical modulator that needed to be integrated on the silicon transceiver designed for applications at $\lambda = 1.27 \,\mu m$. The simulation method and the variables considered for the optimization are presented in the next sections.



Figure 5.5: Schematic diagram of the simplified lumped equivalent circuit a silicon optical modulator fabricated on an SOI substrate.

5.2.2 Electro-Optical Response Simulation

An important step in assessing modulator operation speed is to measure its electro- optical response (small signal frequency response). This measurement is difficult to simulate, as it requires the interaction of both the optical and electrical signals in the device as a function of the frequency. However, the ability to simulate and visualize the modulator's electro-optical response is useful when selecting the dimensions of the CPW electrodes, to guarantee the most optimal response. It is then appropriate to integrate the evaluation of the modulator's electro-optical response into the simulation model of the traveling-wave electrodes with the help of the results obtained from the S-parameters' ADS simulation.

The electro-optical response of a MZM modulator can be calculated from the transmission spectra shift of the signal caused by the effective refractive index Δn_{eff} . It will be shown in Chapter 6 that the high-speed response of the signal is evaluated in a linear region of the spectra. Consequently, the phase variation is assumed to be proportional to the DC transmission spectra difference, as follows:

$$\Delta I = K \cdot \Delta \phi \tag{5.10}$$

where *K* is a proportionality constant. The optical response of the modulator at high-speed operation (dynamic response) is given by

$$\frac{\Delta I(f)}{\Delta I_{DC}} = \frac{\Delta \phi(f)}{\Delta \phi_{DC}} \tag{5.11}$$

where $\Delta I(f)$, and $\Delta \phi(f)$ are the optical transmission and phase variation as a function of the frequency, respectively. For simplification purposes, it is assumed that the refractive index variation Δn_{eff} is proportional to the voltage, such that

$$\Delta n_{\rm eff} = K \cdot V. \tag{5.12}$$

When the frequency increases, the electrical signal attenuates as it travels along the electrodes, the influence of the propagation speed difference between the electrical and optical signals increases, and the amplitude of the modulated voltage ΔV decreases. Due to the interaction of these effects with increasing frequency, it is important to divide the phase shifter (*L*) into *L_i* sections to track down and measure the voltage of each *L_i* section while the signal is propagating along the electrodes (see Fig. 5.6).

Looking at the equivalent circuit schematic of the CPW electrodes shown in Fig. 5.5, the voltage applied in the active region can be measured across C_D . If the length of the modulator L is divided into i number of sections, then, the optical transmission becomes



Figure 5.6: Device length divided into L_i sections to track down the voltage during propagation of the electrical signal.

$$\frac{\Delta I(f)}{\Delta I_{DC}} = \frac{\sum_{i=1}^{N} \Delta \phi_i(f)}{\sum_{i=1}^{N} \Delta \phi_{iDC}}$$
(5.13)

for

$$\Delta\phi_i(f) = \frac{2\pi}{\lambda} \Delta n_{\rm eff\,i}(f) L_i \tag{5.14}$$

Replacing Eq. 5.12 (as a function of the frequency) in Eq. 5.14, we obtain

$$\Delta\phi_i(f) = \frac{2\pi}{\lambda} K \Delta V_i(f) L_i \tag{5.15}$$

where $\Delta V_i(f)$ is the voltage across C_D as a function of the frequency. Considering that the optical and electrical waves propagate in the device at different speeds, the speed of the optical

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wave can be determined from c/n_{OPT} , while the RF wave propagates at a speed c/n_{RF} . The variables n_{OPT} and n_{RF} are refractive indexes of the optical and electrical waves, respectively. The RF refractive index n_{RF} changes with frequency and can be calculated using the phase (ϕ_{S21}) of the RF signal's S-parameter S_{21} (forward transmission) using the following equation

$$n_{RF}(f) = \phi_{S21}(f) \cdot \frac{c}{2\pi f L}.$$
(5.16)

where *L* is the length of the device. A wave that leaves from z = 0 at t = 0 can be found at a distance *z* at a time *t* given by

$$t = \frac{z}{c/n_{OPT}} \tag{5.17}$$

and the voltage can be found from the sinusoidal wave equation

$$V(z,t) = V_0(f) \cdot \cos\left[2\pi f\left(t - \frac{z}{c/n_{RF}}\right)\right]$$
(5.18)

where $V_0(f)$ is the voltage amplitude with respect to the frequency, and the cosine term is the propagation influence of the optical and electrical waves at a distance *z* and time *t*. Assuming the optical and RF waves are in phase at the entrance of the phase shifter, the voltage seen by the optical wave front at a distance *z* is

$$V(z) = V_0(f) \cdot \cos\left[2\pi f\left(\frac{z}{c/n_{OPT}} - \frac{z}{c/n_{RF}}\right)\right]$$
(5.19)

Finally, by replacing the voltage found in Eq. 5.19 in Eq. 5.15, the electro-optical response of the optical modulator can be determined by

$$H(f) = \frac{\Delta I(f)}{\Delta I_{DC}} = \frac{\sum_{i=1}^{N} \frac{2\pi}{\lambda} K L_i \cdot V_0(f) \cos\left[2\pi f \left(\frac{i \cdot L_i}{c/n_{OPT}} - \frac{i \cdot L_i}{c/n_{RF}}\right)\right]}{\sum_{i=1}^{N} \frac{2\pi}{\lambda} K L_i \cdot V_{DC}}$$
(5.20)

The above procedure was followed to simulate in ADS the electro-optical response of silicon optical modulators using the data obtained from the S-parameters' measurements and simulations. The results obtained in ADS were then compared to experimental results to validate the model and our assumptions. All the details about the ADS simulation are presented in the next section, followed by the comparison of the experimental and simulated results.

5.3 ADS Simulation

The main parameters to consider when designing CPW electrodes are the width of the signal line (W) and the gap between the signal line and one of the ground lines (G). Previous theoretical studies developed in [3] for CPW electrodes designed for PIPIN-based optical modulators provided two main W-G dimension pairs. The theoretical analysis determined that a W = 5 μ m, and $G = 25 \mu$ m provides a total device-characteristic impedance Z_D closer to 50 Ω (which is the input impedance seen by other devices connected to the network), at the expense of increasing the transmission loss. Using these criteria, one dimension pair was selected to yield a matched device-characteristic impedance ($Z_D = 50 \Omega$), while the other was selected to deliver the least transmission loss by increasing W. These were the dimension pairs chosen for the silicon optical modulators designed for HELIOS.

The value of G was equally important to maintain these design considerations as this dimension determines the transmission line electrical bandwidth, as will be seen in the experimental results presented in the next section. Table 5.2 lists the dimension pairs chosen for the analysis. The experimental results of the silicon optical modulator based on PIPIN diodes, and designed for applications at 1.55 μ m, presented in the next section, were performed on CPW electrodes with both of these dimension pairs.

Considering the multiple parameters needed to validate the ADS simulation model (such as substrate and device capacitances and resistances, material parameters such as

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permittivity, and metal conductivity, CPW electrode dimensions, and metal dimensions), the study was divided into two parts. It was first important to evaluate the influence of the traveling-wave electrodes on the substrate alone, which means that the CPW electrodes were fabricated directly on an SOI substrate (without the modulator's active region).

Label	W (μ m)	G (µm)	$Z_D\left(\Omega ight)$
G25W05	5	25	50
G085W30	30	8.5	< 50

Table 5.2: RF electrode dimension pairs W and G used for the MZMs designed for applications as $\lambda = 1.55 \ \mu m$.

The characterization results of this batch were used to optimized the ADS simulation model to our substrate and electrodes design. The model was later used to study the influence of the complete device with the characterization results of the silicon optical modulators that delivered a high-speed operation of 40 Gbit/s, which was wafer P04 from batch U532P. *Agilent*'s ADS was used as the simulation tool to verify the model by measuring the S-parameters of the equivalent transmission line of our devices. Figure 5.7 shows the main circuit used for the simulations. The circuit comprised an AC source at the entrance, an input and output 50 Ω -termination, and a number of consecutive boxes. Each box represents a section L_i of the total length of the CPW transmission line, where the CPW element was found in ADS's element library, and it includes the *RCLG* influence from the metal lines on the substrate.

From previous studies presented in [3], it was established that the simulations made with the ADS CPW element are equivalent to those made with the equivalent *RCGL* lumped circuit of a CPW transmission line. For that reason, all the simulations presented here used the CPW element from ADS. The total length of the electrodes determines the number of L_i sections that must be added to the circuit, considering that this length needs to be small enough so the wavelength of the electric signal ignores it. In our design, the maximum allowed L_i length used was below 100 μ m. The zoomed-in view inside one of the boxes of the ADS simulation circuit shows an equivalent circuit of a CPW component and two parallel RC circuits. These RC circuits correspond to the device and substrate influence of the overall signal propagation as discussed in section 5.2.1.



Figure 5.7: One section of the main circuit and elements used to simulate the CPW electrodes.

It had been previously demonstrated in [3] that the substrate's RC has an influence on the transmission bandwidth. To minimize such influence, the resistance R_{Si} is reduced by increasing the substrate's resistivity, while the capacitance C_{Si} is reduced by increasing the thickness of the BOX layer of the SOI substrate (modified substrate). The transmission S-parameter (S_{21}) of a "standard" (low silicon resistivity and BOX thickness of 1 μ m) and a "modified" SOI substrate are depicted in Fig. 5.8. The green curve corresponds to the S_{21} of a standard substrate. The curve shows two important decays of the transmission S_{21} . The first visible decay occurred early on at approximately 3 MHz; this decay corresponds to the RC from the substrate, while the second corresponds to the device's RC.



Figure 5.8: Plot showing the substrate's influence of the transmission S-parameter S_{21} .

The use of a modified substrate shows improved results, as seen in the red curve, where the influence of the substrate's RC is removed almost completely from the transmission measurement. For that reason, to obtain the largest possible transmission bandwidth, the CPW electrodes were fabricated over modified SOI substrates with resistivity of over 1000 Ω ·cm and a silicon dioxide layer thickness of 2 μ m. The device capacitance used for the simulations of the optical modulators based on PIPIN diodes was $C_D = 0.28$ fF/ μ m at -2.5 V, while the device resistance R_D was given by [3].

$$R_D = \left(9.7 + \frac{6G}{23}\right) K\Omega \cdot \mu m \tag{5.21}$$

It can be seen in Eq. 5.21 that R_D depends on the dimension chosen for G, and is therefore responsible for the electrical bandwidth of the modulator. Additionally, considering that the device resistance depends on the electrode design, and because the resistivity of the p dopants is larger than the n dopants, the phase shifter was designed to be located closer to the signal electrode in order to reduce the overall device resistance, as illustrated in 5.9, where the p dopants are represented in blue, and the n dopants in red.

The values of the substrate's resistance and capacitance were determined with simu-

lations performed in ISE simulation software. These values are listed in Table 5.3 for each electrode dimension pair. Complete tables listing all the R_{Si} and C_{Si} values as a function of G and W are presented in Appendix B.

Label	W (μ m)	G (µm)	R_{Si} (M $\Omega \cdot \mu$ m)	C_{Si} (fF/ μ m)	$R_D (K\Omega \cdot \mu m)$	C_D (fF/ μ m)
G25W05	5	25	39.5	0.085	16.22	0.28
G085W30	30	8.5	18	0.355	11.92	0.28

Table 5.3: *R*_S*i*, *C*_S*i*, *R*_D, and *C*_D values used for the ADS simulations of HELIOS' devices.

The units of R_{Si} and C_{Si} are listed in units of μ m so that the total device RC can be easily calculated for any electrode length. All other substrate dimensions relevant for the simulations are shown in Figure 5.9 and the respective dimensions are listed in Table 5.4. The device-characteristic impedance Z_D of a transmission line can be determined from the ABCD parameters matrix (voltages and currents in a general circuit) as follows:

$$Z_D = \sqrt{\frac{B}{C}} \tag{5.22}$$

where *B* and *C* are the m_{12} and m_{21} elements of the 2 × 2 matrix, respectively. The calculation of Z_D is performed in ADS using a built-in function that converts the S-parameters matrix into a chain ABCD matrix, as seen in Appendix C.1.

Furthermore, Table 5.4 shows that the dimension of the metal thickness used for the electrodes with and without the active region was different.

During the course of my research work, increasing the metal thickness of the electrodes to decrease the losses and increase the transmission line bandwidth was a regular topic of discussion. Nevertheless, increasing the height of the metal electrodes is a challenging fabrication process, and therefore a separate test was conducted to study the metal electrodes' influence. The experimental results of the influence of the metal thickness on the transmission line loss and bandwidth are presented in the next section, followed by a comparison of simulated and



Figure 5.9: Schematic cross-sectional view of a PIPIN diode embedded in an SOI substrate showing all the dimensions needed for ADS simulation.

Variable	Layer	Dimension
W	Signal line width	5 μm, 30 μm
G	Gap width between signal and ground lines	$25\mu\mathrm{m}$, $8.5\mu\mathrm{m}$
Wgnd	Ground line width	$200\mu{ m m}$
Т	Metal height	$1\mu\mathrm{m}$, 2.5 $\mu\mathrm{m}^{*}$
Н	Silicon layer height	$0.4~\mu{ m m}$
H _{box}	Silica (SiO ₂) layer height	$2 \mu \mathrm{m}$
H_{Si}	Silicon substrate height	725 μ m

* $T = 1 \ \mu m$ for CPW without modulators, while $T = 2.5 \ \mu m$ for the complete devices from wafers U532P

Table 5.4: Device physical parameters to calculate the equivalent circuit components.

experimental results to validate the ADS model created to design CPW electrodes with and without the influence of the silicon optical modulators.

5.4 Results

5.4.1 Influence of Metal Thickness

The influence of the metal electrodes' thickness was experimentally demonstrated with batch Q301B, where the CPW electrodes were fabricated with different heights on different wafers. A schematic diagram of the electrodes' metal stack and dimensions is shown in Fig. 5.10.



Figure 5.10: Metal layer and dimensions used to fabricate CPW electrodes.

The figure shows metal stacks TiN/TI/AlCu/TiN/Ti, where the AlCu sections vary in height. In wafer P10 the AlCu metal layer of the electrodes was fabricated with a height of 650 nm, while in P12, the AlCu metal layer was fabricated with a height of 1300 nm. The S-parameters were calculated on different devices for both wafers, and the results are plotted in Fig. 5.11. The figure shows the forward transmission S-parameter S_{21} as a function of the frequency, which measures the electrical transmission of the device's transmission line.



Figure 5.11: Comparison of measured S_{21} of CPW electrodes fabricated with different metal heights for MZMs of different lengths.

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The measurements were taken from MZMs of lengths 1 mm, 2 mm, 4 mm, and 6 mm of wafers P10 and P12. The results obtained are consistent for all modulators' lengths, where it is clear that the largest electrical bandwidth is obtained on the devices measured from wafer P12, in which the metal height was the largest.

