Low Cost, Isolated Current Source for LED Strings

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Abstract This article describes the design of an isolated constant current source to drive high power LED's from an AC power line. The constant current source is built using a flyback topology and has an efficiency greater than 80%. Conventional current sources need many components for this. This application is a low-cost solution based on an innovative circuit (patent filed) using a Fairchild Power Switch (FPS). The converter is primary side regulated removing the need for an optocoupler and other additional components.

I. INTRODUCTION

Designers of lighting applications are making wider use of light emitting diodes (LED's) which combine high luminous efficiency with low energy consumption. High Power (high luminosity) LED's are now available in a compact package which is easy to mount in lighting applications. Application examples include traffic lights and street lighting, medical applications and consumer electronics such as cabinet lighting and table lamps. LED's are driven with a constant current. For low power LED's this is generated by a constant voltage source and a series resistor. High Power LED's need a constant current source: resistors would dissipate too much power.

II. CONSTANT CURRENT, CONSTANT VOLTAGE AND CONSTANT POWER SOURCES

A constant current source is a source that always supplies the same output current. The output conductance of the power source is zero. As a result its output resistance is infinitely high. Constant current sources are used if a change of the load must not change the output current.

LED's are diodes which are operated in the forward direction. They are driven with constant current for two reasons. First, this keeps the light output at a constant level. Second, if LED's were to be operated with constant voltage, any temperature rise would increase the mobility of charge carriers within the junction and the current would rise exponentially, destroying the device.

The circuit described in this paper can be used to implement a constant current source, a constant power source and a constant current source with foldback.

For ideal current sources the output current is constant for all output voltages. However for real current sources, the output voltage is limited as shown in the V-I characteristic plot in Figure II-1.

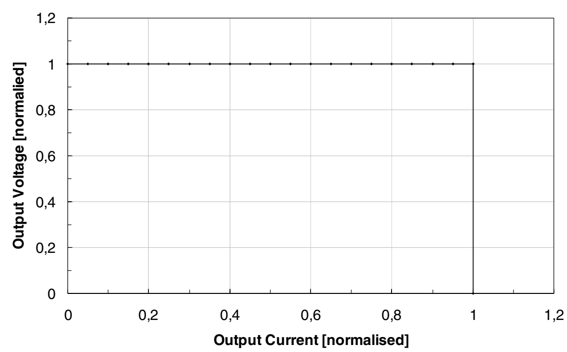

Figure II-1: Current and voltage curve of an idealized current source with a voltage limit

Unlike the output voltage of a constant current source, the output voltage of a constant power source decreases as the current increases. The product of output voltage and output current, or the output power, remains constant. This results in the hyperbolic characteristic shown in Figure II-2.

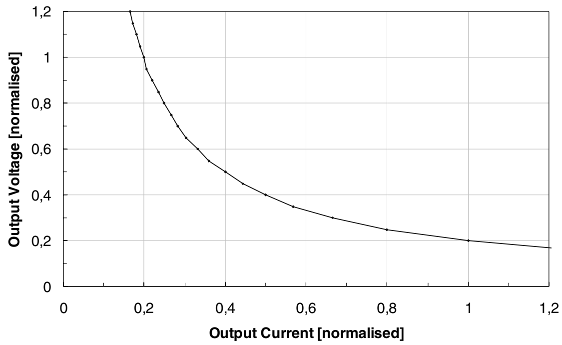

Figure II-2: Constant power characteristics

The following figure shows a typical foldback characteristic (Figure II-3). Once the maximum (nominal) output current is reached, additional reductions of the load resistance lead to reduction of both the output voltage and the output current. If the load is short-circuited, only a minimal current flows in the load.

