

# HIGH POWER LASER DIODE DRIVER WITH PLURAL FEEDBACK LOOPS

by

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B.S., Massachusetts Institute of Technology, (1985)

Submitted to the Department of  
Electrical Engineering and Computer Science  
in Partial Fulfillment of the  
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at the

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## **ABSTRACT**

500 mW semiconductor laser diodes are used in the Polaroid Corporation's Helios medical laser imager. The author has designed and implemented the laser diode drivers that are currently used to pulse the 500 mW lasers at high speed.

The next-generation Helios laser imager will use 750 mW lasers, and a faster diode switching speed is required. The design of a faster, higher-powered laser diode driver is described in this thesis.

A prototype power converter based laser diode driver has been built and tested. An emitter-coupled current switch and a power converter current source are the core of the design. Power conversion allows a higher percentage of power supply power to be delivered to the laser load. The circuit is capable of delivering up to 3 amperes to the laser load, with a 2 ampere 10-90% switching time of less than 20 nanoseconds.

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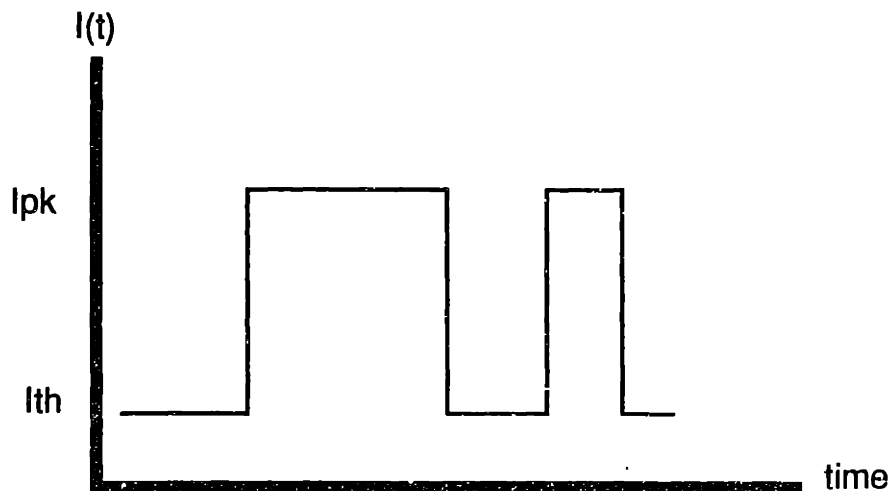
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# Chapter 1

## INTRODUCTION

### 1.1 The Polaroid HELIOS System

The Polaroid Corporation's HELIOS system is a medical laser imager for production of black-and-white grayscale prints on a proprietary transparent medium. The media is exposed by illuminating small areas of its surface by a focused laser beam. The laser beam is scanned across the surface of the media. While scanning, the optical output of the laser diode is direct modulated by switching a current to the laser as shown in **FIGURE 1.1**. The combination of scanning and modulation produces a pixel pattern on the media.



**Figure 1.1 Laser Current Waveform**

The laser current is modulated between two pre-set values. THRESHOLD current  $I_{TH}$  is a current which biases the laser to just below lasing threshold. PEAK current  $I_{PK}$  is the current necessary to



generate a given optical power output. Due to device process variations, the values of  $I_{TH}$  and  $I_{PK}$  vary from laser to laser.

As a result of progress in laser diode long-term reliability, continuous-wave operation of higher-powered laser diodes is now feasible. These devices are well suited for a variety of new applications. The HELIOS medical printer program is one application of this new semiconductor laser technology.

The HELIOS program was initiated at Polaroid Corporation in 1986. Since that time, much work has been done to produce high power, highly reliable semiconductor laser diodes and drive electronics. HELIOS represents Polaroid's first entry into the field of medical laser imaging.

The next-generation HELIOS machine is currently in development. The design, construction, and testing of the prototype laser driver board has been completed in the HELIOS lab at Polaroid Corporation. This system will be used to drive 750 mW semiconductor laser diodes. This board results in a factor of two increase in both switching speed and power delivered to the laser diode from the previous design. The new laser driver may be required to switch laser current at any duty cycle, with a pulse repetition rate of up to 5 MHz, and with a minimum pulse width of approximately 100 nanoseconds.

### **1.1.1 Previous Work**

The existing HELIOS system uses 500 mW GaAs semiconductor laser diodes. These lasers are designed and manufactured by Polaroid's Microelectronics Laboratory. Product requirements have

set the specifications of the current laser diode driver design (TABLE 1.1).

The existing laser driver, designed previously by the author, is required to drive a modulated current with a peak-to-peak amplitude of up to 1.5 amperes, with a risetime and falltime of less than 60 nanoseconds. For this application, 2% settling time is of little importance.

The existing driver (FIGURE 1.2) is built using surface-mount technology.

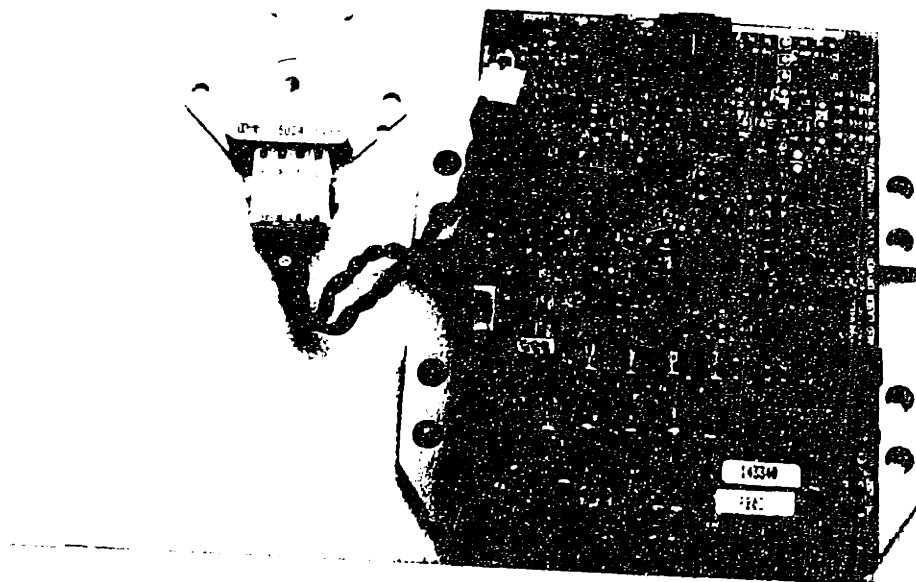


Figure 1.2 Existing Laser Driver

Laser optical power output	500 mW
Laser ON current range	$I_{TH} < I_{PK} < 1.5A$
Laser THRESHOLD current adj.	$50 \text{ mA} < I_{TH} < 300 \text{ mA}$
10-90% switching risetime, falltime	$< 60 \text{ ns}$

**Table 1.1 Existing Laser Driver Specification**

### 1.1.2 Next-generation Effort

The next-generation HELIOS printer will use higher power laser diodes at a higher switching speed. Therefore, the next-generation laser driver will be required to switch more current with a faster risetime and falltime.

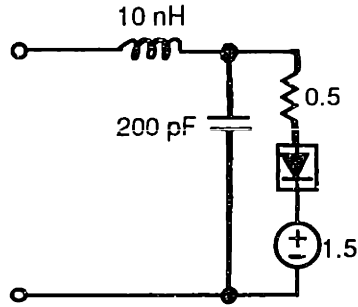
Electrically, the 750 mW laser behaves as a diode with a turn-on "knee" of approximately 1.5 volts and a series resistance of approximately 0.5  $\Omega$ . The laser diode has a series inductance which is due to the bond wires inside the laser package.

As the laser ages, for a constant  $I_{PK}$  there will be less optical power output. Therefore, over the lifetime of the laser  $I_{PK}$  must be adjusted to maintain the required optical power output. Nominal laser characteristics and parameters are as follows (TABLE 1.2):

Laser optical power output	750 mW, max.
$I_{TH}$	300 mA
$I_{PK}$ @ 750 mW, beginning of laser life	1.2 A
$I_{PK}$ @ 750 mW, end of laser life	2.5 A
Laser series resistance	0.5 $\Omega$
Series inductance	10 nH
Shunt capacitance	200 pF
Diode ON voltage	1.5 V

**Table 1.2 Laser Specifications**

These values are approximate, and vary significantly from device to device.



**Figure 1.3 Laser Model**

The next generation laser driver will supply significantly more current with a faster switching speed than the previous design. The following specifications for the next-generation laser driver have been changed (TABLE 1.3):

Laser PEAK current adjustment	$I_{TH} < I_{PK} < 2.5A$
Laser THRESHOLD current adj.	$200 \text{ mA} < I_{TH} < 500 \text{ mA}$
10-90% switching risetime	$< 20 \text{ ns}$

**Table 1.3 Next-Generation Laser Driver Specification**

The same power supply is to be used for the new design as had been used for the 500 mW laser driver design. The power supply specification is shown in TABLE 1.4.

+12V, $\pm 5\%$	500 mA, maximum
-12V, $\pm 5\%$	2 A, maximum

**Table 1.4 HELIOS Power Supply Specification**

The laser diode is to have a grounded anode. A further constraint to the design effort is that the printed-circuit board size

for the next-generation design shall be the same as for the current design (approximately  $3\frac{1}{2}$  in. x  $4\frac{1}{2}$  in.). Also, the new design is to be implemented using surface mount technology wherever possible. Inexpensive components are to be used. This will insure backward-compatibility of the new board with the older HELIOS systems.

## **1.2 Research Efforts**

A number of papers discussing the design of high-speed laser modulators have been surveyed (Refs. 1-10). The circuit topologies presented in these works are applicable to lower power semiconductor lasers (up to approximately 100 mW) where drive currents of less than 500 milliamps are required. For higher-powered lasers, a new topology is considered.

## **1.3 Contribution of this Thesis**

The design, construction, debugging, and documentation of a prototype of the next-generation HELIOS laser diode driver is the focus of this thesis. Following a discussion of specifications and different circuit topologies, this thesis documents the design and test of a prototype laser diode driver. This work also documents the limitations of speed and power for this type of design.

The power converter laser driver, operating at a conversion frequency of 1 MHz, allows the electronics to more efficiently drive lasers than in the current laser driver design. Therefore, the same power supplies can be used, less heatsinking is needed, and the new design is backward-compatible with the existing design.

### 1.3.1 Thesis Contents

Chapter 2 discusses several different circuit topologies that were considered for this design. The strengths and limitations of each approach are qualitatively analyzed. The application of the power converter as a constant current source to drive a semiconductor laser diodes is discussed. Chapter 3 presents the laser driver design in detail, and covers several theoretical design issues. Chapter 4 discusses construction techniques, and Chapter 5 presents measured results. The thesis concludes by summarizing the results of tests on the prototype laser driver, and suggests possible improvements.

Throughout this document, the following schematic symbol has been used for the semiconductor laser diode.



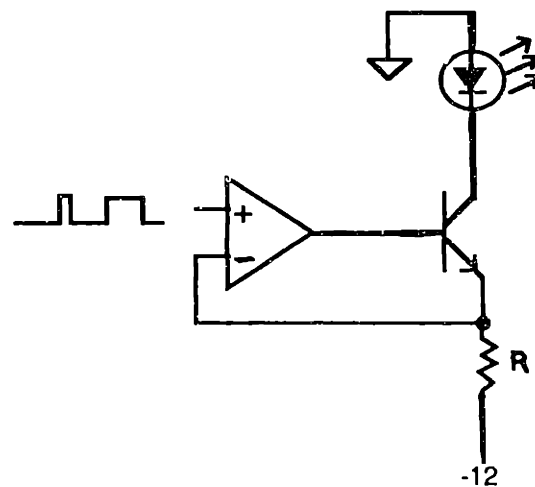
## Chapter 2

### OVERVIEW OF TOPOLOGIES

A brief overview of some possible laser driver circuit topologies is presented.

