Accurate Controllable 325W Laser Diode Driver for Optical Inter-Satellite Links

Martin Persson
Product Line Management
Micross Hi-Rel Power Solutions
Herlev, Denmark
Martin.Persson@micross.com

Truls Mjelde Andersen
R&D Engineering
Micross Hi-Rel Power Solutions
Herlev, Denmark
TrulsMjelde.Andersen@micross.com

Bjarne Soderberg
Product Line Management
Micross Hi-Rel Power Solutions
Herlev, Denmark
Bjarne.Soderberg@micross.com

Steen Brandi
R&D Engineering
Micross Hi-Rel Power Solutions
Herlev, Denmark
Steen.Brandi@micross.com

***Abstract* — A 325W radiation hard, controlled current source is developed and qualified for use in optical inter-satellite communication links. The nature of the application requires accurate control of the current setpoint from 0.5A to 10A as well as adjustable maximum output voltage from 3V to 32.5V to ensure protection of the laser diodes. The wide range of operating conditions pose a challenge to ensure proper gain and phase margin while meeting requirements for transient response to load step and setpoint changes. The control scheme is an augmentation of peak current mode control to include separate setpoints for the voltage control and average current mode control which has proven effective in meeting the application requirements. This paper presents the converter topology, control scheme design and methods for radiation hardening of the circuits. The achieved performance of the current source is demonstrated by measurements on the manufactured hardware.**

Keywords — Power Conversion, Laser Diodes, Optical Communication, DC/DC Topology, Hy-Bridge, Control Scheme, Radiation Hardening.

I. Introduction

Optical inter-satellite communication has become one of the preferred technologies for communication systems that require low latency and high bandwidth. In this application laser diodes are biased by a controlled current source and depending on the diode configuration the resulting voltage across the diodes may be anywhere between 3 V and 32.5 V

For this purpose, controllable power supplies are needed to accurately bias the laser diode circuit. This type of power supplies can simplistically be compared to laboratory power supplies with separate setpoint for both constant current and constant voltage, letting the load determine which control loop is active. The voltage setpoint is used to limit the maximum safe operating voltage for a given laser diode configuration. As for all flight hardware size, mass and power efficiency is of high importance which together with the electrical specification and control features are key design drivers for the power supply presented in this paper.

II. Key Specification

The power supply must provide galvanic isolation between input and output and the key electrical specification for the current source is given in TABLE I.

TABLE I. Key Specification

Parameter	Min	Nom	Max	Unit
Input Voltage	24	28	34	V
Output Voltage Adjust	3		32.5	V
Output Current Adjust	0.5		10	A
Output Over Current Protection	10.5			A
Control Loop, Phase Margin	45			°
Control Loop, Gain Margin	10			dB

III. Topology

The electrical specification calls for a wide output voltage range with a factor of a 10.8 between the minimum and maximum output voltage. Also, the input voltage range is relatively wide with a factor of 1.4 between minimum and maximum input voltage. This combination means that the voltage conversion ratio from input to output spans from 0.09 to 1.35 thus the converter must be able to both step up and step down. This is trivial in a transformer isolated converter and can be accomplished with both Fly-back, Forward, Push-Pull, Half-Bridge and Full-Bridge based topologies. However, for best performance one must consider voltage stress, power dissipation, electromagnetic interference (EMI) and overall size and mass as well.

The Fly-back, Forward and Push-Pull based topologies are quickly discarded since all have poor utilization of the transformer core and winding area as well as high voltage stress on the primary side switch compared to Half-Bridge and Full-Bridge based topologies. With an output power requirement of 325 W the primary side switch current have significant magnitude and to minimize conduction losses, switches with low ON-resistance must be selected. With everything else equal this leads to selecting switches with lowest possible voltage rating.