The results were expected, because an increase of the metal cross-section area leads to a decrease of the access resistance. Nevertheless, it was important to visualize if the height influence was important when measuring the electrical bandwidth. These results show that the influence is indeed important. Looking at the shorter device (L = 1 μ m), the bandwidth decrease from approximately 20 GHz for the highest electrodes, in comparison to 5 GHz for the modulator with the thinnest electrodes. Based on these results, it was important to fabricate electrodes with a rather high metal stack, to avoid the contribution of bandwidth losses due to the metal electrodes. It will be seen in the following section that the height of the metal layer using the CPW electrodes fabricated directly on the SOI substrate (without the active region) was *T* = 1 μ m, as these wafers were only fabricated to validate the simulation model, whereas the electrodes' height of complete devices was *T* = 2.5 μ m, as these devices were meant to function at high operation speeds.

5.4.2 CPW Electrodes Without the Active Region

CPW electrodes with various G-W dimension pairs were fabricated at *CEA-LETI* to analyze the influence of the RF electrodes fabricated on a SOI substrate with the same characteristics as the substrate used to fabricate *IEF*'s silicon optical modulators. The important parameters to study included the electrical response and the transmission line characteristic impedance Z_D , as well as the RF refractive index (n_{RF}). The electrodes were fabricated with lengths L = 1 mm and L = 4 mm to compare the response of the same electrode dimensions for different lengths. These lengths were chosen to analyze the difference of the signal propagation in short and long devices.

The two G-W dimension pairs listed in Table 5.2, G085W30 and G25W05, were of particular interest for the validation of the ADS model, as these were the dimensions of the working modulators. The CPW electrodes of a 4-mm-long MZM were designed with dimensions G = 8.5 μ m, and $W = 30 \mu$ m, to limit additional losses caused by the attenuation of the RF signal during propagation on a long device. The CPW electrodes of a 0.95-mm-long MZM were

designed with dimensions $G = 25 \ \mu m$, and $W = 5 \ \mu m$ to provide an equivalent characteristic impedance of the device ($Z_D = 50 \ \Omega$).

S-parameters' comparison between the simulated and experimental results can be seen in Fig. 5.12 for CPW electrodes L = 1 mm (G25W05), and in Fig. 5.13 for electrodes L = 4 mm (G085W30). The measurements were taken over a frequency range between 10 MHz and 50 GHz. The forward transmission (S_{21}) is shown in Fig. 5.12(a) and Fig. 5.13(a), and forward reflection (S_{11}) is shown in Fig. 5.12(b) and Fig. 5.13(b). The results from Figs. 5.12 and 5.13 show substantial agreement between simulated and measured results, illustrating the model's capability for simulating of the equivalent transmission line circuit of our silicon optical modulators.

The challenge of finding a fitting model in ADS was that the conductivity of the electrodes (σ_m) and the relative dielectric constant (ε_r) were unknown parameters. It was seen in the previous section that the electrodes are made of a stack of different metals, which makes determining a specific value of σ_m difficult. This was also the case for ε_r , because the SOI substrate influences the CPW electrodes' response; thus, ε_r results in values between 3.9 and 11.9, the relative permittivities of SiO₂ and silicon, respectively. It is worth mentioning that finding conductivity and relative permittivity of the CPW is a rather complex topic. However, for our purposes, these values were deduced from trial and error by fitting the ADS model to the experimental results obtained from the CPW electrodes fabricated on a SOI substrate. This method provided a successful way to visualize the influence of each parameter on the response of the equivalent transmission line of the electrodes, namely, the S-parameters S₂₁ and S₁₁, n_{*RF*}, and *Z*_{*D*}, where n_{*RF*} and *Z*_{*D*} were calculated from Eq. 5.16 and Eq. 5.22, respectively.

The metal conductivity (σ_m) and relative dielectric constant (ε_r) were modified in the ADS simulations to fit the experimental results. From this procedure, it was found that σ_m set the loss level of the curve, as well as the slope of the decaying transmission (though to a lesser degree). The conductivity value that fit the experimental results was found to be 3.5×10^7 /m. Additionally, the ADS simulations showed that ε_r had an influence on the first transmission dip, as is easier to seen in Fig. 5.12(a), and on the reflection (S₁₁) of the transmission line. A larger ε_r less pronounced transmission dip, and less signal reflection, while a lower ε_r produced opposite results. By fitting the simulated response to the experimental curve, ε_r was found to be 7. These set of values also delivered an acceptable simulation model that fit for the

longer device, as seen in Fig. 5.13. In this case, the simulated and measured curves of the S_{21} coefficient matched almost perfectly, while the simulated and measured S_{11} curves produced the least (though still significant) correspondence of the all the CPW electrodes measurements without the influence of the active region.

In addition to the validation of the ADS model, the characterization of the CPW electrodes with different G-W dimension pairs provided results of great importance to visualize and understand the relationship among the three dimensions varied in these devices: length, G, and W, as depicted in Figs. 5.14 and 5.15. Figure 5.14(a) shows the forward transmission (S_{21}) of nine 1-mm-long CPW electrodes with different G, and W dimensions, and Fig. 5.14(b) shows the forward reflection (S_{11}) of the same electrodes. Figure 5.14(a) shows that the width of signal line W influences the losses level of the transmission line. The figure shows that the narrowest W (red curve) delivered the largest losses while the widest W (blue curve) delivered the least. Looking at the curves with the same W, it can be seen that the gap between the signal line and the ground electrode G has an influence on the electrical bandwidth, as the largest G provided the fastest attenuation of the signal limiting the electrical bandwidth. Additionally, Fig. 5.14(b) shows that the largest reflections were measured on the electrodes that delivered the largest losses.

However, the losses are not linked solely to W, but also to the mismatching of the transmission line as depicted in the impedance curves shown in Fig. 5.15(a), where a combination of the widest W with the narrowest G produced a matched device impedance $Z_D = 50 \Omega$. A 50 Ω -impedance has a relevant influence on the transmission loss and bandwidth, as seen in the aforementioned figures. Finally, n_{RF} delivered a value that oscillated over a 0.2 of the n_{RF} value, but it does not have a marked change caused by the dimensions of G or W. Consistent results of S₂₁ and S₁₁ were also measured on the longer CPW electrodes (L = 4 mm). These results are shown in Appendix C.



Figure 5.12: Simulated and experimental S-parameters (a) S_{21} and (b) S_{11} of a CPW electrodes of length L = 1 mm (electrodes alone).



Figure 5.13: Simulated and experimental S-parameters (a) S_{21} and (b) S_{11} of a CPW electrodes of length L = 4 mm (electrodes alone).



Figure 5.14: Experimental S-parameters, (a) S_{21} and (b) S_{11} of a CPW electrodes of length L = 1 mm (electrodes alone).



Figure 5.15: Experimental (a) Z_D and (b) n_{RF} of a CPW electrodes of length L = 1 mm (electrodes alone).

5.4.3 CPW Electrodes With the Active Region

The S-parameters from batch U532P of the silicon optical modulators that delivered highspeed operation at 40 Gbit/s were measured to verify the signal propagation on the CPW electrodes, to measure the electrical bandwidth, and to calculate the experimental transmission line Z_D and n_{RF} . A comparison of the S-parameter measurements and ADS simulations are presented in Figs. 5.16 and 5.18 for devices of lengths L = 0.95 mm and L = 4.7 mm, respectively. Figure 5.16 shows a comparison of the experimental results and ADS simulation model of the forward transmission and reflection of modulator MZ11 from batch U532P and wafer P04. The optical modulator MZ11 had a length of L = 0.95 mm and CPW electrodes designed with $G = 25 \ \mu m$ and $W = 5 \ \mu m$. The transmission curve presented in Fig. 5.16(a) shows a reasonable agreement between the simulated and experimental results. However, the measured curve shows a dip in the transmission at approximately 15 GHz which could be caused by a difference in the modal propagation because the two grounds are physically independent. This behavior is related to the fact that the even and odd modes have a propagation-delay difference due to the non-homogenous nature of the CPW electrodes (unconnected ground electrodes); the signals propagate down the CPW lines at different speeds, creating a phase difference that will reach odd multiples of 180 degrees until they cancel each other out. A cancellation of the nodal voltages is observed (dip of S_2 1) at these frequencies and locations, while the reflected waves create the opposite effect and their phases add up constructively. This effect is seen in Fig. 5.16(b), where a sharp peak of the experimental reflection curve is located at the same frequency as the dip measured in the transmission response. This effect is not seen in the simulated curves in Fig. 5.16, as only homogeneous CPW electrodes were considered in the simulations. Despite the dip, it can be seen that the transmission is large enough to allow for high-speed operation, as will be discussed in Chapter 6. Furthermore, the reflection curve corroborates the possibility of high-speed operation, as it shows a maximum reflection wave below -10 dB.

Using the S-parameters obtained, Z_D and n_{RF} were calculated in ADS using the formulas shown in Appendix C.1. A comparison of the experimental and simulated results of the same CPW electrodes' dimensions with and without the influence of the active region is presented in Fig. 5.17. It is seen in Fig. 5.17(a) that Z_D decreased from approximately 100 Ω to a value close to 55 Ω at 25 GHz, when the active region was added to the transmission line equivalent.



Figure 5.16: Simulated and experimental S-parameters, L = 0.95 mm.



Figure 5.17: Experimental (a) Z_D and (b) n_{RF} , L = 0.95 mm.

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The electrodes of the shorter device were designed to provide a matched transmission line impedance. The results show that the G-W dimension pair G25 W05 designed for the CPW electrodes of this MZM delivered a matched impedance. The variation of n_{RF} (Fig. 5.17[a]) showed the opposite effect of Z_D , as its value increased when the active region was added to the transmission line equivalent. A comparison of the measured and simulated transmission and reflection S-parameter coefficients of modulator MZ17 from batch U532P and wafer P04 can be seen in Fig. 5.18. The modulator MZ17 had a length of L = 4.7 mm and CPW electrodes with dimensions $G = 8.5 \ \mu$ m and $W = 30 \ \mu$ m. These dimensions were designed to provide the least losses of all the G-W dimension pairs.

Figure 5.18 also shows a transmission curve whose electrical bandwidth is large enough to provide a high-speed operation of 40 Gbit/s. At the same time, the S_{11} parameter from Fig. 5.18(b) shows reflections below -10 dB, which corresponds to only 10 % of reflected power. This is the standardized reflection limit value for devices of this kind, to allow for high-speed operation. Contrary to the results obtained for the shorter device, the transmission of MZ17 did not have a significant dip in the response. A possible reason for this is that the ground electrodes of the longer devices were designed to be linked, thus providing a common ground point, as shown in the GDS layout of batch T852B (overall, it has the same layout design as U352P) from Fig. 4.1. Additionally, it can also be seen that the ADS simulation model corresponded to the measured response of the transmission and the reflection curves.

It is important to mention that the transmission shown in Figs. 5.16(a) and 5.18(a) were measured without a DC bias applied to the device, which means that the actual 3 dB at the operation point is higher than the one seen in the figure. The measurement of the S_{21} parameter at the operation point (at a DC voltage of ~ 2.5 V) will be seen in Chapter 6.

The comparison between the measured and experimental Z_D and n_{RF} is shown in Fig. 5.19. The transmission line impedance of the CPW electrodes with and without the influence of the active region is depicted in Fig. 5.19(a). The comparison shows the same correspondence found in the shorter device, where the Z_D decreased when the active region was added to the device. In the case of the CPW fabricated with dimensions G085 W30, the matched 50- Ω impedance seen in the previous section (Fig. 5.17), was decreased to approximately 25 Ω at 25 GHz. The variation of the n_{RF} with frequency is depicted in Fig. 5.17(b). The curves show an increase of ~ 2 with the influence of the active region, which is about the same amount of

increase seen in the n_{RF} of the shorter device. The figure also shows a drop of the measured n_{RF} value at approximately 23 GHz, which could be caused by the dip seen in the S_{21} response at the same frequency. Nevertheless, the simulation curve shows the ideal response of this parameter (thin red curve).

In general, the measured parameters S_{21} , S_{11} delivered adequate results to achieve high-speed operation. Additionally, the calculated Z_D and n_{RF} obtained from the measured S-parameters matched the designed values presented in [3]. These results were obtained on different devices of the same length found in chips from wafer P04 of batch U532P. By obtaining the designed parameters experimentally, we were confident in concluding that these silicon optical modulators were expected to reach high-speed operation, and that no additional RF losses from those considered in the design process were expected from the CPW electrodes. Additionally the ε_r was changed to 10 to fit the experimental results regarding the influence of the active region. The increase of the permittivity is explained by the fact that the electric field propagates freely among all materials: air, silicon and silica (from the SOI substrate). However, when the active region is added, the electric field localizes in the silicon overlayer (more specifically, in the diode's space charge region), which means the electric field will not enter the silica layer as freely as before, making ε_r change to a value closer to the pure silicon ε_r .


Figure 5.18: Simulated and experimental S-parameters, L = 4.7 mm.



Figure 5.19: Simulated and experimental (a) Z_D and (b) n_{RF} , L = 4.7 mm (complete device).

5.4.4 Electro-Optical Response Calculation in ADS

After finding a fairly precise ADS model of the silicon modulator's electrical response, it was possible to determine the expected electro-optical response. This process was carried out by adapting the equations developed in section 5.2.2 to calculate the electro-optical response of the modulator using S-parameters. The simulations of the electro-optical response were carried out using an optical refractive index $n_{OPT} = 3.1$. The ADS simulation code can be found in Appendix C.1. The comparison between the simulated and measured results are shown in Fig. 5.20, for both of the lengths used in the previous section. The figure shows that the electro-optical response correctly fits the experimental results for both devices. The electro-optical response of the shorter device, shown in Fig. 5.20(a), shows an excellent consistency between the measured and the simulated results, except for a dip in the experimental results which also occurred at the same frequency as the dip seen for this device at in the electrical forward transmission curve S_{21} (Fig. 5.16). The electro-optical response of the longer device, presented in Fig. 5.20(b), also shows a good fit of the simulated results in comparison with the measured results.

Comments on the measured electro-optical bandwidth of both devices are found in Chapter 6; however, the comparison between the simulated and experimental results are also presented in this chapter to confirm the possibility of obtaining a simulation model that could be used to foresee the electro-optical response of the silicon optical modulators. It is important to mention that, although the ADS model delivered successful results, careful consideration must be made of the multiple unknown parameters up to this point whose values used are just approximations; consequently, the electro-optical model is highly sensitive to any changes in the transmission line.

Contour representation plots of the electrical bandwidth and the electro-optical response of a 1-mm-long silicon optical modulator based on PIPIN diodes are shown in Fig. D.1, and the calculated Z_D and n_{RF} are shown in Fig. D.2 (see Appendix D). These contour graphs were plotted using the ADS simulation model created for the design of the CPW electrodes of the HELIOS silicon optical modulator by simulating all four parameters as a function of all the possible combinations of G and W between 5 μ m and 40 μ m, for every 5 μ m increment. TThe contour plots provide a clear visualization of the influence of the CPW electrode dimensions simultaneously on all the main design parameters. With the help of the ADS model and the contour plots, the CPW electrodes' dimensions of the silicon optical modulator based on PN diodes, designed to be integrated with an optical transceiver for local network applications, were selected. The simulated results are presented in the next section.



Figure 5.20: Simulated and experimental electro-optical response (a) L = 0.95 mm and (b) L = 4.7 mm.