#### 2.1 Linear Amplifier

The first topology considered is a linear amplifier laser diode driver. This is shown in simplified form in **FIGURE 2.1**.



**Figure 2.1 Linear Amplifier Laser Diode Driver**

In the linear amplifier, the power transistor is operated in its linear region. A wideband operational amplifier provides feedback to maintain the proper laser current. The current level through the laser is controlled by an input signal, shown at the non-inverting input of the operational amplifier.

For driving high-power lasers, a difficulty with this topology is achieving a high bandwidth given that there is a power transistor inside the feedback loop. For a given device cost, there is a tradeoff between power-handling capability and switching speed. To achieve a 10-90% risetime of 20 nanoseconds, a loop crossover frequency higher than 15 MHz is required.

Consider operation of this amplifier operating at a laser current of 2 amperes. In order to achieve 15 MHz bandwidth, a bipolar transistor with  $f_T$  of at least 100 MHz is required to achieve a stable feedback loop. At the 2 ampere operating point, the base-emitter capacitance of this transistor is given by:

$$C_\pi \approx \frac{g_m}{2\pi f_T} \quad [2.1]$$

At a high current level, this capacitance will be quite large. The load capacitance seen at the output of the operational amplifier is a fraction of  $C_\pi$ , and depends on collector current and the size of the emitter degeneration resistor. This load capacitance is given by:

$$C_{LOAD} \approx \frac{C_\pi}{1 + g_m R} \quad [2.2]$$

where  $g_m$  is the transconductance of the transistor at its operating point and  $R$  is the emitter resistor value. The capacitance seen by the op-amp can be made smaller by increasing the size of the emitter resistor. However, increasing the size of this resistor also decreases the current gain of the transistor amplifier, and a larger voltage



swing will be required at the transistor base. Therefore, this is not necessarily a good tradeoff, and the load capacitor  $C_{LOAD}$  will be quite large.

A better choice for this application is a power MOSFET. Commercially available power MOSFETs exist with the necessary drain current rating and power rating with gate capacitance less than 1000 pF. In order to drive the MOSFET with the required speed, an amplifier with a large bandwidth, low output impedance at high frequency, and large current output is required. With a loop crossover frequency of 15 MHz, this is a difficult job for inexpensive op-amps.

Another drawback of this topology is its low power delivery efficiency. If we assume that 3 volts is dropped across the laser diode, only 25% of the power drawn from the -12V power supply is delivered to the laser, and there will be significant power dissipated in the power transistor and resistor. This is wasted energy, and a heat sink is likely to be required. A similar design for the new laser driver would require a -12V power supply rated at 3 amperes.

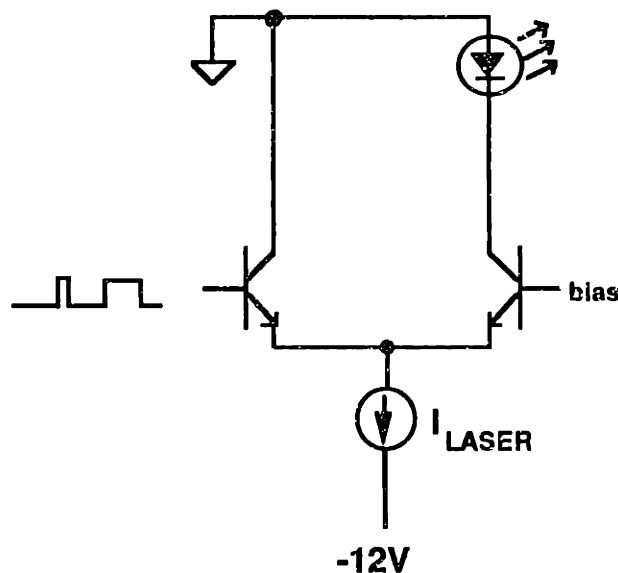
## **2.2 Current Switch**

A second topology is the high-speed current switch (FIGURE 2.2). This topology is similar to that found in commercially available emitter-coupled logic. The laser current is set by the linear current source  $I_{LASER}$ . This current source is switched to and from the laser under control of an external signal.

A feature of the current switch is its high speed. If bipolar transistors are used as the switching device, switching times on the

order of  $1/\omega_T$  of the transistors can be achieved. For transistors rated at an  $f_T$  of 100 MHz, this intrinsic switching time is only a few nanoseconds. Fast switching speeds can also be achieved by using power MOSFETs, if proper gate drives are used.

In the simplest implementation of this design, a linear current source is run from the -12V supply. Unfortunately, there is high power dissipation, as the linear current source is always in operation. Also, power delivery efficiency is only 25% for a laser with a 3 volt drop, and a large heat sink is required.



**Figure 2.2 High-Speed Current Switch Laser Diode Driver**

These two topologies are well suited for driving low power semiconductor laser diodes, where maximum currents are in the 100 mA to 1 A range. High power laser diodes require significantly higher switching currents. At higher currents, power dissipation becomes a limiting factor in achieving a small board size because of the physical size of the components. The additional power dissipation

requires that larger switching devices and more heat sinking are needed. Also, for a given switching speed, a higher-powered device will be more expensive.

Another factor in this design is that only 2 amperes is available from the -12 volt power supply, and a maximum of 3 amperes is needed for delivery to the laser. Therefore, a power converter is needed.

### 2.3 Power Converter Current Source, with Shunt Switch

A semiconductor laser has a relatively small forward voltage drop (a few volts at most). Therefore, it is possible to configure a power converter as a constant current source.

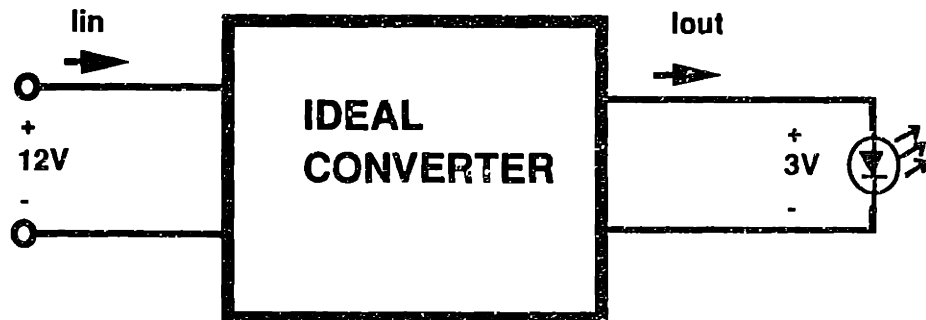


Figure 2.3 Ideal Power Converter

For an ideal converter (FIGURE 2.3) , zero power is dissipated and:

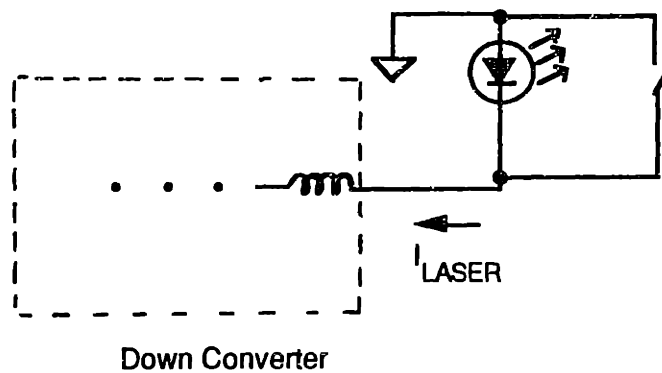
$$V_{in}I_{in} = V_{laser}I_{out} \quad [2.3a]$$

The relationship between laser current and power supply current is:

$$I_{\text{out}} = I_{\text{in}} \frac{V_{\text{laser}}}{V_{\text{in}}} \quad [2.3b]$$

To deliver 2 amperes to a laser with a forward drop of 3 volts would require only 0.5 amperes drawn from a -12V power supply.

Using the power converter concept, the simplest way to modulate the laser current is by using a shunt switch (FIGURE 2.4).



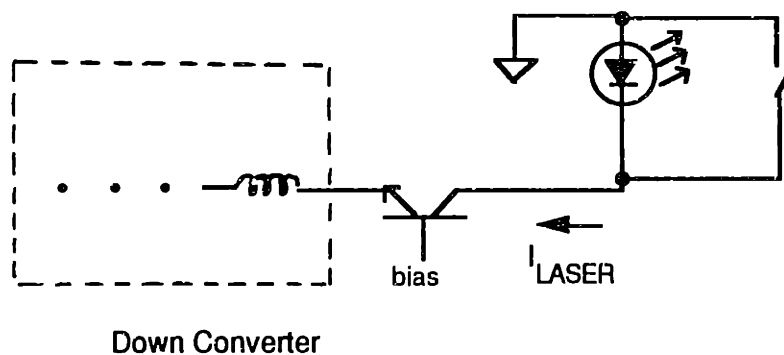
**Figure 2.4 Shunt Switch Laser Driver**

In this topology, a buck converter is operated as a constant current source. To turn the laser ON, current is allowed to flow directly through the laser. To turn the laser OFF, the shunt switch is turned on and current is diverted from the laser.

Ideally, 100% power delivery efficiency could be achieved with this topology, since there are no circuit elements in series with the

current source and laser. However, practical considerations make this topology unattractive for this application.

When the laser is ON, there is a voltage drop of several volts across it. When the laser is OFF, there is a small voltage across it, due to the voltage drop from the ON resistance of the shunt switch. Therefore, the converter inductor will see a voltage swing of a few volts when the laser is switched ON and OFF. This voltage swing will be integrated by the filter inductor to give a transient in laser current. The amplitude of the current transient depends on the bandwidth of the control loop and the size of the filter inductor. It can be reduced by increasing the size of the inductor, by increasing the crossover frequency of the control loop, or by using feedforward in the power converter controller. A simpler approach is to use a series pass element to isolate the filter inductor from the voltage swing, at the expense of some loss in efficiency, as in **FIGURE 2.5**.



**Figure 2.5 Shunt Switch Laser Driver, with pass element**

## 2.4 Hybrid

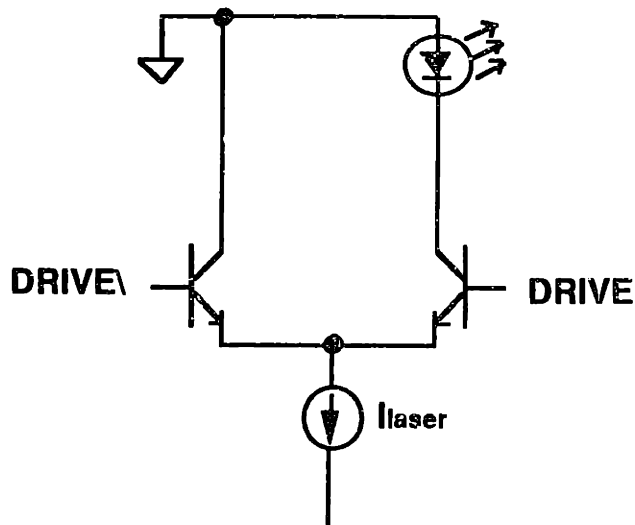
An application of this concept is the use of the fast current switch, driven by a current source that is a derivative of the buck power converter. The voltage drop across the pass transistor will cause some loss of efficiency. It is expected that this design will achieve close to 50 % power delivery efficiency. This will allow delivery of up to 3 amperes to the laser while staying within the power supply specification. This efficiency is compared to that of the previous linear design, which is at maximum 25%. The design of the hybrid laser driver is the subject of the next chapter.