For low input voltages a Full-Bridge based topology is preferred over Half-Bridge because the Half-Bridge only applies half the input voltage across the primary side winding and therefore calls for double the turns-ratio to compensate. As a result, the RMS current in the primary side switch becomes twice that of the Full-Bridge which increases the conduction loss in the switches by a factor of four. In addition to that the transformer may also have increased loss due to poor utilization of the winding area as a result of the need for a large turns-ratio.

With a Full-Bridge on the primary side, several options for secondary side rectifier configurations exist. The well-known Hy-Bridge converter with integrated magnetics offers lowest overall losses as discussed in [1]. The converter design has a secondary to primary turns ratio of 3.33 and an inductance of 80 uH in each of the output chokes. The switching frequency is constant at 100 kHz and the calculated operating conditions, electrical stress and efficiency is as presented in TABLE II.

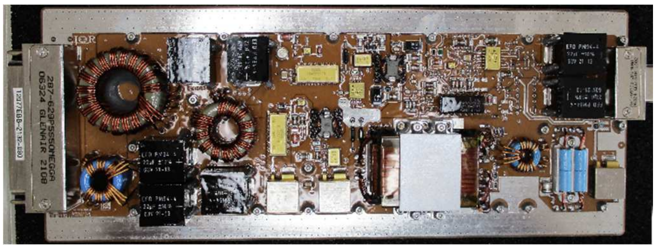

Fig. 1. Photo of manufactured laser diode driver

TABLE II. CALCULATED PERFORMANCE AT 325W OUTPUT

Parameter	Input Voltage			Unit
	24V	**28V**	**34V**	
Duty Cycle	42.1	35.9	29.4	%
Primary Switch Voltage, Plateau	24	28	34	V
Primary Switch Voltage, Peak	25.2	29.4	35.7	V
Primary Switch RMS Current	10.9	10.1	9.2	A
Secondary Rectifier Voltage, Plateau	79.9	93.2	113.5	V
Secondary Rectifier Voltage, Peak	103.9	121.2	147.2	V
Secondary Rectifier RMS Current	7.0	6.8	6.6	A
Transformer Primary RMS Current	16.4	15.5	14.4	A
Transformer Secondary RMS Current	4.9	4.6	4.3	A
Inductor Peak Current	6.20	6.33	6.46	A
Inductor Ripple Current	2.40	2.66	2.92	A
Conversion Efficiency incl. EMI Filters	92.6	92.8	92.9	%

The peak voltage stress across the rectifiers is calculated to be 147.2 V thus a rating of 200 V or more is needed to respect 75 % voltage derating and for that reason diode rectifiers have been selected. While radiation hardened MOSFETs with a voltage rating of 200 V are available the relative high output current means that the voltage drop across R.ds(on) especially at higher temperature may exceed the body diode forward voltage resulting in partial conduction in the MOSFET body diode. This leads to increased reverse recovery losses further heating up the devices to an extend where thermal stability is questionable. Fast recovery diodes designed for switch-mode applications is an attractive alternative that provides competitive power conversion efficiency without the risk of thermal instability. GaN switches is another attractive alternative because of the lack of body diode, no reverse recovery loss and low R.ds(on) compared to MOSFETs. Using GaN rectifiers will allow synchronous rectification and is considered a natural improvement to the chosen design which is based on HFB60HNS20SCS diodes rated for 200 V and 60 A. For the primary side MOSFET switches IRHNS57064SCS are selected rated for 60 V and 75 A.

IV. RADIATION HARDENING

The converter is radiation hard up to a total ionizing dose (TID) of 100 kRad and free from destructive single event effects (SEE) up to a linear energy transfer (LET) level of 60 MeV/mg/cm^2. The TID rating is achieved by using radiation hardened MOSFETs exclusively and radiation hardened ICs where possible. For other semiconductors susceptible to TID, radiation lot acceptance test (RLAT) data is included in the worst-case analysis to justify that performance is met across temperature, aging and radiation. Components susceptible to TID are bi-polar junction transistors (BJT), MOSFETs and ICs including op-amps, PWM controllers and voltage references. The diodes and Zener diodes used for the design are not susceptible to TID levels below 300 kRad.

