200-Gb/s waveform analysis of ultrafast all-optical semiconductor gates towards low-power consumption operation

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Abstract

This thesis is aimed at a fundamental understanding of all-optical signal processing gates based on semiconductor optical amplifiers (SOAs), for their low-power consumption operation in ultrafast frequency region. In particular, modeling of a frequency-dependent electric power consumption in the SOA and suppression of several kinds of output waveform degradations in the ultrafast wavelength conversion are discussed in detail.

After an introduction in Chapter 1, we present a power consumption model of an SOA gate in Chapter 2, to clarify the power consumption dependence on the operation frequency B and other fundamental parameters. The carrier density dynamics in the SOA was modeled using a rate equation including three different carrier-conversion efficiencies of the SOA, which we introduced as the dominant factors. The validity of this rate equation model was tested with measured characteristics of custom-designed SOA samples with different structures. Using the model, we predicted that the power consumption increase with B^2 in the high frequency limit. This B^2 -dependence was experimentally demonstrated for two different SOA samples in the frequency range of 20 ~ 100 GHz. We expect that this initial model will contribute to the designing of SOA power consumption or to the future outlook of SOA gates.

In Chapter 3, we discuss the generation of small sub-pulses from a delayed interference signal-wavelength converter (DISC). This sub-pulse generation is regarded as one potential issue which prevents the practical operation of the DISC. We experimentally revealed the sub-pulse generation in a relatively low frequency region $(10 \sim 25 \text{ GHz})$ through cross-correlation measurements, and verified our sub-pulse generation model with high accuracy $(1 \sim 2 \text{ dB})$. We demonstrated the increase of sub-pulse intensity up to -10 dB with the carrier recovery rate, which will lead to the trade-off between the sub-pulse intensity and the pattern-induced intensity fluctuation. The sub-pulse model verified in this work will be valuable for future solution of this issue.

In Chapter 4, we report on our wavelength conversion experiment for 200 Gb/s, 4992-bit data signal using a DISC gate. For this purpose we developed an original

low jitter, low repetition rate (40 MHz), ultrafast (600 \sim 700 fs) and synchronized sampling pulse generator for the cross-correlator input and succeeded in precise waveform-monitoring of the ultrafast long-pattern signal. The pattern-induced intensity fluctuation (PIF) of the output waveform was systematically measured, and the impact of the holding-beam injection and bandpass filter detuning on its suppression were compared. We systematically exhibited that the filter detuning efficiently suppressed PIF even for 200 Gb/s, long-pattern data signal and it saved the required holding-beam power by ~ 3 dB. We also discovered a new method using the nonlinear polarization rotation in the SOA, and obtained a guideline for use of this method thorough systematic measurements. These results will provide indispensable backgrounds for the practical, ultrafast, low-power operation of the DISC.

In conclusion, we developed the understanding of the SOA gate (DISC) operated in the frequency range of $10 \sim 200$ Gb/s, and considerably facilitated its low-power, small signal-degradation operation and the system design.

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List of acronyms used in this thesis

AL	Assist Light		
ASE	Amplified Spontaneous Emission		
BPF	Bandpass Filter		
cw	continuous wave		
DCF	Dispersion Compensation Fiber		
DDF	Dispersion Decreasing Fiber		
DFB	Distributed Feedback		
DGD	Differential Group Delay		
DISC	Delayed-Interference Signal-wavelength Converter		
DSF	Dispersion Shifted Fiber		
EA	Electro-Absorption		
EDFA	Erbium-Doped Fiber Amplifier		
EO	Electro-optic		
FP	Fabry-Perot		
GVD	Group Velocity Dispersion		
HNLF	Highly Nonlinear Fiber		
ISBT	Intersubband transition		
LD	Laser Diode		
LN	Lithium Niobate		
MLFL	Mode-Locked Fiber Laser		
MZI	Mach-Zehnder Interferometer		
MQW	Multiple Quantum Well		
NPR	Nonlinear Polarization Rotation		
OE	Opto-electronic		
OSA	Optical Spectrum Analyzer		
OTDM	Optical Time-Division Multiplexing		
PBS	Poralization Beam Splitter		
\mathbf{PC}	Polarization Controller		
PD	PhotoDiode		
PD-SMZ	Polarization-Discriminating Symmetric Mach Zehnder		
PIF	Pattern-induced Intensity Fluctuation		
PM-	Polarization Maintaining		

PPG	Programmable Pattern Generator		
PRBS	Pseudorandom Bit Sequence		
RF	Radio Frequency		
RZ	Return-to-Zero		
SCH	Separate Confinement Heterostructure		
SFG	Sum Frequency Generation		
SHG	Second Harmonic Generation		
SOA	Semiconductor Optical Amplifier		
SMF	Single-Mode Fiber		
SMZ	Symmetric Mach Zehnder (structure)		
SPM	Self Phase Modulation		
SSG	Small Signal Gain		
TE	Transeverse Electric		
TM	Transeverse Magnetic		
TOAD	Terahertz Optical Asymmetric Demultiplexer		
UNI	Ultrafast Nonlinear Interferometer		
VDL	Varriable Delay Line		
VOA	Varriable Optical Attenuator		
XGM	Cross Gain Modulation		
XPM	Cross Phase Modulation		
WDM	Wavelength-Division Multiplexing		

Chapter 1

Introduction

The all-optical signal processing has been drawing many researchers' interest for a long time, as a method to overcome the frequency limitation of conventional signal processors. While the operation frequencies of conventional, electric processors are limited by their gate capacitances and load resistances, the all-optical signal processing potentially works at as high speed as that of the optical nonlinear phenomena involved[1]. As the development of highly nonlinear devices reduced the optical signal power required to cause such nonlinear effects, the all-optical signal processing has become recognized as an energy-effective solution for high-speed data-communication systems.

The optical time-division multiplexing (OTDM) [2-4] network is an example of such solution, and supposed to be a cost-effective alternative for the wavelengthdivision multiplexing (WDM) technology. The latter, schematically shown in Fig. 1.1(a), has made considerable success in large-capacity data transmission, indeed. Many data signals with relatively low bit-rate were multiplexed in the wavelength space (together with polarization-multiplexing and differential quadrature phase shift keying), and the aggregated transmission rate reached 25.6 Tb/s[5]. It is not cost-effective, nevertheless, to build a complicated network including functions as routing or signal regeneration, based on the WDM technology. Thus we have aimed to introduce OTDM to the WDM-based network, as schematically shown in Fig. 1.1(b). The function block in the figure is drastically simplified owing to frequency-transparent all-optical signal processors, compared to that in the lowfrequency WDM network. In such OTDM-WDM network, many functions as signalwavelength conversion, 3R regeneration (re-amplification, re-shaping, re-timing), OTDM demultiplexing, header recognition, simple logic-operation, and buffering



Figure 1.1: Schematic drawings of signal processing (a): that in WDM networks, using conventional signal processors. (b): that in OTDM-WDM networks, using all-optical signal processors.

are desired to be realized by all-optical signal processors. Thus many kinds of alloptical gates for these functions have been intensively studied.

Among several kinds of all-optical signal processing, those using the cross-phase modulation (XPM) in the carrier-injected semiconductor optical amplifier (SOA)[6] have been the most practical candidate[7–9]. After the first demonstration of the Terahertz Optical Asymmetric Demultiplexer (TOAD) in 1993[10], many kinds of XPM-based SOA gates such as Symmetric Mach-Zehnder (SMZ, Fig. 1.2(a))[11, 12], Polarization-Discriminating Symmetric Mach-Zehnder (PD-SMZ, Fig. 1.2(b))[13, 14], or Delayed-Interference Signal-wavelength Converter (DISC, Fig. 1.2(c))[15] have been proposed and demonstrated. These interferometric SOA-gates are advantageous against optical gates based on Kerr-nonlinearity in an optical fiber in terms of control pulse energy, environmental stability, polarization sensitivity, and integrability[16–19]. They are also advantageous against XGM-based SOA gates in terms of chirp characteristics and signal extinction ratio. Research efforts mostly from commercial companies have been devoted to the high-frequency operation and practical application of these gates, and all-optical demultiplexing at 336 Gb/s with SMZ[20], 3R regeneration at 84 Gb/s with PD-SMZ[21, 22]^{*)}, and wavelength conversion at 160 Gb/s with DISC[23, 24] have been demonstrated. Even 320- to 640-GHz operation of all-optical SOA gates using ultrafast chirp dynamics has recently been demonstrated[25–27]. Among these gate functions, the wavelength conversion[28] has been one of the most developed and successful functions. SOA-based wavelength converters have been already used as indispensable components in more practical experiments[29, 30].

While many practical demonstrations have been successfully achieved, fundamental understanding of these SOA-gates has not been well advanced especially in the frequency range of over 100 Gb/s. Particularly, minimum requirements on the operating condition of these gates have been unknown. It has been difficult to answer how much electric power is essential for the gating operation at certain frequency. This question is practically important in realizing highly-integrated SOA based optical circuits in the future or to reduce the power consumption of the whole system. Also it has been practically recognized that many other factors as well as the electric power consumption affect the quality of the output signal from SOA gates. One common degradation of the signal is the pattern-induced intensity fluctuation (PIF), which results from imperfect recovery of the carrier depletion inside the SOA. Well-known solution for that is an injection of a strong cw light into the SOA (holding-beam effect, [31], or assist-light effect [32-39]). Also detuning of the bandpass filter (BPF) after the SOA [40–43], or similar and advanced approach [44], are known to be effective in the suppression of PIF. These techniques contributed to the demonstration of the past ultrafast gating for practically-short $(2^7 - 1)$ length data signal. As we found through the experiment, nonlinear polarization rotation (NPR) in the SOA can be also used for its suppression. Systematic evaluation and comparison of these effects in the frequency range of $\sim 200 \text{ Gb/s}$ (and long data pattern) has not been reported, however, to the author's knowledge. For other kinds of waveform degradation as sub-pulse generation, hardly anything has been studied in

^{*)}The assessment of the regeneration property of SOA delayed interference configuration (SOA-DI) at 160 Gb/s was also reported[24]. The structure of the SOA-DI is identical to DISC, whereas its operation condition principle was different in two senses: The delay time Δt in the MZI was almost equal to the bit spacing (6.25 ps). 2.1-ps clock signal was used instead of the cw light, which determined the output pulse width rather than the MZI.



Figure 1.2: Various structures of XPM-based SOA gates. (a): Symmetric Mach-Zehnder (SMZ) (b): Polarization-Discriminating Symmetric Mach-Zehnder (PD-SMZ), or Ultrafast Nonlinear Interferometer (UNI) (c): Delayed-Interference Signal-wavelength Converter (DISC), and (d): Terahertz Optical Asymmetric Demultiplexer (TOAD).

detail. Characteristics of these degradation must be clarified to design and actually operate the all-optical signal-processing system based on the SOA gates.

In this thesis, we discuss these issues both theoretically and experimentally. In Chapter 2, we explain our development of a practical model that scientifically describes the dc power consumption of an SOA-based all-optical gate as a function of its working frequency up to 160 GHz. In Chapter 3, we discuss our first observation of the sub-pulse generation out of the DISC gate, and its theoretical analysis. In Chapter 4, we discuss our results on waveform measurements of the DISC gate at 200 Gb/s, measured PIF, and several methods to suppress it. These research activities will contribute to the practical, low-power consumption, and ultrafast operation of the DISC and other SOA gates, and the realization of the future OTDM-WDM networks. In Chapter 5, we discuss the overall conclusion of this thesis and its impact to the other activities.

Chapter 2

Frequency-dependent electric power consumption model including quantum-conversion efficiencies in ultrafast SOA gates

2.1 Introduction

As mentioned in Chapter 1, all-optical signal-processors based on the XPM in SOAs are promising for future OTDM networks, for their capability of ultrafast signal processing and low power consumption. The SOA gates used in the past demonstration of ultra-high frequency demultiplexing, 3R regeneration, and wavelength conversion typically consumed 0.4- to 1.0-W electrical power, while the latest electrical demultiplexer consumes 5.5 W in 100-GHz operation[45]. Even though these SOA gates need additional power consumption for stabilizing the operating temperature $(0.1 \sim 1 \text{ W in the ideal cases (Fig. 2.1) and more depending on the system cooling})$ efficiency), they should be much advantageous compared to their electric counterparts. As the capabilities of SOAs for ultrafast gating have become clear, the lower limit of electric power consumption and its origin have become important design issues. A practical engineering model that describes dc power consumption has, however, not yet been reported to the best of the authors' knowledge. Because such a model has not been available, it has been very difficult for system researchers to design the total power consumption of the various large-scale 160-GHz OTDM systems now under research.

The electrical power consumption of an SOA in an all-optical gate is dominated by two factors: the amount of optical-pulse-induced modulation as the optical phase shift in the case of XPM, and the recovery time of the depleted carrier density after optical modulation. It has long been recognized that the so-called optical holding



Figure 2.1: Ideal power consumptions of several commercial thermoelectric coolers as functions of their cooling powers, calculated from specification data of each product.

beam effectively accelerates the carrier's recovery time [31–39], while the holding beam itself consumes a significant part of the injected carriers via the stimulated recombination process and consequently reduces the amount of optical phase shift in the co-propagating optical signal component. It has also been qualitatively well known that to accelerate the recovery time while maintaining the amount of optical phase shift in the 0.3- to-1.0 π range, the carrier injection rate (*i.e.*, the injection current density) must be increased, and this increases in the dc power consumption.

To discuss the power consumption, a detailed understanding of the quantum conversion efficiency from the number of injected carriers to the number of opticalpulse-induced stimulated recombinations in the SOA is needed. It is reasonably expected that an SOA with low conversion efficiency requires a large current injection for the gating operation. Furthermore, an increase in the injection current density might slightly degrade the quantum conversion efficiency. This quantum efficiency was not taken into account in most previous studies. Some researchers discussed SOA gain within continuous wave (cw) regime and carrier recovery using one conversion efficiency parameter[46, 47], but no one has studied how conversion efficiency is influenced by the SOA properties. Furthermore, no study of the conversion efficiency in ultrafast optical gating has been reported, even though this has turned out to be significantly lower than that for cw-light amplification through

present research.

In this chapter, we explain our development of a practical engineering model that scientifically describes the dc power consumption of an SOA-based all-optical gate as a function of its working frequency up to 160 GHz. A model of the holding beam's contribution, which is a part of our power-consumption model, is verified through a series of measured results. In these verification processes, measured parameter values are put into most of the independent model parameters, such as several types of quantum efficiencies. The observed dependence of the measured quantum efficiencies on the SOA structural parameters are also discussed.

2.2 Our model of carrier conversion efficiency

We define three different carrier conversion efficiencies, and use them to evaluate different carrier losses in the SOA. Figure 2.2 shows the loss model for injected carriers.

2.2.1 Definition of carrier conversion efficiency

We start from the total carrier number N and carrier density n in the active region. When n is below the population inversion threshold n_0 , the SOA has a negative gain and shows no band-filling nonlinearity. Thus, the injection current $I_{\rm op}$ should be larger than the transparent current I_0 (where $n = n_0$) and the number of carriers available for the gating should be reduced to the excess carrier number $N_{\rm ex}$, where $N_{\rm ex} = N - N_0$. For this process, we define the first conversion efficiency η_1 as $\eta_1 = N_{\rm ex}/N$.

Next, we introduce the carrier number $N_{\rm cw}$, which is the maximum number of stored carriers available for amplification of saturating cw light. This can also be understood as the number of carriers available for optical gating at the low frequency limit. $N_{\rm cw}$ is supposed to be smaller than $N_{\rm ex}$ defined above, because of the effects of ASE, Auger recombination, carrier overflow, intervalence band absorption, and free carrier absorption. The second conversion efficiency η_2 is defined as $\eta_2 = N_{\rm cw}/N_{\rm ex}$. In this paper, we will not discuss the individual loss contributions, because of the experimental difficulty of determining these.

Finally, we introduce the carrier number N_{pulse} , which is the maximum number of stored carriers available for amplification of ultrafast optical pulses. N_{pulse} is supposed to be smaller than N_{cw} , and the third efficiency η_3 is defined as $\eta_3 = N_{\text{pulse}}/N_{\text{cw}}$. Its potential causes are spectral hole burning and carrier heating[49– 53], or increases in ASE recombination caused by the higher average-carrier density compared with the cw-light amplification case, and so on. Note that this difference in SOA reactions to cw light and ultrafast pulses only appears well above the small signal regime, where practical gating operation is carried out. We consider N_{pulse} to have the largest influence on ultrafast gating performance. Therefore, we define the total carrier-conversion efficiency η_{T} for ultrafast gating as $\eta_{\text{T}} = N_{\text{pulse}}/N = \eta_1 \eta_2 \eta_3$. If the dominant factors of each efficiency can be revealed, we will be able to maximize the total efficiency by improving each contribution separately.



Figure 2.2: Loss model of injected carriers. η_1 to η_3 stand for conversion efficiencies.