5.4.5 CPW Electrodes Design of a Modulator Based on a PN Diode

The ADS simulation model depicted in Fig. 5.7 was used to design the CPW electrodes of the optical modulator to be integrated in a silicon transceiver used for applications at $\lambda = 1.3$ μ m. A comparative study of all possible G-W dimension combinations from 5 μ m to 40 μ m was initially carried out on devices of lengths L = 1 mm and L = 2. A set of contour figures is presented in Figs. 5.21 to 5.23 to visualize the variation of the four main parameters needed for the design of the CPW electrodes: electrical bandwidth, electro-optical response, device-characteristic impedance Z_D and RF refractive index variation n_{RF} . These four parameters are presented as a function of the CPW electrodes' dimensions G and W. The curves shown in Figs. 5.21, and 5.22 are the results of ADS simulations for a device of length L = 1 mm, whose active region is based on the PN diode designed in Chapter 3. The device resistance R_D was calculated from Eq. 3.3 (in $K\Omega \cdot \mu$ m), and the device capacitance used was the calculated $C_D = 0.39$ fF/ μ m at -2.5 V. The R_D and C_D were calculated using the optimized doping concentrations found in Chapter 3 $p = 5 \times 10^{17}$ cm⁻³ and $n = 7 \times 10^{17}$ cm⁻³.

Figure 5.21(a) illustrates the variation of the 3 dB electrical bandwidth of the simulated forward transmission S_{21} as a function of G and W. The figure shows that the largest bandwidth was measured for G values between 10 μ m and 20 μ m and for W values between 5 μ m and 15 μ m. As the G and W values increase, the electrical bandwidth decreases. Nevertheless, the electrical bandwidth still provides large values if G and W are designed with the same dimensions, or within 5 μ m of the difference between the two. By doing so, the electrical bandwidth decreases by approximately 1.5 GHz (for the largest simulated G=W=40 μ m) from the maximum value of 18 GHz. Similar results to those of the electrical bandwidth were obtained for the simulated electro-optical response of the complete device as a function of G and W, as shown in Fig. 5.21(b). The electro-optical response was measured at the -3 dB level from the maximum transmission measured at the lowest frequency. This figure shows that a silicon optical modulator could in theory reach up to approximately 27 GHz, which is a promising electro-optical bandwidth for the high-speed operation required by the SILVER project. Furthermore, Fig. 5.22(a) shows the contour representation of the calculated n_{RF} at 25 GHz as a function of G and W.



Figure 5.21: Contour representation of the simulated (a) electrical BW and (b) electro-optical response of a modulator based on a PN junctions as a function of G and W of the CPW electrodes, L = 1 mm.

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The results show that the maximum n_{RF} values are obtained for low W dimensions, regardless of the value of G, while it provides the minimum n_{RF} values for low G dimensions regardless of the value of W. Finally, Fig. 5.22(b) shows the characteristic impedance of the device Z_D (in Ω) measured at 25 GHz as a function of G and W.

The figure illustrates that a larger resistance is obtained for the smallest W and the largest G, while the lowest resistance was obtained for the largest W and the lowest G. Comparing these results to those obtained in Fig. 5.21, it is easy to visualize the challenge of choosing the CPW electrodes' dimensions because, in the previous case, the maximum bandwidth was obtained when the values of W and G are equal, or almost equal. Overall, the lookup figures are a didactic way to illustrate the influence of G and W on the different parameters, so that a set of dimensions can be selected accordingly, while making compromises among all the design parameters.

The same procedure was followed to illustrate the simulation results of a device of length L = 2 mm. The results of Z_D and n_{RF} did not change significantly; hence, Fig. 5.22 can be also used for this length. However, although the change of length was not significant, the electrical bandwidth and electro-optical response did provide different results, as illustrated in Fig. 5.23. It can be seen that the influence of G and W on both of these parameters starts changing as the device length increases. The change shows that the electrical bandwidth and the electro-optical response are no longer proportional. The reason is that the transmission line begins to be sensitive to signal attenuation due to the mismatch between the electrical and optical signals. The electrical bandwidth is evaluated at the end of the line, while the electro-optical bandwidth depends on the sum of the phase difference during propagation, as discussed in section 5.2.2.

It can be seen in Fig. 5.23(a) that the largest electrical bandwidth is now obtained for larger G and W dimensions than in the case of a 1-mm-long device. Furthermore, the electro-optical response seen in Fig. 5.23(b) shows that the largest response occurs for small values of G but large values of W, which are the opposite dimensions needed to obtain a match impedance, as seen in Fig. 5.22(b). The results obtained from the simulation showed that the G-W dimension pair combination $G = 20 \ \mu m$ and $W = 20 \ \mu m$ provided a fair compromise among all four parameters: electrical bandwidth (3 dB of S_{21}), electro-optical response, characteristic impedance Z_D , and variation of the RF refractive index n_{RF} , for short and long devices.



Figure 5.22: Contour representation of the simulated (a) Z_D and (b) n_{RF} , of a modulator based on a PN junction as a function of G and W of the CPW electrodes, L = 1 mm.



Figure 5.23: Contour representation of the simulated (a) electrical BW and (b) electro-optical response of a modulator based on a PN junction as a function of G and W of the CPW electrodes, L = 2 mm.

Additionally, the S-parameter (S_{11}) and the losses were also sway parameters that needed consideration for the selection of the CPW electrode dimensions G and W; e.g., Fig. 5.21 shows that the largest electrical bandwidth and electro-optical response occur for the smallest W values. The signal reflection was also considered, as it needs to be below -10 dB to guarantee better impedance matching, which leads to high-speed operation. All of the selected simulation parameters delivered results that meet the SILVER project requirements, which means that a bandwidth of at least 10 GHz is expected.

The dimensions of G and W were chosen to simulate the complete device, and to visualize its response for the different lengths. ADS simulation results for modulators of length 1 mm, 2 mm, and 2.5 mm are presented in Fig. 5.24. The forward transmission S_{21} shows electrical bandwidths of 10.7 GHz, 12 GHz, and 17.2 GHz, for 1-mm-, 2-mm-, and 2.5-mm-long devices, respectively, as depicted in Fig. 5.24(a), while the forward reflection S_{11} (Fig. 5.24[b]) from the electrodes delivered less than -10 dB in all three devices. Moreover, the simulated electro-optical response delivered 15.1 GHz, 17.7 GHz and 26.4 GHz for 1-mm-, 2-mm-, and 2.5-mm-long devices, respectively, as seen in Fig. 5.24(c). These results produced bandwidths that are large enough for all lengths, to ensure that the CPW electrodes do not limit the high-speed operation of the silicon optical modulator.

I was able to measure the S-parameters of the completed SILVER devices a few days before completion of this document. The first SILVER batch was label Z824J. Wafer P10 was randomly selected for these measurements. A comparison of the measured and simulated results of a complete device of length L = 1 mm is depicted in Fig. 5.25. The forward transmission S_{21} (Fig. 5.25[a]) shows a 3 dB electrical bandwidth over 30 GHz. The measured results show a larger influence of the RC substrate, which is seen by the wavy decaying of the transmission. The frequency dip of the forward transmission and the peak in the forward reflection (Fig. 5.25[b]) are also seen at the same frequency as the HELIOS modulators of the same length. It is important to mention that the doping concentrations of wafer P10 were $p = 1 \times 10^{18}$ cm⁻³ and $n = 1 \times 10^{18}$ cm⁻³, which explains why $R_D = 7$ K Ω and $C_D = 0.18$ fF, according to the resistivity vs. doping concentration equations shown in Appendix A and the equations of R_D and C_D developed in Chapter 3. The results of Z_D and n_{RF} from the measured and simulated S-parameters are presented in Fig. 5.26. The figure shows that a characteristic impedance Z_D of approximately 34 Ω is expected at 25 GHz.



Figure 5.24: Simulated (a) S_{21} , (b) S_{21} and (c)electro-optical response, of an optical modulator based on a PN junction with $G = 20 \ \mu m$ and $W = 20 \ \mu m$, L = 1 mm.



Figure 5.25: Contour representation of the simulated (a) Z_D and (b) n_{RF} , of a modulator based on a PN junction as a function of G and W of the CPW electrodes, L = 1 mm.



Figure 5.26: Experimental and simulated (a) Z_D and (b) n_{RF} of the SILVER modulator with CPW electrodes designed with $G = 20 \ \mu m$ and $W = 20 \ \mu m$, L = 1 mm.

Conclusion

This chapter presented a study of the influence of the RF electrical signal propagation on the silicon optical modulator characteristics, and all the challenges faced in the design of CPW electrodes. The topic of transmission lines is already a complex topic itself, as the response of the propagation of the electrical signal depends on a combination of variables and effects. The chapter presented a summary of the main theoretical challenges encountered in the design of CPW electrodes for our particular application, and how we approached these challenges using various approximations to obtain a fairly accurate simulation model that was able to predict the main design parameters that influence the overall performance of CPW electrodes. *Agilent's* ADS simulation software allowed us to simulate the transmission line equivalent of our silicon optical modulators by converting the complete device into a distributed lumped equivalent circuit.

The measurement of the simulated S-parameters in ADS turned out to be an extremely useful way of studying and designing the CPW electrodes of our silicon optical modulators. By extracting the simulated S-parameters of the device's equivalent transmission line, we were able to calculate and predict the electrical bandwidth, the device's characteristic impedance Z_D , the RF refractive index variation n_{RF} , the signal reflection S_{11} , the losses, and the electro-optical response. The ADS simulation model was improved throughout my research, and the simulation model proved to be acceptable for the design of CPW electrodes by using the measured S-parameters of the working silicon optical modulators based on PIPIN diodes from wafer P04 in batch U532P. The ADS simulation model was validated by showing matching results between the experimental and simulated results of the silicon optical modulators, not only from the parameters obtained directly from the simulations, such as the forward transmission and reflection (S_{21} and S_{11} , respectively), but also from the parameters that were obtained from equations created within ADS (Z_D , n_{RF} , and the electro-optical response).

The ADS simulation model produced different results for the influence of the most critical dimensions of CPW electrodes: namely, G and W. The results showed that the width of signal line W influences the losses level of the transmission line, where a narrow W delivered the largest losses and the widest W delivered the least. If W is kept constant, then G shows an influence on the electrical bandwidth attenuation, where the largest G delivered higher attenuation which led to a smaller 3 dB electrical bandwidth. It was also seen that the widest

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W and the narrowest G produces a matched device-characteristic impedance. However, equal or closer values of G and W produce the largest electrical bandwidths. Furthermore, it was seen from the contour distributions that the electro-optical response of the silicon optical modulators was obtained for smaller values of W, which is inversely related to the losses.

Using the combination of all these results, it is clear that compromises must be made in designing CPW electrodes. These compromises can be made according to the targeted application of the device. Because the high-speed requirement of the silicon optical modulator designed for applications at 1.27 μ m (SILVER) was not as challenging to achieve as the silicon modulator designed for applications at 1.55 μ m (HELIOS), the dimensions of the CPW electrodes of the SILVER modulator were chosen so that all the parameters deliver acceptable results, none of them being the largest or lowest simulated value obtained. For the SILVER modulator, the selected CPW electrodes have dimensions $G = 20 \ \mu$ m and $W = 20 \ \mu$ m, to deliver an electro-optical bandwidth between 15 GHz and 26.4 GHz for devices of length $L = 2.5 \ mm$ and $L = 1 \ mm$, respectively; $Z_D \approx 32 \ \Omega$ at 25 GHz, and $n_{RF} \approx 4.1$ at 25 GHz. These simulation results allow for high-speed operation of the silicon modulator design for the SILVER project.

6 Characterization

To evaluate the general response and speed operation of silicon optical modulators, a comprehensive set of electrical, optical, and electro-optical measurements must be taken in order to study individual and combined responses. These tests not only help us to asses the maximum response and speed operation of each device, but they are also useful to find possible problems, limitations, and areas of improvement in our designs. Generally, the current-voltage (I-V) response is the first measurement carried out on the modulators to evaluate the behavior of the diodes that make up the active region. From there, additional measurements include the S-parameters measurement, to calculate the electrical bandwidth, electro-optical response, RF refractive index variation, and device characteristic impedance. The S-parameters measurements helps us to asses the electrical behavior of the equivalent transmission line. The modulator's optical transmission is evaluated to verify the modulation ability, and to calculate figures of merit (FoM) such as extinction ratio (ER), insertion loss (IL), effective refractive index variation efficiency (V_{π} ·L_{π}), and most importantly, to visualize and choose a wavelength where the most optimal operating point is located.

Furthermore, the modulator's high-speed operation is determined by way of eye diagrams. These measurements evaluate the actual modulation capability of the device according to sequence length and speed. The eye diagrams are measured by simultaneously applying both the electrical and the optical signals on the modulator, thus, providing additional losses and ER measurements (in comparison to the DC transmission measurement), due to the presence of the RF signal. The robustness of the modulator can also be analyzed with the Bit-Error Rate (BER) measurement to determine the modulator's potential in transferring in-

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formation. Finally, the modulator's electro-optical response is also evaluated to determine the electro-optical bandwidth and signal reflection of the device, due to the interaction between the optical and the electrical signals while they propagate along the traveling-wave electrodes.

The characterization bench is composed of different tools and equipment to implement all measurements mentioned above. This chapter presents the characterization results of the multiples silicon optical modulators that were fabricated in the last three years. The results presented in this chapter came from the silicon optical modulators fabricated for high-speed applications at a wavelength of 1.55 μ m. Two different batches fabricated at *CEA-LETI* namely, T582B and U532P—were mostly used. The first batch of silicon optical modulators designed for applications at 1.27 μ m arrived late in the redaction process of this dissertation; consequently, only the S-parameter characterization results presented in Chapter 5 were able to be included.

The next section describes the different kinds of measurements required to characterize high-speed silicon optical modulators. For each measurement, the experimental setup will be presented, including software, equipment and process. Finally, the purpose and importance of each measurement taken will be explained. This section is presented in order to provide a complete view and understanding of the processes used and the results obtained in all the high-speed silicon optical modulators described in the rest of the chapter.

6.1 Measurement Setups and Methods

The characterization of silicon optical modulators requires the measurement of different parameters, such as the I-V response, the static (or DC) optical transmission, the high-speed performance, and electrical and electro-optical responses.

6.1.1 I-V Response

The first step in the characterization process is to measure the current-voltage (I-V) response of each of the diodes embedded in the active regions of the different modulators. The I-V response is measured ensure that the active region response does in fact correspond to a diode. This step is crucial to rule out modulators with problems such as short circuits, open circuits, or low doping concentration, among others. In fact, a characterization method to determine the doping concentration in the active region by measuring the avalanche voltage was studied in [3]. The measurement of the I-V response requires the least amount of tools of all the measurements. The setup used a *Keithley 2400 Boroad Purpose SourceMeter*, to provide precision voltage and current sourcing and measurement, a set of DC needle probes to access the electrode pads of the modulators, and an in-house *LabVIEW* program to automatically obtain I-V behavior, as shown in Fig. 6.1.



Figure 6.1: Schematic diagram of the setup used for I-V measurements.

