

High Power Laser Diode Driver Based on Power Converter Technology

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Abstract— This paper describes the design of a high speed semiconductor laser diode driver designed for driving 500 mW to 1.5 W diodes at full optical power modulation up to frequencies of 10 MHz. The duty cycle of the modulation may be varied. A switching power-converter based current source allows a higher power delivery efficiency to the diode than in previous designs, allowing for a more modest power supply and dissipation requirements. A dynamic ripple cancellation circuit reduces the power converter output current ripple to less than 1% of full-scale current. The circuit is capable of delivering up to 2.5 A to a laser load, with a 10–90% switching risetime from laser threshold to full on of less than 20 ns.

I. INTRODUCTION

AS A RESULT of progress in semiconductor laser technology, laser diodes are now available with continuous-wave (CW) optical output power greater than 1 W. Due to the inherent high speed of the lasing process, direct modulation of the laser current is an attractive method of high speed modulation of the laser power output, and high speed current drivers for these devices are needed. This allows high power lasers diodes to be used in new applications in communications and medical imaging.

Previous designs for high-speed modulation of lasers are suitable for driving low power diodes (less than 50 mW optical power output) [1], [3], [5]–[8], [10]–[12], [17], [19], [21], [24], [26]. These designs are based on linear amplifier techniques. For these lower powered lasers, switching currents are on the order of 100 mA or less and hence power handling requirements are minimal. For driving higher-powered lasers, the low efficiency of these topologies become important. Also, more care is needed in the circuit design because of the higher current slew rate required. A high speed, lower-loss topology is considered in this paper.

In this paper, a high speed current driver for direct modulation of laser diodes is described. An adjustable dc current source is generated from the power supply voltage using a switching down-converter technique. This current is switched to and from the laser under control of an externally applied TTL signal. A second adjustable dc current source pre-biases the laser diode to the lasing threshold. An active ripple reduction circuit compensates for the ripple current at the converter switching frequency.

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The laser driver is designed to operate in the Polaroid Corporation's Helios Medical Laser Imaging System. The circuits described were built using standard PC board and surface-mount component technologies. The work is further described in U.S. Patent #5 444 728 (8/22/95).

II. SYSTEM OVERVIEW

Electrically, a semiconductor laser behaves as a diode with a voltage turn-on knee greater than 1 V. When the laser is biased with a sufficient current so that lasing occurs, a further increase in injected current results in a proportional increase in laser optical output. For representative laser diodes, the frequency response of optical power output due to a change in laser current is flat out to a resonance at a very high frequency (on the order of 0.2 to 2 GHz) [2], [9], [18], [20], [29]. This resonance is due to laser relaxation processes.

For the particular application considered in this paper, the diode is to be operated at a maximum modulation frequency of a few megahertz, with a risetime (depending on laser current set-point) of approximately 10–20 ns. The laser diode will be square-wave modulated at any duty cycle, controlled by a TTL input signal. This modulation frequency is much lower than the laser's resonant frequency, so the resonance is not excited significantly. The light output response exhibits little overshoot with a diode intrinsic risetime on the order of a nanosecond. The ultimate limitations in achievable switching speed are due to the parasitic inductances and capacitances of the laser package and the risetime of the current driver.

In order to maintain constant optical power output, it is desirable to make the injected laser current independent of temperature. Therefore, the laser is driven with a high impedance current source. The laser is also fitted with a thermo-electric (TE) cooler so as to maintain constant laser chip temperature so that optical power remains constant at constant current [13]. Optical feedback is carried out by using a photodetector and relatively slow feedback loop to detect and compensate for optical power output variations as the laser ages [4], [14], [16], [22], [25]. The details of the TE cooler design and optical feedback are not considered here.

The laser current is modulated between two pre-set values I_{th} and I_{pk} . The threshold current I_{th} bias the laser to a point where it just begins lasing. I_{pk} is the current necessary to generate a given optical output power. For a 1 Watt semiconductor laser, representative values for I_{th} and I_{pk} are 300 mA and 1.2 A, respectively. Due to device manufacturing process variations, these values vary considerably from device to device and change somewhat as the device ages.