The SEE rating is achieved by selecting suitable components that have been tested for heavy ion performance. The in-circuit operating conditions are analyzed and derating is applied in accordance with ECSS-Q-30-11A. SEE consists of several types of events. Single event upset (SEU) only occurs for digital ICs which are not used in this design. Single event burn-out (SEB) and single event gate rupture (SEGR) is relevant for MOSFETs. SEGR occurs before SEB for radiation hardened MOSFETs and to mitigate the effects the operating conditions for each individual MOSFET are analyzed and kept within the SEE rating provided in the MOSFET datasheets.

Single event latch-up (SEL) is a non-destructive effect that can occur for ICs. In the event of a SEL the converter may shutdown and cycling the input power is needed to reset the converter. The design includes ISL78845ASEH which is susceptible for SEL at LET levels below 60 MeV/mg/cm^2. The added failure rate is calculated using Omere to be 0.1 FIT (failure per 10^9 hours) based on a critical cross section of 10^{-7} cm^2 and a depth of the critical cross section of 5 µm using GEO integral LET spectrum.

Single event transients (SET) are non-destructive events where the output of an IC is pulled high or low for a short duration typically between 1 µs and 25 µs. In-circuit tests are performed for SET by emulating the actual transients observed during irradiation with heavy ions of the individual components. This is done by injecting a signal or momentarily shorting the relevant circuit node to the supply voltage of the IC in question. In practice an external BJT can be used to emulate the behavior observed during heavy ion exposure. Propagation of the transients is mitigated by filtering such that the transients are contained as much as possible. Some transients cannot be fully mitigated and may affect the output of the converter. This includes transients in the control loop error amplifier which can cause the PWM duty-cycle to change un-intended. By carefully tuning the control loop component values, an acceptable trade-off between control loop performance and SET induced output disturbances is found. It is important to notice that there are no latching effects caused by SET events.

V. CONTROL LOOP AND REGULATION SETPOINT

The converter regulation is based on the UC1825 peak current mode PWM controller which in a typical application regulates the output current based on voltage feedback. This application requires that the power converter can meet performance requirements over a large range of output voltages and output currents as wells as supply an adjustable constant current or an adjustable constant voltage.

Effectively the control loop has two setpoints and need to switch seamlessly between constant current and constant voltage mode in response to load and setpoint changes. In addition to the normal voltage feedback loop and error amplifier an average current mode control loop is implemented with a separate error amplifier. This control scheme enables full control of both voltage and current across the operating range and still includes the cycle-by-cycle current limitation provided by the peak current mode PWM controller to ensure fast limitation of the output current in the event of an overload.

To obtain the required accuracy for the adjustable output current a sense resistor is used to measure the output current directly. The measured signal is compared to the adjustable setpoint and the resulting error is feed into the PWM peak current mode regulator. The average current mode regulator is a type-2 or Proportional-Integral (PI) controller which ensures no static error.

The voltage regulator measures the output voltage directly via a resistor network. The output voltage is compared to the adjustable voltage setpoint and the resulting error is fed into the current regulator such that the output current setpoint is reduced if the output voltage exceeds the voltage setpoint.

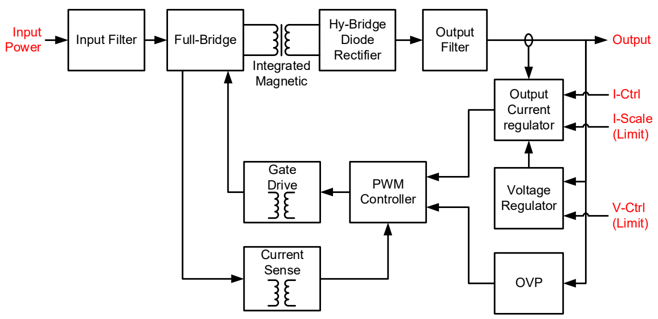

Fig. 2. Topology Block Schematic

The voltage regulator is a type-3 or Proportional-Integral-Derivative (PID) regulator which ensures that the static voltage error becomes zero and the derivative part allows for an increased loop gain at high frequencies to prevent or limit overshoot of the output voltage during turn-on or during rapid load current reduction. To minimize output transients when the converter switches between being controlled by the voltage and current loops clamp circuits are implemented to avoid integrator windup.