### 2.4.1 Prototype Topology

The topology chosen for the laser driver was an emitter-coupled transistor switch, shown in simplified form in **FIGURE 2.6**. Laser current is set by a power converter current source  $I_{\text{LASER}}$ . The emitter coupled switch offers high switching speed.

The switching transistors are driven differentially. This ensures that there will be only a small transient voltage swing at the emitter of the switching element when the laser is switched ON or OFF. The ripple in laser current due to this transient voltage will not exceed the laser current ripple specification.

In order to handle the large laser current while maintaining a sufficiently fast switching speed, an array of multiple small-signal transistors in parallel is used as the switching element. The parallel topology minimizes parasitic inductances that limit switching speed.



**Figure 2.6 Prototype Topology**

The current source is derived from the buck, or down converter topology. Use of the power converter allows more current to be delivered to the laser with the same size power supply, at the expense of switching noise and ripple in the current. The ripple can be made arbitrarily small by sizing the filter inductor. However, in this design, size constraints and the desire to use surface mount components dictate the maximum inductor value.

A relatively small filter inductor was chosen. This resulted in a current ripple that exceeded the specified maximum value. In order to meet the ripple specification, a circuit that senses current ripple and cancels it was designed. This is described in Chapter 3.

#### **2.4.2 Power Converter Design Specification**

The following are design specifications for the power converter laser current source (TABLE 2.1). These specific values are used for the design given in the next chapter.

Input voltage	-12V nominal
Switching frequency	1 MHz
Max. output current	2.5 A
Output ripple	less than $\pm 10$ mA
Loop crossover freq.	100 kHz

**Table 2.1 Power Converter Specifications**



# Chapter 3

## CIRCUIT DESIGN

### 3.1 Overview

Refer to the laser driver block diagram (FIGURE 3.1). The heart of the design is a power converter current source, which includes the pulse-width modulation section, switch, and filter inductor. The power stage generates a DC current that can be adjusted from zero to approximately 2.5 A. The power stage is implemented as a standard down converter, and laser current is drawn from the -12V power supply.

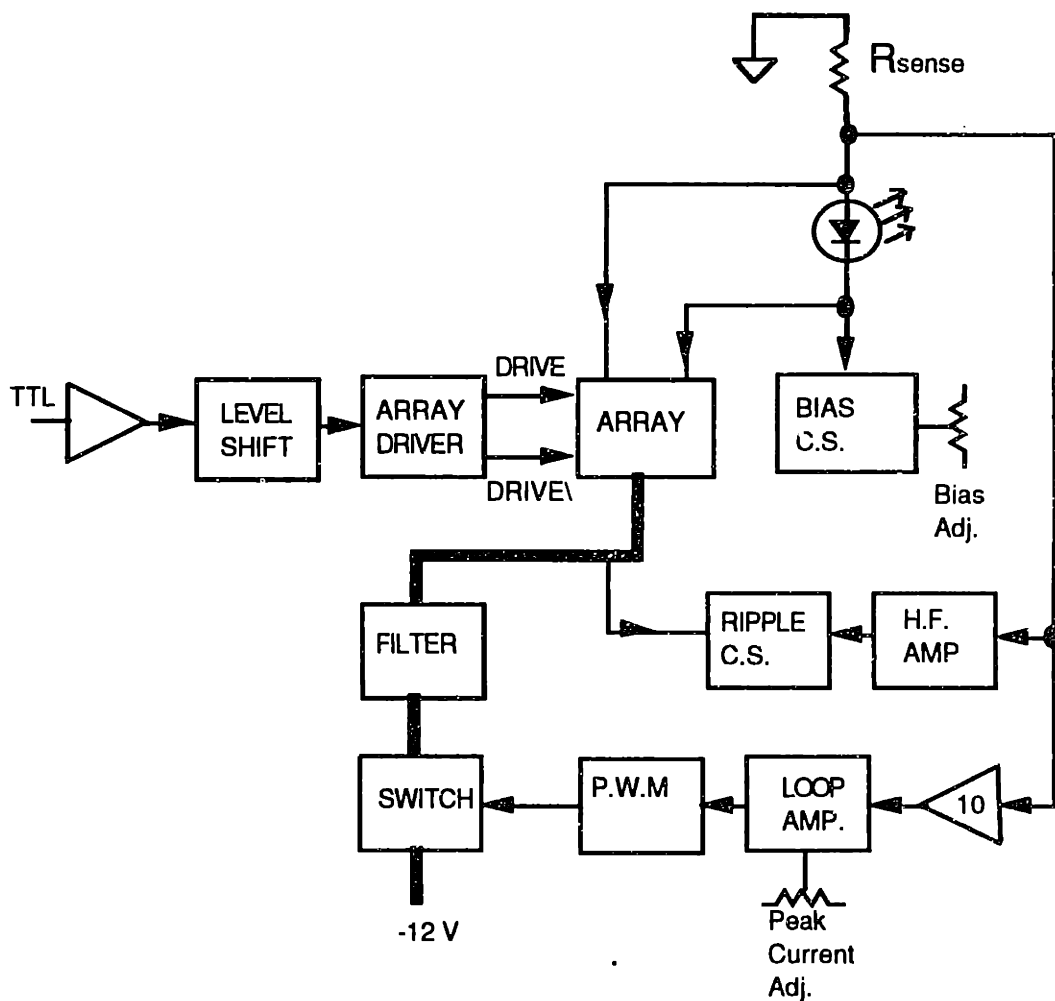
Laser current is under control of a TTL input signal. If the TTL signal is LOW, laser current is that which is set by the bias current source. If TTL = HIGH, the power converter current is sent to the laser.

Total current is sensed by resistor  $R_{SENSE}$ . This voltage is used to close a feedback loop which keeps the average value of laser current constant.

Due to the switching action of the power converter, there is ripple in the output current that would exceed specifications. The purpose of the active ripple reduction circuit is to sense the AC ripple, and generate a current which cancels this ripple.

The transistor array is an ECL-type switch that switches the current to the laser. When the laser is OFF, current is diverted from the laser. The array driver provides sufficient current to charge and discharge the large input capacitance of the array transistors.

An adjustable bias current source biases the diode to just below lasing threshold. Its adjustment range is 50 mA to 500 mA. This results in a maximum laser current of 3 amperes.



**Figure 3.1 Laser Driver Block Diagram**

## 3.2 Operation in Detail

### 3.2.1 Power Stage

Refer to schematic A.1 in the APPENDIX. The power stage is implemented as a standard down converter, configured as a constant current source which drives the current switch. The pulse-width modulation control for the converter operates at a frequency of approximately 1 MHz.

The primary constraint in the design of the power stage is the inductor size. A surface-mountable inductor less than 0.5 inch in height is required. The Coilcraft DT720310-470 is a 47  $\mu\text{H}$  inductor, specified to carry 4 amperes dc, that is in a surface mount package.

Nominal inductance	47 $\mu\text{H}$
Dc resistance (measured)	0.063 $\Omega$
Self-resonant frequency (measured)	8.3 MHz

**Table 3.1 Inductor Specification**

From the self resonant frequency specification, a lumped parameter model for the inductor would be a 47  $\mu\text{H}$  inductor with an 8 pF capacitor in parallel. Due to the voltage switched across the inductor, there is a small transient current through this capacitance. If we assume a 12 volt transition across the inductor with a risetime of 20 nanoseconds, this current transient will be approximately 5 milliamps.

The LM339 comparator generates a triangle wave at 1 MHz. A control voltage from the control amplifier controls the duty cycle via

comparator LM311. This comparator controls the ON and OFF switching of the MOSFET switch.

The DS0026 gate driver can supply up to 1.5 A peak current to supply the necessary charge to turn the MOSFET on and off. This device is specified to have a risetime of 20 ns with a 1000 pF load. In order to charge and discharge the MOSFET gate efficiently, the loop inductance from the output of the DS0026 to the gate-source of the MOSFET must be minimized.

The voltage ratings of the power MOSFET and flyback diode are greater than the maximum input voltage, including any transient voltage. The IRLR014 N-channel MOSFET was chosen because it offers an attractive tradeoff between maximum drain current, ON-resistance, and input capacitance in a D-PAK surface-mount package. The 6CWQ03F is a dual Schottky in a D-PAK package, specified for a maximum DC current of 6.6 A and a maximum reverse voltage of 30 V.

Continuous drain current, $T_{case} = 100^{\circ}C$	6 A
$R_{DS(on)}$	< 0.2 $\Omega$
Maximum $V_{DS}$	60 V
$C_{iss}$ input capacitance	400 pF typ.
$C_{oss}$ output capacitance	170 pF typ.
$C_{rss}$ reverse transfer capacitance	42 pF typ.

**Table 3.2 IRLR014 Specification**

Max. DC reverse voltage	30 V
Max. DC forward current	6.6 A
Max. forward voltage	0.55 V @ 3A
Junction capacitance	180 pF typ.

**Table 3.3 6CWQ03F Specification**

The switching frequency of 1 MHz was chosen as a tradeoff between current ripple and switching losses and switch noise. The junction capacitance of the Schottky diode and the parasitic wiring inductance form a resonant circuit with very little damping. A snubber circuit was required to damp this oscillation and to protect the Schottky diode from excessive reverse voltage. The snubber capacitance (360 pF) was kept as small as possible, and achieved the required damping of the switching oscillation, resulting in a reverse voltage across the Schottky of approximately 16 volts. Given a switching frequency  $f_{sw}$  of 1 MHz, a snubber capacitor  $C_{snubber}$  of 360 pF, and a voltage  $V$  of 16 volts, loss in the snubber circuit is given by:

$$P_{diss} = (2)\left(\frac{1}{2}\right)C_{snubber}V^2f_{sw} \approx 90 \text{ mW} \quad [3.1]$$

This snubber does not add significant loss, and keeps the reverse voltage across the Schottky at a safe level.

Consider the operation at a laser current of 2 amperes: The voltage on the base of one side of the array transistors is -3.3 volts. At a collector current of 250 milliamps, a  $V_{BE}$  of 1 volt is expected for the 2N2222. The voltage at the emitters of the ON transistors will be approximately -4.3 volts. There is a further drop across each of the 2.2  $\Omega$  emitter resistor of approximately 0.55 volts. Therefore, the

voltage on the array side of the inductor is approximately -5.0 volts. The down converter will operate at a duty cycle of approximately  $\frac{5}{12}$ , or 0.4.

At this operating point, the current drawn from the -12V supply will ideally be 0.8 A. Power delivery efficiency is defined as:

$$E = \frac{\text{Power delivered to load}}{\text{Avg. power drawn from supply}} \times 100\% \quad [3.2a]$$

If we assume that the laser is modelled as a 1.1 ohm load, the power delivery efficiency is:

$$E = \frac{(2A)(2.2V)}{(0.8A)(12V)} \approx 46\% \quad [3.2b]$$

This analysis assumes an ideal down converter and ignores all static and dynamic losses. The efficiency would be much higher if not for the voltage dropped across the switching transistor and emitter resistors.

The output current of the down converter will have a DC value with some ripple at the switching frequency. For the ideal down converter, the peak-to-peak ripple in inductor current is given by:

$$\Delta i = \frac{V_{\text{supply}}DT}{L} \quad [3.3a]$$

where D is the duty cycle, T is the switching interval, and L is the inductor value. At a switching frequency of 1 MHz, using a 47  $\mu\text{H}$

inductor, and with a converter duty cycle of approximately 0.4, expected output current peak-to-peak ripple is:

$$\Delta i = \frac{(12)(0.4)(10^{-6})}{47 \times 10^{-6}} = 100 \text{ mA} \quad [3.3b]$$

An output ripple of  $\pm 50$  mA is 5 times too large for our specification. From EQN. 3.3a, the ripple can be made smaller by increasing the switching frequency or by increasing the inductor value. Inductors 5 times as large in surface mount packages and DC current carrying capability of 5 amperes are not readily available. Increasing the switching frequency would also increase switching losses and circuit noise. Since the ripple current from the down converter is relatively small, an active ripple reduction circuit was designed and added to the laser driver circuit.