Different modulators provide a different I-V response even if the modulators have the same dimensions and are fabricated on the same chip. To illustrate this, two plots are shown in Fig. 6.2. The first plot (Fig. 6.2[a]), shows an example of the I-V measurements obtained for all the diodes of a single chip. It can be seen that the diodes have different breakdown voltages (within a range of 7 V), even if they are the same type of diode and have the same dimensions. Modulators MZ07, MZ09, and MZ11 had a length of 0.95 mm, and were all designed with PIPIN diodes with the same active region dimensions. Even though modulators MZ04 and MZ12 were also designed with PIPIN diodes in the active region, modulator MZ04 was 1.8 mm long, and while MZ12 also had a length of 0.95 mm, this modulator was designed with its heavily doped regions etched down to match the waveguide slab height of the rib waveguide across its entire width.

It was mentioned above that the breakdown voltage describes the doping concentration of the diodes. However, although all the diodes should have had the same doping concentration (because the measurements were taken on the same chip), they must have ended up having slight alignment variations of the p -doped slit location during fabrication.



Figure 6.2: I-V measurements of (a) all the MZMs of chip P01-L14C14 of batch T582B, (b) modulator MZ04 for chips located on different places throughout the wafer.

To make an equivalent comparison, modulator MZ04 was measured in five different chips chosen specifically in different places within the wafer: top, bottom, left, right and center. The results are shown in Fig. 6.2(b). In this case, the results provided breakdown voltages within a 1-V range, demonstrating a reasonable repeatability of the electrical characteristics of the same MZM from chip to chip. A small plot inside Fig. 6.2(b) shows the same I-V measurements when the current is converted to log scale. This method is sometimes preferred, as small current changes are easier to see than in the linear form.

6.1.2 Optical Transmission (Static Characteristics)

The optical transmission, or static characteristic measurement, is the first optical response of the modulator as a function of the wavelength and the reverse bias, as seen earlier in the operation principle sections of the optical structures in Chapter 2 (Figs. 2.8 and 2.12). With the first measurement of the static response it is possible to determine if a modulator works optically, by analyzing the shape of the output transmission spectra. Possible outcomes of the optical transmission measurement include the inability to detect a modulated output due to extremely high loss, noisy transmission due to Fabry-Perot Resonator (FP) resonances on the chip's facets, or the measurement of an optical transmission with random spectra due to multimode propagation of the light in the waveguide. If the initial optical transmission is correct, the refractive index variation can be determined from subsequent measurements of the optical transmission with increasing reverse voltage, as seen in Fig. 2.12.



Figure 6.3: Schematic diagram of the setup used to measure the optical transmission of silicon optical modulators. The setup includes a tunable laser diode, and EDFA, a polarized controller, a focal lens system, and an optical component tester to acquire the optical transmission spectra.

The optical transmission is also essential to assess the first FoM of a modulator, such as ER_{DC} and IL, the modulation efficiency $(V_{\pi} \cdot L_{\pi})$, as well as the free spectral range (FSR), Q factor, finesse and FWHM of modulators based on RRs. The setup to measure the optical transmission of a silicon modulator shown in Fig. 6.3 consists of an *Anritsu Tunics Tunable*

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Laser Diode source for wavelength measurements between 1520 nm and 1640 nm, followed by an EDFA, which is added because the optical laser source is not powerful enough (3 mW) for the light to be detected by the high-speed measurement tools, and a polarization controller to filter out the TM polarization component of the light beam.

Light is butt-coupled to the modulator's Device Under Test (DUT) entrance waveguide by means of a lensed fiber with a diameter of approximately 2 μ m. The coupled light propagates in the device and the light is collected at the modulator's output waveguide using a lens system to focus and detect the modulated light using an *Anritsu MT9820A All-Band Optical Component Tester* to allow acquisition of the transmission spectra in a computer using a General Purpose Interface Bus (GPIB) interface. The optical transmission measurements enable us to find the modulators that deliver the best FoMs so that they can be tested for high-speed operation.

6.1.3 High-Speed Performance (Dynamic Characteristics)

The high-speed performance quality of the optical modulators is determined by measuring eye diagrams. An eye diagram is a graphical representation of the quality of electrical and optical signals from the appearance of an eye [68]. An eye diagram measurement is taken by overlapping all possible one-zero (*hi-lo*) combinations on the oscilloscope of repeated signals samples. The performance of an eye diagram can be obtained from the rise/fall times, data eye height, jitter, and noise amplitude as depicted in Fig. 6.4. The vertical and horizontal eye openings are important characteristics of the eye diagram to quantify the signal quality. The vertical opening is the ER, while the horizontal opening is measured with rise and fall times, which is the time the modulator takes to achieve a *hi* voltage level starting from a *lo* logic level (rise time), or vice versa (fall time).

The eye openness also depends on the sequence length of the Pseudo-Random Binary Sequence (PRBS), which is typically between 2^7 and $2^{31} - 1$ (number of bits sent by the pulse pattern before the PRBS repeats itself). If the optical modulator has a low-frequency cutoff, the eye diagram will tend to close with increasing pattern length. Thus, the sequence length should also be specified along with the results of an eye diagram measurement.

The eye diagram is used to determine the dynamic ER and IL at an optical modulator's operating point. The measurement is taken by choosing a wavelength or operating point



Figure 6.4: Eye diagram.

from the optical transmission where the phase shift (modulation) is large enough to obtain an important ER. Usually, the largest ER is measured near the minimum transmission spectra level where losses are considerably high, thus, difficult to be detected by the measurement equipment and not ideal for our particular application.

The best modulators chosen from the optical transmission measurements are tested for high-speed operation. The schematic diagram of the setup for this measurement is shown in Fig. 6.5. The figure shows the optical connections in blue, and the electrical connections in red. The optical part of the setup consists of the same optical input as the optical transmission setup: a tunable diode laser source, an EDFA, a polarization controller, and a lensed-fiber to butt-couple the light in the modulator's entrance waveguide. The modulated output light is collected using a lens system that sends it to a second EDFA for reamplification of the signal for detection. Because the optical transmission is weak when it enters the second EDFA, the amplified spontaneous emission induces noise to the input signal [3]; for that reason, a selective filter is added at the output of the EDFA, and it must be adjusted according to the wavelength set for the measurements. The filtered light output is connected to a *Agilent* DCA-J 86100C oscilloscope with an 86106B optical module (for high-speed measurements up to 10 Gbit/s, or an 83484A 50 GHz module (for high-speed measurements over 10 Gbit/s).

The electrical part of the setup consists of a *Centellax* TG2P1A 10 Gbps or TGP4A 40 Gbps (for high-speed measurements ranging from 21 Gbit/s to 40 Gbit/s) Non-Return-to-Zero (NRZ)



Figure 6.5: Schematic diagram of the high-speed performance measurement setup. The setup includes a simultaneous combination of optical and electrical connections to the silicon optical modulator under test.

PRBS source. The bit rate clock of the PRBS is connected to the trigger of the oscilloscope to ensure that the display of the signal is centered on the screen, while its output is connected to a *Centellax* OA4MVM3 three-stage broadband driver amplifier, and a *Centellax* TE1B bias board to fine-tune the drain gate and bias supply of the three-stage amplifier module, that delivers approximately $6 - 7 V_{pp}$. The amplified output is connected to a bias tee that adds a reverse

DC voltage to guarantee reverse bias operation of the optical modulators. The electrical signal is connected to one end of the silicon modulator's CPW electrodes using *Cascade Microtech* 40 GHz or 50 GHz RF probes. The output of the transmission line signal (propagation of the RF signal thought the CPW electrodes) is terminated by connecting another RF probe to the other end of the electrodes, along with a bias tee and a $50-\Omega$ load to avoid signal reflection.

6.1.4 S-Parameters (Electrical Performance)

The electrical performance of silicon optical modulators is evaluated by measuring the Sparameters (scattering parameters) to quantify how an RF signal propagates through a twoport network, as described in Chapter 5. A schematic diagram of the setup used to measure the S-parameters is illustrated in Fig. 6.6. The setup consists of a network analyzer to supply an RF signal that is applied to the silicon modulator.



Figure 6.6: Schematic diagram of the electrical performance (S-parameters) setup. A network analyzer is used to supply and recover on the microwave RF signal to analyze the electrical response of the silicon optical modulator.

In MZMs, the signal propagates along the modulator's CPW electrodes and is recovered at the output port of the network analyzer where the microwave transmission and reflection characteristics of the RF signal are analyzed. A DC bias is also connected to evaluate the electrical properties of the device under bias. In RRs modulators, only the reflection coefficients S_{11} and S_{22} can be measured, as this type of modulator does not have RF traveling-wave electrodes.

6.1.5 Electro-Optical Performance (Small-Signal Characteristics)

The small-signal measurement technique is based upon the same characterization concepts of the S-parameters measurement. In this case, the response of the electrical and optical signals applied to a modulator are evaluated as a function of the frequency; e.g., in a MZM, the electrical signal propagates in the traveling-wave electrodes at microwave velocities while, at the same time, the optical guided waves propagate with an optical speed. The modulation is thus created through the frequency-dependent electro-optical interaction of a traveling microwave with a traveling optical guided wave. Figure 6.7 shows a the general representation of an electro-optical modulation measurement, showing input modulating electrical current and output optical modulation power.

As it was mentioned in Chapter 5, the S-parameters are used to describe the transmitted and reflected signal flow within a device or network, described by the transmission coefficients S_{21} and S_{12} and the reflection coefficients S_{11} and S_{22} . In the electro-optical modulation measurements, the individual transfer function is defined in terms of responsivity because it uses signals in both the optical and electrical domains, and optical signals do not lend themselves conveniently to S-parameter definitions [69]. If the input and output impedances are assumed to be Z_0 , then S_{11} and S_{22} are zero, then

$$R_s = \frac{\Delta P_{OUT}}{\Delta I_{IN}} \quad \text{E/O source responsivity}$$
(6.1)

where R_s is the source responsivity is expressed in Watts per Amp. In decibels, R_s is related to the conversion efficiency of 1 W/A, so that



Figure 6.7: EO measurement. The network analyzer measures input modulating current and output modulation power and displays the ratio of the two.

$$R_s(dB) = 10\log_{10}\frac{R_s(dB)}{W/A}$$
(6.2)

$$R_s(dB) = 10\log_{10} \frac{\Delta P_{OUT}}{\Delta I_{IN}}$$
(6.3)

The characterization setup is shown in Fig. 6.8. The setup is similar to the high-speed measurement. Electrical and optical signals are sent to the silicon modulator using an *Agilent* N4375B Lightwave Component Analyzer with an *ECal* calibration module. The optical part of the setup is the same as that used in the high-speed measurements, except that the modulated output is connected to the optical receiver in the lightwave component analyzer.

For the electrical setup, an RF signal is supplied by one of the network analyzer's ports for a frequency sweep up to 50 GHz. The output is connected to a bias tee where DC reverse voltage is also connected to guarantee carrier depletion operation of the devices. The output signal from the bias tee is attached to the modulator's input port of the CPW electrodes using 50-GHz RF probes. The output port of the device is terminated by connecting another RF

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probe capped with a bias tee and a 50- Ω load to avoid signal reflection.



Figure 6.8: Schematic diagram of the high-speed performance measurement setup.

The optical receiver in the lightwave component analyzer is used to measure the transmitted signal; it measures the electro-optical modulation response while the network analyzer yields a measurement of the reflection S-parameter S_{11} . The measurements are evaluated at the modulator's small-signal linear transmission characteristic curve, where the maximum optical response is obtained, as seen in Fig. 6.7.

6.2 Results

The following section presents the results obtained on the high-speed silicon optical modulators designed at *IEF* for applications at 1.55 μ m. It includes results of MZMs based on PIPIN diodes, as well as RRs modulators based on PIPIN diodes and interleaved PN junctions. Although the fabricated SILVER modulators arrived at *IEF* a few weeks before completion of this document, the devices were not optically characterized in time to be able to include the results in this dissertation. Nevertheless, in view of the wavelength application requirement of the SILVER project, several modulators designed for applications at 1.55 μ m were also tested at a wavelength of 1.27 μ m.These characterization results are presented and discussed at the end of this section.

6.2.1 Mach-Zehnder Modulator at 10 Gbit/s Based on a PIPIN Diode

During my research at *IEF*, the first characterized MZMs made part of batch T582B. These modulators were fabricated using the standard-aligned process flow, as described in Chapter 4. All the MZMs characterized from this batch were based on PIPIN diodes embedded in the arms of an asymmetric MZI with voltage applied to one of the arms, and designed $1.55-\mu$ m applications. The best working modulator was found in chip L14C14 of wafer P01, a chip located near the center of the wafer. MZMs of lengths 0.95 mm, 1.8 mm, 4 mm, 4.7 mm, and 6 mm were characterized, but it was modulator MZ04 with a length *L* = 1.8 mm that delivered the best modulation results of all the modulators characterized from this batch. All the results presented in this section were obtained from the measurements taken from modulator MZ04.

A Electrical Measurements

The I-V measurements of the T582B devices delivered reverse currents in the order of a couple of nA; specifically, modulator MZ04 delivered -0.2 nA at -4 V, as depicted in Fig. 6.9. The curve shows proper rectifying behavior and breakdown voltage around -11 V.

Studies to estimate the doping concentration of the least-doped region of one-sided abrupt junctions from the diode's breakdown voltage have been demonstrated previously [42]. At the same time, this estimation can be used to determine the depletion region width at breakdown voltage. The results of this study are shown in Appendix E, where plot Fig. E.1 depicts the relationship between breakdown voltage and doping concentration, and Fig. E.2 depicts the relationship between depletion region width and doping concentration. The plots show the results for one-sided junctions in Si, Ge, and GaAs. These plots were used to estimate the characteristics of MZ04's PIPIN junction.



Figure 6.9: I-V curve of MZ04, *L* = 1.8 mm, from T582B.

Looking at the I-V response, breakdown occurred at approximately -11 V, which means that the doping concentration of the *p*-doped slit found in Fig. E.1 corresponds to approximately 1.5×10^{17} cm⁻³, and that the depletion region width at breakdown in ~ 0.3 μ m. These results showed from the onset that the actual doping concentration in the waveguide was lower than the targeted concentration, at least in the *p*-slit, which is highly responsible for the variation of refractive index in the waveguide. Furthermore, a depletion region width of 0.3 μ m gives evidence of complete junction depletion at this voltage, considering that the *p*-doped slit width was 0.1 μ m. Using the experimental *p* doping concentration, Eq. 2.29 and Eq. 2.30, the experimental capacitance of the PIPIN diode was calculated to be $C_D = 0.20$ fF/ μ m.

The I-V results provided an initial insight to the following experimental measurements, as it was known at this point that the active region was under-doped. This information raised the possibility that the modulator would not work at the highest speed for which it was designed. Nonetheless, the diode's proper I-V characteristics offered promising optical modulation results.

B Optical Transmission

The optical transmission spectra as a function of the MZ04 wavelength at zero bias applied to the active region, and of the nearest passive waveguide to MZ04, are plotted in Fig. 6.10. The optical spectrum of the passive waveguide is used to normalize the optical transmission of MZ04. This normalization method separates on-chip loss caused by light coupling from the fiber into the rib waveguide, as well as by additional propagation loss in the silicon waveguide.



Figure 6.10: Optical transmission of a passive waveguide and MZ04 of chip P01-L14C14 batch T582B, L = 1.8 mm.