2.2.2 Characterization of the conversion efficiency

We suppose that the carrier numbers and conversion efficiencies defined above can be characterized from fundamental SOA parameters as follows. When the carrier loss rate is $1/\tau_{\rm C}$ ($\tau_{\rm C}$ represents the carrier lifetime) and the total carrier number is N, the carrier loss $N/\tau_{\rm C}$ should balance the injection flow $I_{\rm op}/q$ (q: elemental charge). Thus, the total carrier number N can be estimated as

$$N = \frac{I_{\rm op}}{q} \tau_{\rm C} \tag{2.1}$$

 η_1 and the excess carrier number are supposed to be given by the transparency current I_0 as

$$\eta_1 = \frac{N_{\rm ex}}{N} = \frac{I_{\rm op} - I_0}{I_{\rm op}}$$
(2.2)

Here, we should mention that N and N_{ex} obtained from these relations contain certain underestimations when the dependence of the carrier loss rate on the injection current is large. We will discuss it in §2.3.

 $N_{\rm cw}$ is supposed to be obtained from the gain-saturation measurement of the SOA for cw light. As the intensity of the input light increases, the SOA chip gain G decreases from the small signal gain (SSG) G_0 . For an SOA with saturation power $P_{\rm cw}^{\rm sat}$ and $G_0 \gg 1$, G is suppressed by 3 dB when the output power reaches $P_{\rm 3dB} = P_{\rm cw}^{\rm sat} \cdot \ln 2[55]$. Taking the conventional discussion [54, 55] regarding the gain

Table 2.1: List of SOA samples and their structure					
		Active region	Active region	Active region	Confinement
Sample	Type	length,	width,	thickness,	factor,
		$L_{\rm eff}~(\mu{ m m})$	$w_{ m A}~(\mu{ m m})$	$d_{ m A}~(\mu{ m m})$	Γ
A#1	Bulk	1000	2	0.1	$0.1 \sim 0.2$
A#2	Bulk	700	2	0.1	$0.1 \sim 0.2$
A#3	Bulk	300	2	0.1	$0.1 \sim 0.2$
A#4	Bulk	700	2	0.1	$0.1 \sim 0.2$
B#1	MQW	1100^{*1}	1.25	0.038^{*2}	0.2
B#2	MQW	700^{*1}	1.25	0.038^{*2}	0.2
B#3	MQW	500^{*1}	1.25	0.038^{*2}	0.2
B#4	MQW	1100^{*1}	1.25	0.038^{*2}	0.2
C#1	Bulk	700	1.2	0.2	$0.3 \sim 0.4$
D#1	Bulk	1500	0.6	0.35	0.5

(*1) Effective length, considering 50 % contribution from taper regions

(*2) Sum of well thickness

saturation into account, we relate $N_{\rm cw}$ to $P_{\rm cw}^{\rm sat}$, $G_0^{\rm cw}$ (G₀ for the cw light) and $\tau_{\rm C}$ as

$$\eta_2 = \frac{N_{\rm cw}}{N_{\rm ex}}, N_{\rm cw} = \frac{P_{\rm cw}^{\rm sat} \tau_{\rm C}}{h\nu} \ln G_0^{\rm cw}.$$
(2.3)

Similarly, N_{pulse} is supposed to be obtained from the gain-saturation measurement of the SOA for ultrafast pulses. If the injected pulse train has a much longer interval than τ_{C} , the net gain of the pulse will be as follows[55]:

$$G^{\text{pulse}}(E_{\text{out}}) = \frac{E_{\text{out}}}{E_{\text{pulse}}^{\text{sat}}} \frac{1}{\ln\{\exp\left(E_{\text{out}}/E_{\text{pulse}}^{\text{sat}}\right) + G_0 - 1\} - \ln\left(G_0\right)}$$
(2.4)

The pulse saturation energy $E_{\text{pulse}}^{\text{sat}}$ can be obtained from the 2.35-dB suppression point, and we relate it and G_0^{pulse} (G_0 for the ultrafast pulse) to N_{pulse} as follows:

$$\eta_3 = \frac{N_{\text{pulse}}}{N_{\text{cw}}}, N_{\text{pulse}} = \frac{E_{\text{pulse}}^{\text{sat}}}{h\nu} \ln G_0^{\text{pulse}}.$$
(2.5)

The intrinsic difference between $N_{\rm cw}$ and $N_{\rm pulse}$ seems to be overlooked in the conventional gain-saturation theory. In the new rate equation model (see §2.4.1), both cw and pulse saturation are modeled as independent characteristics.



Figure 2.3: Measured ASE spectra of SOA samples at $I_{\rm op} = 200$ mA.

2.3 Conversion efficiency measurement and results

To evaluate the carrier conversion efficiency of actual SOAs, we prepared several custom-designed SOA chips and commercial modules from different manufacturers (A to D) and measured their fundamental parameters. Attributes of the SOA samples and examples of ASE spectra are shown in Table 2.1 and Fig. 2.3, respectively. Most notable classification of the SOAs is whether they have bulk structures or multiple quantum well (MQW) structures. While MQW structures have shown many characteristics advantageous for the LD application (low propagation loss, small threshold current, low α parameter, high modulation frequency), the superiority of MQW SOAs for the xpm-based gating application has not been demonstrated, to the author's knowledge. Strained, polarization-insensitive MQW SOAs have often been believed to be unsuitable, on the contrary, because their active layer thicknesses were limited below the critical thicknesses and large confinement factors cannot be achieved as a consequence. Therefore the direct comparison of bulk and MQW structures should be meaningful. For series A and B, we prepared chips with different (effective) active-region lengths L_{eff} . (The structural parameters were defined as schematically shown in Fig. 2.4.) The B#4 sample is the SOA used throughout the 200-Gb/s experiment in §4 and will provide good comparison with B#1 sample. For each custom-designed chip, except for the shortest one (A#3), more than two wires per electrode were bonded as shown in Fig. 2.5(a) to enable injection currents over 500 mA. We manufactured SOA-chip mounts with temperature controllers in view of the dimensions of supplied chip submounts (one example shown in Figs. 2.5(b)



Figure 2.4: Drawing of typical SOA-structures and definitions of dimensional parameters. (a): Cross-sectional view, Bulk active region. (b): MQW active region. (c): Top view, without taper and window regions. (d): Top view, with taper and window regions.

to (d)). Temperatures were stabilized at close to 25° C during the measurements. It was unfortunate that the cooling efficiencies of these mounts were not quite good, and the heat from the SOA chips sometimes caused thermal expansion to the whole mount. Therefore experiments with larger injection current than 400 ~ 500 mA were difficult due to rapid decreases in the chip-fiber coupling efficiencies.

The typical setup for SOA-chip characterization is shown in Fig. 2.6. We coupled lensed fibers to the input and output facets of the chip using precision stages (as shown in Figs. 2.7(a) and (b)), and obtained 2- to 5-dB coupling losses for each facet. To accurately control the polarization angle of the light in the later chapters, we used a polarization-maintaining dispersion-shifted (PM-DS) lensed fiber fabricated by Namiki Precision Jewel co. ltd. (Fig. 2.7(c)). The ASE intensities from both facets were continuously monitored by power meters, and through careful adjustment of the couplings fluctuations were kept below 0.2 dB.

First we measured the cw-gain using a DFB-LD ($\lambda_2 = 1548.5 \text{ nm}$) and an optical spectrum analyzer, from careful comparison between the spectral intensities of the input and output light. Figure 2.8 plots the measured SSG of each sample against The transparent currents I_0 were obtained from these results. Figure 2.9(a) $I_{\rm op}$. shows typical gain-saturation profiles. We obtained $P_{\rm cw}^{\rm sat}$ as a function of $I_{\rm op}$ from the 3-dB suppression points. As can be seen in Fig. 2.10(a), $P_{\rm cw}^{\rm sat}$ increased almost linearly with $I_{\rm op}$ without convergence. The SOA gains for ultrafast pulses were measured for each sample. To obtain ultrafast optical pulses with lower frequency than $1/\tau_{\rm C}$, we used a mode-locked fiber laser (MLFL: Pritel Inc., UOC-3) with an external LiNbO₃ modulator. Pulses with a 2-ps width, $\lambda_1 = 1555$ nm, and 10.5-GHz frequency from the MLFL were modulated down to a 0.65-GHz pulse train. We monitored the auto-correlation trace of the pulse train so that we could keep its extinction ratio above 20 dB. Typical gain-saturation profiles are shown in Fig. 2.9(b). We see that for each $I_{\rm op}$, the SSGs for the cw input are almost the same as for ultrafast pulses. We fitted the measured profiles to Eq. (2.4), and they showed good agreement up to $E_{\rm out} \sim 3000$ fJ. The difference at $E_{\rm out} > 3000$ fJ was probably caused by the slight residual of the 10.5-GHz pulses. From the fit results, we obtained $E_{\text{pulse}}^{\text{sat}}$ of each SOA as a function of I_{op} (Fig. 2.10(b)). In contrast to the case of $P_{\rm cw}^{\rm sat}$, they converged as $I_{\rm op}$ increased.

The carrier recovery rates $1/\tau_{\rm C}$ of each SOA at several injection currents were determined from XGM measurements using a cross-correlator (Femtochrome Research Inc., FR-103XR) and mode-locked pulses modulated down to 1.3 GHz. External variable delays and an elaborate calibration process were used to expand the scan



Figure 2.5: Photographs of the SOA samples and drawings of the SOA mounts. (a): Microscopic images of the custom-designed SOA samples. (b): The mount for A-series SOA chips, side view. (c): Top view. (d) Photograph.



Figure 2.6: Experimental setup for SOA chip characterization. VOA: variable optical attenuator, Pol: polarizer, PC: polarization controller, PBS: polarizing beam splitter, BPF: band-pass filter, MLFL: mode-locked fiber laser

range to 600 ps. The intensity of the probe cw light was typically set to -25 dBm to suppress the holding beam effect. Pulse energies into each SOA chip were adjusted within 10 ~ 500 fJ so that no gain modulation due to residual 10.5-GHz pulses appeared. Figure 2.9(c) shows examples of measured XGM waveforms. We can see two recovery processes after gain depletion. From the slower process, we obtained $1/\tau_{\rm C}$ (Fig. 2.10(c)). In one exception, we assumed $\tau_{\rm C} \sim 100$ ps for sample A#3 without fitting, since its small gain seriously degraded the S/N ratio of the output signal and prevented a precise determination of $\tau_{\rm C}$. From Fig. 2.10(c), we see that the SOA samples can be categorized into two types: the recovery rates of B#1, B#4 and D#1 increased rapidly to 30 GHz with $I_{\rm op}$, while those of the others increased slowly around 5 to 10 GHz.

Using the measured fundamental parameters, we calculated the conversion efficiencies (Fig. 2.11). The similarity of the results for the sample B#1 and those for B#4 indicates that our measurements were accomplished precisely. The efficiency values for each sample at $I_{\rm op} = 200$ mA were $\eta_1 = 0.68$ to 0.87, $\eta_2 = 0.17$ to 1.30, $\eta_3 = 0.29$ to 0.53, and $\eta_{\rm T} = 0.07$ to 0.40. These values indicate that total efficiency $\eta_{\rm T}$ strongly depends on the SOA structure. This dependence resulted mostly from η_2 . The B-series MQW samples had a much larger η_2 than the other bulk samples. Among the A-series, the longer chips had a larger η_2 , whereas the B-series showed the opposite η_2 dependence on $L_{\rm eff}$. The reason for this difference is unclear. η_1



(c)



Figure 2.7: Photographs and a drawing of the experimental setup, SOA samples and lenzed fibers. (a): Lensed fiber coupling stages and the SOA chip mount. (b): Microscopic view of the lensed fiber coupling. (c): Design of the PM-DSF lensed fiber.



Figure 2.8: Measured small-signal gains of SOA samples versus current. Sample details are given in Table 2.1. Gains were measured with cw light, $\lambda_2 = 1548.5$ nm. Input cw power into each chip was kept under -30 dBm.

depends on I_0 , and decreases with L_{eff} since I_0 is proportional to L_{eff} . η_3 was significantly smaller than unity. The results in Fig. 2.11(c) indicate that η_3 decreases as I_{op} goes above the optimum condition, and that longer SOAs tend to have larger η_3 values at large I_{op} . The total efficiency η_{T} of longer SOAs also tends to be higher at large I_{op} . Sample B#1 (and B#4) will be the most economical device for $I_{\text{op}} \sim 400$ mA ($\eta_{\text{T}} \sim 0.34$), while B#3 has the best efficiency for $I_{\text{op}} \sim 150$ mA ($\eta_{\text{T}} \sim 0.46$). We consider that the superiority of the MQW structure found in η_2 has a different origin from the well-known MQW superiority for LD application. The latter is typically attributed to the low internal loss ($\sim 5cm^{-1}$) or larger differential gain at low carrier density, but these will not explain the observed high efficiency at high carrier density. High cw-saturation power of MQW SOAs have been noticed by SOA researchers, but they do not provide convincing explanation since they do not precisely evaluate carrier recovery rate values.

There appears to be a problem with the definition of N_{ex} using Eqs. (2.1) and (2.2), since the calculated η_2 sometimes exceeds unity. This underestimation of N_{ex} will be reduced if we use another definition as

$$N_{\rm ex} = \int_{I_0}^{I_{\rm op}} \frac{\tau_{\rm C}(I)}{q} dI,$$
(2.6)

taking into account that the carrier lifetime $\tau_{\rm C}$, which is related to the differential carrier injection rate, decreases with the injection current. Determination of $N_{\rm ex}$



Figure 2.9: Typical SOA chip characteristics (a) Gain-saturation profiles for cw light. (b) Gain-saturation profiles for ultrafast pulses (2-ps width, 0.65-GHz repetition). The dashed lines in (b) show the theoretical fit using Eq. (2.4). (c) Typical XGM profiles measured with a cross correlator. The cw-probe intensity and pulse energy into the chip were set to -25 dBm and 10 fJ, respectively.



Figure 2.10: Measured SOA parameters used for evaluation of the conversion efficiencies (a) Saturation power of SOA chips versus current for cw light. (b) Saturation energy for ultrafast pulses (2-ps width, 0.65 GHz). (c) Carrier-recovery rate $1/\tau_c$.





Figure 2.11: Measured dependence of SOA sample conversion efficiency on $I_{\rm op}$ (a): η_1 , (b): η_2 , (c): η_3 , according to Eqs. (2.2) to (2.5), and (d): total efficiency $\eta_{\rm T} = \eta_1 \times \eta_2 \times \eta_3$.

using Eq. (2.6) is difficult, however, because it requires precise measurement of $\tau_{\rm C}$ at small $I_{\rm op}$. Then we think using Eqs. (2.1) and (2.2) is not a concern and appropriate since the idea is to get a rough understanding of the SOA-characteristics.

Thus, our interpretation of the dependence of the efficiencies on the SOA structure can be summarized as follows. First, the MQW structure is advantageous for increasing η_2 . A longer active region decreases η_1 , increases or decreases η_2 , and increases both η_3 and total efficiency at large $I_{\rm op}$. We could not obtain conclusive results for width, thickness or confinement factor, though some of these factors seemed to significantly affect efficiencies. Although these results are useful, we will need to perform more systematic characterizations before we can completely understand the SOA-structure effects.

2.4 Estimation of SOA electrical power consumption of all-optical gating

Here we propose a model for determining the electrical power requirements of SOAs used in ultrafast all-optical gates. Our first target is to understand the dependence of the power requirement on the operation frequency. The other target is the power consumption estimation based on the SOA fundamental parameters. First, we discuss the carrier dynamics of SOAs. Then, we estimate the power consumption from the calculated carrier dynamics. Finally, we compare the calculated power consumption with measured results.

2.4.1 Rate equation model including conversion efficiencies

In our preceding analyses [9, 56, 57], we used a single rate equation to calculate the carrier-density modulation in one SOA module. This is a simple but powerful tool to explain the gating characteristics in the time domain, as we will see in Chapter 3. The three kinds of carrier conversion efficiencies had not been taken into account, however, in the previous model. To account for the effect of the three different conversion efficiencies, we expanded the rate equation to the following form:

$$\frac{dn_{\text{pulse}}}{dt} = \frac{I_{\text{op}}}{qV} \eta_1 \eta_2 \eta_3 - \frac{n_{\text{pulse}}}{\tau_{\text{C}}} - \eta_3 \left\{ \exp\left(\Gamma L_{\text{eff}} \frac{dg^{\text{cw}}}{dn} n_{\text{pulse}}\right) - 1 \right\} \frac{P_{\text{cw}}}{h\nu V} - \left\{ \exp\left(\Gamma L_{\text{eff}} \frac{dg^{\text{pulse}}}{dn} n_{\text{pulse}}\right) - 1 \right\} \frac{P_{\text{pulse}}}{h\nu V} \quad (2.7)$$

 $(n_{\text{pulse}} = N_{\text{pulse}}/V$: carrier density available for ultrafast gating, P_{pulse} and P_{cw} : input light intensities, L_{eff} and $V = L_{\text{eff}} w_{\text{A}} d_{\text{A}}$: active region effective length and volume, Γ : confinement factor, dg/dn: differential gain). Note that η_3 is multiplied by the cw-light term, in view of the difference between n_{pulse} and n_{cw} (the carrier density available for cw light). Both cw gain and pulse gain were supposed to be determined by n_{pulse} , since we did not observe much difference between them unless gain saturation occurred (Fig. 2.9). We also supposed there are reserved carriers $(n_{\text{reserve}} = n_{\text{cw}} - n_{\text{pulse}})$ which do not explicitly affect gain, and $(1-\eta_3)$ of the photons are converted from reserved carriers through intraband interactions in the case of cwlight amplification. This rate-equation model can explain the experimental results regarding the gain saturation for ultrafast pulses quite well, and those regarding

Table 2.2 :	able 2.2: Estimated refractive index change of each sample, and related parameters.				
	Refractive index,	Carrier density for	Differential gain,	Confinement	
Sample	change $dn_{\rm r}/dn_{\rm pulse}$	gating, n_{pulse}	$dg_{ m cw}/dn_{ m pulse}$	factor,	
	(10^{-20} cm^3)	$(10^{18} \text{ cm}^{-3})$ (*2)	(10^{-17} cm^2) (*2)	Γ (*3)	
A#1	3.8	0.17	142	0.2	
A # 2	3.8(*1)	0.11	196	0.2	
A#3	3.8(*1)	0.15	136	0.2	
B#1	1.4	0.43	76	0.2	
B#2	1.2	1.0	42	0.2	
B#3	0.7	2.0	30	0.2	
(***	1 0 1 1	C 1 A // 4			

(*1) Estimated from the value for sample A#1.