In addition to the voltage and current control inputs, a third control signal is used to scale the current setpoint by effectively change the gain of the current control signal. This is accomplished by a simple resistive divider which in combination with a user selectable external resistor changes the gain of the current control signal. This allows the user to match the laser diode voltage-current characteristic with the same control signal range and also set a maximum safe drive current for a given diode configuration

The block schematic in Fig. 2 shows the main converter and the top level control scheme. The implemented housekeeping circuits such as internal supply voltages, ON/OFF command and telemetries is not shown in the block diagram. The manufactured hardware is shown in Fig. 1.

VI. Performance

The control scheme objective is to provide accurate and well-behaved response to changes in setpoint and load changes. The open loop transfer functions for the voltage loop and current loop are shown in Fig. 3 and Fig. 4. The loop bandwidth, gain and phase margin are summarized in TABLE III. The measured gain and phase margin is greater than 20 dB and 60 ° which is well within the requirements and normally accepted stability criteria.

In addition to measuring the open loop transfer function and determining gain margin and phase margin a number of transient tests are performed. Fig. 5, Fig. 6 and Fig. 7 present the transient response to load changes with the voltage control loop active, the current control loop active and when transitioning from voltage loop to current loop. Both control loops respond in an acceptable manner to load changes without any ringing or sign of underdamped system response. In voltage mode a 50 % current load step results in a 3.56 V voltage transient. In current mode a resistive load step from 3 Ω to 5 Ω results in a 2.5 A current transient. The current transient is caused by charge and discharge of the output filter capacitance in response to the changed load resistance. During the transition to voltage mode it can be seen that the derivative part of the PID controller prevents significant voltage overshoot.

TABLE III. Control Loop Bandwidth and Stability Margin

Parameter	Voltage Mode Loop	Current Mode Loop	Unit
Bandwidth	4.47	7.62	kHz
Phase Margin	60.1	90.3	°
Gain Margin	20.4	21.7	dB

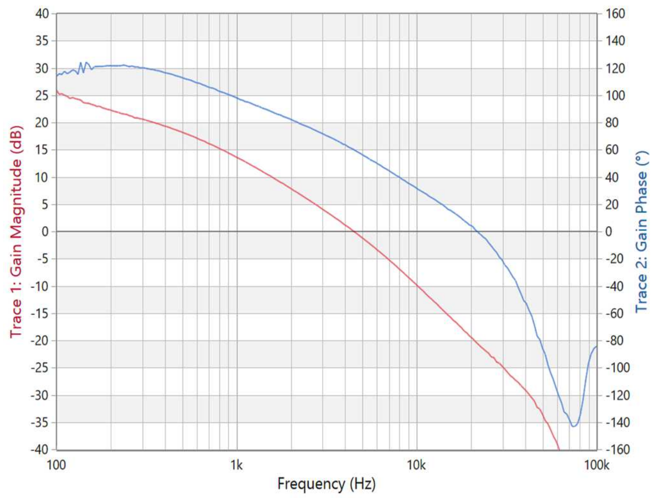

Fig. 3. Voltage mode open loop response at 28 V input, 20V output into 4 Ω resistive load. Red: Gain, Blue: Phase