### 3.2.2 Active Ripple Reduction Circuit

Refer to schematic A.2 in the APPENDIX. The requirement is that the output current ripple be reduced by a factor of 5. The function of the active ripple reduction circuit is to reduce the ripple in the laser current to an acceptable level. A circuit with a loop transmission magnitude of at least 5 at 1 MHz is needed.

A simplified schematic of the ripple reduction circuit is shown in FIGURE 3.2a. The voltage across the laser current sense resistor  $R_{\text{sense}}$  is sensed and amplified. This voltage is applied to the base of 4 2N2222 NPN transistors in parallel. The resulting transistor

collector current cancels the ripple current due to converter operations. Each transistor is biased to approximately 50 mA, resulting in the small-signal parameters in TABLE 3.4.

The low frequency loop-transmission is found by considering the small-signal model of this circuit (FIGURE 3.2b):

$$LT = \frac{i_{out}}{i_{in}} = -4R_{sense}A_o g_m \frac{r_{\pi}}{r_x + r_{\pi}} \quad [3.4]$$

The gain element is implemented with an LM733, which is configured for a single-ended gain of 50 and a -3 dB point of 100 MHz. With the values in TABLE 3.2, and with  $A_o = 50$ , the low-frequency loop transmission is approximately -16.

The ripple reduction circuit also performs the additional function of cancelling the current due to effects of the transient voltage seen at the emitter of the array. This transient voltage occurs when the laser is turned ON and OFF, and is integrated by the filter inductor to give a change in laser current. The cause of this transient voltage is shown in FIGURE 3.3.

In order to turn the laser ON,  $v_{b2}$  is brought high, and  $v_{b1}$  is brought low. These transitions occur simultaneously, with a risetime of approximately 20 nanoseconds. The emitter voltage  $v_e$  shows a transient of approximately 20 nanoseconds in width, with an amplitude of approximately 5 volts.

The average value of this transient voltage depends on the switching frequency of the TTL input signal. If changes in the TTL signal occur at a frequency higher than the loop bandwidth, this



voltage could result in a change in laser current as the filter inductor integrates the voltage.

The maximum current change due to this phenomenon is calculated as follows. For a bandwidth of 100 kHz and a phase margin of 50 degrees, it will take the loop approximately 3.5 microseconds to respond to a change in duty cycle. Therefore, the incremental voltage is integrated by the down converter inductor for this period of time. The current change due to a change in the emitter voltage  $\Delta v$  is approximately given by:

$$\Delta i \approx \frac{\Delta v \tau_R}{L} \quad [3.5]$$

where  $\tau_R$  is the risetime of the loop. The voltage that is integrated by the filter inductor depends on the duty cycle. If we assume a maximum pulse rate of 5 MHz, with a transient duration of 10 nanoseconds and an amplitude of 5 volts, this transient voltage is:

$$\Delta v \approx \frac{1}{2} \frac{(5V)(2)(10ns)}{200ns} \approx 0.25 \text{ V} \quad [3.6]$$

There is a factor of 2 is due to the fact that there is a transient voltage on both the rising and falling transition. By substituting this value into EQN. 3.5, the transient current change is found to be:

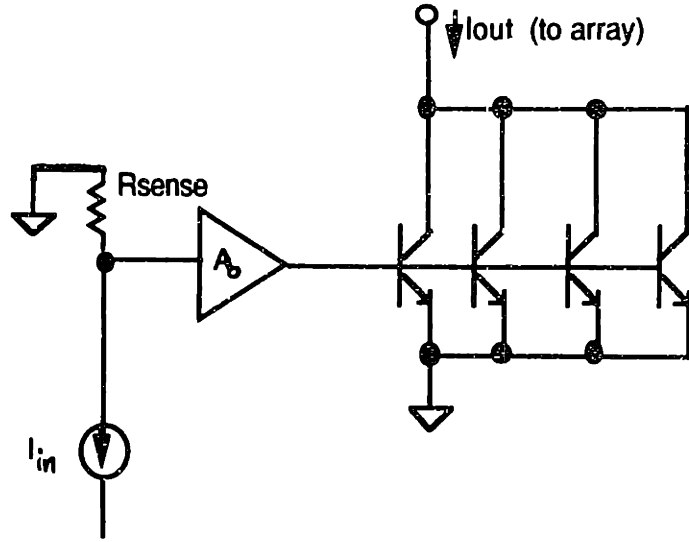
$$\Delta i \approx \frac{(0.25)(3.5 \times 10^{-6})}{47 \times 10^{-6}} \approx 18 \text{ mA} \quad [3.7]$$

The ripple reduction circuit reduces this current change to less than a few milliamps.

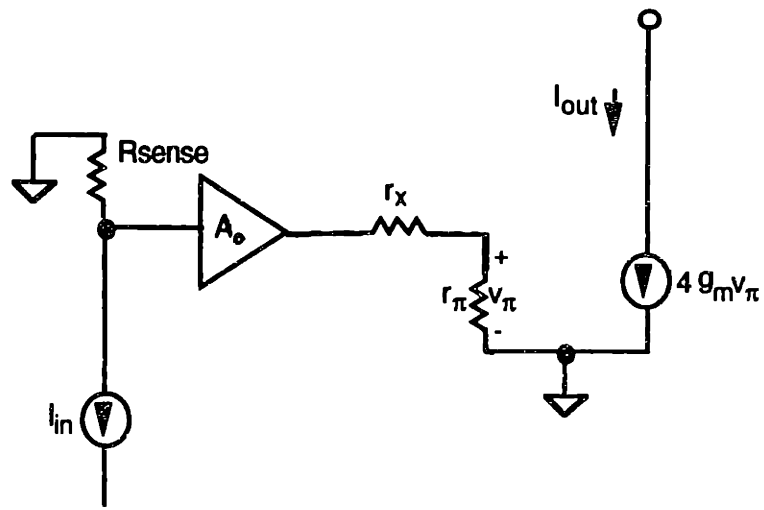
To maintain stability of the feedback system, a dominant pole is created from the  $C_{\pi}$  of the transistors and the  $47 \Omega$  source resistance in series with the base spreading resistance. This gives a dominant pole at approximately 400 kHz.

$I_C$	50 mA
$f_T$	300 MHz
$h_{fe}$	150
$g_m$	$\approx 2 \text{ ohm}^{-1}$
$r_{\pi}$	$75 \Omega$
$C_{\pi}$	1000 pF
$r_x$	$\approx 50 \Omega$

**Table 3.4** 2N2222 Small-signal parameters

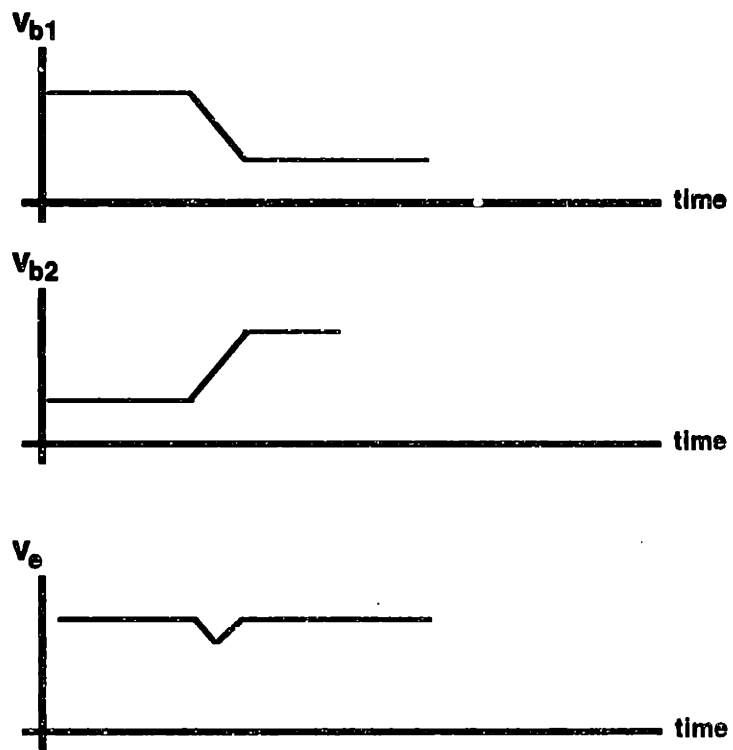
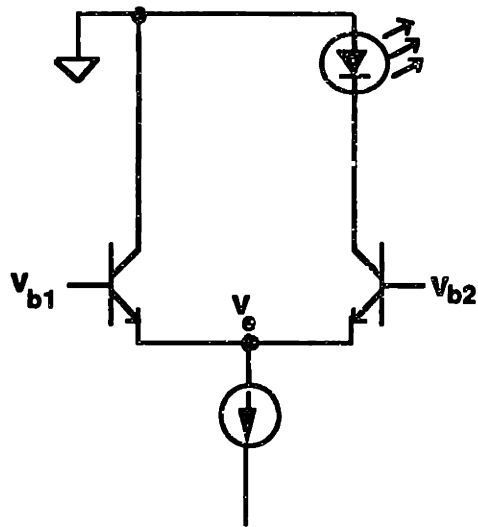


(a)



(b)

**Figure 3.2 Active Ripple Reduction Circuit**



**Figure 3.3** Array Switching Transient

### 3.2.3 Current Switch and Driver

Refer to schematic A.3 and A.4 in the APPENDIX. The current switch is implemented as an array of 16 2N2222 switching transistors. The 2.2  $\Omega$  emitter resistors force the laser current to divide up equally between the ON transistors. Choice of this particular device and topology offers several advantages.

Each ON transistor is required to carry up to 300 mA. The  $f_T$  of this transistor operating at a collector current of 300 mA is approximately 100 MHz. Therefore, the intrinsic switching time of the current switch is only a few nanoseconds.

Another advantage in using multiple devices is that during switching, there are 8 parallel paths in which switching current flows. Therefore, the effects of inductance due to the switching loops are greatly reduced.

Another important consideration for this design is that the 2N2222 is capable of carrying hundreds of milliamps, and is available in a 16-pin surface mount SOIC package, the MMPQ2222,A.

The laser driver is required to interface with standard TTL levels. Laser current is controlled by the TTL input signal as follows:

TTL input	Laser Current
HIGH	Laser ON ( $I_{out} = I_{PK}$ )
LOW	Laser at THRESHOLD

**Table 3.5 Laser Current Truth Table**

The TTL input signal is buffered by the 74F14 buffer. This signal is level shifted by the emitter-coupled 2N3906 pair. Another

74F14, powered by -3.3V and -8.3 volt power supplies, drives the input of the DS0056 gate driver. The DS0056 provides sufficient transient current to drive the transistor array.

Due to the high  $f_T$  of the switching devices, the primary limitation in laser current switching speed is the rate at which charge can be supplied to the base region of the transistors. The switching speed is approximated using the charge control model, found in Reference 11.