The normalized transmission spectra of MZ04 for reverse bias from 0 to 8 V depicted in Fig. 6.11 shows an IL (maximum transmission value) at -4 dB. The red shift (shift towards longer wavelengths) of the optical transmission spectra with increasing reverse bias is caused by the effective refractive index variation n_{eff} , which is produced by hole depletion in the longer arm of the asymmetric MZI [4]. However, Fig. 6.11(a) shows that a maximum shift of the optical transmission spectra, or π shift, does not occur between 0 V and -8 V, which means that the targeted doping concentrations and distribution inside the waveguide were probably not achieved during fabrication. To find the experimental phase shift, the FSR and the $\Delta\lambda$ were evaluated from the optical transmission curves.



Figure 6.11: (a) Optical transmission spectra of MZ04 for reverse applied voltage between 0 V and -8 V and (b) inset of the optical shift showing the IL and ER at the operating point of 1551 nm.

The plot shows measurements of FSR = 13 nm, and $\Delta\lambda \approx 2$ nm at -8 V; thus, the phase shift can be calculated as follows

$$\Delta \phi = \frac{2\pi \Delta \lambda}{FSR} = 0.3\pi \tag{6.4}$$

As the length of MZ04 is 1.8 mm, the modulation efficiency $V_{\pi}L_{\pi} = 6$ V·cm. The same $V_{\pi}L_{\pi}$ was evaluated for reverse voltages larger than 4 V, and is reduced to 4 V· cm at 2 V. The theoretical $V_{\pi}L_{\pi}$ value had been calculated to be 1.7 V· cm, which may corroborate the theory that the targeted doping concentrations were not reached inside the waveguide. Another possible explanation of this result is unwanted dopant diffusion within the waveguide regions. The inset of the optical transmission shift between 1558 nm and 1563 nm is seen in Fig. 6.11(b). The figure shows that at 1561 nm, the chosen operating point, the IL is 6 dB, and the ER is 8 dB.

C High-Speed Performance

The high-speed performance of MZ04 was evaluated with the 10 Gbit/s NRZ PRBS using a sequence length of $2^7 - 1$. The measured eye diagrams are shown in Fig. 6.12, where the ER as a function of the wavelength was evaluated in order to find an operating point with an optimal compromise between ER and IL, as depicted in Fig. 6.13. The optical transmission curve shifted approximately 3 nm in the high-speed measurements. This shift is a result of device heating caused by dissipation of the electrical signal when the capacitance is charged and discharged by the RF signal. The largest ER was found at 1564 nm, as shown in Fig. 6.12(a), where the minimum ER corresponds to the maximum level of the DC transmission spectrum.

Furthermore, the signal saturation of the *hi* bits shown in the eye diagram measurement from Fig. 6.12(a) corresponds to a modulation close to the maximum transmission spectra of the MZM. The saturation decreases when the wavelength increases as the signal moves to an operating point at a lower transmission level. However, the ER measurements at larger wavelengths are limited by the sensitivity of the experimental setup as it gets closer to the minimum transmission level. The optical loss was calculated at the maximum transmission

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level; thus, considering the red-shift, at λ = 1564 nm, IL= 6 dB.



Figure 6.12: 10 Gbit/s eye diagram measurements of MZ04 showing increasing values of the measured ER when choosing progressive operating points: (a) at 1561 nm, ER = 1.8 dB, (b) at 1562 nm, ER = 3.6 dB, (c) at 1562, ER = 6.2 dB, and (d) at 1564 ER = 8.1 dB.

Using the driving voltage from this measurement, $V_{RF} = 6.5 V_{pp}$, and the calculated device capacitance C_D , it was possible to calculate the dynamic energy consumption per bit, using the following equation [70]

$$\Delta E_{\rm bit} = \frac{C_D V_{\rm pp}^2}{4} \tag{6.5}$$

The total energy consumption per bit of modulator MZ04 (L = 1.8 mm) was calculated to be 3.8 pJ/bit, whose value is comparable to the energy consumption of devices of similar lengths such as the ones listed in Table 2.1 (columns 2 and 4).



Figure 6.13: $V_{\pi}L_{\pi}$ measurements of MZ04, L = 1.8 mm, from T582B.

D Bit-Error Rate (BER)

The BER of a commercial lithium niobate (LiNbO₃) electro-optical modulator and the silicon optical modulator T582B MZ04 were measured at different received optical power levels (Fig. 6.14). The transmitted output signals were detected by a *NEL* 10 Gbit/s optical receiver (Module SS21B) and injected into an *Anritsu* MP1794A bit error detector for evaluation.



Figure 6.14: 10 Gbit/s BER comparison between MZ04 from T582B and a commercial LiNbO₃ electro-optical modulator.

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The measured results of the BER versus the received optical power for both modulators can be seen in Fig. 6.14. The plot shows the error-free operation as well as the resulting power penalty of MZ04 in comparison to the LiNbO₃ modulator. A power penalty of approximately 4 dB is observed, which is probably due to a lower ER than the LiNbO₃. The BER analysis of the silicon optical modulator was completed in collaboration with *III-V Labs*.

6.2.2 Mach-Zehnder Modulator at 40 Gbit/s Based on a PIPIN Diode

The MZMs that delivered operation speeds up to 40 Gbit/s were characterized from batch U532P. These devices were fabricated using the self-aligned process flow described in Chapter 4. This fabrication process offered the most accurate alignment between the modulator's waveguide and the p-doped slit, and guaranteed the best modulation efficiency in the middle of the rib waveguide.

A total of nine wafers were fabricated on batch U532P. The results of the MZMs presented from this batch had the same design parameters as the modulators from batch T582B; e.g., the modulators were based on PIPIN diodes embedded in the arms of an asymmetric MZI, and designed for applications at 1.55 μ m. In general, U532P delivered a fair quantity of working modulators from wafers P03 and P04. The working modulators were found in chips chosen from columns located in the middle of the wafer: specifically, columns C09 to C13. In terms of rows, the modulators with high-speed operation were located between rows L03 and L06, considering that only the top-half section of all the wafers from U532P was received from *CEA-LETI*. The best-working modulators were found in chip L05C11 of wafer P04. The results of two working MZMs are presented next: a modulator of length 0.95 mm labeled as MZ11 in the fabrication mask, and a modulator of length 4.7 mm labeled as MZ17.

A Electrical Measurements

The I-V measurements of the U532P devices delivered reverse currents in the order of a couple of tenths of μ A, two orders of magnitude larger than the current delivered by MZ04 from T582B. Larger current values, and breakdown voltages of approximately -10 V and -9 V, for MZ11 ($L = 0.95 \mu$ m) and MZ17 ($L = 4.7 \mu$ m), respectively, in comparison to -11 V from the previous batch, indicated slightly higher doping concentrations in the active region, which was one of the limitations for high-speed operation above 10 Gbit/s of the modulators T582B. The I-V plot also shows that the measured current of MZ17 was approximately one order of magnitude

larger than the measured current of MZ11, which is due to device length difference.

The doping concentration of the *p*-slit was estimated using the plots from Appendix E. Breakdown voltage occurred between -10 and -9, which corresponded to a *p* doping concentration between $1.5 \times 10^{17} - 2 \times 10^{17}$ cm⁻³. These concentrations are still considerably lower than the targeted *p* doping concentration, yet higher than those estimated from batch T852B. These results showed that the expected speed operation of these devices could be slightly higher than in the previous modulator (See Fig. 6.15).



Figure 6.15: I-V curve of MZ11 (L = 0.95 mm) and MZ17 (L = 4.7 mm) from U352P.

The S-parameters of MZ11 (L = 0.95 mm) and MZ17 (L = 4.7 mm) are shown in Fig. 6.16. In the S-parameters' measurement of MZ11, the transmission parameter S_{21} experiences a dip at approximately 15 GHz (Fig. 6.16[a]). The dip in S_{21} is matched by a sharp increase of the reflection parameter S_{11} at the same frequency, as was shown in Chapter 5. Disregarding the transmission dip, an S_{21} of 3 dB bandwidth of approximately 25 GHz was measured for the shorter modulator MZ11, when a DC reverse bias of -2.5 V was applied to the device. Furthermore, when evaluating the S_{21} parameter of MZ17 (Fig. 6.16[b]), the influence of the substrate is noticeable as a slight dip can be seen from approximately 2 GHz to 10 GHz. This dip is a result of an increase in the substrate's capacitance due to the considerably longer length of MZ17 in comparison to MZ11. The measured 3-dB bandwidth of the S_{21} of MZ17 was approximately 2.5 GHz. Nevertheless, an S_{11} parameter below 10 dB up to 50 GHz is seen for both devices, indicating that high-speed modulation operation is possible in both devices.


Figure 6.16: Schematic view of the possible problems encounter with RF probes when measuring S-parameters. Measured forward transmission and reflection parameters S_{21} and S_{21} and S_{11} of (a) MZ11 (L = 0.95 mm) and (b) MZ17 (L = 4.7 mm), from U532P.

B Optical Transmission

Figure 6.17 depicts the measured optical transmission spectra as a function of the wavelength, for different applied voltages between 0 V and -8 V. The measured spectra were normalized using the transmission spectrum of a neighbor passive waveguide that traveled the same distance in the chip as the MZMs, as was done for the transmission spectra of MZ04 from T582B. To measure the actual loss from the modulator, the losses due to the passive waveguides need to be added. These losses were removed when normalizing with respect to neighboring waveguides, and are usually negligible in comparison to other loss contributions. However, because the devices from batch U352P delivered considerably weaker losses than the modulators from the previous batch, the loss contribution caused by passive waveguides are no longer negligible; therefore, calculation of the overall device loss is important. The passive losses were determined by measuring the transmission spectra of passive waveguides of different lengths available in the chip. These losses were calculated to be less than 0.5 dB/mm. By adjusting the calculated passive losses, the plotted optical transmission of MZ11 (L = 0.95mm) showed an IL of -2.5 dB (Fig. 6.17[a]), while the IL of MZ17 (L = 4.7 mm) was evaluated at -6 dB (Fig. 6.17[b]). These values represent the entire on-chip IL from the optical modulators, including losses in both the star coupler-based splitters and combiners, and the phase shifters. From these values, the loss of each coupler was extracted as 0.75 dB, and phase shifter loss of 1 dB/mm was deduced [5].

Looking at the transmission spectra of both modulators, we noticed that the ER_{DC} of MZ11 was larger than the ER of MZ04 from T582B. Modulator MZ11 delivered an ER_{DC} larger than 10 dB, as shown in the inset of Fig. 6.17, in comparison to the ER_{DC} of MZ04 of T582B, which was measured at 8 dB. Knowing that the length of MZ11 was almost half that of MZ04 from T582B, this measurement helped us to verify that the optical response of the modulators from U532P had improved.

Replicating the same procedure followed to estimate the $V_{\pi}L_{\pi}$ of MZ04 from T582B, a value of 3.5 V·cm was calculated for both modulators of batch U532P. This modulation efficiency is an improvement from the previous batch, however, the experimental $V_{\pi}L_{\pi}$ continues to be moderately larger than the theoretical value 1.7 V·cm, showing the first possibility of improving the modulator's performance by optimizing the dopant concentrations inside the waveguide.



Figure 6.17: (a) Optical transmission spectra of MZ11 (L = 0.95 mm) for reverse applied voltage between 0 V and -8 V, and inset of the optical shift showing ER at an operating point. (b) Optical transmission spectra of MZ17 (L = 4.7 mm) for reverse applied voltage between 0 V and -8 V, showing a complete π -shift at -8 V.

C High-Speed Performance

Eye diagrams at 40 Gbit/s were measured using a PRBS with a $2^{15} - 1$ sequence length and a driver amplifier that delivered an RF of 6.5 V_{pp}. The maximum ER obtained for the shortest modulator MZ11 was 3.2 dB, as shown in Fig. 6.18(a), while the longest modulator MZ17 reached a maximum ER of 7.57 dB, as seen in Fig. 6.18(b).

The operating point of MZ17 was chosen at the maximum optical transmission level delivered by the optical modulator in such way as there were no additional optical loss contributions associated with the optical modulator working at a lower optical transmission level. The operating point was also seen in the optical transmission spectra from Fig. 6.17(b), where a full π -shift occurred between the optical transmission curves at 0 V and at -8 V. The measurement of the eye diagram at the maximum optical transmission indicated that the total IL measured is 6 dB. Additionally, saturation of the signal can be seen at the bottom of the eye diagram from Fig. 6.18(b), which corresponds to the noise limitation due to the residual spontaneous emission of the EDFA, indicating that the ER value of 7.5 dB was not limited by the modulator itself, but by the experimental tools [5]. This result was consistent with the optical transmission spectra of MZ17, in which an ER_{DC} larger than 15 dB was expected, with ideal measurement conditions. The operating point chosen to measure the eye diagram of MZ11 did not correspond exactly to the maximum optical transmission level of the modulator, and 2 dB of additional loss was tolerated to obtain an ER of 3.2 dB. In fact, in the case of MZ11, the phase shift was not as large as MZ17, making it more challenging to find an optimal operating point where the ER eye diagram was open enough for it to be measured. Adding the 2 dB loss from the high-speed measurement, MZ11 delivered an IL of 4.5 dB.

A rise /fall time of approximately 25 ps is seen in Fig. 6.18(c), which corresponds to the eye diagram measurement of the signal delivered by the driver amplifier. The figure shows that the rise/fall time is approximately the same rise/fall time seen in the eye diagrams measured from the modulators, especially from modulator MZ17 (Fig. 6.18[b]). This observation corroborates once again that the measurements obtained are, in fact, limited by the experimental equipment. Following the same procedure used for modulator MZ04 from T582B, the experimental capacitance C_D of each device was estimated as 0.22 fF/ μ m and 0.21 fF/ μ m, for MZ11 and MZ17, respectively, using the *p*-doping concentration found in Appendix E.



(a) Eye diagram of MZ11, L=0.95 mm.



(b) Eye diagram of MZ17, L=4.7 mm.



(c) Eye diagram of the RF driver amplifier.

Figure 6.18: Eye diagram measurements of (a) MZ11 (L = 0.95 mm) showing an ER= 3.2 dB, (b) MZ17 (L = 4.7 mm) showing an ER= 7.5 dB, and (c) the output signal delivered by the RF driver amplifier.

The dynamic electrical consumption per bit was calculated from Eq. 6.5 as 2 pJ/bit and 10 pJ/bit for MZ11 and MZ17, respectively. The results show similar energy consumption of MZ11 as for the previous modulator, provided that MZ11 is almost half the length of MZ04 from T852B. Additionally, the energy consumption of modulator MZ17 is among the lowest when compared to other long MZM (> 3 mm) listed in Table 2.1 and Table 2.2. Low energy consumption is caused by the use of a PIN junction in the active region, as it produces a smaller C_D than a PN junction.

D Electro-Optical Response

The electro-optical responses of modulators MZ11 and MZ17 as a function of the frequency for an applied reverse voltage of -2.5 V, are seen in Fig. 6.19. The curves were normalized to the response at the lowest frequency measured. The electro-optical measurement of modulator MZ11 seen in Fig. 6.19(a) shows a trend with a somewhat constant decaying optical response with increasing frequency. A similar transmission dip was seen in Chapter 5 in the S-parameter measurements (Fig. 5.16) and the electro-optical response (Fig. 5.20), which was expected, as these measurements involve the propagation of the RF along the traveling-wave CPW electrodes. It can be seen, though, that the transmission dip seen in Fig. 6.19(a) occurs at approximately 22 GHz, whereas in the previous case the measurements occurred at ~ 15 GHz. The shift in the dip is probably caused by the applied DC voltage, as the measurements shown in Fig. 6.19 were taken at -2.5 V, while the electro-optical measurements shown in Chapter 5 were taken at zero bias. Nevertheless, it can be seen that the 3-dB bandwidth measured of MZ11 is larger than 40 GHz.