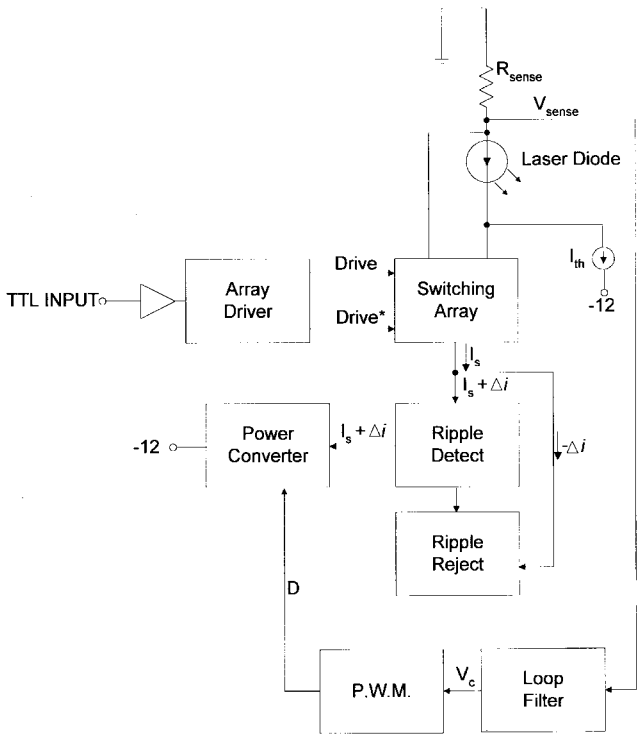


Fig. 1. Laser driver system block diagram.

Referring to the block diagram (Fig. 1), the laser modulation is under control of an input TTL signal. This signal causes rapid switching of the switching array. When the laser is on, laser current I_{pk} flows through the diode. When the laser is off, only the pre-bias current I_{th} flows through the diode, with I_{pk} being shunted around the laser.

The average value of total laser current is sensed. The loop filter and pulse-width modulator maintain the necessary average current. The ripple rejection circuit measures and compensates for the ripple in the current source output.

Reasons for biasing the diode at a preset threshold level I_{th} in the off state are as follows [23].

- 1) The bias current reduces the turn-on time due to charging of parasitic diode capacitances.
- 2) The pre-bias current reduces the thermal variations that the laser chip undergoes.
- 3) It reduces any laser relaxation oscillations.

Although biasing the laser at I_{th} does result in some spontaneous emission of laser light, this level of prebias current is acceptable since the stimulated emission at a current level of I_{pk} is much greater.

The current pulses to the laser need fast rise and fall times and need to have well controlled amplitudes with little overshoot over any duty cycle. A difficulty in achieving this lies in the fact that there is inductance in the high-speed current switching path. This inductance is comprised of package inductance inside the laser, inductance of the interconnection between PC board and laser, and parasitic printed-circuit board inductance. This total distributed inductance will interact with the parasitic capacitances at the collectors of the switching transistors. Since the resistance of the laser when on is relatively low ($<1 \Omega$) the damping of this resonant circuit is

 TABLE I
 LASER DRIVER SPECIFICATIONS

Power supplies	+12 V, -12 V \pm 5%
Threshold current	$0 < I_{th} < 500$ mA
Laser peak current	750 mA $< I_{pk} < 2.5$ A
10–90% switching risetime	$\tau_R < 20$ ns
Converter switching frequency	800 kHz
Laser current ripple	< 20 mA, <i>p-p</i>
Laser ON resistance	$< 0.5 \Omega$
Laser ON voltage	< 3 V @ full current

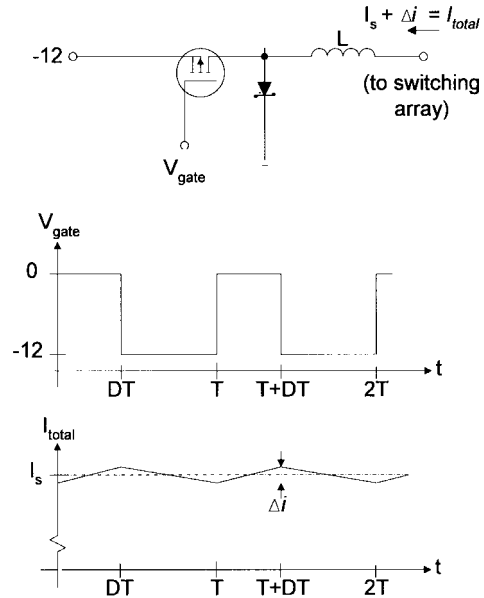


Fig. 2. Power converter arranged as a current source.

low. Therefore, every effort is made to reduce the inductance of the printed-circuit board routing and of the interconnection path.