(*2) Typical values measured at $I_{\rm op} = 200$ mA are shown.

(*3) Assumed in view of the values in Table 2.1.

the cw-gain saturation up to 3-dB suppression points (Fig. 2.12(a)). For stronger cw light, other theories may provide better predictions of the gain suppression[58]. We also mention that the sub-pulse intensities calculated using this model showed less agreement with measured results, than those described in Chapter 3. Thus we limit the use of this model within this chapter, until more comprehensive model is devised.

The rate equation can be used to calculate the optical carrier modulation and nonlinear phase shift. Injection of P_{pulse} causes carrier recombination Δn_{pulse} , and the nonlinear phase shift is given by

$$\Delta \Phi = -\frac{\alpha}{2} \Gamma L_{\text{eff}} \frac{dg}{dn} \Delta n_{\text{pulse}} = k_0 dn_{\text{r}} / dn \Gamma L_{\text{eff}} \Delta n_{\text{pulse}}$$
(2.8)

(α : α parameter, k_0 : wave number in vacuum, n_r : refractive index).

To verify the model, we simulated the SOA characteristics under holding-beam injection. We used the measured values of η and $\tau_{\rm C}$ at each $I_{\rm op}$, and extracted $dg_{\rm cw}/dn$ and $dg_{\rm pulse}/dn$ from the corresponding SSG values. Figure 2.12(b) shows measured and calculated results for the effective carrier recovery rate of sample B#3 accelerated by the holding-beam. Figure 2.12(c) shows measured and calculated results for pulse-gain saturation with and without a holding beam. We see that both results approximately agree with each other.

We also compared measured and calculated nonlinear phase shifts for samples A#1 and B#1 to #3. These can be experimentally determined from XPM spectra out of the SOA[9]. The chosen pulse frequency was either 2.6 GHz or 10.5 GHz,



Figure 2.12: Comparison of measured and calculated SOA properties under holdingbeam injection. (a) Cw-gain saturation, (b) effective carrier recovery rate $1/\tau_{\text{eff}}$, (c) gain saturation for a 2-ps pulse, and (d) nonlinear phase shift $\Delta\Phi$ caused by ultrafast pulses.

depending on the carrier recovery rate of the SOA. dn_r/dn was estimated for each sample from the measured phase shifts at $I_{\rm op} = 200$ mA (Table 2.2). The difference between samples seemed to be negatively correlated to the typical carrier densities. The SOA simulation could approximately reproduce the measured nonlinear phase shifts at various $I_{\rm op}$, $P_{\rm cw}$ and $P_{\rm pulse}$ values (Fig. 2.12(d)), and our model seems feasible for power-consumption simulations.

2.4.2 Electrical power consumption

We calculated the correspondence between the electrical power consumption $P_{\rm op}$ and the operating frequency B for an SOA used in an XPM-based all-optical gate as follows. First, we fixed the control pulse energy $E_{\rm pulse} \equiv \int_{-\infty}^{\infty} P_{\rm pulse}(t) dt$ at a realistic value (340 fJ). Then, for each $I_{\rm op}$ condition we calculated the induced $\Delta \Phi$ as a function of the holding beam intensity. The required amount of $\Delta \Phi$ depends on the function of the gate: for 3R regeneration or wavelength conversion it is ~ 0.3 π [56]. Therefore, for each $I_{\rm op}$, the equation $\Delta \Phi = 0.3\pi$ provides the maximum available holding-beam intensity $P_{\rm cw}$. The carrier recovery rate $1/\tau_{\rm eff}$ for this condition can be interpreted as the approximate frequency limit B of the gate. The actual B for a specific gate may depend also on other factors, as BPF detuning in Chapter 4, but we do not discuss such detail in this chapter. $P_{\rm op}$ was calculated from $I_{\rm op}$ using empirical V-I characteristics.

Before discussing the power consumption of actual SOAs, we will look at calculated values using non-empirical parameter sets to find their dominant factors. Figure 2.13 shows our results. Here, the dependences of dg/dn, η and $\tau_{\rm C}$ on $I_{\rm op}$ were ignored, in contrast to the cases shown in Figs. 2.12 or 2.14. Calculations were done down to nearly the low power limits, where $\Delta \Phi = 0.3\pi$ can be satisfied only when $P_{\rm cw} = 0$. We found that the maximum frequency is almost proportional to the total conversion efficiency $\eta_{\rm T}$ and dg/dn, but not sensitive to the intrinsic carrier recovery rate $1/\tau_{\rm C}$. As for the conversion efficiencies of each process, η_3 affects the required holding-beam intensity as well as *B*. As long as we keep the SOA parameters constant, $I_{\rm op}$ increases almost linearly with *B*, so $P_{\rm op}$ increases with B^2 .

Figure 2.14 shows the calculated power consumption for the SOA samples. We used the SOA parameters from §2.4.2. The B-series shows better performance than the A-series. When $P_{\rm op}$ is small, B#3 has the best performance around 20 GHz. As the target frequency increases to over 100 GHz, B#1 becomes preferable. The power consumption of B#1 at 160 GHz is about 750 mW. This is comparable to the



Figure 2.13: Calculated power consumption of SOAs versus their maximum operating frequencies when SOA parameters are independent of I_{op} (Table 2.3).

Table 2.5. Tarameters assumed for Fig. 2.15				
Symbol	Description	Value		
$L_{\rm eff}$	active region length	$1000 \ \mu \mathrm{m}$		
$w_{ m A}$	active region width	$1~\mu{ m m}$		
d_{A}	active region thickness	$0.05~\mu{ m m}$		
Γ	confinement factor	0.2		
$dg_{ m cw}/dn$	differential gain for cw	$20.0 \times 10^{-17} \text{ cm}^2$		
$dg_{ m pulse}/dn$	differential gain for pulse	$20.0 \times 10^{-17} \text{ cm}^2$		
$ au_{ m c}$	carrier recovery rate	$100 \mathrm{\ ps}$		
	without acceleration			
$dr_{\rm r}/dn_{\rm pulse}$	refractive index change	$0.4\times10^{-20}~{\rm cm}^3$		
$\eta_{ m T}$	total carrier-conversion efficiency	1.0		

Table 2.3: Parameters assumed for Fig. 2.13
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Figure 2.14: SOA electrical-power consumption versus carrier recovery rate under holding-beam acceleration. (a): Calculated results using the measured parameters for each sample. (b): Measured results for sample B#3 and B#4. A control pulse with a 2-ps width and 340-fJ energy was used.

power consumption, about 400 mW, of the latest XPM-based 160-Gb/s wavelength conversion[23] or wavelength conversion using ultrafast chirp dynamics[25]. For the A-series, the longer sample had a higher frequency, and the shortest sample A#3 could not achieve $\Delta \Phi = 0.3\pi$ under the injection-current limit. These characteristics result from those of $\eta_{\rm T}$ as discussed in §2.3. Except for the difference in scale, most of the $P_{\rm op}$ -B profiles in Fig. 2.14 resemble the corresponding $P_{\rm op}$ - $1/\tau_{\rm C}$ profiles. This is due to the complicated correlations of SOA parameter changes. Consequently, $P_{\rm op}$ of A#1 and B#1 increase with B^2 within the range of about 300 to 1000 mW, and the corresponding values of the others likely increase faster than that. Further discussion requires measurement of the efficiencies at larger $P_{\rm op}$.

To test the validity of our power-consumption model, we measured $P_{\rm op}$ and $1/\tau_{\rm eff}$ at $\Delta \Phi = 0.3\pi$ for sample B#3 (Fig. 2.14). We obtained good agreement between the measured and calculated results within a frequency range of 20 ~ 40 GHz. We also tried to measure that for sample B#4 and compared with the calculated results for B#1. In this case, we could only reduce $\Delta \Phi$ down to ~ 0.38π even with our strongest holding beam (+11 dBm into chip). Thus the each plot for this series indicates only lower limit of the bandwidth at each power consumption. For this SOA, somewhat larger discrepancy appeared between measured and calculated results. Thus we consider that our model has to be improved to provide absolute value estimation in the frequency range around and over 100 GHz. As an initial power-consumption model, nevertheless, we believe that our model is valuable enough. Especially, the B^2 law has become more convincing due to the measured results.

2.4.3 High-frequency limit

To discuss the future of photonic networks, it is useful to express the B- $P_{\rm op}$ relation with a simple formula. In this section, we therefore derive an analytical expression for the high-frequency limit.

First, we assume that both the injection current $I_{\rm op}$ and holding-beam power $P_{\rm cw}$ are large. $P_{\rm cw}$ is chosen according to $I_{\rm op}$ so that $\Delta\Phi$ stays at a fixed value, as in §2.4.2. Without control pulses the carrier density reaches an equilibrium at $n_{\rm pulse} = n_{\rm eq}$. The small difference in $dg_{\rm cw}/dn$ and $dg_{\rm pulse}/dn$ is ignored for simplicity, and the saturated SOA gain is expressed as

$$G_{\rm eq} = \exp\left(\Gamma L_{\rm eff} \frac{dg}{dn} n_{\rm eq}\right).$$
(2.9)

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Note that we can approximately suppose that G_{eq} does not change with I_{op} , because of the constraint on $\Delta \Phi$. The reason is as follows: The number of carrier-to-photon conversions when a control pulse is injected into the SOA in the steady state is given from the net gain of the pulse G_{pe} and E_{pulse} :

$$\Delta n_{\rm pulse} \approx \frac{G_{\rm pe} - 1}{h\nu V} E_{\rm pulse} \tag{2.10}$$

as long as the pulse width is short enough compared to the carrier relaxation time. Then, $G_{\rm pe}$ can be roughly estimated from $G_{\rm eq}$. Since $\Delta \Phi$ is given from $\Delta n_{\rm pulse}$ by Eq. (2.8), we can transform Eq. (2.10) into

$$G_{\rm eq} \approx G_{\rm pe} \approx 1 + \frac{h\nu V\Delta\Phi}{E_{\rm pulse}\Gamma L_{\rm eff}k_0 dn_{\rm r}/dn}.$$
 (2.11)

^{*)} Both $\Delta \Phi$ and $E_{\rm pulse}$ are regarded as constant when we obtain the B- $P_{\rm op}$ relation. We also assume that $dn_{\rm r}/dn$ does not depend on $I_{\rm op}$. Therefore, we can approximately regard $G_{\rm eq}$ as constant. For example, $\Delta \Phi = 0.3\pi$ and $E_{\rm pulse} = 340$ fJ result in $G_{\rm pe} = 2.5$ for sample B#1.

By assuming a steady state and taking only terms including I_{op} or P_{cw} in Eq. (2.7), we obtain the following relation from Eqs. (2.7) and (2.9):

$$\frac{P_{\rm cw}}{h\nu V} \approx \frac{\eta_{\rm T} I_{\rm op}}{\eta_3 \left(G_{\rm eq} - 1\right) q V} \tag{2.12}$$

The operating frequency limit can then be obtained as follows: After first-order expansion of Eq. (2.7) in terms of n_{pulse} around n_{eq} , the relaxation of the carrier density deviation $\Delta n(t) = n_{\text{pulse}}(t) - n_{\text{eq}}$ after the control-pulse injection can be expressed as

$$\frac{d\left(\Delta n\left(t\right)\right)}{dt} = -\frac{\Delta n}{\tau_{\rm C}} - \eta_3 \frac{G_{\rm eq} \Gamma L_{\rm eff}}{h\nu V} \frac{dg}{dn} P_{\rm cw} \Delta n$$
$$= -B \cdot \Delta n \tag{2.13}$$

After substituting Eq. (2.12) into the above, we obtain

$$B \approx \frac{1}{\tau_{\rm C} \left(I_{\rm op}\right)} + \frac{G_{\rm eq}}{G_{\rm eq} - 1} \frac{\Gamma L_{\rm eff}}{qV} \frac{dg}{dn} \left(I_{\rm op}\right) \cdot \eta_{\rm T} \left(I_{\rm op}\right) \cdot I_{\rm op}.$$
(2.14)

^{*)}Note that this is a fairly rough approximation.

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When $1/\tau_{\rm C}$ converges to a finite value as $I_{\rm op}$ increases, the second term will dominate. The applied voltage $V_{\rm op}$ approaches $RI_{\rm op}$ (R: resistance). Then, when we assume both dg/dn and $\eta_{\rm T}$ are independent of $I_{\rm op}$, $P_{\rm op}$ is expressed as

$$P_{\rm op} \approx R \left(\frac{G_{\rm eq} - 1}{G_{\rm eq}} \frac{qV}{\Gamma L_{\rm eff}}\right)^2 \left(\frac{dg}{dn}\right)^{-2} \eta_{\rm T}^{-2} B^2$$

= constant × B^2 . (2.15)

Therefore, the B- $P_{\rm op}$ relation will be quadratic under the above assumption. This explains the Fig. 2.13 results. The total conversion efficiency has an inverse-square contribution to $P_{\rm op}$ for the actual SOA, so it should be elaborately optimized to reduce the dc power consumption.

In a limiting case when we can use arbitrary-high energy control pulses, we can obtain even simplified expression of the operation frequency limit. In this case, Δn_{pulse} is equal to n_{eq} . Using this relation and Eq. (2.8), we obtain

$$n_{\rm eq} = \frac{2\Delta\Phi}{\alpha\Gamma L_{\rm eff}\frac{dg}{dn_{\rm pulse}}}.$$
(2.16)

From Eq. (2.9) and Eq. (2.16), we obtain

$$G_{\rm eq} = \exp\frac{2\Delta\Phi}{\alpha} \tag{2.17}$$

instead of Eq. (2.11). The derivation of Eq. (2.15) is still valid, and after substituting Eq. (2.17) into Eq. (2.15) we obtain

$$P_{\rm op} \approx \frac{B^2}{M},$$

$$M \equiv R^{-1} \left(\left(1 - \exp\left(-\frac{2\Delta\Phi}{\alpha}\right) \right) q \frac{d_{\rm A}w_{\rm A}}{\Gamma} \right)^{-2} \left(\frac{dg}{dn}\right)^2 \eta_{\rm T}^2$$
(2.18)

The factor M in this equation depends only on the SOA parameters (except for $\Delta \Phi = 0.3\pi$), and we can regard this quantity as a figure of merit of the SOA in terms of the power consumption. Table 2.4 shows M values for the measured SOA samples. For simplicity, the dependence of each parameter values on the injection current was ignored. This figure can be used for crude comparison of SOA samples, though it does not properly reflect decrease in the efficiency at large injection current.

Table 2.4: Figure of merit values of the SOA samples						
	α parameter,	Total	Differential gain,	Figure of		
Sample	α	conversion efficiency	$dg_{ m cw}/dn_{ m pulse}$	Merit, M		
		$\eta_{ m T}$	(10^{-17} cm^2)	$(10^2 \text{ GHz}^2/\text{mW})$		
A#1	2.2	0.15	142	0.29		
A # 2	1.5	0.11	196	0.15		
A#3	2.3	0.084	136	0.03		
B#1	1.5	0.34	76	2.43		
B#2	2.3	0.41	42	1.88		
B#3	1.9	0.34	30	0.45		

2.5 Conclusion

We developed a new model of the electrical power consumption of SOAs used in ultrafast all-optical gates. The dominant factors of this model are three different carrier-conversion efficiencies and the differential gain of the SOA.

To predict power consumption, we measured the conversion efficiencies of current SOAs with our original method. They were $\eta_{\rm T} = 0.07$ to 0.46 in total. A systematic study of the efficiencies of SOAs with different structures revealed that a longer SOA with an MQW structure will have high conversion efficiency at a high injection current. For high-frequency operation above 100 GHz, we consider that the SOA effective length of more than 1100 μ m will be preferable. Optimum SOA-length will exist depending on the target frequency, which should be clarified by further study. Also further study is needed with this method to understand the dependences on the SOA width, thickness, and confinement factor, and through such studies we hope to design a more efficient SOA in the near future.