The same consistency occurs for the electro-optical response of MZ17. In this case, the electro-optical response does not have a sudden dip, but it does have slight ripples all along the measurement, which are also more pronounced in the S-parameter measurements. The electro-optical bandwidth measured for this modulator was ~ 20 GHz. Regardless of the dips and ripples in the electro-optical responses, the electro-optical bandwidths obtained indicate the high-speed operation capability of the modulators, which is partly due to the design of the RF electrodes, as well as the substrate chosen, because minimum electrical loss contribution came from using a 2 μ -thick buried oxide silica and high resistivity silicon on the SOI substrates.



Figure 6.19: Small-signal electro-optical response silicon optical modulators (a) MZ11 (L = 0.95 mm) showing a bandwidth of > 40 GHz and (b) M17 (L = 4.7 mm) showing a bandwidth of ~ 22 GHz.

Throughout the last two sections of this chapter, the first high-speed silicon optical MZMs designed at *IEF* were presented. The experimental results obtained from the MZMs provided, for each batch, the highest operation speed of the silicon optical modulators designed for applications at a wavelength of $1.55 \,\mu$ m. Table 6.1 presents a summary of the results obtained in three MZMs, including modulator MZ04, the first high-speed-working modulator, as well as modulators MZ11 and MZ17, the second generation of MZMs delivering the highest speed operation. The results show an improvement in all the FoM of MZ11 and MZ17 in comparison to MZ04, showing the importance of recognizing the issues of previous devices. The FoM delivered by the latest modulators show comparable FoM as similar MZMs, but they delivered fewer losses and energy consumption per bit than the others, mainly due to the PIPIN diode used as active region. Additionally, experimental results show the possibility of improvement, especially if the targeted doping concentrations are properly implanted.

Specification	Modulator MZ04	dulator MZ04 Modulator MZ11	
Wavelength (λ_0)	$1.55\mu{ m m}$	$1.55\mu\mathrm{m}$	$1.55\mu{ m m}$
Batch	T852B	U352P	U352P
Length	1.8 mm	0.95 mm	4.7 mm
Speed	10 Gbit/s	40 Gbit/s	40 Gbit/s
Breakdown Voltage	$-11\mathrm{V}$	$-10 \mathrm{V}$	$-9 \mathrm{V}$
Operating V_{DC}	-2.5 V	-2.5 V	-2.5 V
V_{RF}	$6.5\mathrm{V_{pp}}$	$6.5\mathrm{V_{pp}}$	$6.5\mathrm{V_{pp}}$
$V_{\pi}L_{\pi}$	4 V·cm	3.5 V⋅cm	3.5 V⋅cm
f _{3dB}	_	~ 25 GHz	~ 2.3 GHz
E/O Bandwidth	_	>40 GHz	20 GHz
ER _{DC}	> 15 dB	~ 25 dB	~ 25 GHz
ER _{RF}	8.1 dB	3.2 dB	7.5 dB
IL	6 dB	4.5 dB	6 dB
Δ Energy/bit	3.8 pJ/bit	2 pJ/bit	10 pJ/bit

Table 6.1: Characterization results comparison among the MZM designed for applications at $\lambda = 1.55 \ \mu m$.

6.2.3 Ring Resonator Modulator based on Interleaved PN Junctions

In addition to the MZM results obtained from the wafers of batch T582B, various RRs delivered high-speed operation up to 10 Gbit/s. One type of working RR was designed by integrating interleaved PN junctions along the perimeter of a ring of radius 50 μ m, except for the ring section that passes in front of the passive straight waveguide, where the interleaved junctions were embedded in 292 μ m of the 314 μ m of the ring's perimeter (Fig. 4.3). The lengths of the P and N regions were 400 nm and 300 nm, respectively. The passive waveguide was designed 200 nm away from the edge of the ring.

The RRs discussed in this section were found in wafer P01, in chips located close to chip L14C14, the chip where the working MZM was found. In fact, all the working RRs were found in chips located in column C14 and rows L12 to L15. This location corresponds to the center of the wafer. The results of RR64 from L12C14 are presented next.

A Optical Transmission

The normalized optical transmission spectra as a function of the wavelength are plotted in Fig. 6.20(a) for reverse bias from 0 V to -10 V, for Transverse Electric (TE) polarization. Using Eq. 2.15, the Q factor can be calculated. Looking at the transmission curve at 0 V from Fig. (b), the FWHM is ~ 78 nm, and $\lambda_0 = 1552.35$; thus, $Q = \lambda_0 / \Delta \lambda_{FWHM} \approx 20,000$. The optical transmission for Transverse Magnetic (TM) polarization was also measured, and the Q factor was calculated to be approximately 25,000. The ER ratio was ~ 14 dB for TE polarization and ~ 10 dB in TM polarization.

The effective index variation $\Delta n_{\rm eff}$ was calculated from the optical transmission spectra using the relationship

$$\frac{\Delta n_{\rm eff}}{n_g} = \frac{\Delta \lambda}{\lambda_0} \tag{6.6}$$

where $\Delta \lambda$ is the wavelength shift, λ_0 is the wavelength where the optical transmission reaches a minimum value, and n_g is the group index, which could be calculated using Eq. 2.13. To calculate Δn_{eff} for the TE polarization, we can refer to Fig. 6.20(b). The center frequency λ_0 is located at $\lambda = 1552.35$ nm, and the measured FSR is 1.79 nm; thus, n_g is calculated from

$$n_g = \frac{\lambda_0^2}{FSR \cdot L} = 4.28\tag{6.7}$$

where $\Delta\lambda$ between 0 V and -10 V is $\Delta n_{\text{eff}} = 1.6 \times 10^{-4}$. A value of $\Delta n_{\text{eff}} \approx 2 \times 10^{-4}$ was estimated for TM polarization, and Δn_{eff} and $V_{\pi}L_{\pi}$ were calculated for all the resonance shifts at each voltage and for both polarizations. The results are plotted in Fig. 6.21.

Figure 6.21(a) shows a larger Δn_{eff} for TM polarization. This change in refractive index is translated into a better V_πL_π (smaller values), as seen in Figure 6.21(b). V_πL_π varies from 2.5 V·cm to 4.9 V·cm and from 2.5 V·cm to 4.2 V·cm for TE and TM polarizations, respectively [49]. The largest modulation efficiency for TM polarization is due to a larger confinement of the optical mode in the rib waveguide, where carrier depletion occurs. The FF of the optical mode had been calculated to be 88% for the TM mode and 83 % for TE.

The $V_{\pi}L_{\pi}$ of the RR modulator based on interleaved PN junctions was simulated to be 1.4 V·cm for TE polarization and 1.2 V·cm for TM polarization. However, Fig. 6.21(b) shows that the best $V_{\pi}L_{\pi}$ obtained is 2.5 V·cm for both polarizations. To understand the discrepancy in experimental results, the S-parameters measurements of the RR were measured to evaluate the capacitance as a function of the voltage from 0 V to -10 V. A capacitance varying from 54 to 38 fF was evaluated. The theoretical and experimental capacitances were compared, and the results indicated that the real doping concentrations were lower than the targeted values of a *p*-doping concentration of 4×10^{17} cm⁻³ and an *n*-doping concentration of 1×10^{18} cm⁻³. If the modulator had had the designed doping concentrations, the capacitance varied from ~ 220 fF to 64 fF. As seen in the previous devices, the on-chip IL was estimated to be ~ 6 dB, including ~ 5 dB from the 5-mm-long passive waveguides used to guide the light on the chip, and ~ 1 dB excess loss from coupling losses. The specific amount of loss from the ring resonator and the passive waveguides was evaluated using the same procedure explained in the MZM sections.



Figure 6.20: Optical transmission spectra of RR64 from T582B, a RR modulator based on interleaved PN junctions, measured in TE polarization.



Figure 6.21: (a) $\Delta n_{\rm eff}$ and (b) $V_{\pi}L_{\pi}$ of RR64, a RR modulator based on interleaved PN junctions for TE and TM polarizations.

B High-Speed Performance

The high speed performance of RR64 was measured at 10 Gbit/s for a PRBS sequence length of $2^7 - 1$. The setup used for MZMs was changed to fulfill the need of the RR's electrode pads. The RF electrodes of the RR modulators were not designed to be able to carry a 50 Ω -load. For that reason, the output of the RF driver amplifier was connected to a power divider, where a 50- Ω load was attached to one of its outputs to prevent damage from the driver amplifier. This setup requires a 50- Ω termination, and a bias tee was attached to the other output. The bias tee was used to apply the DC bias to the ring to guarantee continuous carrier depletion in the active region. The presence of the power divider decreased the peak-to-peak voltage to less than 5 V_{pp}.

The measured eye diagram for both TE and TM polarizations can be seen in Fig. 6.22. An ER of 4 dB and 2.7 dB at was measured at 10 Gbit/s for TE and TM polarizations, respectively. The ERs were slightly decreased in comparison with the ER_{DC} , which can be attributed to the presence of the 50- Ω load connected in parallel to the RF signal applied to the ring.



(a) Eye diagram for TE polarization.

(b) Eye diagram for TM polarization.

Figure 6.22: Eye diagram measurements of RR64 from T582B at 10 Gbit/s for (a) TE polarization, and (b) TM polarization.

The RR's power consumption was calculated from Eq. 6.5 using the modulator's measured capacitance $C_D \approx 46$ fF. Knowing that $V_{pp} = 5$ V, the calculated energy consumption per bit was 287 fJ/bit.

6.2.4 Ring Resonator Based on a PIPIN Diode for Applications at $\lambda = 1.55 \,\mu m$

The last type of working RR (from batch T852B) was designed with a PIPIN diode along the perimeter of a ring of radius 50 μ m. The diode was also suppressed from the section of the ring that passes in front of the passive straight waveguide. The passive waveguide also was designed 200 nm away from the edge of the ring. Various RRs based on PIPIN diodes produced an acceptable optical transmission and high-speed operation up to 10 Gbit/s on the same chips where the RRs based on interleaved junctions had been found. The results presented in this section correspond to ring RR01 (mask layout) from chip L13C15 .

A Optical Transmission

The characterization of the RR was carried out for both TE and TM polarizations. The normalized optical transmission spectra as a function of the wavelength for reverse bias from 0 to -10 V is shown in Fig. 6.23 for both polarizations. The optical transmission was optimized using the same process of the MZMs.

The transmission spectra of the RR for TE polarization show an IL lower than 1 dB, and an ER larger than 9 dB between 0 V and -10 V (see Fig. 6.23[a]), while transmission spectra for TM polarization show an IL of ~ 2 dB and an ER of ~ 17 dB between 0 V and -10 V (see Fig. 6.23[b]). The ER difference between both polarizations indicates that in TM polarization, the RR is close to critical coupling [56].

The effective index variation $\Delta n_{\rm eff}$ was calculated using the same procedure developed for the RR based on interleaved PN junctions. $\Delta n_{\rm eff}$ of 2.3×10^{-4} and 2.4×10^{-4} were calculated for TE and TM polarizations, respectively. From $\Delta n_{\rm eff}$, the calculated modulation efficiency $V_{\pi}L_{\pi}$ was ~ 3 V·cm for both polarizations.



Figure 6.23: Optical transmission spectra of RR01 from T582B, a RR modulator based on a PIPIN diode.

B High-Speed Performance

The high speed performance of RR01 was measured using the same setup used to measure the eye diagrams of the RRs based on interleaved PN junctions. The measured eye diagrams for both polarizations can be seen in Fig. 6.24. ER of 5.9 dB and 5.3 dB were measured for TE and TM polarizations, respectively. A reduction of the ER is seen in comparison with the ER_{DC} measured in the static response, which can be attribute to signal attenuation. However, it can also be explained by a reduction in the voltage applied to the diode due to impedance mismatch, and by the fact that the general characteristics of the optical transmission spectra of RRs strongly depend on the light coupling between the resonator and the passive straight waveguide. It was mentioned before that RRs are sensitive to temperature changes. The application of the RF signal causes a temperature increase of the device, leading to a change of the refractive index in the active region, and modifying the RR's coupling condition.



(a) Eye diagram for TE polarization.

(b) Eye diagram for TM polarization.

Figure 6.24: Eye diagram measurements of RR01 from T582B at 10 Gbit/s for (a) TE polarization, and (b) TM polarization.

C Bit-Error Rate

The BER of modulator RR01 was measured at different received optical power levels. The same measurements were carried out for both polarizations in comparison with the BER of a commercial LiNbO₃ electro-optical modulator, as seen in Fig. 6.25. A BER lower than 10^{-9} was measured for both light polarizations. The figure shows that at low received power values, the

measurement is limited by the silicon modulator's ER. It can also be seen that at high power values the curves do not follow a linear pattern. Possible explanations include thermal drift, as the temperature of the RR was not controlled during the measurements, and the spontaneous emission noise of the EDFA, whose input power was approximately –30 dBm due to large coupling losses.



Figure 6.25: 10 Gbit/s BER comparison between a commercial $LiNbO_3$ electro-optical modulator and RR01 from T582B for TE and TM polarizations.

Table 6.2 summarizes the FoM obtained for the RR silicon optical modulators characterized during my Ph.D. work. The results presented show the versatility that these kinds of modulators offer in providing similar results with the same electrical structure yet different interferometric structure, such as the case of RR01, which was also designed with a PIPIN diode in the active region. Moreover, the results of modulator RR64 were the first experimental results published on silicon optical modulators using interleaved PN junctions in the active region, opening new possibilities for future generations of RR modulators. It is important to mention that these results were obtained from the first characterized batch T852B, as no RR modulators were designed in batch U352P. Optimization of these devices to improve operation speed and FoM will soon be made.

Specification	Modulator RR01		Modulator RR64	
	TE	ТМ	TE	TM
Wavelength (λ_0)	1.55 <i>µ</i> m	1.55 <i>µ</i> m	$1.55\mu{ m m}$	$1.55\mu{ m m}$
Electrical Structure	PIPIN diode	PIPIN diode	Interleaved PN diodes	Interleaved PN diodes
Batch	T852B	T852B	T852B	T852B
Length	292 µm	292 µm	$292 \mu \mathrm{m}$	292 µm
Speed	10 Gbit/s	10 Gbit/s	10 Gbit/s	10 Gbit/s
Operating V_{DC}	$-2.5 \mathrm{V}$	$-2.5\mathrm{V}$	$-2.5\mathrm{V}$	$-2.5\mathrm{V}$
V _{RF}	5 Vpp	5 Vpp	$5\mathrm{V}_\mathrm{pp}$	5 Vpp
$V_{\pi}L_{\pi}$	∼ 3 V·cm	~ 3 V·cm	2.7 V⋅cm	2.4 V·cm
Q Factor	25,800	-	20,000	25,000
ERDC	~ 9 dB	~ 17 dB	~ 14 dB	~ 10 dB
ER _{RF}	5.9 dB	5.3 dB	4 dB	2.7 dB
IL	< 1 dB	~ 2 dB	6 dB	6 dB
Δ Energy/bit	-	-	287 pJ/bit	287 pJ/bit

Table 6.2: Characterization results comparison among the RRs designed for applications at $\lambda = 1.55 \ \mu m$.