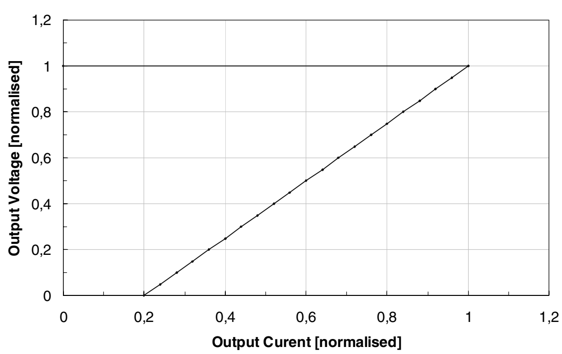

Figure II-3: Foldback characteristics

III. THE FAIRCHILD POWER SWITCH: FPS

The FPS is a solution for switch mode power supplies, requiring few external components. It integrates the pulse width modulator (PWM) and SenseFET into one package. The PWM block works in current mode, by measuring the current through the SenseFET and using this information to regulate the pulse width. The peak current through the SenseFET is proportional to the voltage at the feedback pin. Here the peak current is the maximum value of the current through the primary side of the flyback transformer, specifically ignoring capacitive turn-on effects masked by the leading-edge blanking circuit.

The isolated power supply presented in this paper uses the adjustable current limit feature of the FPS. An external resistor can be used to adjust the maximum permissible peak current in the integrated SenseFET.

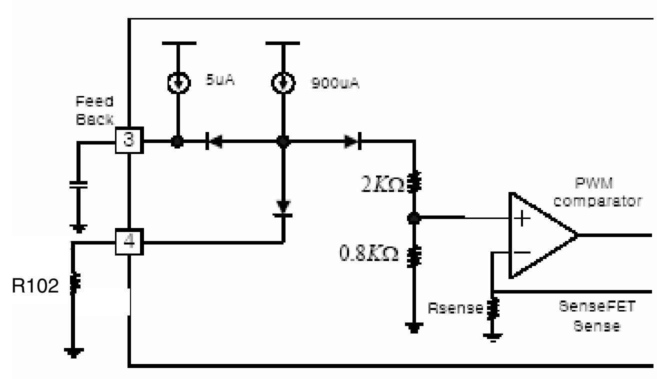

Figure III-1: Block diagram of current limit and feedback section of FPS

With reference to Figure III-1, the non inverting input of the PWM comparator is connected to an internal voltage divider network, which is fed by an internal current source. An additional resistance at the current limit pin thus forms a parallel connection to the internal resistance network. The voltage drop over this network caused by the current of the internal current source affects the PWM comparator. The latter compares the signals of the current limit pin (and feedback pin) with the current through the SenseFET and regulates the pulse width. The value of the resistance can be determined with the following equation:

$$R102 = \frac{R_{FPS} \times I_{LIMIT}}{I_{LIMITFPS} - I_{LIMIT}} \qquad (1)$$

Here I_{LIMIT} is the current limit desired by the designer, $I_{LIMITFPS}$ is the current limit specified in the FPS datasheet and R_{FPS} is sum of the two internal resistances, specified in the FPS datasheet. For example, R_{FPS} is 2kohm + 0.8kohm = 2.8kohm for the device shown in Figure III-1 which is the FSDH321BM. Both $I_{LIMITFPS}$ and R_{FPS} are device specific.

The FPS has other features relevant to this application. First, it has an internal start-up circuit, removing the need for an external one. Second, the switching frequency is modulated over a limited range, reducing peak EMI values as the spectrum of the conducted noise is spread over a range of frequencies. Further features include under-voltage lock out (UVLO), leading edge blanking (LEB), thermal shutdown (TSD), abnormal over current protection (AOCP). Compared with the controller and MOSFET solutions for switch mode power supplies, solutions using the FPS reduce the number of additional components, decreasing the design size and weight, and increasing efficiency, productivity and reliability. At light loads, the FPS operates in burst mode to improve efficiency.

IV. FLYBACK CONVERTERS

The flyback converter (Figure IV-1) is the most frequently used topology for switch mode power supplies in household and consumer electronics.

The main elements on the primary side are an input rectifier (D101, C102), a transformer (T1) to transfer energy and a switch (integrated in the FPS, IC101).

R101, C103 and D102 form a snubber network to limit the voltage spikes on the FPS drain generated by the leakage inductance of the transformer when the FPS MOSFET is switched off. The snubber reduces the voltage stress on the drain.

The FPS bias voltage comes from an auxiliary winding of the transformer, D105 being the output diode and C105 the output capacitor on the primary side of the transformer.

The signal measuring the output state is fed back to the primary side using an optocoupler.

The energy is transferred from the magnetizing inductance to the secondary side during the OFF state of the SenseFET. D201 is the flyback output diode. C201 is the output capacitor.