III. DESIGN SPECIFICATIONS

The laser driver is designed with standard small-outline integrated circuit (SOIC) packages, with the emphasis that no special or expensive components be used.

IV. OPERATION IN DETAIL

A. Power Converter Current Source

The laser diode, when it is biased in its linear region at maximum laser current, is specified to have less than 3 V across it. This voltage is a combination of the diode turn-on voltage and the IR drop across it. Since the laser is powered from a -12 V supply, a linear amplifier would have significant power loss during modulation. Substantial power is saved by using a switching down-converter (Fig. 2) to generate a dc current at a lower voltage.

A converter switching frequency of 800 kHz was chosen. The choice of this frequency was a tradeoff between maximum ripple current, system noise, switching losses, and available inductor size. One important constraint for this design was the availability of surface-mount inductors. A commercially avail-

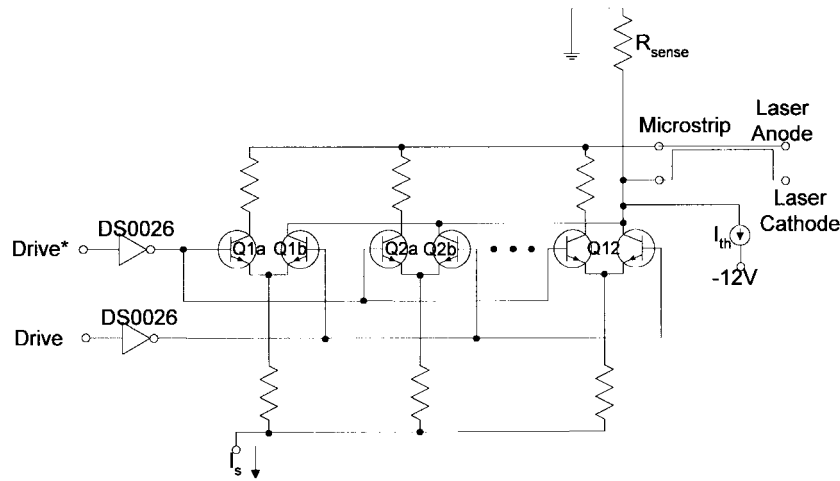


Fig. 3. High-speed switching transistor array designed to modulate laser.

able Coilcraft surface mount $47 \mu\text{H}$ inductor with sufficient current-carrying capability was chosen.

As shown in Fig. 2, the output current of the converter has a dc value with some ripple at the switching frequency. For the ideal converter, the peak-to-peak ripple in inductor current is given by

$$\Delta i = \frac{V_s D}{f_{sw} L} \quad (1)$$

where V_s is the power supply voltage, D is the converter duty cycle, and f_{sw} is the converter switching frequency. The duty cycle is approximately 0.4, due to 3 V maximum across the laser and 1.8 V across the array transistors and ballast resistor.

The open-loop behavior of the converter is dominated by a single pole at approximately 4 kHz, which is due to an L/R time constant of the converter inductor interacting with various circuit resistances. A control loop which includes an integrator with a zero allows for zero dc current error, with a loop transmission crossover of 60 kHz and a phase margin of approximately 50 degrees.

B. Current Switch

The current switch (Fig. 3) forms the high speed switching element which modulates the laser current under control from an externally applied TTL signal. An inherent feature of the nonsaturating current switch is its high switching speed. With bipolar transistors, switching speeds on the order of $1/\omega_T$ are readily achieved for the emitter-coupled switch with proper base drive. The open-loop current switch was chosen over a closed-loop current source because of the trade-off between bandwidth and power-handling capability in closed-loop designs, with larger devices necessarily having higher input capacitance.