We calculated the power consumption $P_{\rm op}$ as a function of operating frequency limit *B*, based on our power consumption model and measured SOA parameters. We predicted that the dc power consumption of SOA gates $P_{\rm op}$ increases with their operation frequency *B* as B^2 , or even faster, depending on the decrease in the conversion efficiencies and differential gain. Measured results for two SOA samples on the power consumption showed clear B^2 dependence within the frequency range of 20 ~ 100 GHz. Our first power-consumption model will be a fundamental for developing the power-consumption models in in the frequency range above 100 GHz.

Chapter 3

Observation of sub-pulse generation from the DISC-type wavelength converter

3.1 Introduction

As we introduced in Chapter 1, several types of all-optical SOA gates have been intensively studied for use in future OTDM-WDM networks and systems. Notably, DISC and its variants have led to some of the most successful achievements in the last decade. 168 Gb/s error-free wavelength conversion [23] and 40-Gb/s error-free ultralong-distance 3R transmission [29] have been demonstrated using DISCs in their original configuration, and a 320 Gb/s wavelength conversion has also been demonstrated using transient cross-phase modulation in the SOA[26].

It has been noticed, however, that the eye diagram of the DISC output signal often suffers from serious degradation, which could not be ascribed to the well-known pattern-induced intensity fluctuation (PIF). Some modeling research studies[57] have been focused on explaining this phenomenon, and one fundamental issue has been recognized as the potential source: the generation of small sub-pulses between RZ-formatted output pulses.

In this chapter, we explain our first observation of sub-pulse generation through a gate experiment and its theoretical analysis. In addition, we discuss measured results on the theoretically predicted increase in the sub-pulse intensity with the carrier recovery rate of the SOA. From the measured and calculated results, we conclude that our model is applicable for use in a wide recovery rate range. Finally, we discuss the trade-off between the sub-pulses and PIF associated with the recovery rate change in a wide range. This result illustrates that sub-pulse generation can be a significant limiting factor for practical DISC operations.

3.2 Sub-pulse generation model

Two phenomena have been suspected as potential sources of sub-pulse generation. One is the exponential recovery profile of the SOA carrier, and the other is the ultrafast SOA response due to the carrier-heating effect. In this section, we explain the sub-pulse generation of the first kind using our DISC-gate model. Generation of the latter sub-pulses is beyond the scope of this research.

3.2.1 DISC-gate Operation model based on the SOA Carrier Dynamics

A DISC gate in its original concept consists of a single SOA as an optical nonlinear element, and a Mach-Zehnder Interferometer (MZI), which determines the output timing window (Fig. 3.1(a)). The input pulse train with a wavelength of λ_1 is combined with continuous-wave (cw) light (λ_2) and injected into the SOA. The alloptical modulation process can be modeled based on the rate equation of the excess carrier density in the SOA as

$$\frac{d}{dt}\overline{n_{\rm c}(t)} = \frac{\eta_{\rm T}I_{\rm op}}{qV} - \frac{\overline{n_{\rm c}(t)}}{\tau_{\rm C}} - \frac{1}{V}\left(G(\overline{n_{\rm c}(t)}) - 1\right)\left[\frac{|\mathbf{E}_{\rm pulse}(t)|^2}{h\nu_{\rm pulse}} + \frac{|\mathbf{E}_{\rm cw}|^2}{h\nu_{\rm cw}}\right].$$
 (3.1)

Here, $\overline{n_{\rm c}(t)}$ is the average $n_{\rm c}(z,t)$ over the active region length. $n_{\rm c}(z,t)$, the carrier density which contributes to the SOA gain, is diminished from the total carrier density, as represented by the quantum conversion efficiency $\eta_{\rm T}$. $I_{\rm op}$ is the injection current, $\tau_{\rm C}$ is the carrier lifetime, V is the active region volume, and $\mathbf{E}_{\rm pulse}(t)$ and $\mathbf{E}_{\rm cw}$ are the amplitudes of the input optical signals. This model is almost same as that used in the past analyses[9, 56, 57], except for one point that the total carrier conversion efficiency is explicitly included. Meanwhile, the three different conversion efficiencies which we introduced in §2.2.1 is not included at the moment. The first, second, and third terms on the right-hand side of eq. (3.1) represent the carrier injection, carrier relaxation, and stimulated recombinations, respectively. $\mathbf{E}_{\rm pulse}(t)$ for a clock operation is expressed as

$$\mathbf{E}_{\text{pulse}}(t) = \sum_{m=1}^{\infty} \sqrt{\frac{E_{\text{pulse}}}{2T_{\text{W}}}} \operatorname{sech}\left(\frac{t-mT}{T_{\text{W}}}\right),\tag{3.2}$$



Figure 3.1: The schematic and operation principle of the DISC gate: (a): Schematic of the gate. (b): Calculated example of the time profile of the SOA carrier density during optical modulation. (c): Time profiles of the nonlinear phase shifts for the split optical components in the MZI, and the phase difference between them.

where E_{pulse} is the pulse energy and T_{W} is a parameter related to the full-width at half-maximum pulse width T_{FWHM} as

$$T_{\rm W} = T_{\rm FWHM} / \left(2\ln\left(1 + \sqrt{2}\right) \right). \tag{3.3}$$

G, the chip gain of the SOA, is expressed as

$$G(\overline{n_{\rm c}(t)}) = \exp\left(\Gamma L_{\rm eff} \frac{dg}{dn} \overline{n_{\rm c}(t)}\right),\tag{3.4}$$

where Γ is the confinement factor, L_{eff} is the effective length of the active region, and $\frac{dg}{dn}$ is the differential gain of the active region. $\overline{n_{\text{c}}(t)}$ decreases through stimulated recombinations as each control pulse incidents on the SOA (Fig. 3.1(b)), and this decrease causes an ultrafast change in the refractive index n_{r} of the SOA through the band-filling effect. Then the co-propagating cw signal suffers nonlinear phase modulation expressed as

$$\Delta \Phi(t) = k_0 \Gamma L_{\text{eff}} \frac{dn_{\text{r}}}{dn} \Delta \overline{n_{\text{c}}(t)}.$$
(3.5)

Here, $k_0 = 2\pi \frac{\nu_{\rm cw}}{c}$ is the wave number of the light in vacuum. $\Delta \Phi(t)$ increases within less than one picosecond after each pulse injection, but recovers slowly according to the effective carrier relaxation time $\tau_{\rm eff}$. The delayed interference scheme has been adopted to inhibit this slow response as follows: The gain-modulated and phasemodulated cw light is split into two components at the MZI, which are combined after suffering a time difference Δt and a phase difference (phase bias) $\Delta \Phi_{\rm B}$ at the end of the MZI. Then the amplitude of the combined signal is expressed as

$$\mathbf{E}_{\text{out}}(t) = \frac{1}{2} \{ \sqrt{G(t)} \exp\left[i \left(\Delta \Phi(t) + \Delta \Phi_{\text{B}}\right)\right] + \sqrt{G(t - \Delta t)} \exp\left[i \Delta \Phi(t - \Delta t)\right] \} \mathbf{E}_{\text{cw}},$$
(3.6)

and the output signal intensity $P_{out}(t)$ is expressed as

$$P_{\rm out}(t) = |\mathbf{E}_{\rm out}(t)|^2 = \left[\sqrt{G(t)G(t-\Delta t)}\cos^2(\delta\Delta\Phi(t)/2) + \frac{1}{4}\delta G(t)^2\right] |\mathbf{E}_{\rm cw}|^2$$
(3.7)

$$\delta \Delta \Phi(t) \equiv \Delta \Phi(t) - \Delta \Phi(t - \Delta t) + \Delta \Phi_{\rm B}$$
(3.8)

$$\delta G(t) \equiv \sqrt{G(t)} - \sqrt{G(t - \Delta t)} \tag{3.9}$$

Thus, the output signal intensity is dominated by the phase difference $\delta\Delta\Phi(t)$. It is largely shifted from $\Delta\Phi_{\rm B}$ only within $t_0 < t < t_0 + \Delta t$, where t_0 is the pulse incident time, and the output gate window is shortened to Δt (Fig. 3.1(c)). Since the input data pulses at λ_1 do not pass the band-pass filter (BPF), we obtain wavelengthconverted signals from the DISC gate.

3.2.2 Sub-pulse generation model

There remains a problem with this scheme, however, since the recovery of $\Delta \Phi(t)$ is not linear-function-like but exponential with t expressed as $\exp(-t/\tau_{\text{eff}})$. Figures 3.2(a) to 3.2(d) show exaggerated examples of the phase modulation profiles and and the output waveform. Since the degree of the phase modulation for the two components $\Delta \Phi(t)$ and $\Delta \Phi(t - \Delta t)$ at a certain time position is not equal, the phase difference $\delta \Delta \Phi$ temporarily goes below $\Delta \Phi_{\text{B}}$ after $t > t_0 + \Delta t$. Transient cancellation is possible if we properly detune the $\Delta \Phi_{\text{B}}$ from π , but complete cancellation of the phase shift after $t > t_0 + \Delta t$ is not possible since $\frac{d\Delta \Phi(t)}{dt}$ and $\frac{d\Delta \Phi(t-\Delta t)}{dt}$ do not balance as it is. Therefore there occurs a sub-pulse outside the gate window, potentially causing a bit error. Note that if we can induce asymmetric phase shifts for the two components as $\phi_1 \sim \exp(-t/\tau_{\text{eff}})$ and $\phi_2 \sim f_{\text{phase}} \cdot \exp(-(t - \Delta t)/\tau_{\text{eff}})$, we can suppress sub-pulse generation by selecting $f_{\text{phase}} = \exp(-\Delta t/\tau_{\text{eff}})$. As we will show in §3.4.2, this sub-pulse generation becomes a serious issue as we increase the SOA carrier recovery rate compared to Δt and so on.

3.3 Experimental Setup for Sub-pulse Measurement

3.3.1 Overview

The generation of sub-pulses had not been revealed clearly before our measurement owing to the lack of monitoring resolution in previous experiments. Precise, quantitative investigation of sub-pulses requires waveform monitoring in a high dynamic range with a high time resolution over a long time span. Thus, we employed a cross correlator (Femtochrome Research Inc., FR-103XR) that had a dynamic range exceeding 27 dB. Its time resolution was determined by the width of optical sampling pulses, and 2.2 ps could be achieved using pulse trains from a frequency-tunable mode-locked fiber laser (MLFL: Pritel Inc., UOC-3).



Figure 3.2: Time profiles of the (a) nonlinear phase shifts, (b) and (c) phase difference, and (d) gate output signal when τ_{eff} is short.

Table 3.1: Measured characteristics of the SOA#10 used in §3.4.1 at $I_{\rm op} = 250$ mA.

Parameter	Description	Value
G_0	small signal gain	23 dB
	coupling loss against pigtail fibers	2 dB
$P_{\rm cw}^{\rm sat}$	saturation power for cw light	$14 \mathrm{mW}$
$E_{\rm pulse}^{\rm sat}$	saturation energy for 2-ps pulse	600 fJ
$ au_{ m C}$	carrier lifetime	$200~\mathrm{ps}$

(a)



Figure 3.3: Setup of the DISC-gate experiment. (a): Setup for §3.4.1. VOA: variable optical attenuator, MZI: Mach-Zehnder interferometer, SOA: semiconductor optical amplifier, Q: quarter-wave plate, H: half-wave plate, P: polarizer, MLFL: Mode-locked fiber laser, MUX: optical time-division multiplexer. (b): Modified DISC-gate setup for §3.4.2.



Figure 3.4: Measured characteristics of the SOAs. (a): Small-signal gains for cw light, $\lambda = 1548.5$ nm. (b): Gain saturation profiles for temporally-isolated pulses. Dashed lines show fit results with a theoretical function. (c) and (d): Spectra of the cross-phase-modulated cw outputs from the SOAs. Dashed lines indicate the cw-light wavelength.

The setup for the waveform monitoring in $\S3.4.1$ is shown in Fig. 3.3(a). In this case, we used a commercially available SOA module #10 (InPhenix, Inc., IP-SAD1503, drive current = 250 mA) in the DISC. Some of the measured characteristics of this SOA are shown in Table 3.1 and Figs. 3.4(a) and 3.4(b). Particularly, precise characterization of the gain saturation for temporally isolated optical pulses was indispensable for the accurate sub-pulse calculation in §3.4.1. In addition, the degree of nonlinear phase shift for particular optical inputs was estimated from the XPM spectrum, as shown in Fig. 3.4(c), to determine $\frac{dn_r}{dn}$ later. We used a set of high-precision, high-extinction-ratio polarization-control devices $Q_1 \sim P_2$ (Optoquest Inc.) in order to construct the MZI. The rotating quarter-wave plate Q_1 and the rotating polarizer P_1 were used to adjust the polarization of the cw light to exactly 45° off the principal axis of the calcite crystal C₁ (DGD, $\Delta t = 5$ ps). The next quarter-wave plate Q_2 and the rotating polarizer P_2 were used to control $\Delta \Phi_{\rm B}$ precisely. Details of this polarization-based MZI will be discussed in §3.3.2. 25 GHz and 12.5 GHz clock pulses were generated from the MLFL and an optical multiplexer to drive the DISC gate. The width of the pulses sent to the DISC was broadened from 2.2 to 3.8 ps with a 100-m-long single-mode fiber, to suppress carrier-heating-related phenomena inside the SOA[49–53]. The properties of the 12.5 GHz pulses are shown in Figs. 3.5(a)-3.5(c). The pedestals of the input pulses were minimized to the best, and we obtained good extinction ratio between the RZ input pulses ($\sim 30 \text{ dB}$). The polarization of cw light from the distributed-feedback laser diode (DFB-LD) as well as the clock signal was aligned to either the TE or TM axis of the SOA at the SOA input facet, to suppress the nonlinear polarization rotation[63, 64].

For the measurement discussed in §3.4.2, where a much faster SOA was required, we used the custom-designed SOA chip B#1 (see §2.3) instead of the previous SOA module. The small-signal gain, the gain saturation profile, and the example of XPM spectrum are shown in Figs. 3.4(a), 3.4(b), and 3.4(d), respectively. Part of the experimental setup was modified as shown in Fig. 3.3(b). Signal injection to the SOA chip was achieved using the PM-DSF lensed fiber as in §2.3, with its PM axis aligned to the TM axis of the SOA. Counter-propagating assist light of $\lambda_{assist} = 1480$ nm was used in some measurements to obtain an effective recovery time as short as $\tau_{\text{eff}} = 11$ ps. The polarization-based MZI was improved so that all the components were directly connected to each other without fiber connections (Fig. 3.6). This reduced the drift of the polarization-transfer property inside the MZI.



Figure 3.5: Measured properties of the 12.5-GHz input pulses. (a) Auto-correlation trace (solid) of the 2.2-ps pulses from the MLFL. Dashed curve shows a deconvoluted waveform (two-component *sech*-pulses). (b) Cross-Correlation trace of the DISC input (3.8 ps). (c) Spectrum of the input pulses and cw light.



Figure 3.6: Improved polarization-based MZI

3.3.2 Phase bias of polarization-based MZI

As we discussed in §3.2, the MZI in the DISC plays a critical role in wavelength conversion and sub-pulse generation. Since the sub-pulse waveform is quite sensitive to the MZI phase bias as we will show in §3.4.1, precise determination of the bias value $\Delta \Phi_{\rm B}$ is essential for comparison of the measured and calculated results. In this subsection, therefore, we discuss how the set of wave plates, calcite and polarizers acts as an MZI, how we can determine $\Delta \Phi_{\rm B}$ from the wave-plate settings, and the precision of $\Delta \Phi_{\rm B}$.

In the first step, cw light is evenly split into two orthogonally polarized components aligned to the C_1 axes, by adjusting the Q_1 , P_1 , and C_1 angles. Seen from one of the C_1 axes, the Jones vectors of these components are expressed as

$$\mathbf{E}_{\text{in1}} = \begin{pmatrix} 1\\0 \end{pmatrix}, \mathbf{E}_{\text{in2}} = \begin{pmatrix} 0\\1 \end{pmatrix}.$$
(3.10)

The Jones matrices of a calcite with an optical path length difference of Δl , a quarter-wave plate, a polarizer, and a rotation operation are expressed as

$$M_{\rm C} = \begin{pmatrix} \exp\left(2\pi i \frac{\Delta l}{\lambda}\right) & 0\\ 0 & 1 \end{pmatrix}, \quad M_{\rm Q} = \begin{pmatrix} \exp\left(i\frac{\pi}{4}\right) & 0\\ 0 & \exp\left(-i\frac{\pi}{4}\right) \end{pmatrix}, \quad (3.11)$$
$$M_{\rm P} = \begin{pmatrix} 1 & 0\\ 0 & 0 \end{pmatrix}, \quad R(\theta) = \begin{pmatrix} \cos\left(\theta\right) & -\sin\left(\theta\right)\\ \sin\left(\theta\right) & \cos\left(\theta\right) \end{pmatrix},$$

respectively.