In this circuit the regulation is based on a measurement made on the secondary side. The error amplifier is also on the secondary side. IC201 combines an optocoupler and an industry standard FAN431 circuit which in effect combines an error amplifier with a reference.

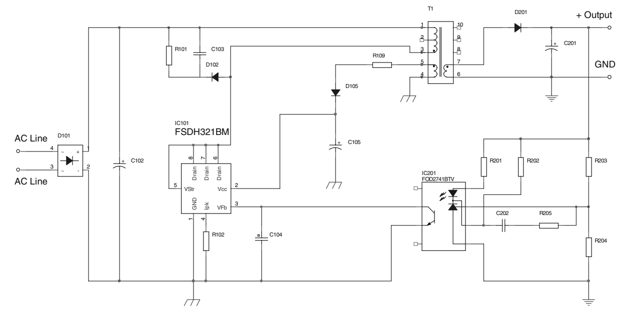

Figure IV-1: Voltage output flyback converter with secondary side regulation

The voltage divider (R203 and R204) sets the output voltage. The optocoupler LED is biased using R201. The reference circuit is biased using R202. The full circuit with the addition of R205 and C202 forms a Type 2 compensator (integrator, gain, and pole-zero pair) to compensate the power supply.

An alternative, lower cost possibility (not shown) is to put a suitably dimensioned Zener diode in series with the optocoupler LED and a bias resistor. However, the high variability of the Zener diode and the absence of a compensator with an integrator term make the system regulation less accurate and less stable.

Primary side regulation can be realized with fewer components. In this case there is no need for an optocoupler, even for a fully isolated power supply.

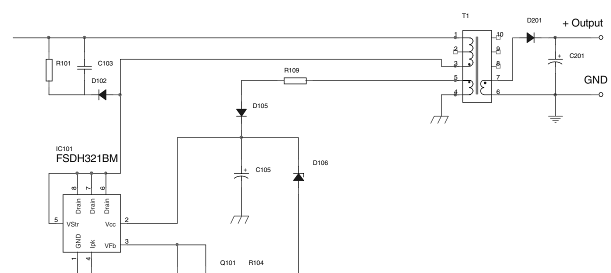

Figure IV-2 Voltage output flyback converter with primary side regulation

Figure IV-2 shows an isolated flyback design using primary side regulation. Winding W3 (between pins 4 and 5) of transformer T1 is used to generate the feedback signal. The voltage on this winding is proportional to the voltage on winding W2 (between pins 7 and 6), as long as the two windings are coupled very well. An increase in the load results in a reduction in the current in D106. This results in a lower base current for Q101. This decreases I_C of Q101, so V_{FB} increases. The controlled peak current in the SenseFET increases, which results in an increase in the duty cycle and therefore the output voltage. R110 is the bias resistor for the Zener diode and R104 is the base resistor for Q101.

C104 reduces the noise on the feedback pin and influences the control loop transfer function.

An additional function of C104 is to set the turn off delay in case of an output overload condition. An overload condition will tend to drive the voltage on the feedback pin high. If the voltage on the feedback pin is 4V (nominal), the PWM controller operates at the maximum duty cycle specified in the datasheet. If the voltage on C104 exceeds this level, the PWM controller continues to operate at maximum duty cycle. Above 4V, C104 is charged by an internal current source of around 5uA. If the overload condition persists, the voltage on C104 will sooner or later exceed the shutdown feedback voltage, around 6V. In this case, the FPS is shut down. By dimensioning the output capacitor, the time during which overload conditions is tolerated can be adjusted. The power supply can support temporary overload conditions but will shut down after longer overload conditions.