A second function of the current switch is to isolate the power converter inductor from the relatively large voltage swings at the output of the laser. The closed-loop bandwidth of the control loop is approximately 60 kHz, but the laser is expected to be switched at frequencies up to several MHz. A large switching voltage seen by the inductor would result in a relatively large transient current change which would be

dependent on duty cycle. Since the transistor array is being switched differentially, the transient voltage "seen" by the power converter inductor is relatively small. This eliminates the need for other mitigating strategies, such as feed-forward, to reduce transient changes in laser current. In this instance, some loss of power delivery efficiency in the emitter coupled switch is sacrificed for control-loop simplicity.

The switch is implemented as an array of 12 MMPQ2222, A switching transistor pairs, connected in parallel. Each transistor is required to carry up to 200 mA at maximum laser current. The f_T of these devices at this collector current is approximately 100 MHz. Therefore, the intrinsic switching time of each array element is only a few nanoseconds. An advantage of using multiple small devices in parallel is that the effects of inductance are minimized, while maintaining significant power handling capability. If one large switching device were used, the few nanoHenries of package inductance ($\approx 10 \text{ nH}$ per pin) would be an ultimate limitation in switching speed. Use of the SOIC package has the effect of several smaller inductances in parallel, so that the effective switching inductance is less than 2 nanoHenries.

A second advantage to using multiple SOIC is power handling capability. To first order, thermal resistance is inversely proportional to the surface area of the SOIC packages. For multiple devices in parallel, this surface area is greater than that of a single power device, and the necessary power is handled without need for a heat sink.

An input buffer (not shown in the figure) level shifts the TTL input signal to a negative reference of -8.3 volts. The bases of the array transistors are driven differentially by a DS0026 gate driver, which operates with power supply voltages of -3.3 V and -8.3 V . Small emitter ballast resistors ensure that current shares equally between the switches.

The DS0026 is a high current output buffer capable of driving up to 1.5 A to a 1000 pF load with a risetime of 20 ns. Due to the high f_T of the switching devices, the primary limitation in current switching speed is the rate at which charge can be supplied to the base regions of the transistors. Using the charge control model, the transistor base and collector currents

in the forward-active region are given by

$$i_B(t) = \frac{q_F}{\beta\tau_F} + \frac{d}{dt}(q_F + q_{ve} + q_{vc}) \quad (2a)$$

$$i_C(t) \approx \frac{q_F}{\tau_F}. \quad (2b)$$

Since the transistors are biased at a relatively high collector current, an estimate of τ_F is found by

$$\tau_F \approx \frac{1}{2\pi f_T}. \quad (3)$$

For total laser current set at 2.5 A, each transistor in the array will support up to $i_C \approx 200$ mA of collector current. This corresponds to approximately $q_F \approx 300$ pC of excess charge storage. With the emitter junction forward-biased, ignoring the effects of recombination in the base, the time required for the collector current to rise to its final value depends on the rate at which charge is supplied by the array driver, given by the integral

$$\int_0^{\tau_R} i_B(t) dt = \Delta q_F + \Delta q_{ve} + \Delta q_{vc} \quad (4)$$

where Δq_{ve} and Δq_{vc} are the charges stored in the space-charge regions of the transistor, and Δq_F is the excess charge needed to support the collector current.

As a rough approximation, the base current drive $i_B(t)$ during the switch transition is limited by the base spreading resistance r_x of the array transistor and the risetime $\tau_T \approx 20$ ns of the voltage swing Δv of the DS0026 buffer. For this system, $r_x \approx 50 \Omega$ and $\Delta v \approx 5$ V. The nonlinear transistor base-emitter C_{be} and base-collector C_{bc} space charge capacitances will be approximated as linear capacitors of 30 pF and 10 pF. The buffer risetime-limited base current is given by

$$i_B(t) \approx \frac{\Delta v}{r_x} (1 - e^{-(t/\tau_T)}). \quad (5)$$

For $I_{pk} = 2.5$ A, $\Delta q_F \approx 300$ pC, $\Delta q_{ve} \approx 85$ pC, and $\Delta q_{vc} \approx 35$ pC.