Consider the side of the array that is OFF, with a laser current of 2 amperes. This analysis considers one transistor in the array. In order to turn this side of the array ON, the base-emitter and base-collector space charge capacitances must be charged, and excess charge  $q_F$  must be supplied to the base. At a large collector current,  $q_F$  will be the dominant charge component, and the effects of space-charge-layer capacitances and recombination in the base may be neglected. Using these approximations, total excess charge is given by:

$$q_F \approx I_C \tau_F \quad [3.8]$$

A rough estimate of  $\tau_F$  is found from data sheet parameters of the transistor by:

$$\tau_F \approx \frac{1}{2\pi f_T} \quad [3.9]$$

$I_C$	250 mA
$f_T$	100 MHz
$r_x$	$\approx 50 \Omega$

**Table 3.6** 2N2222 Small-signal parameters,  $I_c = 250$  mA

The collector current of each transistor in the array will be one-eighth of total laser current. Given a collector current of 0.25 A and a transistor  $f_T$  of 100 MHz, the excess charge is:

$$q_F \approx \frac{0.25}{2\pi(100 \times 10^6)} \approx 0.4 \text{ nC} \quad [3.10]$$

This is the total charge required to turn one transistor in the array ON. With the emitter junction forward biased, the time required for the collector current to rise to its final value depends on the rate at which charge is supplied to the base region from the array driver, given by the integral:

$$\int_0^{\tau} i_B dt = q_F \quad [3.11]$$

To a first-order approximation, the base current  $i_B$  is constant during the switch transition. This current is limited by the base-spreading resistance  $r_x$  of the transistor, which is approximately 50  $\Omega$ . There is a voltage swing from the array driver of 5 volts. The voltage transition point where one side of the array turns ON and the other side turns OFF that is approximately half way between the two ends. The transient base current is:

$$i_B \approx \frac{V_{sw}}{r_x} \approx \frac{2.5V}{50 \Omega} \approx 50 \text{ mA} \quad [3.12]$$

Given a constant base current of 50 mA, the transition time is given by:

$$\tau \approx \frac{q_F}{i_B} \approx \frac{0.4 \text{ nC}}{50 \text{ mA}} \approx 8 \text{ ns} \quad [3.13]$$

This analysis shows that switching time will increase as the laser current level increases, because the stored base charge increases and base current remains approximately constant. For a faster switching speed, more voltage swing may be applied at the transistor bases.

### 3.2.4 Control Loop

Refer to schematic A.5 in the APPENDIX. To design a stable control loop, a model of the switching function has been developed using the methodology given in Reference 12. The small signal model of the power converter and transistor array is shown in FIGURE 3.4. This represents the "plant".

The input to the plant is a voltage  $V_{in}d$ , which is the averaged switched voltage to the filter inductor. This analysis ignores the voltage drop across the MOSFET and the Schottky diode.

$V_{in}$  is the -12V power supply, and  $d$  is the time-varying duty cycle of the switched converter. The duty cycle varies slowly compared to the switching frequency.

The various circuit resistances are:



$R_{sense}$	current sense resistor	0.1 $\Omega$
$R_{on}$	averaged ON resistance of MOSFET and SCHOTTKY	0.2 $\Omega$
$R_{ind}$	DC resistance of inductor	0.063 $\Omega$
$R'_E$	equivalent parallel emitter resistor	0.275 $\Omega$
$R_{OUT}$	equivalent output resistance of array transistors	0.41 $\Omega$ min. 0.58 $\Omega$ max.

**Table 3.7 Circuit Resistances, Power Converter**

The transfer function from perturbations in the duty cycle to the output voltage is given by:

$$\frac{v_o(s)}{d(s)} = - \frac{V_{in} R_{sense}}{R_{on} + R_{ind} + R'_E + R_{OUT} + Ls} \quad [3.14]$$

From this model, it is seen that there is a single pole due to the  $L/R$  time constant of the converter inductor and the various circuit resistances. The output resistance  $R_{OUT}$  of the array transistors depends on  $h_{fe}$ ,  $r_x$ , and  $g_m$ , and varies somewhat as the laser current level is adjusted. The switch resistance  $R_{on}$  is approximated as 0.2  $\Omega$  at all current levels. Using these numbers, the pole location varies from 3.6 kHz to 4 kHz as the laser current is adjusted from maximum to minimum.

With the power converter operating at 1 MHz, a crossover frequency of 100 kHz is achievable. At this crossover frequency, the phase shift from the power converter sampling loop and delay

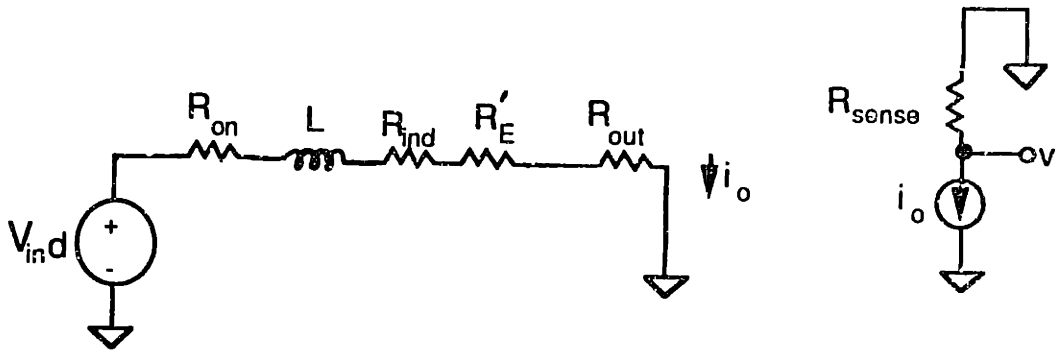
through the comparator is tolerable, and a phase margin of 50 degrees is achieved.

Laser current is adjusted by measuring the voltage at the output of the gain-of-10 amplifier, implemented with a LM101A. The voltage  $V_{\text{sense}}$  is generated across a  $0.1 \Omega$  current sense resistor  $R_{\text{sense}}$ . There is a 1-to-1 ratio between this voltage and laser PEAK current.

The LM101A was chosen for several reasons. First, a gain of 10 with a bandwidth of 1 MHz can be achieved with proper choice of compensation capacitor. This adds only  $5.7^\circ$  of negative phase shift at the 100 kHz crossover frequency. Also, the LM101A has a lower voltage offset than the LM301A. With a gain of 10, and a current level of 2 amperes, a 1% error may be expected due to the 2 mV maximum offset of the LM101A. There is an additional phase shift of approximately  $-5.7^\circ$  from the second amplifier.

A TL431 voltage reference is used to set a stable reference by which PEAK laser current can be adjusted.

An integrator is included in the control loop so that there is zero DC laser current error. A zero is placed at approximately 3 kHz, or a factor of 30 below crossover. This zero insures that there is a one-pole rolloff at crossover.



**Figure 3.4 Power Stage Small-Signal Model**

### 3.2.5 Laser Bias Current Source

Refer to schematic A.6 in the **APPENDIX**. This linear current source supplies the bias (**THRESHOLD**) current for the laser. This source is adjustable from approximately 50 mA to 500 mA.

**THRESHOLD** current is set by measuring the output of the differential amplifier.

# Chapter 4

## IMPLEMENTATION

### 4.1 Component Selection

Due to the high frequency nature of much of the circuitry, all resistors are specified to be carbon composition, carbon film or metal film. In some instances, multiple resistors were used in parallel to reduce parasitic inductance.

High-frequency ceramic and mica capacitors were used in most instances. Where higher values of capacitance were needed, a tantalum capacitor in parallel with a ceramic capacitor was used. In all cases, lead lengths were kept as short as possible. Wherever feasible, surface mount capacitors were used for their low parasitic inductance.

### 4.2 Layout

The prototype was built on a copper-clad vector board (FIGURE 4.1). The copper side of the board was used as a ground plane. Where necessary, other current-carrying planes were used. Care was taken to reduce the loop area of any paths carrying fast switching currents.

In this circuit, parasitic inductance has a stronger affect on circuit operation than parasitic capacitance. Thus, all interconnections that carried large switching current were made as wide and as short as possible, over a ground plane.

### 4.3 Laser Connection

A low-impedance connection to the laser from the printed-circuit board is required. This connection (connected to the collectors of the array transistors) must carry a switching current of up to 2.5 A and a risetime of 20 nanoseconds or less. The laser may be up to 4 inches away from the collectors of the transistors.

For this connection, inductance is the dominant parasitic. To reduce this inductance, the connection to the laser was made by a plane structure. That is, connection was made via two wide current carrying planes separated by a thin insulator. In order not to disturb operation of the switching transistors, an inductance of less than a few nanohenries is required. The structure has the following approximate dimensions:

length	10 cm
width	4 cm
spacing between planes	0.05 cm

**Table 4.1 Laser Connection Dimensions**

The inductance of the structure is approximated using the one-turn solenoid approximation [Ref. 13]. The inductance is given by:

$$L_p \approx \frac{\mu_0 A}{W} \quad [4.1a]$$

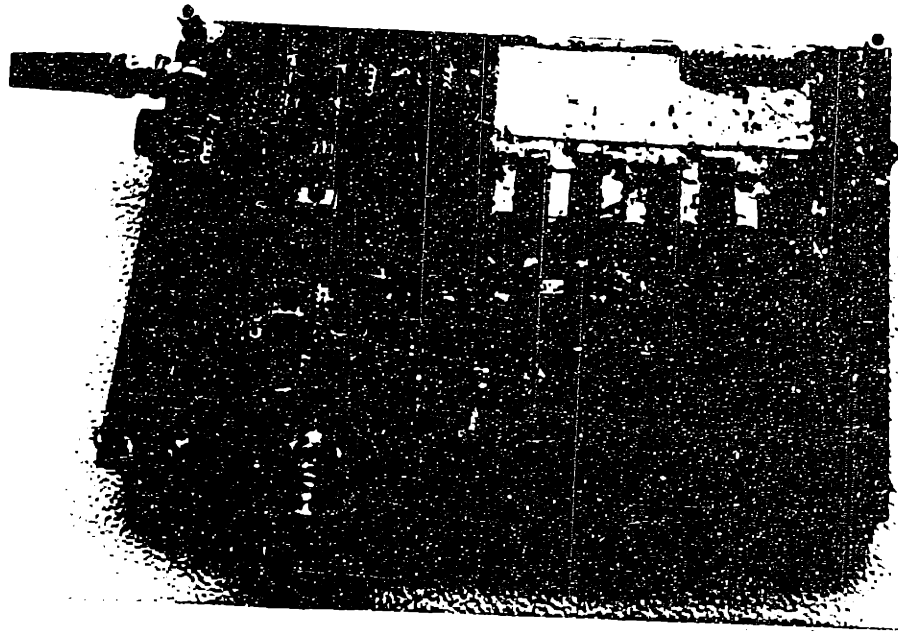
where  $W$  is the width of the conductor, and  $A$  is the cross sectional area of the loop in which the current flows. For this structure, the inductance is:

$$L_p = \frac{(4\pi \times 10^{-9})(10)(0.05)}{4} = 1.5 \text{ nH} \quad [4.1b]$$

Given a laser current of 2 amperes and a switching time of 20 nanoseconds, the amplitude of the transient voltage generated across the structure when laser current is switched is given by:

$$\Delta v = L_p \frac{di}{dt} = \frac{(1.5 \times 10^{-9})(2)}{20 \times 10^{-9}} = 0.15 \text{ V} \quad [4.2]$$

This transient voltage does not significantly affect operation of the array transistors.



**Figure 4.1** Prototype Laser Driver

# Chapter 5

## EVALUATION

### 5.1 Probing

The greatest source of error in measurements on high speed circuits is grounding. The inductance of a probe ground lead forms a series-resonant circuit with the input capacitance of the probe. This resonant circuit can degrade risetime, bandwidth, and transient measurement accuracy. To reduce these effects, a low-impedance connection from the scope probe to the circuit ground is required. All high frequency test points were monitored with subminiature probe tip circuit board test points, available from Tektronix.

A Tektronix 2465B 400 MHz scope and Tektronix P6131 300 MHz probe were used for all high frequency measurements. These elements each have a risetime of approximately 1 nanosecond, and did not significantly affect risetime measurements. A 100 MHz Tektronix current probe was used for current measurements.

The laser load was simulated by paralleling  $22 \frac{1}{2}$  watt,  $22 \Omega$  surface mount resistors in parallel, over a ground plane. The total resistance used for current sensing was  $1.1 \Omega$ . Paralleling reduces the inductance of the sense resistors, so that the total inductance of the structure is less than 1 nH. This was necessary to accurately probe the laser current.