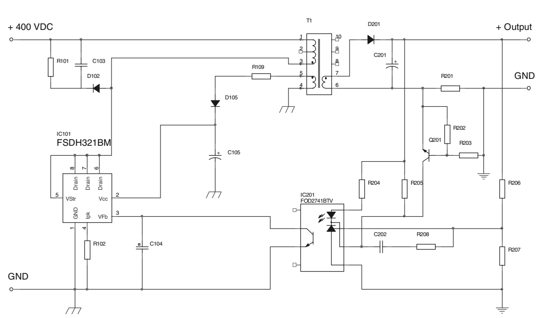

Figure IV-3: Constant current output, secondary side regulated flyback converter

A flyback converter can be used to implement a constant current source as shown in Figure IV-3. This is achieved by introducing a measurement resistor R201. The voltage across this resistor is proportional to the load current. So regulating the voltage across this resistor in a similar way to that of a voltage output flyback circuit results in a regulated output current. Since the V_{BE} of Q1 is around 0.7 V, the power dissipation in R201 (measure resistance) is very high with large output current. Further as Q1's V_{BE} is temperature dependent, an NTC (negative temperature coefficient) resistor (R202) is required. This kind of regulation is not very exact, because of the thermal variations. The efficiency of this circuit depends on the preset output current. In order to achieve a good performance for larger output currents an operational amplifier circuit (Figure IV-4) must be used. We note that both circuits need optocouplers.

As the intention is to describe a simpler circuit, the function of the circuit in Figure IV-4 will not be explained here in detail.

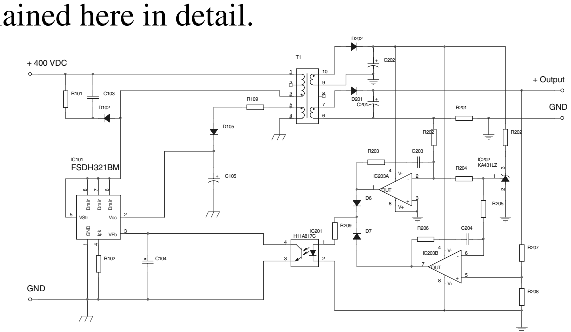

Figure IV-4: High constant current output, secondary side regulated flyback converter