Iteratively, solving the base current charge control equation results in a calculated switching risetime $\tau_R \approx 15$ ns when the effects of the risetime τ_T of the array driver are included. At lower laser current, the response will be quicker. Also, this analysis assumes that the interconnection inductance between driver and laser is small (less than a few nanohenries) so that other transient effects can be ignored.

C. Active Ripple Reduction

At a switching frequency of 800 kHz, with $L = 47 \mu\text{H}$ and with a duty cycle $D \approx 0.4$ the output ripple current is approximately 120 mA. This peak-to-peak ripple current is six times too large for this application. The ripple can be made smaller by increasing the switching frequency or increasing the inductor size. Inductors six times as large in surface mount packages and sufficient dc current carrying capability are not readily available. Increasing the switching frequency would increase MOSFET and driver switching losses and system noise. An alternative solution is to add an active ripple reduction circuit.

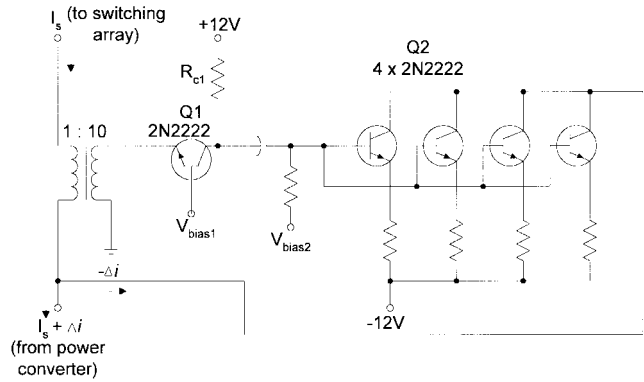


Fig. 4. Ripple reduction circuit.

The function of the active ripple reduction circuit is to reduce the ripple in the laser current to an acceptable level. The system requirement is that the output current ripple be reduced by a factor of six. Current sensing is accomplished with a transformer for low loss. Reference [15] details a similar method which uses a second inductor to provide ac current sensing.

For the circuit shown in Fig. 4, a 1 to 10 ratio transformer senses ac ripple current. Transistor Q_1 provides a low impedance load to the transformer so that the magnetizing inductance does not affect current sensing appreciably at 800 kHz. Transistors Q_2 are biased at approximately 40 mA each, and provide current gain for the high frequency current feedback loop. Also, the effects of transformer leakage inductance are ignored. Given this, and since $R_{c1} \gg r_x + r_\pi$ and R_b , the loop transmission of this circuit for ac signals is approximately given by

$$-LT = -\frac{i_{out}}{i_{in}} \approx -\frac{h_{fe}(\omega)}{N}. \quad (6)$$

For the 2N2222 transistor, current gain $h_{fe}(\omega)$ begins to drop off at approximately 1 MHz. At 800 kHz, for $N = 10$, and for $h_{fe} \approx 100$, the loop transmission is approximately -10 , so the necessary ripple reduction is achieved.

D. Bias Current Source

An adjustable bias current source biases the laser to the threshold of lasing (I_{th}). This is implemented as a simple linear current source, since the power dissipation will be relatively small. For a 500 mW diode, lasing threshold is generally in the range 150–250 mA, and for a 1.5 W diodes, threshold is approximately 300–400 mA. Therefore, this dc current source is adjustable over these ranges.

E. Interconnection to the Laser

It is desirable to be able to connect the laser driver to a laser which may be several inches away from the PC board. To maintain a fast transient response without ringing, the inductance of this path must be kept to an acceptably small level. Assuming that parasitic capacitances are not important and assuming a maximum transient voltage Δv_t that the array transistors can tolerate, an upper bound for allowable

inductance size is found by

$$L_p < \frac{\Delta v_t}{\frac{di}{dt}}. \quad (7)$$

For an allowable transient voltage of 0.5 V, with a 2 A risetime of 20 ns, the maximum allowable inductance value is 5 nH. This design choice assures that the transient voltage will not significantly affect operation of the array transistors.