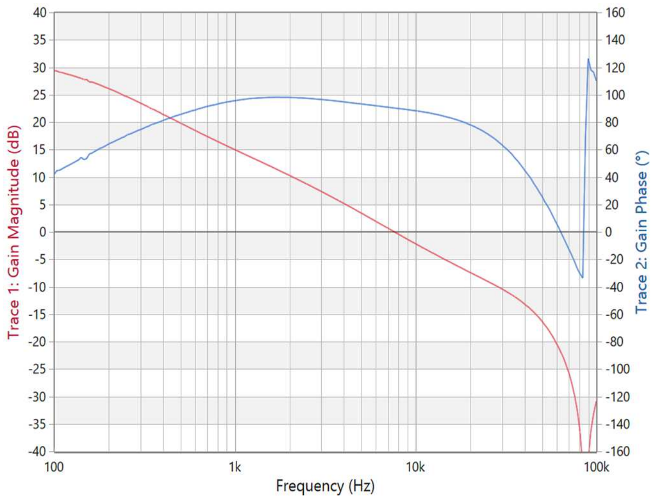

Fig. 4. Current mode open loop response at 28 V input, 5 A output into 4 Ω resistive load. Red: Gain, Blue: Phase

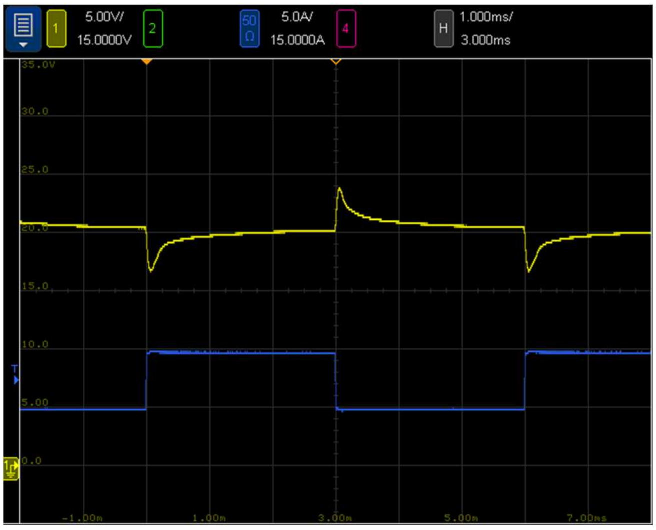

Fig. 5. Output transient response to a current load step from 5 A to 10 A and vice versa – voltage control active at 20 V. CH1: Output Voltage, CH3: Output Current

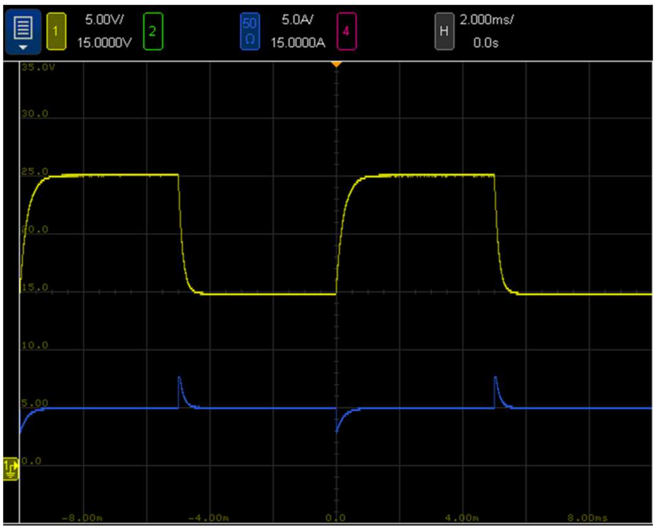

Fig. 6. Output transient response to a resistive load step from 3 Ω to 5 Ω and vice versa – current control active at 5 A. CH1: Output Voltage, CH3: Output Current