In the MZI used in §3.4.1, a dispersion-shifted single-mode fiber was used to connect C_1 and Q_2 . The Jones matrix of a single-mode fiber is generally expressed by a unitary matrix[61] as

$$M_{\rm F} = \begin{pmatrix} \cos\left(\theta_{\rm M}\right) \exp\left(i\phi_{\rm M}\right) & \sin\left(\theta_{\rm M}\right) \exp\left(i\psi_{\rm M}\right) \\ -\sin\left(\theta_{\rm M}\right) \exp\left(-i\psi_{\rm M}\right) & \cos\left(\theta_{\rm M}\right) \exp\left(-i\phi_{\rm M}\right) \end{pmatrix}, \tag{3.12}$$

where $\theta_{\rm M}, \phi_{\rm M}$, and $\psi_{\rm M}$ are parameters to be measured separately.

Thus, the two components \mathbf{E}_{in1} and \mathbf{E}_{in2} are converted to

$$\mathbf{E}_{\text{out1}} = M_{\text{P}}R\left(-\left(\theta_{P_{2}} - \theta_{Q_{2}}\right)\right)M_{\text{Q}}R\left(-\theta_{Q_{2}}\right)M_{\text{F}}R\left(\theta_{C_{1}}\right)M_{\text{C}}\mathbf{E}_{\text{in1}},$$
$$\mathbf{E}_{\text{out2}} = M_{\text{P}}R\left(-\left(\theta_{P_{2}} - \theta_{Q_{2}}\right)\right)M_{\text{Q}}R\left(-\theta_{Q_{2}}\right)M_{\text{F}}R\left(\theta_{C_{1}}\right)M_{\text{C}}\mathbf{E}_{\text{in2}},$$
(3.13)

where θ_{C_1} , θ_{Q_2} , and θ_{P_2} are the angles of the corresponding devices' axes. Then the complex ratio of these parallel vectors determines $\Delta \Phi_{\rm B}(Q_2, P_2)$ and the real amplitude ratio $r(Q_2, P_2)$ as

$$\mathbf{E}_{\text{out2}} = \mathbf{E}_{\text{out1}} \times r \exp\left(i\Delta\Phi_{\text{B}}\right). \tag{3.14}$$

The output signal of the MZI is given by $\mathbf{E}_{out} = \mathbf{E}_{out1} + \mathbf{E}_{out2}$. We call r the mixing ratio. \mathbf{E}_{out} becomes zero when r = 1 and $\Delta \Phi_{\rm B} = \pi$. This can be achieved by adjusting θ_{Q_2} and θ_{P_2} properly, so that Q_2 eliminates the ellipticity of the combined signal and P₂ extinguishes its outcome. Slight detunings of θ_{Q_2} and θ_{P_2} from these conditions lead to a shift in $\Delta \Phi_{\rm B}$. Figure 3.7 shows the calculated dependence of $\Delta \Phi_{\rm B}$ for the MZI used in §3.4.1 on $\Delta \theta_{Q_2}$, while keeping $\Delta \theta_{P_2} - \Delta \theta_{Q_2}$ constant $(\Delta l/\lambda = -973.392, \theta_{C_1} = 68^{\circ}, \theta_{\rm M} \sim 0.85 \text{ (rad)}, \phi_{\rm M} \sim 1.98 \text{ (rad)}, \psi_{\rm M} \sim 2.76 \text{ (rad)},$ $\theta_{Q_2} \sim 131^{\circ} + \Delta \theta_{Q_2}, \ \theta_{P_2} \sim -64.5^{\circ} + \Delta \theta_{P_2})$ Keeping $\Delta \theta_{P_2} - \Delta \theta_{Q_2}$ constant was an approximate means of reducing r shift, and it was shifted from 0 to $+2^{\circ}$ in order to slightly compensate for the r shift at around $\Delta \theta_{Q_2} = -3 \sim -6^{\circ}$. Then $\Delta \Phi_{\rm B}$ shifted linearly with $\Delta \theta_{Q_2}$, and $\frac{d\Delta \Phi_{\rm B}}{d\Delta \theta_{Q_2}} = -7.3 \times 10^{-3} \pi \text{ (rad/deg)}$. Thus, we could accurately control $\Delta \Phi_{\rm B}$ (~ $\pm 0.004\pi$ (rad)) through precise (~ $\pm 0.5^{\circ}$) adjustments of wave-plate angles. Here we note that the actual fiber transfer matrix $M_{\rm F}$ slightly drifted with time. To compensate for this issue, we periodically adjusted θ_{Q_2} and θ_{P_2} to angles where $\Delta \Phi_{\rm B}$ became π in the case in §3.4.1, and carefully compensated for the angle drifts.

3.4 Results

3.4.1 Sub-pulse measurement and calculation with systematic MZI-bias shifts

In this subsection, we explain our first observation of sub-pulses. First, we operated the DISC at 25-GHz operation frequency, with clock pulse input (+12.3 dBm, $\lambda_1 =$ 1560 nm) and cw input (+9 dBm, $\lambda_2 =$ 1548.5 nm) into the SOA module #10, and observed its output waveform. Figure 3.8(a) shows a typical cross-correlation trace of the DISC output that we measured after maximizing the extinction ratio by adjusting Q₂ and P₂. As shown in Fig. 3.8(a), the extinction ratio was limited to 25 dB by small sub-pulse-like components between the output pulses. Then, to observe the waveform of the remaining components in more detail, we dropped the



Figure 3.7: Calculated dependence of the MZI-bias (solid) and the mixing ratio (dashed) on the quarter-wave plate angle.

pulse frequency from 25 GHz to 12.5 GHz and carefully optimized the wave plate angles once again. The average power of the 12.5 GHz pulse train was +13.5 dBm. At this frequency, the extinction ratio was limited to about 18 dB by several types of relatively large sub-pulse-like components depending on the Q_2 and P_2 settings (Figs. 3.8(b) - 3.8(d)).

After that, we calculated the DISC output waveform to demonstrate that the above sub-pulse-like components actually result from the operating principle modeled in §3.2. First, we chose the parameter values for use in the simulation, taking measured chip gain, saturation energy, and the gain recovery time of the SOA into account (Table 3.1). The saturation energy was determined to be 600 fJ from the pulse-gain saturation measurement (Fig. 3.4(b)). $\frac{dn_r}{dn}$ was chosen so that the amount of the cross-phase-modulation in the 25 GHz operation becomes 0.5π (Fig. 3.4(c)). The chosen parameter values are summarized in Tables 3.2 and 3.3. Then we calculated output waveforms using eqs. (3.1)~(3.9). For the numerical integration of the differential equation eq. (3.1), we employed second-order Runge-Kutta method. Fig. 3.9 shows the flow chart of our waveform simulation, including long-bit-pattern waveform handling scheme to calculate pattern-induced intensity fluctuation in the later section.

We could obtain sub-pulse waveforms quite similar to the measured data, as



Figure 3.8: Measured cross-correlation traces of the DISC output with different Q_2 and P_2 settings: (a) with +12.3-dBm, 25-GHz input pulses, (b)-(d) with +13.5-dBm, 12.5-GHz input pulses ((b) $\Delta\theta_{Q_2} = -3$ (deg), $\Delta\theta_{P_2} = -1$ (deg), (c) $\Delta\theta_{Q_2} = -6$ (deg), $\Delta\theta_{P_2} = -4$ (deg), (d) $\Delta\theta_{Q_2} = -8$ (deg), $\Delta\theta_{P_2} = -6$ (deg).) Calculated phase bias values are shown in each graph.



Figure 3.9: Flow chart of our DISC-gate waveform simulation.



Figure 3.10: Calculated cross-correlation traces of the DISC output: (a) with 25-GHz input pulses, (b)-(d) with 12.5-GHz pulses. ((b) $\Delta \Phi_{\rm B} = 1.030\pi$, (c) $\Delta \Phi_{\rm B} = 1.060\pi$, (d) $\Delta \Phi_{\rm B} = 1.100\pi$). Other parameter values are summarized in Tables 3.2 and 3.3.

Table 3.2: Parameters used in the DISC-gate simulation							
Parameter	Description	Value	Value	Value			
		(Fig. 3.10(a))	(Figs. $3.10(b)-(d)$)	$(\S{3.4.2})$			
f	Pulse repetition frequency	$25~\mathrm{GHz}$	$12.5~\mathrm{GHz}$	10.5 GHz (50 GHz)			
$T_{\rm FWHM}$	Input pulse width (FWHM)	$3.8 \mathrm{\ ps}$	$3.8 \mathrm{\ ps}$	$2.0 \mathrm{\ ps}$			
$E_{\rm pulse}$	Input pulse energy (chip)	$429~\mathrm{fJ}$	1129 fJ	$30 \mathrm{fJ}$			
$P_{\rm cw}$	Input cw power (chip)	$5 \mathrm{mW}$	$5 \mathrm{mW}$	$0.3 \mathrm{mW}$			
Δt	Delay time in the MZI	$5.0 \mathrm{\ ps}$	$5.0 \mathrm{\ ps}$	$5.0 \mathrm{\ ps}$			
$T_{\rm probe}$	Correlator probe width	2.2 ps	2.2 ps	$2.0 \mathrm{\ ps}$			
$ u_{ m cw}$	Frequency of cw light	$193.7 \mathrm{~THz}$	$193.7 \mathrm{~THz}$	$193.7 \mathrm{~THz}$			
$ u_{\mathrm{pulse}} $	Frequency of pulsed light	$192.1 \mathrm{~THz}$	192.1 THz	192.8 THz			

shown in Figs. 3.10(a) to 3.10(d), after choosing $\Delta \Phi_{\rm B}$ values properly. The calculated relative peak intensities of the sub-pulses were -24 dB in Fig. 3.10(a) and -18 dB in Fig. 3.10(b).

For more systematic comparison of the measured and calculated results, though, we studied how the relative intensities of sub-pulse-like components behave as functions of the MZI phase bias $\Delta \Phi_{\rm B}$ (Fig. 3.11). The relative intensity was taken at three time positions ($t = t_0 + 15$ ps, $t_0 + 40$ ps, and $t_0 + 65$ ps, where t_0 is the primary pulse's position). The calculated behavior was quite similar to the measured results, taking the ~ 27-dB dynamic range of the cross-correlator into account. The mismatch in the intensity was small, 1 ~ 2 dB for $t = t_0 + 15$ ps series, at $\Delta \Phi_{\rm B} = 1.00\pi \sim 1.08\pi$. We are aware that the phase-bias dependence and extinction ratios are dominated by many parameters, and future improvement in the gate model may solve the remaining small mismatch. Regarding sub-pulse intensity, we consider that the gate model in this chapter provides the most reliable predictions currently available.

Thus, as shown in Figs. 3.8-3.11, we observed a good match between the measured and calculated waveforms and the phase-bias dependence. Therefore, we presume that our sub-pulse generation model has been experimentally verified for the particular recovery rate, and that we can predict sub-pulse intensity with $1 \sim 2$ dB accuracy using this model.



Figure 3.11: Relative intensities of sub-pulse-like components in 12.5-GHz DISC output: (a) from experimental results, (b) from calculation. These were obtained at three timing positions: $t = t_0 + 15$ ps (solid), $t_0 + 40$ ps (dashed), and $t_0 + 65$ ps (dotted).

3.4.2 Trade-off regarding sub-pulse generation and patterninduced intensity fluctuation

The sub-pulse intensity of the DISC output discussed in §3.4.1 could be suppressed by tuning the MZI bias down to ~ -20 dB, which was negligible for practical applications. Nevertheless, we have regarded sub-pulse generation as a major issue in the case of random-data operation, because of the trade-off relationship between sub-pulse generation and PIF. If the input signal is not a continuous clock but has a random pattern, the instantaneous carrier density in the SOA varies upon each signal injection depending on the preceding input-signal pattern. Then the intensities of the output signals become pattern-dependent. This effect can be reduced by accelerating the carrier recovery with current injection or assist light injection. Our sub-pulse model indicates, however, that a sub-pulse becomes large as we increase the carrier recovery rate. In this section, we will discuss the measured and calculated results for this effect.

We systematically measured sub-pulse waveforms for several carrier lifetimes, using the customized SOA chip B#1, as explained in §3.3.1. Figure 3.12(a) shows the gain-recovery profiles of the SOA for several injection currents, which we showed in Fig. 2.9(c) in Chapter 2. The recovery time $\tau_{\rm C}$ was reduced from 280 to 36 ps as we increased the injection current $I_{\rm op}$ from 100 to 400 mA, and it was approximately inversely proportional to $I_{\rm op}$ above 150 mA. The effective recovery time $\tau_{\rm eff}$ was further shortened to 11 ps after we injected 80 mW assist light with a nearly-



Figure 3.12: DISC output waveforms for several carrier recovery rates. (a): XGM profiles of the SOA B#1 for several operation currents, indicating the recovery acceleration. The cw-probe intensity and pulse energy into the chip were set to -25 dBm and 10 fJ, respectively. Note that each plot is translated intentionally in vertical direction. (b): 10.5-GHz DISC output waveforms. $\Delta \Phi_{\rm B}$ was set to be 1.0π .

transparent wavelength (1480 nm) into the chip. Using this SOA, we measured the DISC output waveforms on each $I_{\rm op}$. The operation frequency was 10.5 GHz. The input pulse width, wavelength, and energy into the chip were 2 ps, 1555 nm, and 30 fJ, respectively. The cw light wavelength and power into the chip were 1548.5 nm and -5 dBm, respectively. The delay time of the MZI was set to 5 ps. The MZI phase bias was fixed at 1.0π throughout the measurement, since the optimum phase bias depends largely on $\tau_{\rm C}$ ($\tau_{\rm eff}$). The measured waveforms are shown in Fig. 3.12(b). We found that the relative intensity of the sub-pulse increased up to -10 dB, as we increased the carrier recovery rate $1/\tau_{\rm C}$ (Fig. 3.13(a)).

Meanwhile, we could predict this result qualitatively using our gate model. For simplicity, we assumed that $\tau_{\rm C}$ was inversely proportional to $I_{\rm op}$ and other fundamental parameters did not change with $I_{\rm op}$. The adopted values are summarized in Tables 3.2 and 3.3. The values in Table 3.3 were determined in view of measured characteristics of SOA B#1 at $I_{\rm op} = 400$ mA. Then the measured and calculated sub-pulse intensities agreed quite well for most $1/\tau_{\rm C}$ values (Fig. 3.13(a)). We regarded this matching as convincing evidence that our model is applicable in a wide $1/\tau_{\rm C}$ range. Some discrepancy at $\tau_{\rm C} = 11$ ps probably originated from the difference between the acceleration mechanism with current injection and that with assist-light injection. This result clearly indicates that we can suppress sub-pulse generation as long as we keep the carrier recovery rate small.

As we mentioned above, this method of sub-pulse suppression leads to an increase

Description	Value in $\S3.4.1$	Value in $\S3.4.2$
Injection current	250 mA	400 mA $\times \alpha$
Active region length (effective)	$700~\mu{ m m}$	$1100~\mu{ m m}$
Active region volume	$1.4 \times 10^{-10} \ (\mathrm{cm}^3)$	$5.2 \times 10^{-11} \ (\mathrm{cm}^3)$
Confinement factor	0.2	0.2
SOA carrier lifetime	200 ps	$35 \text{ ps} / \alpha$
SOA differential gain	$2.17 \times 10^{-15} \ (\mathrm{cm}^2)$	$6.36 \times 10^{-16} \ (\mathrm{cm}^2)$
Carrier-conversion efficiency	0.079	0.34
Refractive index change	$-1.87 \times 10^{-19} \ (\mathrm{cm}^3)$	$-1.4 \times 10^{-20} \ (\mathrm{cm}^3)$
Maximum-available carrier density	$1.76 \times 10^{17} \ (\mathrm{cm}^{-3})$	$5.72 \times 10^{17} \ (\mathrm{cm}^{-3})$
SOA small-signal gain	23.4 dB	34.8 dB
Saturation energy for pulse	600 fJ	$473~\mathrm{fJ}$
Phase shift at complete carrier depletion	$+6.0\pi$	$+2.3\pi$
	Description Injection current Active region length (effective) Active region volume Confinement factor SOA carrier lifetime SOA differential gain Carrier-conversion efficiency Refractive index change Maximum-available carrier density SOA small-signal gain Saturation energy for pulse Phase shift at complete carrier depletion	DescriptionValue in §3.4.1Injection current 250 mA Active region length (effective) $700 \ \mu \text{m}$ Active region volume $1.4 \times 10^{-10} \ (\text{cm}^3)$ Confinement factor 0.2 SOA carrier lifetime $200 \ \text{ps}$ SOA differential gain $2.17 \times 10^{-15} \ (\text{cm}^2)$ Carrier-conversion efficiency 0.079 Refractive index change $-1.87 \times 10^{-19} \ (\text{cm}^3)$ Maximum-available carrier density $1.76 \times 10^{17} \ (\text{cm}^{-3})$ SOA small-signal gain $23.4 \ \text{dB}$ Saturation energy for pulse $600 \ \text{fJ}$ Phase shift at complete carrier depletion $+6.0\pi$

Table 3.3: SOA fundamental parameter values used in the calculation, and derived quantities

in PIF. We calculated the DISC output waveforms for a 50-Gb/s input pulse train with a pseudorandom bit sequence (pattern length = $2^{15} - 1$), and figured the PIF from the intensity ratio of the largest and smallest signals in the output pulse train (Fig. 3.13(b)). The other parameter values were kept unchanged from the former calculations. The dotted line in Fig. 3.13(a) shows the dependence of PIF on the recovery rate. Evidently PIF could be suppressed by increasing recovery rate, which lead to the unwanted increase in sub-pulse intensity.