6.2.5 Ring Resonator Based on a PIPIN Diode for Applications at $\lambda = 1.3 \,\mu m$

In view of the need for future characterization of silicon optical modulators at 1.27 μ m, we decided to test the available silicon optical modulators at this wavelength. Because the devices had not been optimized at 1.27 μ m, it was difficult to find a working modulator. No MZMs were found in either of the wafers of batches T852B and U532P, but a few RRs delivered optimal optical transmission spectra as presented next. The working RRs were found in wafer P04 of U532P in chips L06C02, L06C03 and L06C04, which are locations near the top edge of the wafer.

Four RRs produced an important shift of the optical transmission when bias was applied. Two of the rings had the same dimensions as the ring resonators presented in the previous section. In particular, the rings had a radius of 50 μ m, and a 200-nm gap between the ring and the passive waveguide. Additionally, ring resonators of radius 50 μ m and gaps 250 nm, up to 400 μ m delivered acceptable results. The length of the active region was 292 μ m for all the RRs characterized at 1.27 μ m.

A Optical Transmission

The DC transmission spectra as a function of the wavelength are shown in Fig. 6.26. The curves were plotted for different reverse voltages from 0 V to -10 V. Only the results of one of RRs modulators is shown for simplification: Fig. 6.26(a) shows the optical transmission of modulator RR11, a RR of radius 50 μ m and gap 200 nm, and Fig. 6.26(b) shows the calculated $V \cdot L\pi$ of the same RR. The transmission curve was normalized with respect to an adjacent passive waveguide. At 1.27 μ m, the modulator delivered a moderate shift when looking at the transmission curve at 0 V and 10 V. Using the $\Delta\lambda$ from modulator RR11 (Fig. 6.26[a]) the calculated $V_{\pi}L_{\pi}$ ~ varies from 3 to approximately 5.75 V·cm for an applied reverse voltage from 0 to -10 V (Fig. 6.26[b]). It is important to note, however, that the usual DC voltage used during characterization is approximately -2.5 V, which means that the results obtained at $\lambda = 1.27$ μ m show a modulation efficiency that is considerably lower than the RRs characterized at 1.55 μ m, whose V_{π}L_{π} was estimated to be ~ 3 V·cm for both polarizations. Furthermore, the transmission plot shows that at 1.27 μ m the losses are considerably larger than those obtained at 1.55 μ m. However, despite these results, the measurements obtained at 1.27 μ m helped us to find a starting point with a set of dimensions that could be optimized for future RRs silicon optical modulators that operate at this wavelength.



Figure 6.26: Optical transmission spectra of RRs from U532P, characterized at $\lambda = 1.27 \ \mu m$. (a) RR11 with $r = 10 \ \mu m$ and gap= 200 μm . (b) Calculated V·L_{π} for an applied reverse voltage from 0 to -10 Vs.

Conclusion

This chapter presented the characterization process followed throughout my work, starting with the first experimental demonstration of high-speed silicon optical modulators based on PIPIN diodes and interleaved PN junctions working at 10 Gbit/s. The modulation response obtained from batch T582B helped us to discover design and fabrication issues that were addressed in the next generation of silicon optical modulators U352P. Throughout this chapter, a variety of characterization techniques to determine the efficiency of silicon optical modulators was presented. These techniques showed the challenges faced when testing modulators, where electrical and optical characterization procedures are required in separate and integrated configurations.

The first working MZM at 10 Gbit/s, provided an ER as high as 8.1 dB with losses of approximately 6 dB. However, it was found that the modulation efficiency limited the device's efficiency for higher operation speeds. Most characterization measurements gave evidence of this limitation, starting with the DC transmission measurement, where the transmission spectra did provide the expected shift, and thus, Δn_{eff} , with the applied reverse voltage. However, the results obtained for this modulator left several clues behind for the improvement of future devices. In fact, the MZM fabricated in the next generation, U532P, gave evidence of the improvements made to the initial design. The newer MZM provided an ER that is comparable with other available silicon optical modulators based on carrier depletion, but it was the IL that verified the advantage of using PIPIN diodes as the active region, because the lower IL allowed for an improved SNR of the signal. The silicon optical modulator delivered an appreciably low IL (6.5 dB) for a rather long device (L = 4.7 mm), which proves the effectiveness of the PIPIN in reducing the optical losses in the active region. This loss level is comparable with 1-mm-long silicon optical modulators found by other research groups around the world. It was also shown that the results obtained for the MZM that operated at 40 Gbit/s were limited by the experimental equipment, and not by the device. AAt the same time, it was shown that the PIPIN diode works for different structures, as it was demonstrated that it also works when implemented on RRs, which could be useful modulators for applications in which a reduced footprint may be required.

The first working modulator based on interleaved PN junctions showed promising results. The modulator delivered high-speed operation at 10 Gbit/s, leaving behind many

clues not only for design improvement for RRs, but also for MZMs. Nevertheless, interleaved PN junctions have a potential for operation at higher speeds. Furthermore, the successful characterization of the silicon optical modulators at 1.27 μ m represented a starting point in the design and fabrication of devices for applications at this wavelength. Obtaining some results at 1.27 μ m on devices that were not optimized at this wavelength offers a positive outlook for future projects. The characterization process also highlighted the importance of each of the designed parameter details, as they show up in the results. However, it was also seen throughout the chapter that certain design parameters are not easy to control (e.g., dopant implantation), thus the reason for implementing different fabrication techniques that will improve the outcome of the devices. Furthermore, it was also seen that the experimental equipment is a limiting factor in the characterization process; therefore, maintenance of a clean and organized experimental bench is essential, where cables and measurement equipment are in an good working condition, using the short connections and the least possible adaptors, as signal attenuation has a great influence on characterization results. It was also important to check the quality of the signal delivered to the modulators in order to identify problems and make possible changes to solve them.

Overall, the characterizations of the multiple silicon optical modulators presented in this chapter showed promising results, by contributing two unusual designs of the active region for future modulator generations. Finally, it is important to mention that having complete hands-on access to all the characterization methods and processes of the HELIOS modulators was very useful in providing awareness, understanding, and appreciation of all the details in the design of the silicon optical modulators needed for the conception of the silicon transceiver at $\lambda = 1.27 \ \mu$ m required for the SILVER project.

7 Conclusion and Future Work

7.1 Conclusion

This dissertation has presented a walk through the complete conception of a silicon optical modulator, including the design, simulation, and optimization of the main optical and electrical structures, the fabrication processes, the design of the IC layout GDS file used to fabricate the device, and the characterization of the completed devices. The purpose of this work was the design of a silicon optical modulator that can be integrated into a silicon transceiver to be used for local network application at a 1.27 μ m wavelength to fulfill the parameters of *ANR*'s SILVER project.

The modulator design was conceived from scratch, as this was the first silicon optical modulator meant to be used for 1.27- μ m-wavelength applications. The initial model was selected by making a series of comparisons with a previously designed silicon optical modulator for 1.55- μ m-wavelength applications (HELIOS project). Nevertheless, due to important differences between both modulators—namely, the operation wavelength, and the more relaxed speed operation requirements of the SILVER project—a PN diode was selected as the main electrical structure of the active region. This structure was selected because it is relatively easier to fabricate than other electrical structures, and because the optimization of the PN diode met all of the project requirements. The PN diode was designed to be fabricated Inside a rib waveguide of height H = 440 nm, width w = 360 nm, and slab height h = 100 nm, with doping concentrations $p = 5 \times 10^{17}$ cm⁻³ and $n = 7 \times 10^{17}$ cm⁻³.

The design process also included the simulation and optimization of the traveling-wave electrodes that drive the RF signal of the modulator. Coplanar-waveguide (CPW) electrodes

in a GSG configuration were selected for this task. A simulation model was developed in *Agilent*'s electronic design automation software ADS with the purpose of finding the most appropriate CPW electrode dimensions according to the envisioned application. It was shown that the design of CPW electrodes depends mainly on two dimensions: the width of the signal electrode line W, and the gap between the signal and one of the ground electrode lines G. The optimization showed the influence of these two dimensions on the modulator's forward transmission (electrical bandwidth), characteristic impedance Z_D , variation of the RF refractive index, signal reflection, electro-optical response, and losses; it also showed how a compromise must be made among the affected parameters.

The simulation results showed that (W) determines the losses on the electrodes, where the widest W delivers the lowest loss, while G is partly responsible for the transmission signal attenuation (S_{21} bandwidth), although the entire response of the traveling-wave electrodes is determined by a combination of these two dimensions. Moreover, different G and W dimensions showed that the characteristic impedance of the modulator's equivalent circuit Z_D depends on the ratio between W and G. Consequently, the main compromises to be made in designing CPW electrodes are the characteristic impedance, the losses, and the bandwidth. Using all the considerations described above, the CPW electrode dimensions selected for the SILVER modulators are $G = 20 \ \mu m$ and $W = 20 \ \mu m$.

The characterization of devices designed for applications at $\lambda = 1.55 \ \mu m$ (HELIOS project) and fabricated with standard and self-aligned processes provided useful results for future generations of silicon optical modulators. These two processes were selected to verify the effectiveness of each method in creating the PIPIN diode with the design dimensions, guaranteeing the best overlap between the optical mode and the depleted region inside the rib waveguide. Because multiple batches of the HELIOS modulators were completed during my research, we were able to assess the advantages and disadvantages of each process. Some advantages included the fabrication of the waveguide with different widths from column to column. This technique allowed us to add some wiggle room to allow for differences between the simulation and experimental results caused by physical dimensions of the waveguide. Additionally, it was clearly seen that the self-aligned fabrication process did indeed provide a better overlap between the optical mode and the depleted region, as the modulators fabricated with this technique delivered optimal results at high-speed operation. The evaluation of the

fabrication processes raised additional problems that end up being a limiting factor for highspeed operation of the devices, however, such as the unevenness of the waveguide's flanks, and failure to completely etch off the silicides around the waveguide.

Taking into consideration the fabrication results obtained with the HELIOS modulators, a different process was created for the fabrication of the SILVER silicon optical modulators, not only to avoid the recurrence of the previous devices' problems, but also because the active region of the SILVER modulators was designed with a different electrical structure. All the design considerations taken for the generation of the IC GDS layout mask for the SILVER modulators were explained, with special consideration given to the dimensions used on all the different parts of each modulator: device length, passive structures, optical structures, electrical structures, and connection pads.

The HELIOS modulators designed for 1.55 μ m-wavelength applications delivered highspeed operation in two of the characterized batches. Batch T852B delivered three types of silicon optical modulator: a 1.8-mm-long asymmetric MZM based on lateral PIPIN diodes, a RR modulator based on a lateral PIPIN diode, and a RR modulator based on interleaved PN diodes. Both RRs were designed with a radius of 50 μ m and a gap between the ring and the passive straight waveguide of 200 nm. All three modulators provided high-speed operation up to 10 Gbit/s. The MZM delivered a dynamic ER of 8.1 dB and losses of 6 dB at the operating point. The largest dynamic ERs measured were 5.9 dB and 4 dB for the RR modulators based on a PIPIN diode and PN interleaved junctions, respectively. Batch U532P provided asymmetric MZMs of lengths 0.95 mm and 4.7 mm that worked at 40 Gbit/s. The 0.95-mm-long MZM delivered an ER of 3.2 dB and IL of 4.5 dB, while the 4.7-mm-long MZM delivered an ER of 7.57 dB and IL of 6 dB. Additionally, electro-optical responses of > 40 GHz and ~ 20 GHz for the 0.95-mm-long and 4.7-mm-long devices, respectively, in combination with forward reflection coefficient S_{11} below -10 dB, proved the high-speed operation of these two modulators. Furthermore, it was shown that the limitations of the modulators of this batch lay not in the devices themselves, but in the experimental setup.

The conception process of a silicon optical modulator studied in this dissertation exposed many of the challenges faced in the design of these devices. However, throughout this document, different techniques used and compromises made to circumvent the challenges proved to be successful for all the different parts that make up a high-speed silicon optical modulator. The results showed the possibility of designing devices that are able to fulfill the application requirements, such as speed, losses, and modulation efficiency, among others. The HELIOS modulators fulfill all the project requirements using different types of modulators, showing the versatility of the different structures within them. The modulators presented in this dissertation were mainly based on the PIPIN diode, an electrical structure designed to reduce the losses in the rib waveguide, where the modulation occurs. This structure proved to be successful for such a task.

The characterization results obtained for the HELIOS modulators helped me learn more about silicon optical modulators, including their critical design and fabrication parameters as well as their characterization limitations. The work done in HELIOS allowed me to put the acquired knowledge into practice for the conception of the SILVER silicon optical modulators, which were designed to be integrated on silicon transceivers meant for use in local network applications at a 1.27 μ m wavelength. The results obtained from the simulation of the SILVER modulators produced promising results to fulfill the project's requirements. If the SILVER modulators and transceiver deliver successful results at 10 Gbit/s, the results could be among the few 1.27- μ m-wavelength silicon optical modulators to be published. Successful results for the SILVER modulators would also place designers and engineers one step closer to replacing the last leg of existing electronic connections in current local Fiber to the Home (FTTH) networks.

7.2 Future work

A stand-alone silicon optical modulator is only one step in the quest of creating high-speedperformant photonic integrated circuits (PICs). For that reason, plans for future work on silicon optical modulators include the characterization of the silicon transceivers designed for SILVER, as the fabricated wafers include complete devices (laser and modulator), allowing the study of the response of a more complex system. Furthermore, for the modulator to be applied to existing applications in optical interconnections, the silicon optical modulator needs to deliver low power consumption, a large extinction ratio, and low losses. The results presented in this dissertation left room for improvements in all these areas for future generations of silicon optical modulators. Additional improvements include lowering the driving voltage to be able to integrate the optical devices with electronic circuits. Current optical transmission systems work at 10 Gbit/s using Wavelength-Division Multiplexing (WDM), but the increasing demands of information capacity have led to a search for the integration of multiple high-speed transmitters, where various amplifiers and tunable laser modulators, working in multilevel modulation schemes, are integrated in a single PIC. For a modulator to be able to work in a multilevel modulation scheme such as Quadrature Phase-Shift Keying (QPSK), it requires a more complex structure and 2π -shift [71]. *ANR*'s m<u>ULTI</u>level <u>Modulation IntegrAted TransmittEr</u> (ULTIMATE) project has increased the complexity of the silicon optical modulator, and this will be the future task that *IEF* needs to address.

Finally, the ultimate and main goal of this research was to create versatile silicon optical modulators that can be integrated in PICs that can be used in optical interconnects. The silicon optical modulator proved to be a feasible solution for the change from electronic to optical interconnects, hopefully, in the foreseeable future.