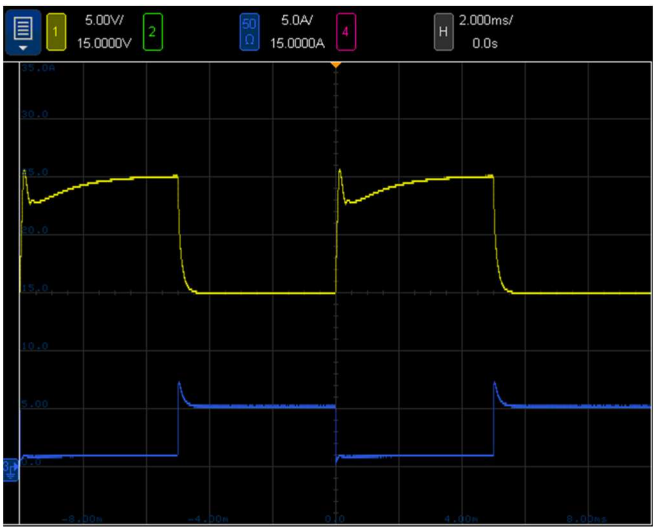

Fig. 7. Output transient response to a resistive load step from 30 Ω to 3 Ω and vice versa – converter transitions between current control mode at 5 A and voltage control mode at 25 V. CH1: Output Voltage, CH3: Output Current

Fig. 8. Output voltage setpoint change with diode load (fifteen BYV32F in series). Output current is uncontrained. CH1: Output Voltage, CH3: Output Current

Fig. 9. Output current setpoint change with diode load (fifteen BYV32F in series). Output voltage is uncontrained. CH1: Output Voltage, CH3: Output Current

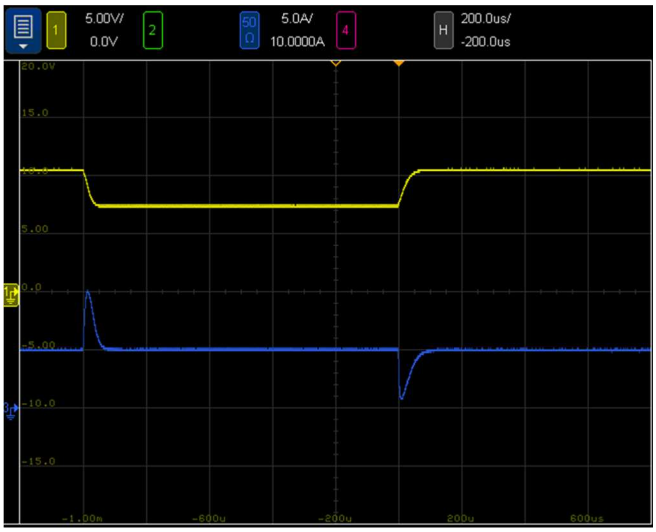

Fig. 10. Load step with diode load (fifteen BYV32F in series), five diodes shorted. Output current setpoint 5 A, Output voltage is uncontrained. CH1: Output Voltage, CH3: Output Current

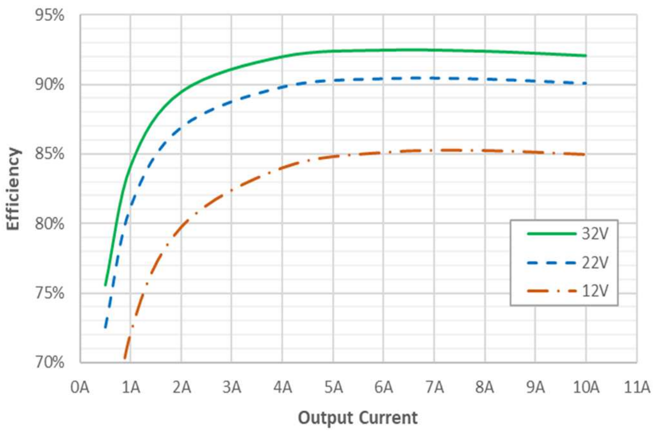

Fig. 11. Measured Power Efficiency vs Output Current at 28 V input, Parametric with Output Voltage.