In this way, our model shows trade-off between sub-pulse generation and PIF. Although what we have shown is only a particular example, we consider that this trade-off relation arises fairly universally. We will encounter sub-pulses as large as -7 dB, in the present case for example, as we suppress the PIF down to 1 dB (or $-10 \sim -12$ dB when PIF is 4 dB). The situation may become even worse depending on other operating conditions such as signal bit rate, signal duty ratio, and the control pulse energy. Therefore, we consider that a new gating scheme to coping with this issue is necessary, and that sub-pulses should be evaluated using the current model, or an expanded model in the near future.

Here we mention a few things. In some reports on the results of the ultrafast wavelength-conversion experiments [23, 26], we cannot find such large sub-pulses. We suppose several possible reasons for this.

1. PIF was not highly suppressed, as indicated by the eye-diagrams (2 \sim 3 dB ?).



Figure 3.13: Trade-off between the pattern-induced intensity fluctuation (PIF) and the sub-pulse noise. (a): Measured (solid) and calculated (dashed) sub-pulse intensities against the carrier recovery rate $1/\tau_{\rm C}$ at 10.5 GHz, and calculated PIF for the 50 Gb/s signal (dotted). (b): Calculated eye diagrams of the DISC output for 50 Gb/s pseudo-random signal, for two carrier recovery rates. Cross-correlation process was omitted for simplicity. Parameter values are summarized in Tables 3.2 and 3.3.

- 2. The MZI phase bias values were detuned from π , and the sub-pulses were suppressed to some extent (Fig. 3.12).
- 3. Unexpected contribution of the nonlinear polarization rotation (as in Chapter 4 or as in [57]).
- 4. Unexpected contribution of the carrier heating [62] and the BPF detuning (Chapter 4).

Since there were many potential factors in the past experiments, it is difficult to provide clear answer to them. In Chapter 4, however, we still observe unexpectedly-small sub-pulses when no BPF-detuning is employed. We are not sure this is accidental or intrinsic. We speculate that the contribution of the carrier heating is quite large in the frequency region over 160 GHz, and some modification in our model will reduce the mismatch in such high-frequency region.

3.5 Conclusion

We have studied sub-pulse generation of the first kind (originated from the exponential recovery profile of the SOA carriers) in the wavelength-converted signal from a DISC gate operated at $10 \sim 25$ GHz, both theoretically and experimentally. We observed that the DISC output contained small sub-pulses in addition to the wavelength-converted output pulses. Good matching (1 ~ 2 dB) was obtained between the measured and calculated sub-pulse intensities. The measured dependence of sub-pulse waveforms on the MZI setting was quantitatively explained with our DISC model.

In addition, we demonstrated that our model is applicable in a wide range of SOA carrier recovery rates. The impact of carrier-recovery acceleration (from $\tau_{\rm C} = 280$ ps to $\tau_{\rm eff} = 11$ ps) on sub-pulse waveforms was systematically studied using a DISC with a customized SOA chip. We obtained good agreement between the measured and calculated sub-pulse intensities for the majority of the recovery rate range. We consider that our model is quite reliable as long as the influence of the carrier-heating in the SOA is small.

Using our sub-pulse model, we discussed the trade-off between sub-pulse generation and PIF. Our calculations indicated that significant sub-pulses may emerge when we suppress PIF, which may lead to difficulty in the practical high-speed operation of the DISC. This result suggests that part of the DISC structure should be improved to overcome this trade-off. We are in the process of solving this issue using new gating schemes, for example, using an expanded DISC including an optical spectral synthesizer[62].

We believe that the success in such modeling will be valuable not only for the practical applications of the DISC but also for the detailed understanding of other kinds of all-optical gates using SOAs, delayed interferometers, and so on.

Chapter 4

200-Gb/s wavelength conversion using DISC with long-pattern waveform monitoring

4.1 Introduction

Through chapter 2 and 3, we have investigated the characteristics of SOA-based alloptical gates and DISCs. We have tried to understand their characteristics in the frequency range above 100 Gb/s, as well as in the lower frequency range as ~ 10 GHz. Since we had to analyze the properties of the SOAs and the DISCs as accurate as possible, we used the experimental results precisely measured at the lower frequency range. For the comprehensive understanding of the gate characteristics, however, systematic study on the high-frequency operation of all-optical gates will be also desired.

In this chapter, we discuss our results on the output waveform measurements of a DISC gate at 200 Gb/s. As we have shown through Chapter 3, the waveform monitoring is a powerful method to understand the fundamental dynamics of all-optical gates. So far error-free operations of DISCs at 160 Gb/s have been reported by a few groups[23–26], and waveform monitoring has been performed in several ways. Each of the methods had their own drawbacks, however. The most common method was to use a set of optical time-division demultiplexing gates (DEMUXs), OE converters and electric sampling scopes. In such measurements, detailed information of the waveforms was lost due to the bandwidth limits of the electric devices. Costly streak-cameras were sometimes used, but their time resolutions were also limited to ~ 4 ps. Eye-diagram monitoring of the output waveform has been reported for the extended DISC operated at 160 Gb/s and 320 Gb/s using an optical sampling scope. Though it has an excellent time resolution[78], it is based on (what is called

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) "asynchronous sampling" [79, 80] and not suited for the characterization of each pulses in a patterned data signals. Result on the systematic measurement has not been reported from them, neither. Finally, we measured and reported the influence of ultra-fast effects on the patterning effect of an SOA gate (similar to DISC) at 105 Gb/s, through the cross-correlation method[43]. The repetition-rate of the bitpattern allowed in this method was restricted by the frequency of the probe pulse (10.5 GHz), however, and the adopted bit-pattern ('11110000001111000000...') was far from realistic digital data. We performed a 1/16-demultiplexing of the probe pulses using an LN modulator and a programmable pattern generator (PPG) in Chapter 2. This method requires, however, an exclusive use of the costly PPG. The extinction ratio of the demultiplexed pulse was not quite high, either. Therefore we developed a low-frequency (40 MHz) sampling-pulse generator with low additional cost and used it as a probe source to the cross correlator. This enabled us to acquire the waveform of very-long ($T_{\text{period}} = 25000 \text{ ps}, 4992 \text{ bits at } 200 \text{ Gb/s}$) bit-pattern digital signal. Through this long-pattern waveform measurement, we aimed to investigate the pattern-induced intensity fluctuation (PIF) of the output signal and other waveform distortions, more strictly than in the preceding short-bit [42, 43] or $2^7 - 1$ -bit[23, 25, 26] experiments.

4.2 Experimental setup for the 200-Gb/s wavelength conversion

Figure 4.1(a) shows the overview of our experimental setup. In the following sections, we first discuss the data-multiplexing part (Fig. 4.1(b)), the sampling-pulse generator (Fig. 4.1(d)), and the measured waveform of 200-Gb/s input-data signal. Then we provide measured results on the 200-Gb/s operation of the DISC gate (Fig. 4.1(c)).

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(d)



(e)



Figure 4.1: The setup of the 200-Gb/s DISC-gate experiment. (a): Experimental setup overview. (b): Diagram and photograph of the optical time-division multiplexing part (c): DISC gate (d): Simplified diagram of the 40-MHz sampling-pulse generator (e) Photograph of the core part of the sampling-pulse generator

4.2.1 Optical time-division multiplexing of mode-locked pulses up to 200 GHz for the gate input

The time-division multiplexing of RZ-formatted optical signals is possible by splitting them into two components, applying proper temporal shift to one component, and combining them after that. We multiplexed the repetition rate of mode-locked pulse trains ($\lambda = 1555$ nm) from 12.48 GHz up to $12.48 \times 2^4 = 199.68$ GHz, using four-stage multiplexers (MUXs). The first and second MUXs have been fabricated by ourselves using commercially-available fiber-optic devices, while the latter two were prefabricated MUX products (Pritel Inc., OCM-4). All the devices in our MUXs were pig-tailed by PM-DSFs. Since the third and fourth MUX had certain amount of velocity dispersion, though, we had to use a dispersion-compensation fiber ($\beta_2 = -84$ ps/nm·km, $\beta_3 = -0.26$ ps /nm²·km) of 1 m after multiplexing.

Since our target frequency was as high as 200 GHz, we performed adiabatic compression[81] of the 12.48-GHz pulses before multiplexing, using a dispersion-decreasing fiber (DDF) module. Figures 4.2(a) and 4.2(c) show auto-correlation traces of the pulses before and after compression in the best case. We see that the pulse width was shortened from 1.6 ps to 260 fs, when the average input power was +21.3 dBm. The waveform of the compressed pulse generally depends on the status of the input mode-locked pulses, and hence we had to tune the operating condition of the MLFL carefully from day to day. Note that the correlation trace of the neighboring pulses in the 12.48-GHz pulse train, as shown in Fig. 4.2(d), was slightly broadened from that of the identical pulses, probably due to an instability of our MLFL. This broadening was one cause of the bit-rate limitation in our cross-correlation measurements.

Figures 4.3(a) to 4.3(d) show auto-correlation traces of the multiplexed pulse trains. Periodic traces were obtained after fine adjustment of the delay length in each MUX stage. In general, we had to perform this adjustment on each day, at least for the 1st and 2nd MUX stages. The group velocity dispersion, which broadened the pulse width at 100 GHz to 1.36 ps, was partially compensated by the DCF at 200 GHz. The resultant pulse-width of 800 fs was suitable for 200-Gb/s operation of the DISC gate, in view of the finite time resolution of our sampling system. Figure 4.4 shows the cross-correlation trace of the 200-GHz clock pulses and 12.48-GHz compressed clock pulses. The intensities of the pulses in the sixteen OTDM channels were adjusted with the variable attenuators, so as to be almost equal with each other.



Figure 4.2: Auto-correlation traces and optical spectra of the 12.48-GHz modelocked pulses before and after compression. (a), (b): Before compression, around t = 0 ps and t = -80 ps. (c), (d): After compression, around t = 0 ps and t = -80 ps. (e) and (f): Typical spectra of the pulses before and after compression.



Figure 4.3: Auto-correlation traces of the multiplexed pulse trains. (a): at 25 GHz. (b): at 50 GHz. (c): at 100 GHz. (d): at 200 GHz.

Bit-pattern encoding was performed at 12.48 GHz through an LN modulator (#1). Since the frequency of our new sampling-pulse generator was restricted to 40 MHz, we chose a bit-pattern with a pattern length of 312 as shown in Fig. 4.5 and synchronized this pattern to the sampling pulses. The bit-pattern in this stage was simply determined through a Monte Carlo method. To avoid the temporal correlations of multiplexed bits in a short span, we inserted an additional ~ 12.5-ns delay in the first MUX and a ~ 6.25-ns delay in the second stage. The delay values in the third and fourth stages were pre-adjusted for 10-Gb/s, $2^7 - 1$ -bit PRBS input signal and could not be changed.



Figure 4.4: Cross-correlation trace of the 200-GHz pulse train with 12.48-GHz compressed probe pulses.



Figure 4.5: Adopted encoding pattern at 12.48 Gb/s.
4.2.2 Generation of low-jitter sampling pulses at 40 MHz

A low frequency, low-jitter, distortion-free, ultrafast and high-power pulse source is inevitable for the optical sampling of long-period, high-bit-rate lightwave signals. Several commercially-available MLFLs satisfy some of these requirements. In fact, we have a custom-designed MLFL (Calmar Optocom Inc., FPL-02CFTUEC12) which generates low-distortion, ultrafast (FWHM = 200 fs) optical pulses at 40-MHz repetition frequency (Fig. 4.6(a)). It is not easy in general, however, to synchronize such low-frequency pulses to other ultrafast signals with small timing jitter. Figure 4.6(b) shows a measured example of several cross-correlation traces of compressed pulses from the 12.48-GHz MLFL (MLFL #1) and pulses from the 40-MHz MLFL (MLFL #2), acquired in a short period. The experimental setup was as shown in Fig. 4.6(c). The correlation traces show that the timing jitter between pulses from two MLFLs exceeds a few picoseconds. Furthermore, temporal position of the pulses from MLFL #2 drifted in a few hours by several hundred picoseconds, presumably due to the thermal expansion of the laser cavity. Thus we had to develop another low-frequency pulse generator which can be easily synchronized to the 12-GHz system.

The setup of our sampling-pulse generator was as shown in Fig. 4.1(d). This system requires a branch of the 12.48-GHz pulse train and a synchronized electric signal for the input. The pulse train was electrically demultiplexed down to 40 MHz, through a 10-Gb/s LN modulator (#2) and a 40-Gb/s EA modulator (Oki Electric Industry, OM5642W-30B). To generate high-speed electric gate signals for these modulators, we used the 40-MHz MLFL and a 16-GHz photo-diode (Discovery Semiconductor, Inc., DSC20H). Figure 4.7(a) shows the OE-converted signal of the mode-locked pulse, acquired with a 20-GHz electric sampling scope. The averaged input power to the photo-diode was typically adjusted to -14 dBm, though Fig. 4.7(a) was exceptionally acquired at -20 dBm. This gate signal was split into two by an Wilkinson divider, and amplified by broadband ($\sim 12 \text{ GHz}$) RF amplifiers as shown in Figs 4.7(b) and 4.7(c). The two amplifiers had fairly large difference in their performances. The first inverting amplifier #2 (Cernex Inc., CBMV3142019) seemed to have poor broadband-gain-flatness, and the generation of the after-pulse could not be avoided. Since the EA modulator did not accept inverted gate pulses, this amplifier was used for the LN modulator. The output signal from the second amplifier (Picosecond Pulse Labs. Inc., 5865) was less distorted, and used for the EA modulator. Figures 4.7(d) and 4.7(e) show the gate windows of the LN and EA

modulator driven by these electric pulses, measured using cw-light input, a 35-GHz photo-diode, and the sampling scope. The width of the windows, broadened by the band-width limits of the photo-diode, amplifiers, and modulators, were much larger than the timing jitter of the MLFL #2, and still smaller than the interval of the 12.48-GHz pulse train. The FWHM length of the cascaded windows measured using the auto-correlator was about 30 ps. As the 12.48-GHz pulse train went through these two modulators, the pulses outside the cascaded windows got suppressed with an estimated extinction ratio of ~ 30 dB. The modulated waveform after LN #2 was monitored by the sampling scope during the experiment, so that easy tuning of the DC bias for LN #2 and the timing of the gate windows was possible.

The pulses right after the demultiplexing were fairly weakened due to the insertion losses of the EA modulator and other devices. Therefore we amplified them twice with dispersion-compensated EDFAs. After then we performed adiabatic compression using another DDF module (#2), and amplified the compressed pulses again. Figure 4.8(a) and 4.8(b) show the auto-correlation traces of the sampling pulses right after the third EDFA. The pulse width was compressed down to 440 fs. This is almost the shortest value we could achieve. The suppression ratio of the peak at t = -80 ps was 19 dB. This probably resulted from the after-pulses in the gate windows, and we expect that most of the pulses in the 12.48-GHz pulse train were suppressed with higher ratios. Figure 4.8(c) shows the spectrum of the generator output after the third EDFA. The broad peak around $\lambda = 1540 \sim 1570$ nm indicate the spectrum of the compressed 40-MHz pulses, which was broadened by self-phase modulation inside the DDF, while the un-broadened peak at $\lambda = 1555$ nm partially resulted from residual 12.48-GHz pulses. To reduce the contribution from the residual 12.48-GHz component, we used a 9.2-nm (or 7.5-nm in §4.3) bandpass filter. The autocorrelation traces and optical spectra in these cases were as shown in Fig. 4.8(d) to 4.8(f). Though the pulse width was broadened to 620 fs (BPF 9.2 nm) or 740 fs (BPF 7.5 nm), the extinction ratio was improved to at least 27 dB. This extinction ratio is comparable or larger than the demultiplexing ratio (12480/40 = 312), and is enough for the cross-correlation measurements of patterned signals in the following sections.



Figure 4.6: The auto- and cross-correlation traces of the mode-locked pulses from MLFL #2 and a schematic diagram of this measurement. (a): Auto-correlation trace. (b): Several cross-correlation traces of pulses from MLFL #1 and #2, taken within a few seconds. (c): Schematic diagram of the cross-correlation measurement in (b).

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Figure 4.7: Electric gate-signal generation in our sampling system. (a): OEconverted signal of the pulse from MLFL #2. (b),(c): Signals amplified by the RF amplifiers. (d),(e): Gate windows generated with the LN and EA modulators. Superposed gray curves indicate the 12.48-GHz pulse train to be modulated.