Connection to the laser was made with a microstrip cable. The cable was constructed of two parallel copper planes, closely spaced, separated by a dielectric with $\epsilon \approx 3\epsilon_o$. A crude approximation to the maximum frequency that the microstrip line is required to carry under transient excitation can be derived from risetime measurements

$$f_{\max} \approx \frac{0.35}{\tau_R} \approx \frac{0.35}{10 \text{ ns}} \approx 35 \text{ MHz}. \quad (8)$$

The parasitic inductance L_p of the microstrip is approximated using the magneto-quasistatic one-turn solenoid approximation, and is given by

$$L_p \approx \frac{\mu_o A}{W} \quad (9)$$

where W is the width of the conductor, and A is the cross sectional area of the loop in which the laser switching current flows. For the structure chosen, $A = 0.125 \text{ cm}^2$ and $W = 1 \text{ cm}$ corresponding to a characteristic impedance $Z_o \approx 5 \Omega$. With microstrip line length $l = 8 \text{ cm}$ there is a resultant parasitic inductance $L_p \approx 2 \text{ nH}$. The approximation is valid since the laser on resistance $R_L \ll Z_o$ for the microstrip, and the length of the cable is much less than a wavelength at the 35 MHz maximum operating frequency. This small inductance does not significantly affect the transient response of the laser current.

V. RESULTS

The prototype circuit and test results are described in more detail in [27].

A. Switching Speed

The switching speed of the circuit, as shown in Fig. 5, is approximately as calculated, with rise and fall times less than 20 ns while switching up to 2 A. As expected, switching times increase as the total switched current is increased. This is due to the fact that there is a limited current available to charge the array transistors. Risetime measurements were made with an active laser set to a threshold current $I_{th} = 500 \text{ mA}$; measurements were made with a fast (response time less than 1 ns) photodetector (see Table II).

B. Power Supply Load

Using this circuit, it is possible to deliver more current to the laser than is drawn from the -12 V supply. Current drawn from the -12 V power supply was measured with $I_{th} = 0$ (see Table III).

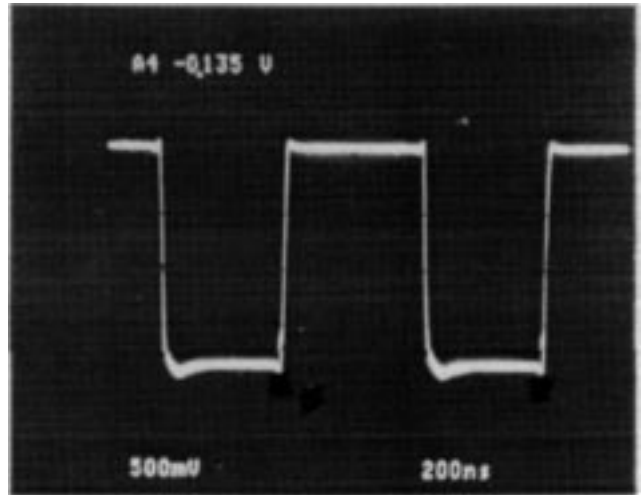


Fig. 5. Measured laser output power using photodetector. Vertical: 2 W full scale, Horizontal: 200 ns/div.