## 5.2 Measurements

### 5.2.1 Switching Speed

The switching speed of the circuit is approximately as calculated, with 10-90% rise and falltimes less than 20 ns while switching up to 2 amperes. If a faster response is required, more voltage could be applied to the bases of the array transistors. This would increase the current available to charge and discharge the base regions of the transistors.

**FIGURE 5.1a** shows the waveform switching a laser current of 2 amperes peak-to-peak, switching at 1 MHz, 50% duty cycle.

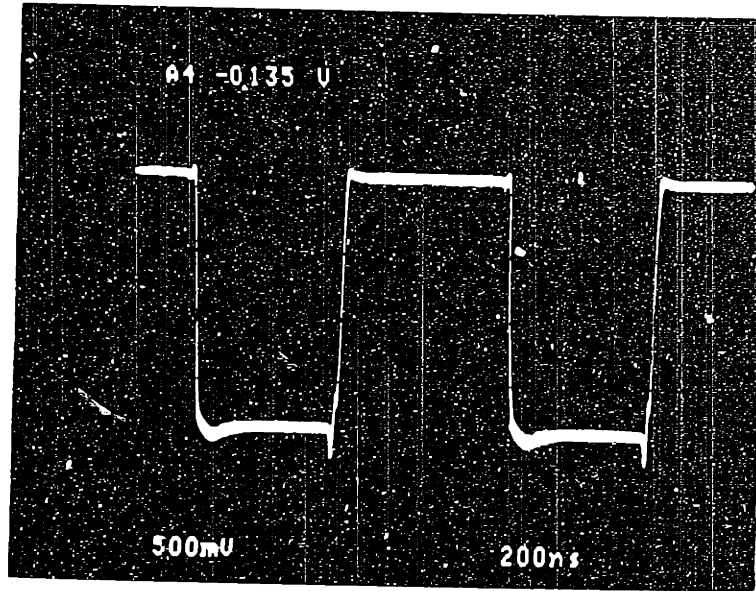
**FIGURE 5.1b** shows the detailed laser current risetime, which is approximately 19 ns.

**TABLE 5.1** shows tabulated risetimes and falltimes for various current settings. Note that total laser current is the value set by the BIAS current source (up to 500 mA) plus the peak-to-peak switching current.

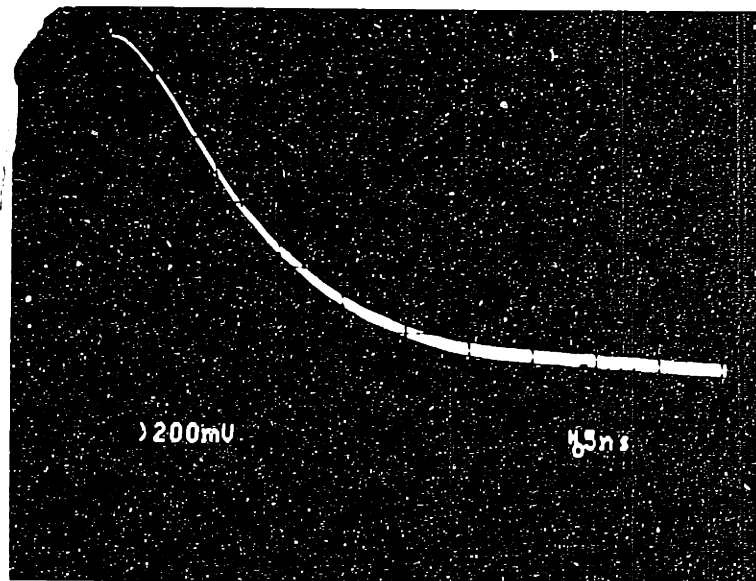
P-P	Switched Current	10-90 %	Risetime/Falltime
1 A			8ns/15ns
2 A			19ns/15ns
2.5 A			34ns/20ns

**Table 5.1 Measured Risetime/Falltime**

As expected, switching times increase as the current is increased. This is due to the fact that there is a limited current available to charge the array transistor input capacitances.



(a)  
200 ns/div.



(b)  
5 ns/div.

Figure 5.1 Laser Driver Output Waveform

### 5.2.2 Efficiency

The power delivery efficiency of the new prototype laser driver is compared to the previous laser driver design. For the previous design, the maximum current is approximately 2 amperes. All efficiency measurements were made with a resistive load of 1  $\Omega$ .

Output Power	Input Power	Power Dissipation	Efficiency
1A @ 1.1 V	1A @ -12 V	10.9 W	9%
2A @ 2.2 V	2A @ -12 V	19.6 W	18%

**Table 5.2** Projected Efficiency (previous driver design)

At the maximum current setting, almost 20 watts is dissipated as heat in the previous driver design. A large heat sink and several power resistors were used to dissipate the power.

For the new laser driver design, dissipation is significantly lower. The first measurement was made with the laser fully on.

Output Power	Input Power	Power Dissipation	Efficiency
1A @ 1.1 V	0.55A @ -12.10 V	5.6 W	17%
2A @ 2.2 V	1.00A @ -12.05 V	7.6 W	37%
2.5A @ 2.75 V	1.27A @ -12.01 V	8.4 W	45%

**Table 5.3** Measured Efficiency, Prototype (at DC)

Dissipated power at the 2 ampere level is 7.6 watts, compared to 19.6 watts for the previous design.

Efficiency was also measured with the input TTL signal pulsed at 1 MHz, 50% duty cycle. The efficiency is slightly lower than at DC operation due to losses in the array driver and active ripple reduction circuit.

Output Power	Input Power	Power Dissipation	Efficiency
1A @ 1.1 V	0.57A @ -12.10 V	5.8 W	16%
2A @ 2.2 V	1.06A @ -12.05 V	8.4 W	34%
2.5A @ 2.75 V	1.37A @ -12.01 V	9.6 W	42%

**Table 5.4 Measured Efficiency, Prototype, pulsed at 1 MHz**

At the 2 ampere current level, measured efficiency is 9 % less than that calculated in section 3.2.1. This is reasonable, because that calculation assumed that the power converter had zero loss, and did not consider current flowing in the control circuitry.

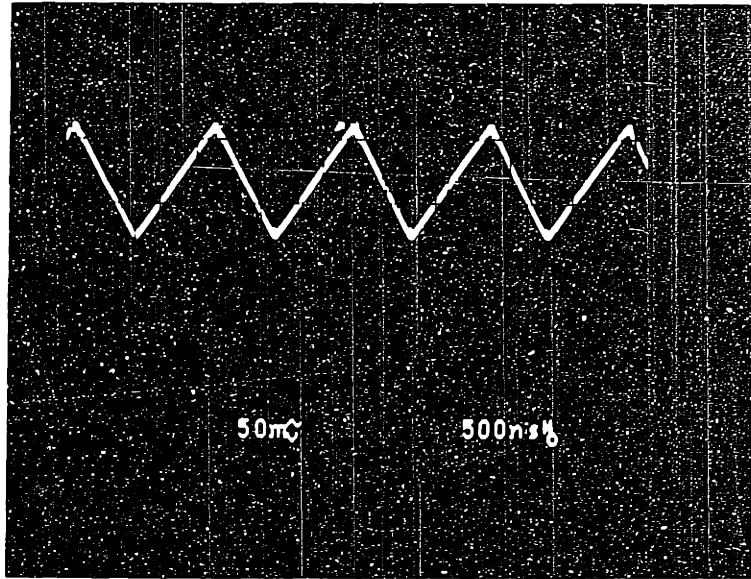
The primary losses are static losses in the MOSFET and Schottky diode. The  $R_{DS(on)}$  of the IRFR014 MOSFET is 0.2  $\Omega$ , maximum. The losses in the Schottky are comparable. The series resistance of the inductor is approximately 0.063  $\Omega$ . Using these numbers, total static losses total approximately 1.04 Watts, or 8.6% of input power.

For this circuit, efficiency is not of prime importance. It is important to note that 2.5 amperes may be delivered to the laser, while staying within the power supply specification. The heat sink has been eliminated on the prototype.

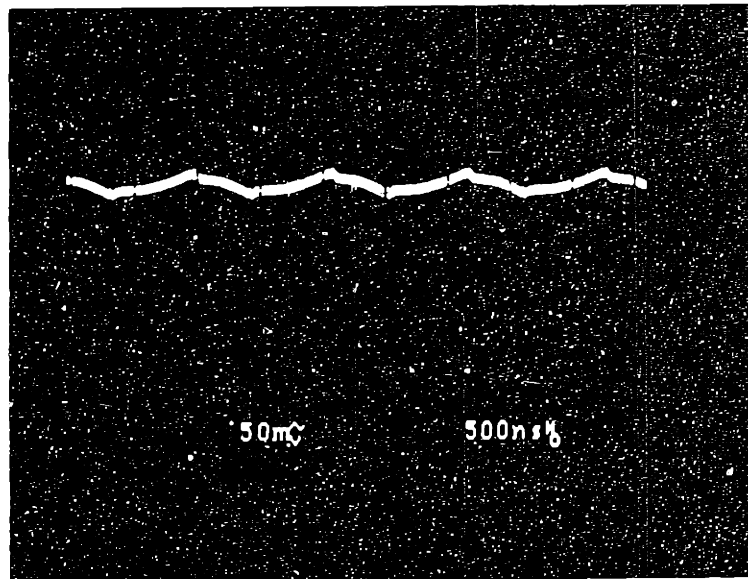
### **5.2.3 Laser Current Ripple**

As can be seen in **FIGURE 5.2a**, laser output current ripple due to the power converter is approximately 85 mA peak-to-peak.

**FIGURE 5.2b** shows performance of the driver when the active ripple reduction circuit is operating. Measured ripple is less than 15 milliamps peak-to-peak.



(a)  
50 mA/div



(b)  
50 mA/div

**Figure 5.2 Laser Driver Output Current Ripple**

## **Chapter 6**

### **CONCLUSIONS**

This thesis has demonstrated that a pulse circuit, suitable for driving high-power semiconductor laser diodes can be constructed using inexpensive components. The laser current is adjustable up to 3 amperes. Furthermore, the power efficiency of the circuit is significantly higher than previous linear designs. The risetime and falltime is a factor of 3 faster than the previous design.

Higher efficiency allows elimination of a large heat sink that is required on the previous design. Also, the power converter design allows 3 amperes to be delivered to the laser using a -12V power supply rated at 2 amperes. The new design will be manufactured using the same printed-circuit board form factor as the original design, with the elimination of the large heat sink.

If higher laser current is required, additional array transistors could be added in parallel. Since the dominant parasitic element affecting circuit operation is inductance, it is expected that high speed could be maintained at the higher current level.

If still higher switching speed is required, the power supply to the array drivers could be adjusted to give a larger voltage swing at the transistor bases..

Care must be taken in the layout of the printed-circuit board. A 4-layer PCB with a dedicated ground plane is anticipated. Hopefully, noise can be reduced through the use of ground plane techniques and high frequency surface-mount components.