Figure 4.8: Auto-correlation traces and optical spectra of the ultrafast pulses obtained from the sampling pulse generator. (a) to (c): without BPF after the last EDFA, (d) to (f): with 9.2-nm BPF (solid) and 7.5-nm BPF (dashed). (a),(d): Short span, linear scale. (b),(e): Long span, logarithmic scale. (c),(f): Optical spectrum measured before the last band-pass filter. Resolution was 0.5nm.

4.2.3 Waveform monitoring of long-pattern 200-Gb/s optical signal

Before we perform 200-Gb/s operation of the DISC, we tested our sampling system with the 200-Gb/s input signal.

Figure 4.9 shows the cross-correlation trace of the 200-GHz clock pulses with 40-MHz sampling pulses. From the width of each pulse in the trace (1.5 ps), we estimated the time resolution of our measurement system to be slightly better than 1.5 ps. Constant background degraded the extinction ratio to ~ 6.5 dB (from 12.2 dB in the case of Fig. 4.4), but still we could observe a clear waveform.

Then we measured the patterned waveform of the 200-Gb/s signal, encoded at 12.48 GHz. Since the time span of our cross-correlator was limited to ~ 150 ps, we could measure only a short division of the 25000-ps-long pattern waveform in a single scan. By applying a bit shift to the encoding bit-pattern at 12.48 GHz for every single scan, we successfully monitored the long-pattern waveform over the whole period. The entire waveform will be found in the appendix. Figures 4.10(a) to 4.10(g) show typical examples of the patterned waveform in the long-pattern signal. In some sections, '1' bits and '0' bits appeared alternately. In some other sections, successive '0' or '1' appeared. Maximum length of the successive '0' in the whole period was eleven, and that for '1' was eight. Figure 4.11 shows the (temporal) correlation of the bit-pattern, defined as

$$C(\Delta k) \equiv \frac{\sum_{k=0}^{n_{period}} \left\{ (B(k) - \overline{B}) \times (B(k + \Delta k) - \overline{B}) \right\}}{\sum_{k=0}^{n_{period}} (B(k) - \overline{B})^2}$$
(4.1)

where B(k) represents the on-off of the k-th bit and \overline{B} represents the average of B. This result indicates that the bit-pattern we used was close to random, though not completely random as in the case of the widely-used 'pseudo-random' bit pattern. Note that we paid high attention to less-random parts (e.g. '00000000001') in the whole bit-sequence, rather than the randomized parts, in order to evaluate PIF.

Figure 4.10(h) shows an eye-diagram-like expression of the input signal, obtained using 158 piece of the patterned correlation traces *). Careful compensation of the

^{*)}Not the actual eye-diagram, since each dot in this diagram was obtained after averaging many data.



Figure 4.9: Cross-correlation trace of the 200-GHz clock pulses with 40-MHz sampling pulses. The gray line shows a background output, which was acquired by injecting only sampling pulses into the correlator.

timing drift between the traces has been achieved through data analysis, and we successfully obtained clear eye-opening.



Figure 4.10: Cross-correlation traces of the 200-GHz patterned signal with 40-MHz sampling pulses. (a) \sim (g): Examples of the traces measured in single scans, with certain bit-shifts. (h): Eye diagram of the 200 Gb/s input signal, calculated using the measured correlation traces.



Figure 4.11: Correlation of the multiplexed bit-pattern.

4.3 Wavelength conversion with DISC at 200 Gb/s

The DISC gate we investigated was almost same as that we used in §3.4.2. In view of the short pulse-width required for the 200-Gb/s operation, we replaced the 3nm bandpass filter after the SOA to that with 9.2-nm width, and the calcite with 5-ps DGD to that with 1-ps DGD. The wavelength of the cw light was changed from 1548.5 nm to 1540 nm because of the larger signal bandwidth. We spared the B#1 SOA sample for this experiment and used another sample with same structure instead (B#4). A half-wave plate was necessary before the MZI (fig. 4.1(c)), since our calcite module in this experiment was not rotatable as in the case of Chapter3. The MZI bias was set around π .

We operated the DISC gate with the 200-Gb/s signal described in the previous section, and monitored its output signal with the 40-MHz sampling system. Figures 4.12(a) and 4.12(c) show typical input and output waveforms. We see that the on-off patterns of those two were same, while the output waveform shows large PIF. Figures 4.12(b) and 4.12(d) show the input and output optical spectra. The center wavelength of the input signal was shifted to 1558 nm from that of the initial mode-locked pulses (1555 nm), through self phase modulation in the DDF. The center wavelength of the output was 1540 nm, determined by the cw light from the DFB-LD and the bandpass filter detuning. While the input signal spectrum



Figure 4.12: Typical input and output signals of the DISC. (a) Part of the input waveform (60 bits out of 4992 bits). (b): Spectrum of the input signal. (c) Corresponding part of the output waveform. The SOA injection current was 200 mA, the cw power into the chip was -5.3 dBm, and the pulse energy into the chip was 3.3 fJ. (d) Spectrum of the output signal.

looks continuous due to unstable characteristics of the MLFL, the output spectrum clearly shows discrete components separated by 200 GHz. 2.2-THz up-conversion of the carrier frequency was achieved.

Figure 4.12(c) indicates that the primary issue regarding the DISC operation is the PIF. Therefore we will focus on it from now. Our definition of PIF in this section is as follows: First, we assumed that the strongest output pulse appears around the bit sequence of '000010000000001'. (Fig. 4.13(a)) In this sequence, only one pulse per 75 ps was injected before the last bit. The measured intensity of the last output pulse is denominated by A. Second, we assumed that the weakest output pulse appears in the bit sequence of '11111111'. (Fig. 4.13(b)) The measured intensity of the weakest output pulse is denominated by B. Finally, we obtained the PIF from A/B. For comparison, '0-dominant' part and '1-dominant' part of the standard PRBS signal with $2^7 - 1$ -bit length are shown in figs. 4.13(c) and 4.13(d). We consider that our data-pattern should show more realistic pattern-dependence. Figure 4.14(a) shows crude estimation results of the output PIF dependence on the data pattern length for PRBS inputs. We calculated these using the same method in §3.4.2 for several cw-light intensities (and effective recovery rates). The SOA injection current, operation frequency, pulse width, and the MZI delay were chosen to be 400 mA, 200 GHz, 1 ps and 1 ps, respectively. Pattern length of $2^7 - 1$ seems enough when τ_{eff} is shorter than 10 ps. When τ_{eff} is around 20 ps, much longer data pattern with nine successive '0's will be preferable. Since the shortest recovery time of the SOA B#4 using a holding-beam injection at $I_{op} = 300$ mA was about 15 ps (Fig. 4.14(b)), we consider that the data pattern we are using is more suitable than the $2^7 - 1$ PRBS data.

The signal extinction ratio was similarly obtained from B/C, where C represents the signal background level. The background C in our measurements was fairly high. It was because we could not cut large ASE noise from the EDFA just before the cross-correlator input owing to a short of BPFs. We expect that the extinction ratio of of the DISC output before amplification was much higher.

In the following subsections, we will show systematically-measured results of the PIF and other properties.

4.3.1 Impact of the holding-beam effect

We first studied to how extent the intensity of the cw light affects PIF through the holding-beam effect. Figures 4.15(a) to 4.15(f) show measured output waveforms



Figure 4.13: Our definition of the pattern-induced intensity fluctuation (PIF) in this section. (a) '0-dominant' part of the DISC output waveform. (b): '1-dominant' part of the DISC output waveform. PIF is obtained by A/B. (c) and (d): Standard PRBS signal waveform with $2^7 - 1$ -bit length.



Figure 4.14: Crude estimation of the pattern-length effect on the PIF. (a) calculated PIF dependence on the length of the input PRBS data pattern. (b): measured example of the gain-recovery profile of the SOA B#4 ($\tau_{\rm eff} \sim 15 \text{ ps}$).

and spectra at three cw powers of -10.1 dBm, +3.8 dBm, and +11.2 dBm. The injection current I_{op} was 300 mA, and the control pulse energy into the chip was 7.7 fJ. The amount of the nonlinear phase shift was estimated to be 0.2π (for +11.2dBm cw input). The inset of each figure shows the eye-diagram-like expression, each containing 32 bits. We could suppress the PIF down to 2.7, using as high cw power as +11.2 dBm (fig. 4.15(e)). The increase of sub-pulse intensity, which we predicted in Chapter 3, seemed to cause broadening of the output pulses to ~ 3.4 ps. Honestly speaking, the impact of the sub-pulses generation seemed somewhat different from what we had expected through Chapter 3. This probably resulted from large contributions of the carrier-heating related phenomena, which were intentionally suppressed in the previous measurement. Since we could still distinguish each bit from the output waveform, the result of this effect was not serious in this case. Figures 4.16(a) and 4.16(b) show the dependences of the PIF and the average output power from the DISC on the cw input power, at $I_{op} = 300 \text{ mA}$, 200 mA and 150 mA. We see that the impact of the cw power is extremely large. This explain why as strong cw power as +16 dBm (into module) was required in the past 168-Gb/s wavelength conversion. When the injection current is 200 mA, output power was 4-dB lower than that for $I_{\rm op} = 300$ mA, and when the injection current is 150 mA, output power was 10-dB lower than that for $I_{\rm op} = 300$ mA.

4.3.2 Impact of the filter detuning

Unless we also use other effects to suppress PIF than the holding-beam effect, we have to consume considerable electric power for the holding-beam amplification in the ultrafast optical gating. Therefore we investigated such effects.

Recently, some researchers have come to utilize detuning of the center wavelength of the bandpass filter in ultrafast gating. This filter-detuning technique was theoretically proposed by Nielsen *et. al.*[40] Our collaboration team studied the systematic change of the waveform on the detuning in 2005, and explained the measured patterning to a certain degree [42, 43]. The operating condition of this experiment was, however, fairly limited in some senses.

1. The frequency and the pulse width of the optical signal were only 43 Gb/s and 2 ps in ref. [42], respectively. Those in ref. [43] were 105 Gb/s and 1.8 ps, respectively. More contribution of ultrafast phenomena in the SOA should be observed at higher frequency.



Figure 4.15: DISC output signals on several holding-beam intensity. Solid and dashed curves show the '1'-dominant and '0'-dominant part of the waveform. (a): -10.1 dBm, (b): +3.8 dBm, (c): +11.2 dBm. The insets show eye-diagram-like expressions. (d) to (f): corresponding optical spectra.



Figure 4.16: Systematic results on the impact of the holding-beam injection. (a): Dependence of the PIF on the holding-beam intensity, at $I_{\rm op} = 300$ mA (solid) and 200 mA (dashed). (b): Dependence of the average output power on the the holding-beam intensity.

- 2. The discussion in ref. [43] is focused on 'SOA+detuned filter' gate rather than the standard DISC. (SOA+MZI+detuned filter). Their characteristics are similar to each other, but not identical. The latter one has been used more widely.
- 3. The adopted bit-patterns were '0011' in ref. [42] and '0000001111' in ref. [43], only.
- 4. Strong holding-beam was not used, and hence two '1111' might not be separated in ref. [43].

Similar and advanced approach has been also investigated [44], but the operation frequency has been still limited to 40 Gb/s.

Apart from our fundamental study, Liu *et. al.* demonstrated error-free 160-Gb/s and 320-Gb/s wavelength conversion using this technique[25, 26]. They succeeded in the wavelength conversion of $2^7 - 1$ -bit PRBS signal, with much lower cw power than that used in ref. [23]. Systematic, detailed waveform information including PIF, sub-pulse generation and so on, however, has not been reported.

Thus we studied the impact of the filter detuning on the output waveform, in practical operation conditions. In view of the result of §4.3.1, we operated the gate on two conditions: with strong (+11.2 dBm) cw power and moderate (+3.8 m)dBm) cw power. We detuned the 9.2-nm bandpass filters, one just after the MZI and another after the EDFA, from the wavelength of the input cw-light (1540 nm). Figures 4.17(a) and (b) show the output waveforms and spectra on several detuning conditions, with +11.2-dBm cw-light power. When the filters were detuned to the blue side by 4.8 nm, the PIF was suppressed from 2.7 to 1.6. In this case, the cw component was just around the edge of the passband, and diminished by 3.5 dB on each filter. The decrease of the average output power by this detuning was only 1.8 dB. Farther detuning lead to sub-pulse generation and output degradation, and hence not advantageous. While detuning to the blue side did not distort the waveform much, detuning to the red side caused complicated waveform distortion. It was surprising that only 0.7-nm detuning caused large sub-pulses as shown in the figure. When the detuning to the red side reached 4.2 nm, the waveform distortion disappeared and slight improvement of the PIF was observed. This effect will not be as useful as that for the blue shift, since it caused larger decrease of the output power (4.2 dB) and smaller signal-wavelength conversion. Figures 4.18(a) and (b) show the output waveforms and spectra on several detuning conditions, with +3.8dBm cw-light power. The effect of the filter detuning on the output waveform was quite similar to that in the previous case. Detuned to the blue side by 4.8 nm caused the suppression of the PIF from 6.6 to 2.2, with 1.6-dB output power loss. This result indicates that the filter detuning technique largely contributes to the reduction of the cw-power requirement, and the total energy consumption. Figures 4.19(a), 4.19(b), and 4.19(c) show the dependence of PIF, average output power, and extinction ratio of the output signal on the filter detuning. Then the impact of the filter detuning on the PIF was demonstrated in more practical, high-frequency condition than in ref. [42]. The optimum filter detuning in terms of PIF suppression was 4.8 nm. In such case, the wavelength of the cw light lies just on the edge of the BPF transmission window. The optimum detuning in terms of the extinction ratio was slightly shifted to the red side, and the optimum detuning to minimize bit errors will be somewhere between them. Figures 4.20(a) to (c) show the dependence of the PIF on the HB intensity when the BPF is blue-shifted by 4.8 nm. These results show that required HB power was reduced by ~ 3 dB compared to the case without BPF detuning. In each case the reduction of the output power associated with BPF detuning was small (Fig. 4.20(d)).

4.3.3 Impact of the nonlinear polarization rotation

In this section, we show that there is another effective method to suppress PIF of the DISC. This method is based on the nonlinear polarization rotation (NPR) of the cw light in the SOA[63–67]. This effect has been used for some all-optical signal processing at relatively low frequency[68–77], but nobody has suggested that the NPR can be utilized for much higher optical gating when combined with the delayed-interference scheme.

In Figs. 4.21(a) to 4.21(b), the original scheme of the DISC operation and that using the NPR are shown in contrast. When the polarization of the cw light is aligned to TE (or TM) axis of the SOA at the input facet, each control pulse does not induce any NPR to the cw light (Fig. 4.21(a)). The polarizer in front of the MZI acts only for reduction of the ASE from the SOA. The output waveform of the DISC is approximately proportional to $\sqrt{G(t)G(t-\Delta t)}\cos^2(\delta\Delta\Phi(t)/2)$ (see eq. (3.9) in §3.2), and PIF is induced mostly by the gain modulation, $\sqrt{G(t-\Delta t)}$. Actually, the polarization of the cw light is not aligned to TE in most cases, and it rotates on every pulse injection. If the polarizer in front of the MZI is properly adjusted, this NPR should cause instantaneous increase in the transmittance of the polarizer and suppress PIF (Fig. 4.21(b)).



Figure 4.17: DISC output signals on several BPF detunings, with strong (+11.2 dBm) holding-beam injection. (a): Monitored waveforms. Solid and dashed curves show the '1'-dominant and '0'-dominant part of the waveform. (b): Optical spectra. The dashed line shows the transmittance spectrum of the 9.2-nm BPF.

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Figure 4.18: DISC output signals on several BPF detunings, with moderate (+3.8 dBm) holding-beam injection. (a): Monitored waveforms. Solid and dashed curves show the '1'-dominant and '0'-dominant part of the waveform. (b): Optical spectra. The dashed line shows the transmittance spectrum of the 9.2-nm BPF.



Figure 4.19: Systematic results on the impact of the BPF detuning. (a): Dependence of the PIF on the BPF detuning, with strong (+11.2 dBm) and moderate (+3.8 dBm) holding-beam injection. (b): Average output power. (c): Extinction ratio.



Figure 4.20: Reduction of HB power for PIF suppression by the BPF detuning. (a) to (c): PIF dependence on the HB intensity with and without BPF, at $I_{\rm op} = 150 \sim 300$ mA. (d): Dependence of the average output power on the HB intensity, with and without BPF detunings.



Figure 4.21: Schematic images of the suppression mechanism of PIF by NPR in the DISC. (a): case with no NPR. (b): case when NPR suppresses PIF.