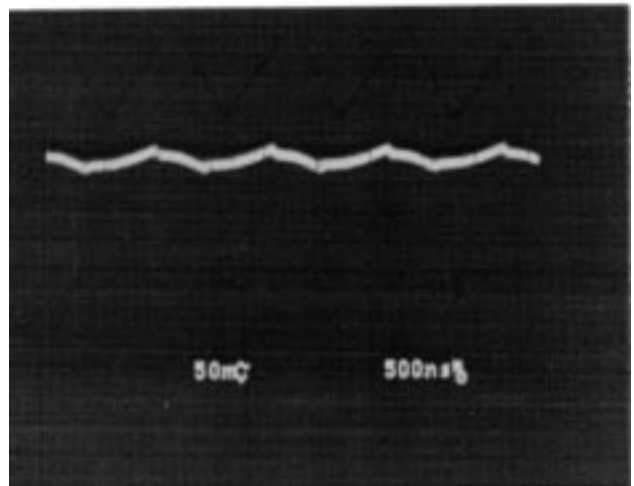
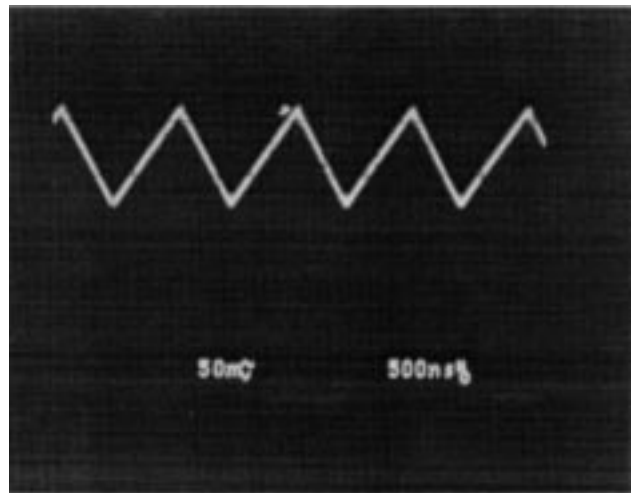


Fig. 6. Ripple reduction circuit results: (a) without ripple reduction, 50 mV/div. and (b) with ripple reduction enabled, 50 mV/div.

TABLE II
LASER SWITCHING SPEED (10-90% RISETIME)

I_{pk}	τ_R
1.0 A	8 ns
2.5 A	17 ns

TABLE III
POWER SUPPLY LOAD

I_{pk}	$I_{(-12)}$
1.0 A	0.55 A
2.5 A	1.27 A

C. Ripple Reduction

As can be seen in Fig. 6(a), laser output current ripple due to the power converter is approximately 120 mA. Fig. 6(b) shows performance of the driver when the active ripple reduction circuit is operating. Measured ripple is less than 15 mA peak-to-peak.

VI. CONCLUSION

This circuit demonstrates the successful application of switching power converter technology to the practical problem of high speed modulation of high power semiconductor laser diodes. The power delivery efficiency of the circuit is significantly higher than that of previous linear designs, resulting in a system which is able to modulate high-power lasers without large switching devices and extensive heat sinking. The finished product has been packaged in a PC board of approximately 10 in^2 , 1 W and greater laser diodes may be modulated at several MHz, with a current risetime less than 20 ns. Furthermore, it is built completely with off-the-shelf components, and no special magnetics are required. The power delivery efficiency is high enough so that the laser may be turned on continuously, and no PC board heat sink is needed.

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REFERENCES

- [1] A. Albanese and H. F. Lenzing, "Video transmission tests performed on intermediate-frequency lightwave entrance links," *SMPTE J.*, vol. 87, Dec. 1978, pp. 821-824.
- [2] *Optical Wideband Transmission Systems*. C. Baack, Ed. Boca Raton, FL: CRC Press, 1986 pp. 22-53.
- [3] F. S. Chen, M. A. Karr, and P. W. Shumate, "Laser transmitters for 70-MHz entrance links," *Bell Syst. Tech. J.*, vol. 58, no. 7, pp. 1617-1629, Sept. 1979.
- [4] F. S. Chen, "Simultaneous feedback control of bias and modulation currents for injection lasers," *Electron. Lett.*, vol. 16, no. 13, pp. 7-8, Jan. 1980.
- [5] M. Chown, A. R. Goodwin, D. F. Lovelace, G. B. H. Thompson, and P. R. Selway, "Direct modulation of double-heterostructure lasers at rates up to 1 Gbit/s," *Electron. Lett.*, vol. 9, no. 225, pp. 34-36, Jan. 1973.