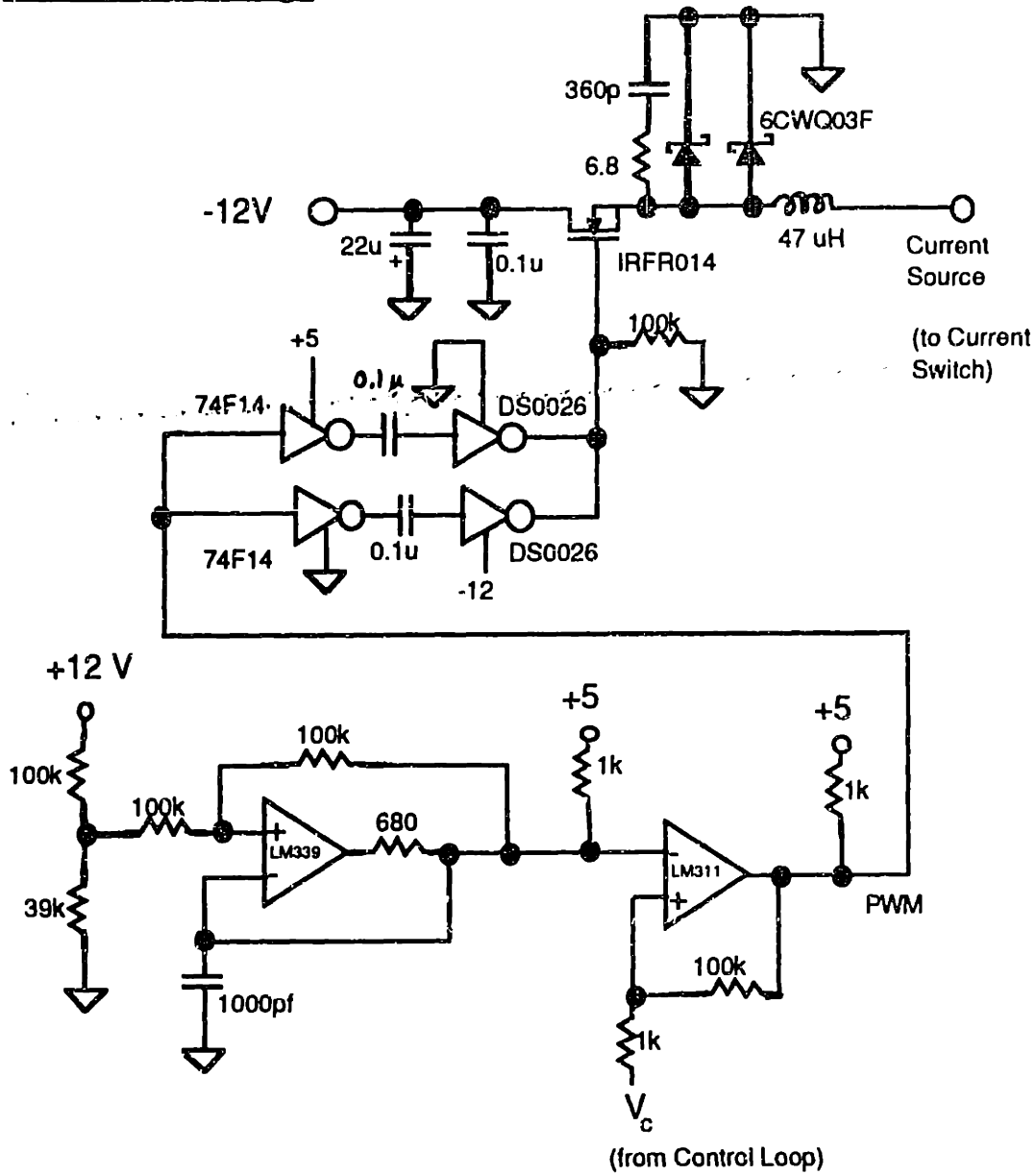
Achieving a fast switching speed while delivering additional power was the primary goal of this work. If higher efficiency is required, the circuit could be re-evaluated with that goal in mind. For instance, the laser bias current source could be derived from the power converter power supply, instead of the -12V supply.