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We demonstrated that such suppression of PIF is actually possible. Figures 4.22(a) to (d) show the waveforms of the DISC output, with different polarization settings. We aligned the polarization of the cw light (and that of the control pulses) at the input facet of the SOA to $+30^{\circ}$ from its TE axis, using the polarizationmaintaining lensed fiber. We also aligned the polarization at P_1 to its axis, using the wave-plates Q_1 and H_1 . The cw-input power, control pulse energy and injection current to the SOA were same as those used for the measurement in Figs. 4.18, and no detuning was given to the BPF. The PIF in this case was 5.1. After that, we the H₁. As the rotation angle $\Delta \theta_{H_1}$ approached to +42°, we observed that the PIF was drastically reduced down to 1.5, as shown in Fig. 4.22(b) to 4.22(c). The DISC output signal was almost extinguished around $\Delta \theta_{H_1} = 45^{\circ}$, and the PIF increased again on farther wave-plate rotation (Fig. 4.22(d)). We systematically measured the dependence of the PIF on $\Delta \theta_{H_1}$, at several input polarization to the SOA. The results are shown in Fig. 4.23(a). $\Delta \theta_{H_1}$ was defined separately for each input polarization series, so that the cw-light without optical modulation went through P_1 most at $\Delta \theta_{H_1} = 0^{\circ}$. When the input polarization was TE or TM, we could not see significant change of the PIF. When the input polarization was $+30^{\circ}$ off from the TE, in contrast, the PIF was reduced down to $1.5 \sim 3$ around $+32 \sim 42^{\circ}$. Similar reduction of the PIF was obtained when the input polarization was $+45^{\circ}$ off from the TE, though within narrower $\Delta \theta_{H_1}$ range. We regard these results as a proof of PIF reduction caused by NPR.

Thus we demonstrated a new method to suppress PIF, in addition to the use of the holding beam and filter detuning. This method will be as important for the practical application of the DISC as the filter detuning technique, since it requires no additional device from the original configuration. Unfortunately, this method alone will not provide a perfect solution for the wavelength conversion due to some drawbacks. One of such drawbacks is the large reduction of the output power, as shown in Fig. 4.23(b), compared to that for the BPF detuning method. When we obtained the waveform in Fig. 4.22(c), the average output power decreased from -15.3 dBm to -26.4 dBm. This decrease is also apparent in the optical spectra, shown in Figs. 4.22(e) to 4.22(h). Nonetheless, the output waveform in 4.22(c) indicates that its signal-to-noise ratio was still good. When we optimized the PIF, the extinction ratio was not seriously degraded (4.23(c)). Therefore this method should be used together with the other methods, in order to optimize the performance of the DISC. For that, farther systematic measurements and theoretical understanding of this effect will be quite valuable.



Figure 4.22: Examples of the waveforms and spectra of the DISC output with suppression of the PIF by NPR. Input polarizations for the SOA are aligned to $TE+30^{\circ}$. (a) to (d): Waveforms with different half-waveplate angles. Solid and dashed curves show the '1'-dominant and '0'-dominant part of the waveform. (e) to (h): corresponding optical spectra.



Figure 4.23: Dependence of the DISC output properties on the half-waveplate angle setting. (a): PIF. (b): Average output power. (c): Extinction ratio.

4.4 Conclusion

We performed 200-Gb/s wavelength conversion of long-bit-pattern optical signal using a DISC, and searched for the requirement on its operating condition to reduce the output waveform degradation. Signal-waveform monitoring was realized using the cross-correlation method and the specially-developed ultrafast (~ 600 fs), low-repetition rate (40 MHz), low-timing jitter, synchronized sampling pulse generator. The time resolution of our monitoring system was about 1.5 ps. We measured the PIF of the DISC output waveform for a practical data sequence, and studied the impact of the holding-beam injection and BPF detuning on its suppression. To the author's knowledge, this is the first systematic report on the impact of BPF detuning in the frequency range over 160 Gb/s. We confirmed that the BPF detuning suppressed the PIF of the DISC at 200 Gb/s, in a similar way to that in the lower frequency case (43 Gb/s)[42].

As long as we use only holding-beam effect, approximately 300 mA (580 mW) dc power consumption and fairly strong holding-beam power ($\sim +11.2$ dBm into the SOA chip) were required to reduce the PIF down to 2.7. This power-consumption value is reasonably close to that calculated for 160-Gb/s operation of the SOA gates in Chapter 2. When we used detuned BPFs after the SOA, we could suppress PIF with ~ 3 dB less HB power. Therefore BPF-detuning technique is promising to reduce the total power consumption of the SOA-based optical gate.

In addition to these two methods, we also discovered that another method using the nonlinear polarization rotation in the SOA is available to reduce PIF. We demonstrated that this technique is available when the input polarization to the SOA is $30 \sim 45^{\circ}$ off from TE axis, and suppression of the PIF down to 1.5 (from 5) could be achieved with no additional power consumption through precise polarization control. The required accuracy of the polarization control will be of the order of $\sim 5^{\circ}$. We believe that this new method should also contribute to reduce the power consumption of SOA-based future all-optical gates. Thus farther study on this effect will be quite worthful.

Chapter 5

Conclusions and outlook

This chapter gives a summary of the main conclusions of this thesis and takes a look at the prospective contributions to the future ultrafast all-optical gating research.

5.1 Conclusions of this thesis

The ultrafast all-optical signal processors based on SOAs have been expected as key elements in future digital networks for a long time, but the lack of their fundamental understanding in many aspects have often caused difficulty in their practical operations and system designs. Especially, minimum requirements on the operating conditions such as the electric power consumption have not been clarified so far. In this thesis, therefore, we aimed to characterize and model the dynamics of SOAs and SOA-based optical gates to facilitate their applications.

We discussed a new model in Chapter 2, which determines the required amount of electrical power consumption of SOAs in XPM-based all-optical gates as a function of the gate operation frequency B (10 ~ 160 GHz). In this model, three different carrier-conversion efficiencies in the SOA were reasonably introduced to account for the different gain-saturation characteristics against low-frequency optical signals and ultrafast optical signals. We predicted that the required amount of the power consumption increases as B^2 (or even faster, depending on the decrease in the conversion efficiencies and differential gain) and decreases with the total conversion efficiency $\eta_{\rm T}$ as $\eta_{\rm T}^{-2}$ in the high-frequency limit. From the systematically-measured results for the actual SOA samples with different structures, we revealed that long SOAs with an MQW structure had high conversion efficiency at high injection current and required low power consumption. The estimated power consumption value for our best SOA (B#1) was ~ 750 mW at 160 GHz. We performed the experimental

verification of the power consumption calculation and obtained some agreement in the frequency range up to 40 GHz, though farther need of improvement for higher frequency region. Thus we made a first step to predict, design, and improve the electric power consumption in the SOA gates. We believe that optimizations of SOAs for arbitrary operation frequencies in various optical gates become possible in future with more comprehensive understanding of the conversion efficiencies.

In Chapter 3, we discussed the waveform degradation issue regarding DISCtype wavelength converters due to sub-pulse generation originated from exponential recovery profiles of the SOA carriers. This issue might have appeared in the past gate experiments and have made the gate operation difficult. We precisely observed and theoretically characterized such generation for the first time, in the frequency range of $10 \sim 25$ GHz using the cross-correlation method. Measured sub-pulse properties on the relative intensity, waveform dependence on the MZI phase bias, and intensity increase with the SOA carrier recovery rate were explained using our DISC-gate model quite accurately. Thus we concluded our present gate model to be a powerful tool to cope with this issue, at least in the frequency range where carrierheating related phenomena do not appear much. Based on this model, we predicted the trade-off between the sub-pulse and the patter-induced intensity fluctuation. Though this issue has not been solved at this moment, we believe that farther advance in the gate structure supported by our gate-model simulation will provide a fundamental solution.

In Chapter 4, we studied the waveform degradation, mostly the pattern-induced intensity fluctuation (PIF), of the output signal from the DISC operated with 200-Gb/s, long-bit-pattern (4992 bits) input signal and several methods to suppress it. In this study, we used our original waveform-monitoring system for the long-patterned signal. When we did not detune the BPF after the SOA (and did not use the nonlinear polarization rotation), fairly strong holding-beam injection and the dc current injection of ~ 300 mA (580 mW) or more was inevitable as estimated in Chapter 2. Then we systematically measured the impact of BPF detuning, and fundamentally confirmed that it effectively suppressed the PIF of the DISC even at 200 Gb/s and for longer bit pattern (than $2^7 - 1$, used in ref. [23], [25] and [26]), in a similar way to those in the lower frequency (43 Gb/s), very short-bit-pattern case [42]. This measurement enabled us to evaluate the capacity of the BPF detuning to mitigate the required holding-beam power (~ 3 dB). Farther study will contribute to the designing of DISC operation condition with low power consumption. We also discovered that another method using the nonlinear polarization rotation (NPR) in

the SOA is available to suppress PIF. Suppression of the PIF from 5 down to 1.5 could be achieved with no additional power consumption. We clarified the way to suppress PIF using this method. This method will be possible with polarization-control accuracy of ~ 5°. We believe that this new method should also contribute to reduce the power consumption of SOA-based future all-optical gates.

Through these activities, we developed the understanding of the SOA gate (DISC) operated in the frequency range of $10 \sim 200$ GHz, and considerably facilitated its low-power, small signal-degradation operation and the system design.

5.2 Future outlook

Several kinds of all-optical signal processing have been investigated widely to realize low power, ultrafast signal processing. It has been widely approved that those based on SOAs are in the nearest position to the practical use. Much potentiallyfaster nonlinear devices as quantum-dot SOAs[82–85], photonic crystal/quantumdot waveguides[86], coupled quantum-well waveguides utilizing intersubband transition (ISBT)[87–90] and so on have been also expected and intensively studied, however, for later use. The future of the SOA gates and other gates is, therefore, an interesting topic for us.

In Chapter 2, we calculated the dc power consumption in SOA gates. We roughly deduced that power consumption increases with B^2 . This means that the energy consumption per one data-bit increases with B, as shown in Fig. 5.1(a). Meanwhile, those of the other optical gates obey different relations. For optical gates using highly-nonlinear fibers, the only but large source of the power consumption is the generation of high-intensity optical signals. Since the nonlinear effect in fibers depends on the peak intensity of the optical signal, the average power of the signal required for the gating depends rather on the duty ratio than B. This means that the energy consumption of the fiber gates per one data-bit is inversely proportional to B (as long as we do not shorten the fiber length), in contrast to that for the SOA gates. We show the estimated power consumption of fiber gates also in Fig. 5.1(a). Here the required signal (and probe) power is assumed to be +29 dBm (and +18dBm) at 80 \sim 160 GHz (ref. [16, 17], HNLF length = 1 m, nonlinearity $\gamma = 1100$ $W^{-1}km^{-1}$). The dc electric power consumption value to generate this optical signal is conservatively estimated to be 8.6 W, in view of the power conversion efficiency $(\sim 10\%)$ of a typical EDFA (Fig. 5.1(b)). Then the estimated power consumption curves for both kinds of gates cross around 600 GHz. SOA-gates seem to keep



Figure 5.1: Comparison of the power consumption in all-optical gates, and an EDFA property. (a): Energy consumption per one data-bit against operation frequency. (b): Typical dependence of EDFA output power on the dc power consumption.

advantages for a while, but the power consumption in SOAs is already substantial and should be seriously considered. Our result will motivate the researches on other passive ultrafast nonlinear devices, which require constant energy per one bit in the gating. For comparison, the estimated energy consumption of the wavelength converter based on the ISBT in the InGaAs/AlAs/AlAsSb coupled quantum wells[87] was also plotted in the figure. So far the required power consumption is fairly high compared to SOA gates or fiber gates. If reduction of the power consumption by factor 10 can be achieved, this gate will be competitive enough.

For the activities to realize much faster SOAs or active nonlinear devices, our result suggests certain review. So far most studies on the optimization of the SOA structure have aimed for the increase of the recovery rate[39, 48, 91, 92]. Our conclusion based on eq. (2.15) is, however, that the increase in the recovery rate itself does not contribute to reduce the power consumption^{*}). We expect that more efforts will be spent on the improvement of the SOA power consumption efficiency after our research.

Except for Chapter 2, we focused only on the DISC-type wavelength converters. Nonetheless we consider that our success in the accurate modeling of the gate output waveform is quite valuable for researchers of other SOA gates. It is true that there have been plenty of numerical simulation researches for many kinds of SOA gates. Some researches employed much complexed models in the calculation. The present

^{*)}Roughly speaking, an SOA with a fast recovery time and also a high gain is likely to require low power consumption.

author has an impression, however, that most calculated results have not been accurately verified. For some cases, the present author expect that much improved results can be obtained by applying our SOA-gate model with properly-estimated carrier density values. The most interesting application will be the analysis of 2R or 3R signal regeneration[56] over 40 GHz.

The all-optical switching devices have not been widely commercialized. For the wavelength converter or the signal regenerator (2R), so far only a few companies have produced integrated SMZ devices with guaranteed operation frequencies of up to 40 GHz[†]). It is partially because the industrial needs have not been arisen for the immaturity of some other gate functions as optical memories, and partially because of the difficulty in the application of these gates themselves. We are confident that our research reduced the difficulties in realizing much faster (~ 160 GHz) or much low-power commercial switching devices. For the ultrafast wavelength converters, integration of the cw-light sources with proper polarization alignment will be effective. This will support the future realization of large capacity OTDM-WDM mesh networks and associated systems (wireless systems beyond 3G, etc.).

 $^{^{\}dagger)} \rm Examples$ are: Alphion. Inc., QLight^{\rm TM} ISM, Center for Integrated Photonics Inc., 40G-2R-ORP, etc.

APPENDIX A. ENTIRE WAVEFORM OF THE 200-GB/S INPUT SIGNAL

Appendix A

Entire waveform of the 200-Gb/s input signal










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List of publications

• Journal papers related to this thesis

- J. Sakaguchi, M.L. Nielsen, T. Ohira, R. Suzuki, and Y. Ueno, "Observation of small sub-pulses out of the delayed-interference signal-wavelength converter," Jpn. J. Appl. Phys. vol. 44, no. 44, pp. L1358-1360, Oct. 2005.
- J. Sakaguchi, F. Salleras, K. Nishimura, and Y. Ueno, "Frequency-dependent electric dc power consumption model including quantum-conversion efficiencies in ultrafast all-optical semiconductor gates," Optics Express vol. 15, no. 22, pp. 14887-14900, Oct. 2007.
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- 4. J. Sakaguchi, T. Nishida, and Y. Ueno, "200-Gb/s wavelength conversion using a delayed-interference all-optical semiconductor gate assisted by nonlinear polarization rotation," submitted to Optics Communications.
- Journal papers (co-author)
 - M.L. Nielsen, J. Mørk, R. Suzuki, <u>J. Sakaguchi</u>, and Y. Ueno, "Theoretical and experimental study of fundamental differences in the noise suppression of high-speed SOA-based all-optical switches," Optics Express vol. 13, no. 13, pp. 5080-5086, June 2005.
 - M.L. Nielsen, J. Mork, R. Suzuki, <u>J. Sakaguchi</u>, and Y. Ueno, "Experimental and theoretical investigation of the impact of ultra-fast carrier dynamics on high-speed SOA-based all-optical switches," Optics Express, vol. 14, no. 1, pp. 331-347, Jan. 2006.

- 3. Y. Ueno, R. Nakamoto, <u>J. Sakaguchi</u>, and R. Suzuki, "Optical-spectrumsynthesizer design within an all-optical semiconductor gate to reduce waveform distortion induced by carrier-cooling relaxation at sub-Teraherz frequencies," Optics Express vol. 14, no. 26, pp. 12655-12664, Dec 25, 2006.
- Contributions to international conferences (first author)
 - J. Sakaguchi, M.L. Nielsen, T. Ohira, R. Suzuki, and Y. Ueno, "Observation of small sub-pulses generated in the DISC-type wavelength converter," 10th Optoelectronics and Communications Conference (OECC 2005), Seoul, Korea, July 5-8, 2005, paper 8F1-4.
 - J. Sakaguchi, Y. Ueno, K. Nishimura, and T. Yazaki, "New method for characterizing the injected-carrier-to-photon conversion efficiencies inside the ultrafast all-optical semiconductor gates," 9th International Symposium on Contemporary Photonics Technology (CPT 2006), Tokyo, Jan. 11-13, 2006, paper no. P-26, pp. 130-131.
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- Contributions to international conferences (co-author)
 - 1. M.L. Nielsen, J. Mork, <u>J. Sakaguchi</u>, R. Suzuki, and Y. Ueno, "Reduction of nonlinear patterning effects in SOA-based all-optical switches using

optical filtering", Optical Fiber Communication Conference (OFC 2005), Los Angeles, March 6-11, 2005, paper OThE7.

- R. Suzuki, S. Kobayashi, <u>J. Sakaguchi</u>, and Y. Ueno, "Threshold condition for pulse generation from a DISC-loop-type pulse generator," IQEC/CLEO-PR, July 11-15, 2005, Tokyo, paper CMF1-4, pp. 1522-1523.
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- 4. F. Salleras, M. Honma, J. Sakaguchi, Y. Ueno, N. Ozaki, Y. Kitagawa, K. Asakawa, N. Ikeda, Y. Sugimoto, "Criterion for removing a delayed peak from the ultrafast nonlinear response of photonic crystal/quantum dot waveguides," 1st Nonlinear Photonics Topical Meeting (NP 2007), Optical Society of America, Quebec city, Canada, Sept. 2-6, 2007, paper no. NTuC6.
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