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Acronyms

- **ADS** Advanced Designed System. 128, 130, 132, 135–137, 139, 143, 144, 150, 154, 158–160, 165, 169, 220, 244
- APD Avalanche Photodiode. 46, 105, 106
- BER Bit-Error Rate. 172, 190, 211, 212
- BOX Buried Oxide. 18, 93, 117, 137
- CAD Computer-Aided Design. 3, 105
- **CAE** Computer-Aide Engineering. 3
- CMOS Complementary Metal-Oxide-Semiconductor. 1-5, 9-11, 18, 44, 46, 87
- **CPW** Coplanar Waveguide. 73, 74, 88–90, 109–111, 123–125, 128–131, 133, 135–138, 140–151, 154, 155, 158–165, 167–170, 178, 182, 200, 220, 244, 248, 249
- **DUT** Device Under Test. 175
- EDFA Erbium-Doped Fiber Amplifier. 3, 175–178, 198, 212
- **ER** Extinction Ratio. 26, 27, 41–43, 45, 46, 50, 83–85, 91, 108, 123, 171, 175–177, 186–190, 195, 197–199, 203, 204, 208, 209, 211–213, 216, 221
- FF Filling Factor. 19, 20, 56–58, 205
- FKE Franz-Keldysh effect. 10, 12, 13
- FoM Figures of Merit. 26, 40, 41, 46, 59, 70, 71, 84, 171, 175, 176, 203, 212

Acronyms

- FP Fabry-Perot Resonator. 175
- FSR Free Spectral Range. 23, 24, 27, 83, 175, 186, 205
- **FTTH** Fiber to the Home. 1–3, 5, 6, 45, 46, 222
- FWHM Full-Width at Half Maximum. 23, 24, 175, 204
- **G** Gap Width. 129, 135, 136, 138, 139, 142–145, 151, 154, 158, 160–165, 167, 169, 170, 220, 241, 242, 248, 249
- GDS Graphic Data System File. 49, 87, 105, 106, 115, 154, 219, 221
- GPIB General Purpose Interface Bus. 176
- GSG Ground-Signal-Ground. 123, 124, 220
- HDP High Density Plasma. 94
- HELIOS pHotonics ELectronics functional Integration on CMOS. 44, 49, 50, 85, 107, 124, 129, 135, 158, 165, 170, 217, 219–222
- **HPC** High-Performance Computing. 1
- HTO Hight Temperature Oxide. 93, 111, 118
- IC Integrated Circuit. 1, 6, 87, 93, 107, 109, 110, 113, 121, 219, 221
- IL Insertion Loss. 26, 27, 41–43, 45, 46, 50, 51, 55, 59, 61–66, 71, 81, 83–85, 171, 175, 177, 185–188, 195, 198, 203, 205, 209, 213, 216, 221
- ISE DESSIS TCAD 9.0 Integrated Systems Engineering. 50, 51, 139
- MMI Multimode Interferometer. xii, xvii, 20, 28, 29, 31, 32, 50, 76-83, 108, 112, 116
- MOS Metal-Oxide-Semiconductor. 17, 40
- MPA Multimode Propagation Analysis. 29
- MZI Mach-Zehnder Interferometer. 17, 20, 24, 25, 27, 28, 32, 40–43, 45, 46, 49–51, 76, 81–84, 88, 106, 110, 116, 183, 185, 191

- MZM Mach-Zehnder modulator. 6, 17, 26–28, 70, 82, 83, 87–89, 101, 102, 106–112, 123, 124, 132, 136, 141–143, 151, 174, 175, 178, 180, 182–184, 189, 191, 195, 200, 203–205, 208, 209, 213, 216, 217, 221
- NRZ Non-Return-to-Zero. 178, 188
- **ONU** Optical network unit. 45, 46
- PIC Photonic Integrated Circuit. 3–5, 222, 223
- **PON** Passive Optical Network. 45
- PRBS Pseudo-Random Binary Sequence. 176–178, 188, 198, 208
- **Q** Quality Factor. 23, 24, 204
- QCSE Quantum Confined Stark effect. 10, 14
- **QPSK** Quadrature Phase-Shift Keying. 223
- **RF** Radio Frequency. 5, 6, 26, 36, 37, 40, 42, 43, 49, 71, 72, 85, 91, 110, 123, 124, 126, 128, 134, 136, 142, 143, 155, 160, 162, 169, 171, 172, 178–180, 182, 188, 194, 198–200, 208, 211, 220, 244
- **RR** Ring Resonator. 17, 20, 22–24, 28, 41–43, 49, 51, 76, 88, 90–92, 101, 123, 175, 179, 182, 204–217, 221
- **SEM** Scanning Electron Microscope. 102–104
- SILVER <u>SIL</u>icon based integrated optical transcei<u>VER</u> for local access networks. 44, 45, 49, 50, 53, 70–72, 76, 79–85, 87, 105, 106, 110, 115, 117, 124, 160, 165, 170, 182, 183, 217, 219–222
- SNR Signal-to-Noise Ratio. 88, 216
- **SOI** Silicon-On-Insulator. 11, 18, 71, 72, 93, 98, 117, 129, 131, 136–138, 142, 143, 155, 201, 241, 242
- TE Transverse Electric. 204–209, 211

Acronyms

TM Transverse Magnetic. 204, 205, 207–209, 211

ULTIMATE m<u>ULTI</u>level <u>M</u>odulation IntegrAted <u>TransmittEr</u>. 223

W Signal Line Width. 129, 135, 136, 139, 142–145, 151, 154, 158, 160–165, 167, 169, 170, 220, 241, 242, 248, 249

WDM Wavelength-Division Multiplexing. 223

WG Waveguide. 18, 45, 46, 50

Appendices

A Silicon Resistivity vs. Doping



Figure A.1: Resistivity for *p*-type (boron-doped) and *n*-type (phosphorous-doped) silicon at 300 *K* versus doping density.

For boron-doped Si, the boron density is related to the resistivity by [72], [73]

$$\rho = \frac{1.305 \times 10^{16}}{N_B} + \frac{1.133 \times 10^{17}}{N_B \left[\left(1 + 2.58 \times 10^{19} N_B \right)^{-0.737} \right]} \left[\Omega \cdot cm \right]$$
(A.1)

For phosphorous-doped Si, the phosphorous density is related to the resistivity by

$$\rho = \frac{6.242 \times 10^{18} 10^Z}{N_P} \left[\Omega \cdot cm \right]$$
(A.2)

where

$$Z = \frac{A_0 + A_1 y + A_2 y^2 + A_3 y^3}{1 + B_1 y + B_2 y^2 + B_3 y^3}$$
(A.3)

and

$$y = \log_{10} (N_P)$$

$$A_0 = 3.079$$

$$A_1 = 2.2108$$

$$A_2 = -0.62272$$

$$A_3 = 0.57501$$

$$B_1 = -0.68157$$

$$B_2 = 0.19833$$

$$B_3 = -0.0183769$$

B Substrate Influence: R_{Si} and C_{Si}

		G (μm)							
		5	10	15	20	25	30	35	40
	5	60.00	46.70	43.00	39.70	39.50	39.30	39.00	38.70
(uπ) M	10	37.60	34.70	31.50	29.30	28.80	28.60	28.80	29.10
	15	29.00	26.20	25.50	24.70	24.40	24.00	24.00	23.90
	20	23.60	21.50	21.20	21.00	21.50	21.30	21.50	21.60
	25	20.00	19.50	19.00	18.50	18.00	17.50	17.20	17.00
	30	18.00	16.70	16.90	17.10	17.25	17.40	17.70	18.00
	35	16.50	15.20	15.50	15.90	16.00	16.30	16.40	16.60
	40	15.00	14.50	14.70	14.90	15.20	15.40	15.60	15.90

Table B.1: R_{Si} as a function of G and W of an SOI high resistivity substrate with BOX layer thickness of 2 μ m. R_{Si} in given in M $\Omega \cdot \mu$ m

Appendix B. Substrate Influence: R_{Si} and C_{Si}

		G (μm)							
		5	10	15	20	25	30	35	40
	5	0.062	0.079	0.080	0.036	0.085	0.098	0.100	0.103
	10	0.119	0.118	0.140	0.160	0.165	0.170	0.170	0.171
(mη) W	15	0.165	0.180	0.210	0.225	0.228	0.232	0.238	0.240
	20	0.233	0.262	0.270	0.277	0.281	0.288	0.292	0.299
	25	0.300	0.340	0.330	0.340	0.350	0.360	0.370	0.370
	30	0.355	0.412	0.405	0.400	0.420	0.435	0.438	0.440
	35	0.482	0.477	0.467	0.445	0.455	0.481	0.500	0.512
	40	0.500	0.541	0.530	0.546	0.550	0.557	0.561	0.567

Table B.2: C_{Si} as a function of G and W of an SOI high resistivity substrate with BOX layer thickness of 2 μ m. C_{Si} in given in fF/ μ m

C Characterization of CPW Electrodes

C.1 ADS Model

The following equations were created in ADS to calculate relevant design parameters CPW electrodes that influence the propagation of the RF signal.

$n20 = \{1 :: 20\}$	(C.1)
$i20 = transpose\{n20\}$	(C.2)
<i>nOPT</i> = 3.1	(C.3)

$$Z_0 = 50$$
 (C.4)

 $A = \text{stoabcd}(S, 50) \tag{C.5}$

$$Z_D = \sqrt{\frac{A(1,2)}{A(2,1)}}$$
(C.6)

 $ph21 = -(unwrap(2, 1), \pi)$ (C.7)

$$n_{RF} = \frac{ph21 \cdot c}{2\pi \cdot L \cdot 1 \times 10^{-6} \cdot SP.freq}$$
(C.8)

 $U_0 = \{ \max(\operatorname{var}('Y1.X1.V_Cd')), ..., \max(\operatorname{var}('Y4.X5.V_Cd')) \}$ (C.9)

$$I = U_0 \cos \frac{2\pi \cdot \text{SP.freq}\left(\frac{i20 \cdot Li \cdot 1 \times 10^{-6}}{c/nOPT}\right) - \left(\frac{i20 \cdot Li \cdot 1 \times 10^{-6}}{c/nRF}\right)}{20 \cdot \text{mag}(V_DC[1])}$$
(C.10)

$$L_{dB} = 10 \cdot \log 10(I) \tag{C.11}$$

(C.12)



C.2 CPW on an SOI Substrate without Active Region

Figure C.1: Simulated and experimental S-parameters, L = 4 mm (electrodes alone).

Frequency (GHz)

b

D Appendix: CPW Electrodes Design



Figure D.1: Contour representation of the simulated (a) electrical BW and (b) electro-optical response of a modulator based on a PIPIN junction as a function of G and W of the CPW electrodes, L = 1 mm.



Figure D.2: Contour representation of the simulated (a) Z_D and (b) n_{RF} , of a modulator based on a PIPIN junction as a function of G and W of the CPW electrodes, L = 1 mm.

E Breakdown Voltage vs. Doping



Figure E.1: Avalanche breakdown voltage versus doping concentration for one-sided junctions in Ge, Si, and GaAs [42].



Figure E.2: Depletion region width x_d versus doping concentration at breakdown for one-sided abrupt junctions in Ge, Si, and GaAs [42].

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Book Chapter

 D. Marris-Morini, G. Rasigade, M. Ziebell, P. Chaisakul, J-M. Fédéli, G. Isella,
 D. Chrastina, and L. Vivien. "Modulateurs optiques pour la photonique silicium - ou la révolution des systèmes de communication optiques de demain". Techniques de l'ingénieur. Référence IN143, 2012

Résumé : La photonique silicium est un domaine de recherche en plein expansion, visant à intégrer la photonique avec la microélectronique. Cette technologie est envisagée pour l'obtention de solutions bascoûts dans le domaine des télécommunications optiques grâce à l'intégration de l'optique avec l'électronique de commande des composants. La réalisation d'un lien optique rapide pour transporter une information à haut débit nécessite le développement de blocs de base performants et compatibles avec la technologie CMOS.

Dans ce contexte, les travaux menés durant ma thèse ont porté plus spécifiquement sur le modulateur optique en silicium, pour les applications haut débit, de 10 à 40 Gbit/s. Les résultats présentés portent notamment sur la conception, l'optimisation, la fabrication et la caractérisation de ce composant. Le modulateur est basé sur un effet électroréfractif obtenu par déplétion des porteurs dans des structures de type PN et PIPIN pour obtenir une variation de phase de l'onde optique. La modulation d'intensité est ensuite obtenue par intégration de la région active dans un interféromètre Mach Zehnder ou un résonateur en anneau. Des simulations électriques, optiques et radiofréquences ont été menées sur les différents éléments du composant, afin de concevoir un dispositif pour les applications FTTH (Fiber To The Home), fonctionnant à 1,27 µm. Parmi les points clefs de ces études on peut noter l'optimisation des électrodes RF coplanaires réalisée grâce au développement d'un modèle électro-optique prenant en compte la propagation des ondes électriques et optiques dans la région active. Un procédé technologique compatible avec la filière CMOS a ensuite été mis en place et les masques nécessaires à la fabrication ont pu être dessinés en considérant les tolérances de fabrication et paramètres critiques. Enfin un grand nombre de résultats expérimentaux ont pu être obtenus, sur des composants concus à l'IEF et fabriqués au CEA-LETI. On peut notamment retenir un modulateur Mach Zehnder fonctionnant à 40 Gbit/s utilisant une diode « PIPIN » pour réaliser la variation d'indice, et présentant un taux d'extinction de 7.5 dB pour des pertes de seulement 6 dB. Les futures optimisations des modulateurs, silicium visent à les intégrer avec l'électronique de commande, et à aller vers des formats de modulation plus complexes mais aussi plus performants que la modulation d'intensité sur 2 niveaux considérée jusqu'à présent.

Mot clés : silicium, photonique, modulation, optoélectronique, circuits intégrés, télécommunications.

Abstract: Silicon photonics is a research field in full expansion that works towards the integration of photonics and microelectronic components in a single chip. The creation of a broadband optical link that is able to carry high-speed information requires the development of efficient building blocks compatible with CMOS technology.

The work carried out during my Ph.D. focused specifically on silicon optical modulators for highspeed applications from 10 to 40 Gbit/s. The work presented includes design, optimization, fabrication and characterization of the complete device. The modulator is based on the electro-refractive effect obtained by depletion of carriers in PN and PIPIN diodes to obtain a phase change of the optical mode. Intensity modulation is obtained by integration of the active region in a Mach-Zehnder interferometer or a ring resonator. Electrical, optical and radio frequency simulations were conducted on the various elements of the modulator to design a device for FTTH (Fiber To The Home) applications operating at 127 µm. Additional studies included the design of RF coplanar waveguide electrodes optimized through the development of an electro-optical model that takes into account the propagation of the electrical and optical waves in the active region. Compatible CMOS processes were proposed, and the necessary fabrication masks were designed considering fabrication tolerances and critical parameters. Finally, various experimental results were obtained on components designed at IEF and fabricated at CEA-LETI. We can specially mention a Mach Zehnder modulator operating at 40 Gbit/s that uses a PIPIN diode to obtain an index variation, and having an extinction ratio of 7.5 dB and losses of only 6 dB. The goal of future silicon modulators' generations is to integrate them with the RF driver, and to move towards more complex and efficient modulation formats than the two-level intensity modulation seen so far.

Keywords: silicon, photonics, modulation, optoelectronics, integrated circuits, telecommunications.

Transceiver optique en silicium pour les réseaux d'accès