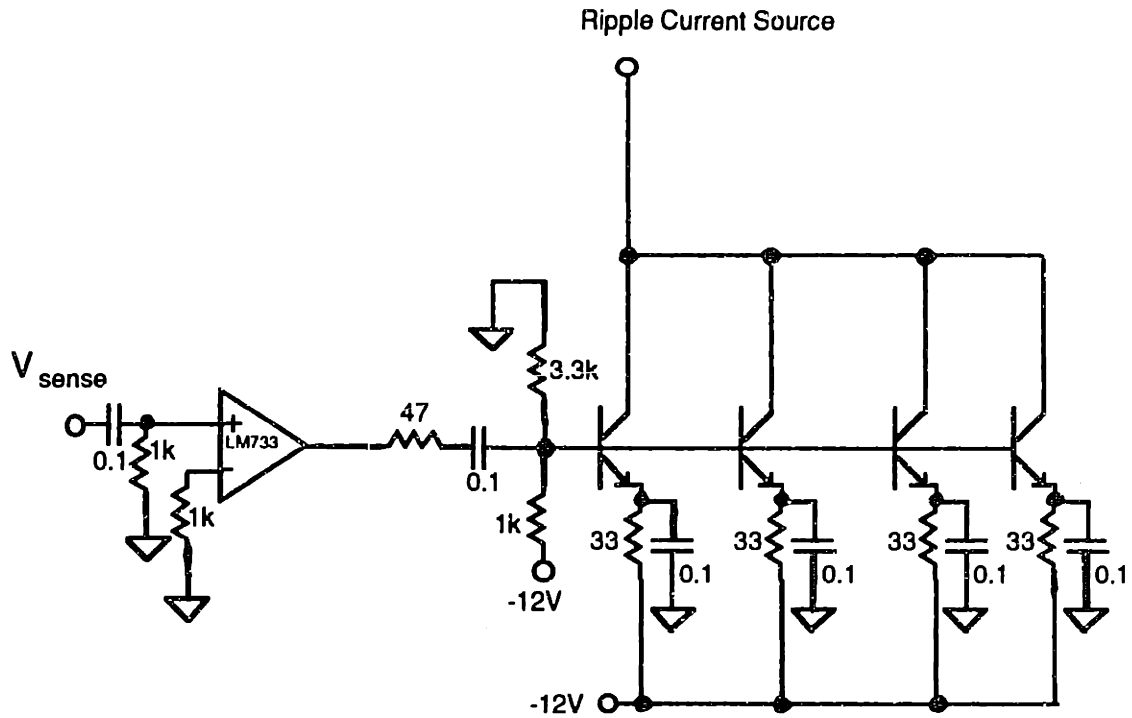
# **APPENDIX**

## **Circuit Schematics**

## A.1 Power Stage

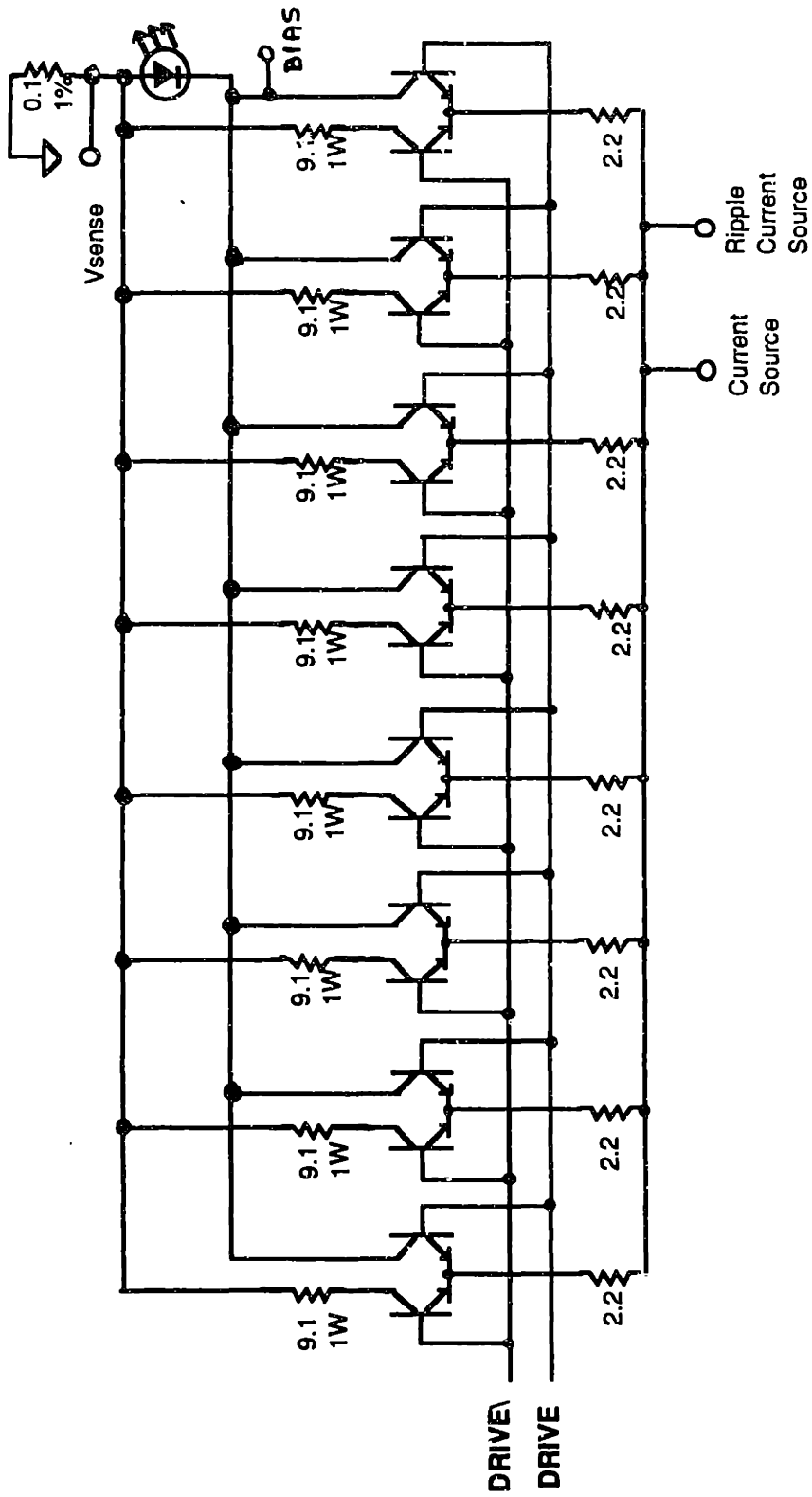


## A.2 Active Ripple Reduction Circuit



NOTE:  
All transistors 2N2222

### A.3 Current Switch



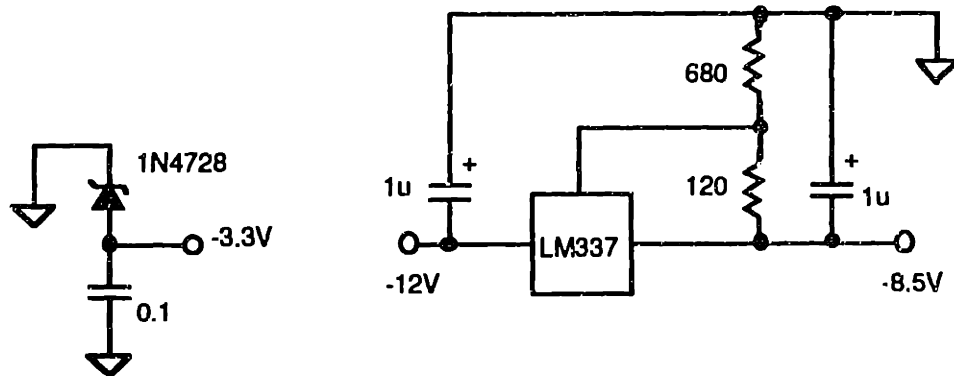
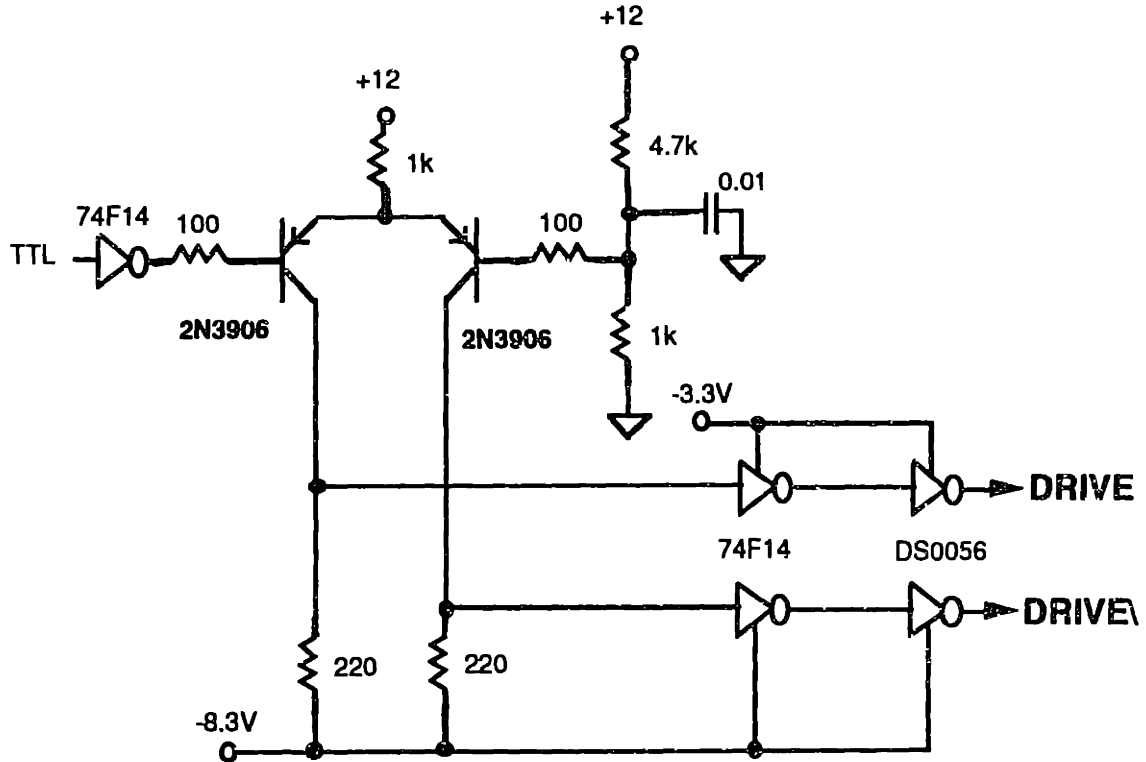
#### NOTES

1. All Transistors  
2N2222

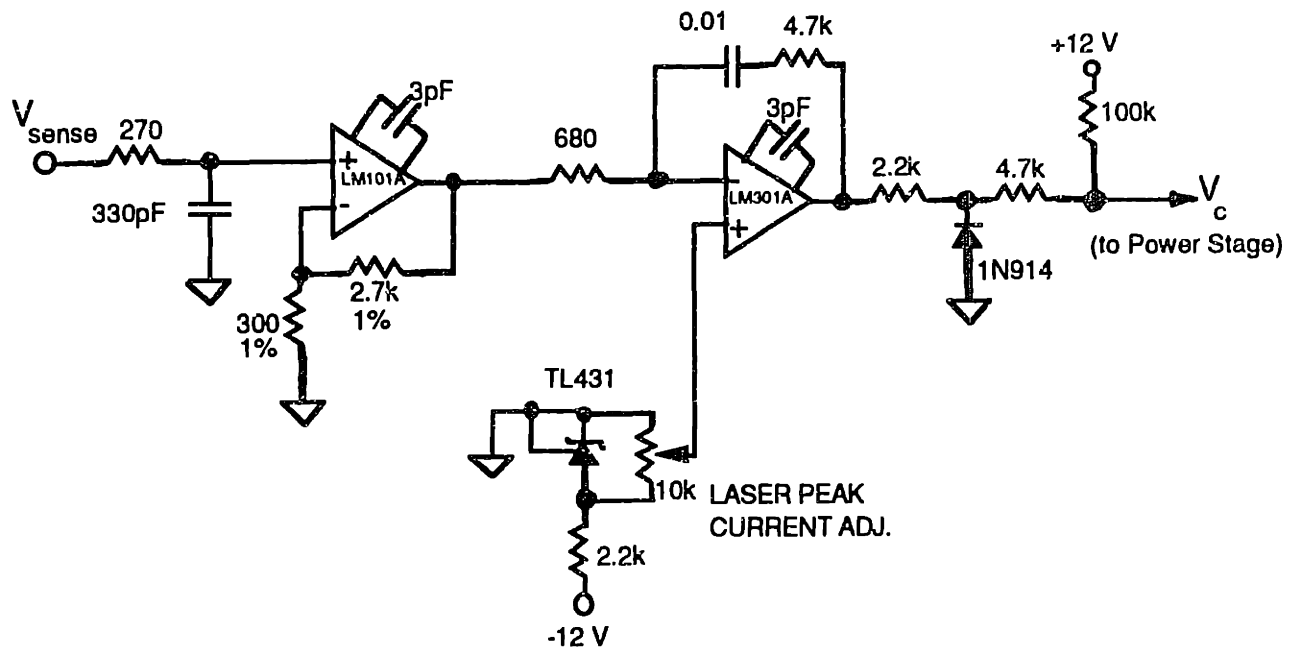
### CURRENT SWITCH

M. Thompson  
8/6/92

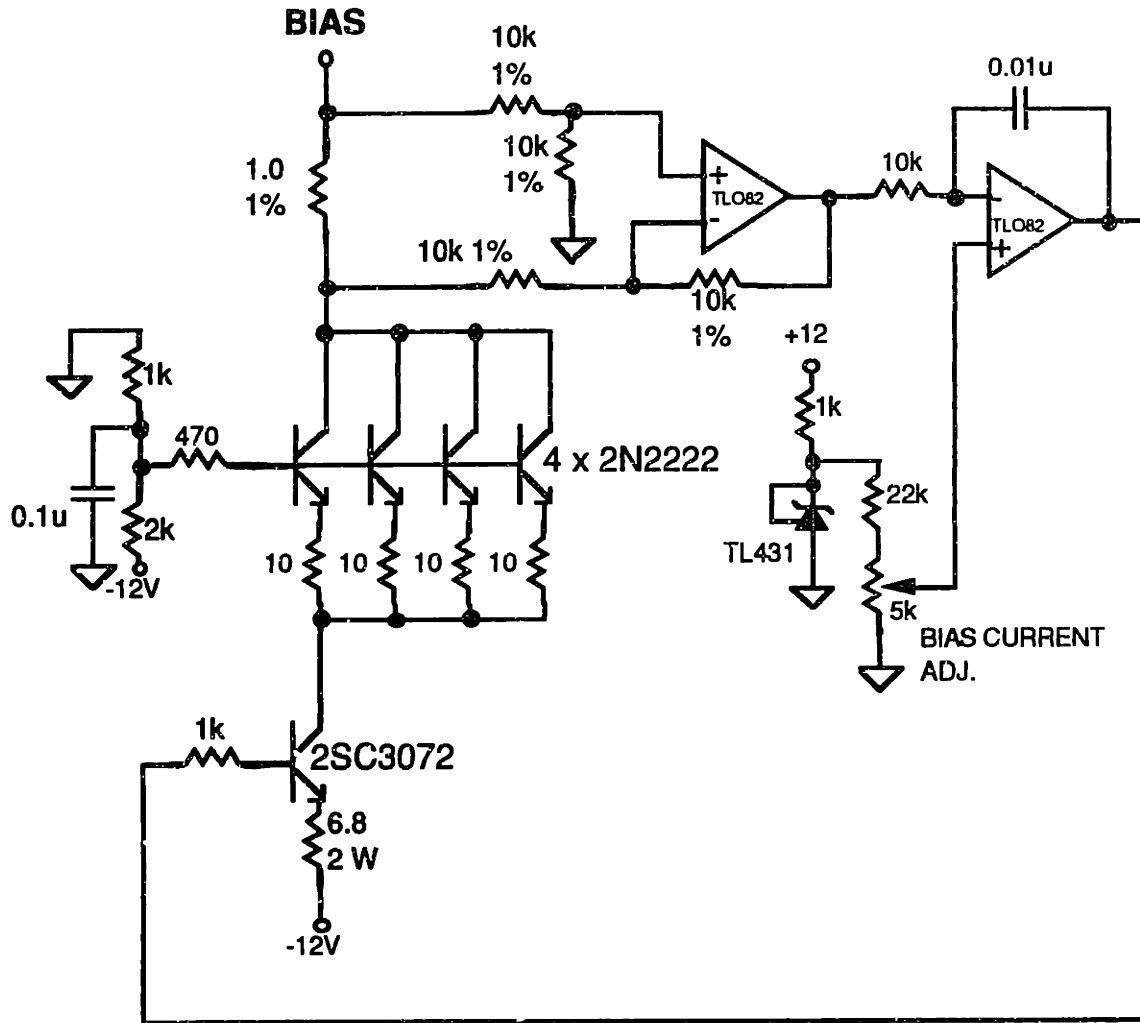
## A.4 Current Switch Driver



## A.5 Control Loop



## A.6 Laser Bias Current Source



## REFERENCES

- [1] Runge, Peter K.  
"An Experimental 50 Mb/s Fiber Optic PCM Repeater"  
IEEE Trans. Comm., Vol COM-24, #4 April 1976 pp. 413-418
- [2] Maione, T.L. and Sell, D.D.  
Experimental Fiber-Optic Transmission System for Interoffice Trunks"  
IEEE Trans. Comm., Vol COM-25, #5 May 1977 pp. 517-522
- [3] Chown, Goodwin, Lovelace, Thompson, Selway  
"Direct Modulation of Double-Heterostructure Lasers at Rates up to 1 Gbit/s"  
Electronics Letters, Vol. 9 #2 Jan. 25, 1973 pp. 34-36
- [4] Herden, W. and Metz, S.  
"A Fast Current Drive Circuit for Injection Lasers"  
IEEE J. Solid-State Ckts, June 1973 pp. 247-248
- [5] Filensky, Klein, Beneking  
"The GaAs MESFET as a Pulse Regenerator, Amplifier, and Laser Modulator in the Gbit/s Range"  
IEEE J. Solid-State Ckts, Vol. SC-12, #3 June 1977 pp. 276-280
- [6] Swartz, Wooley, Voshchenkov, Archer, Chin  
"An Integrated Circuit for Multiplexing and Driving Injection Lasers"  
IEEE J. Solid-State Ckts, Vol. SC-17, #4 Aug. 1982 pp. 753-760
- [7] Cooke, Sumerling, Muoi, Carter  
"Integrated Circuits for a 200-Mbit/s Fiber-Optic Link"  
IEEE J. Solid-State Ckts, Vol. SC-21, #6 Dec. 1986 pp. 909-915
- [8] Chen, Karr, Shumate  
"Laser Transmitters for 70-MHz Entrance Links"  
Bell System Technical Journal, Sept. 1979 pp. 1617-1629
- [9] Gruber, Marten, Petschacher, Russer  
"Electronic Circuits for High Bit Rate Digital Fiber Optic Communications" Systems"  
IEEE Trans. Comm., Vol COM-26, #7 July 1978 pp. 1088-1098
- [10] Shumate, Chen, Dorman  
"GaAlAs Laser Transmitter for Lightwave Transmission Systems"  
Bell System Technical Journal, July-Aug. 1978 pp. 1823-1836



[11] Gray and Searle  
Electronic Principles Physics, Models, and Circuits  
Published by John Wiley and Sons, New York 1969  
pp. 288-309

[12] Kassakian, Schlecht, Verghese  
Principles of Power Electronics  
Pub. by Addison-Wesley, New York 1991  
pp. 270

[13] Casey, Leo  
Circuit Design for 1-10 MHz DC-DC Conversion  
Doctoral Dissertation, Massachusetts Inst. of Technology 1989  
pp. 61