The transient response to changes in the voltage and current setpoint is measured using a load consisting of fifteen BYV32F common cathode diodes connected in series where the two anodes of each package are tied together. This is done to emulate an actual use case of the laser diode driver. Three different scenarios are presented using the diode load. Fig. 8 show the response to a voltage setpoint change with unconstrained current limit. For a voltage change from 8.47 V to 10.53 V the current steps from 2.91 A to 8.09 A without any overshoot or ringing. The equivalent impedance of the diode load is calculated to be 398 mΩ.

In Fig. 9 the current setpoint is changed from 1.56 A to 8.67 A with the voltage loop unconstrained which results in a voltage step from 9.49 V to 12.45 V. The equivalent impedance is calculated to be 416 mΩ. The difference in calculated impedance in the two measurements can be contributed to difference in the diode temperature in the two tests.

In Fig. 10 five of the fifteen diodes are shorted using a MOSFET switch. The current set-point in the measurement is 5A and the output voltage is unconstrained. This results in a voltage change from 10.65 V to 7.36 V and results in a current transient of 5.12 A as the output filter is discharged. All measurements using the diode load show well-behaved transient response with no indication of an underdamped system response or ringing.

Finally, the power convertion efficiency is characterized across output voltage and load current with an input voltage of 28V. Fig. 11 shows the measured efficiency as a function of output current, parametric with output voltage. The peak efficiency is in excess of 92 % from 4 A to 10 A corresponding to 40 % to 100 % of full load. All internal losses, including auxiliary housekeeping circuits and EMI filtering are included in the measurements.

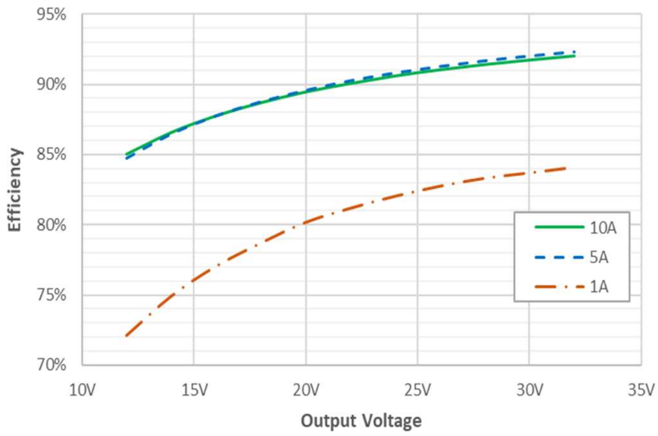

Fig. 12. Measured Power Efficiency vs. Output Voltage at 28 V input parametric with Output Current.

Fig. 12 shows the measured efficiency as a function of output voltage, parametric with output current. It can be seen that the efficiency for all practical purposes is the same for an output current of 5 A and 10 A. The measured efficiency matches the calculated efficiency relatively well. At 325W the measured efficiency is 0.8 %-point lower than the calculated efficiency.

VII. Conclusion

A radiation hard 325W DC/DC power supply specifically designed for driving laser diodes used in optical communication between satellites have been designed, developed and the design methods for radiation hardening the design has been discussed in detail. Key measurements have been presented and test results on the manufactured converter show best in class power efficiency in excess of 92 % including all housekeeping consumption and EMI filtering. Well-behaved system response to load changes in both voltage control and current control mode has been presented using both constant current load and resistive loads. Likewise, it is shown that the system response is well-behaved to setpoint changes and load changes when using a diode array load. It has been demonstrated how conventional peak current mode control can be augmented to include average current mode control and achieve seamless transition between voltage control and current mode control.

References

[1] C. Peng, H. Hanningan and O. Seiersen, "a New Efficient High Frequency Rectifier Circuit", Proceedings of HFPC, pp. 236-243, 1991.