- [6] M. P. Cooke, G. W. Sumerling, T. V. Muoi, and A. C. Carter, "Integrated circuits for a 200-Mbit/s fiber-optic link," *IEEE J. Solid-State Circuits*, vol. 21, pp. 909-915, Dec. 1986.
- [7] W. Filensky, H. Klein, and H. Beneking, "The GaAs MESFET as a pulse regenerator, amplifier, and laser modulator in the Gbit/s range," *IEEE J. Solid-State Circuits*, vol. 12, pp. 276-280, June 1977.
- [8] C. J. Georgopoulos and A. K. Dormakopoulos, "A 1-Mbit/s IR LED array driver for office wireless communication," *IEEE J. Solid-State Circuits*, vol. 21, pp. 582-584, Aug. 1986.
- [9] J. Gowar, *Optical Communication Systems*. London, UK: Prentice-Hall International, 1984, pp. 340-352.
- [10] D. A. Gradi, "250 MHz CCD driver," *IEEE J. Solid-State Circuits*, vol. 16, pp. 100-101, Apr. 1981.
- [11] J. Gruber, P. Marten, P. Reinhard, and P. Ruser, "Electronic circuits for high bit rate digital fiber optic communication systems," *IEEE Trans. Commun.*, vol. COM-26, pp. 1088-1098, July 1978.
- [12] W. Herden and S. Metz, "A fast current drive circuit for injection lasers," *IEEE J. Solid-State Circuits*, pp. 247-248, June 1973.
- [13] G. Keiser, *Optical Fiber Communications*. New York: McGraw-Hill, 1983, pp. 94-117.
- [14] H. Kressel, *Semiconductor Devices for Optical Communications*. Berlin: Springer-Verlag, 1980, pp. 160-200.
- [15] L. LaWhite and M. F. Schlecht, "Design of active ripple filters for power circuits operating in the 1-10 MHz range," *IEEE Trans. Power Electron.*, vol. 3, pp. 310-317, July 1988.
- [16] *Optoelectronic Technology and Lightwave Communications Systems*. C. Lin, Ed. New York: Van Nostrand Reinhold, 1989, pp. 310-333.
- [17] T. L. Maione and D. D. Sell, "Experimental fiber-optic transmission system for interoffice trunks," *IEEE Trans. Commun.*, vol. Com-25, pp. 517-522, May 1977.
- [18] *Optical Fiber Communications*. S. E. Miller and A. G. Chynoweth, Eds. New York: Academic Press, 1979, pp. 563-591.
- [19] V. Ostoich, P. Jeppesen and N. Slaymaker, "Direct modulation of D.H. GaAlAs lasers with GaAs M.E.S.F.E.T.S.," *Electron. Lett.*, vol. 11, no. 2116, pp. 515-516, Oct. 1975.
- [20] T. L. Paoli and J. E. Ripper, "Direct modulation of semiconductor lasers," in *IEEE Proc.*, vol. 58, Oct. 1970, pp. 1457-1465.
- [21] P. K. Runge, "An experimental 50 Mb/s fiber optic PCM repeater," *IEEE Trans. Commun.*, vol. Com-24, pp. 413-418, Apr. 1976.
- [22] *Optical Fiber Communication Systems*. C. P. Sandbank, Ed. New York: John Wiley, 1980 pp. 207-214.
- [23] J. M. Senior, *Optical Fiber Communications*. London, UK: Prentice-Hall International, 1985, pp. 388-479.
- [24] P. W. Shumate Jr., F. S. Chen and P. W. Dorman, "GaAlAs laser transmitter for lightwave transmission systems," *Bell Syst. Tech. J.* vol. 57, no. 6, pp. 1823-1836, July-Aug. 1978.
- [25] M. J. N. Sibley, *Optical Communications*. London, UK: Macmillan, 1990, pp. 58-67.
- [26] R. G. Swartz, B. A. Wooley, A. M. Voshchenkov, V. D. Archer and G. M. Chin, "An integrated circuit for multiplexing and driving injection lasers," *IEEE J. Solid-State Circuits*, vol. SC-17, pp. 753-760, Aug. 1982.
- [27] M. Thompson, "High power laser diode driver with plural feedback loops," S.M. thesis (Electrical Engineering), Massachusetts Institute of Technology, Sept. 1992.
- [28] M. Thompson, "Laser Driver Circuit," United States Patent # 5,444,728 (Aug. 22, 1995).
- [29] A. Yariv, *Optical Electronics*. New York: Holt, Rinehart, and Winston, Inc., 1985, pp. 490-498.



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