V. Constant Current Output Primary Side Regulated Flyback Converter

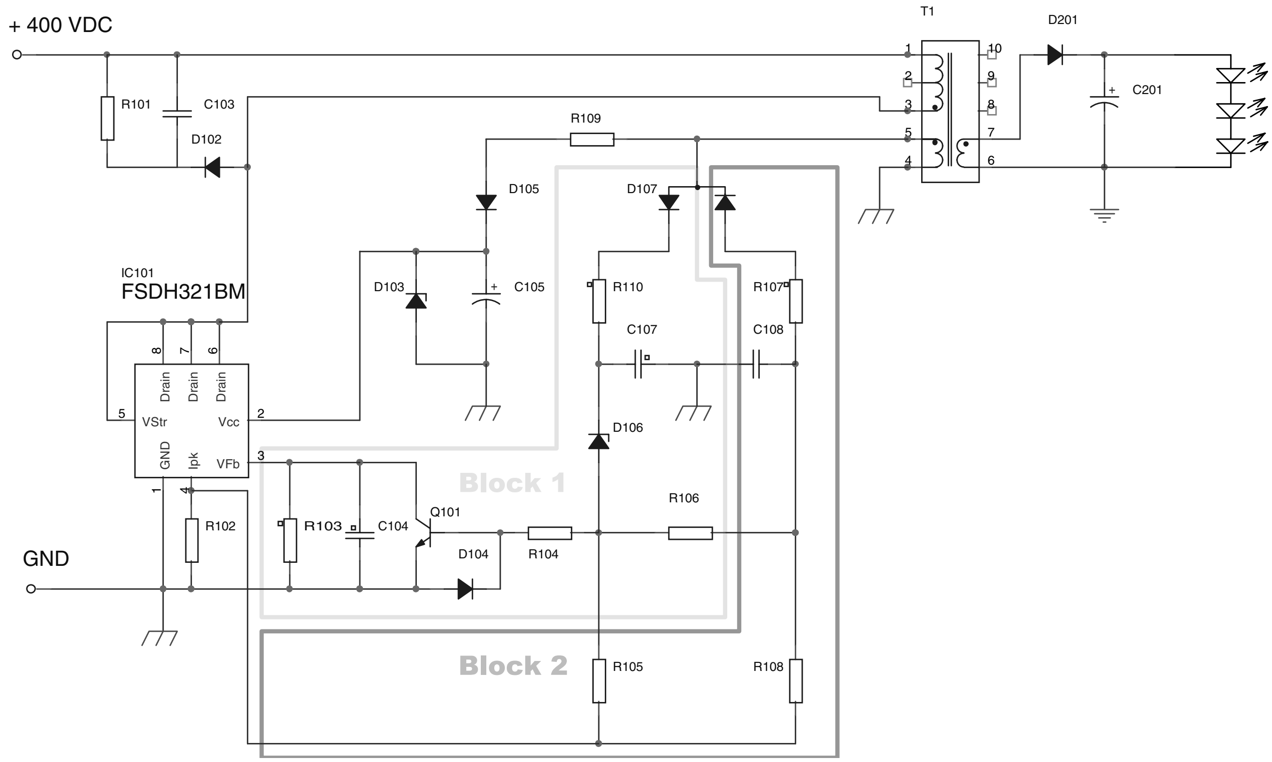

Figure V-1: Constant current output primary side regulated flyback converter

Figure V-1 shows the circuit of a primary side regulated constant current output flyback power supply.

The components required for the primary side regulation can be split into two blocks. Block 1 is similar to the primary side regulation circuit described earlier, with the inclusion of R103 and D104. Block 2 is added to provide the constant current output function.

The function of Block 1 is explained first. For low load currents, the output voltage is constant as described earlier. An increase in load results in a lower output voltage and consequently a lower voltage on winding W3 (pins 4 and 5). This is because W3 is well coupled to the secondary output winding W2 (pins 6 and 7). The voltage on C107 is equal to V_{BE} of Q101 plus the Zener voltage of D106. A reduction in this voltage results in a lower current through D106, leading to a lower base current in Q101. So voltage V_{FB} rises, increasing the duty cycle and ultimately the output voltage.

When the current limit set by R102 is reached, the peak current is constant and the output voltage sinks with increasing load current. This results in a constant power V/I characteristic.

For discontinuous conduction mode (DCM) operation, this can easily be shown. The power transferred from the primary to the secondary (and leakages) is:

$$P_O = \frac{1}{2} L_P I_{PEAK}{}^2 f_{SW} \tag{2}$$

where L_P is the primary inductance, I_{PEAK} is the peak current and f_{SW} is the switching frequency. So if I_{PEAK} is limited, the power transfer is limited, resulting in a constant power characteristic.

The analysis for continuous conduction mode (CCM) operation is more complex and beyond the scope of this paper. The conclusions of the analysis and the experimental results show an almost constant power characteristic.

R103 in Block 1 disables the overload circuit protection described earlier, which is not explicitly needed in a constant current output application. Without R103 in constant power mode, the voltage on the feedback would rise to the shutdown voltage level and switch off the power supply.

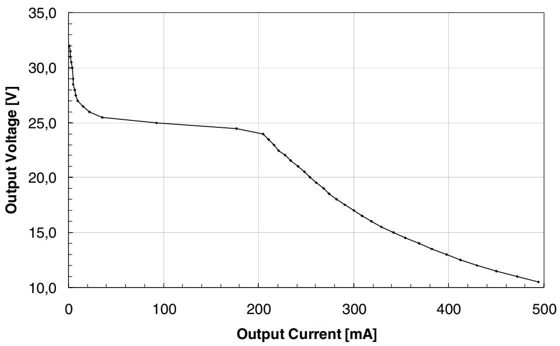

Figure V-2: Primary side controlled constant current output circuit with Block 1 activated (R102 = 1500 ohms)

The above figure shows the effect just described. At low currents, the circuit operates in constant voltage mode. Above 200mA output current, the circuit operates in constant power mode.

To turn the circuit into a constant current source, the components shown in Block 2 are required. Block 2 performs input line regulation and load regulation. The negative part of the voltage Vcc on W3 is rectified by D107 and filtered by R107 and C108. The resulting voltage is negative with respect to ground, and is proportional to the line input voltage. Additionally it generates a negative bias voltage needed for load regulation. As a precaution, a diode D104 is included in Block 1 to protect Q101 against possible negative base bias.

At light loads, the circuit operates in constant voltage mode. At the point where constant power mode is reached, the voltage on the anode of D106 will drop. An additional current therefore flows through the current limit pin through R105, so that the current limit is reduced further. This results in a positive feedback signal. The desired characteristics can be achieved by modifying resistor R105: a large value is used for constant current mode and a lower one for foldback operation.

R108 is used to compensate the output current for variations in the input voltage, which would otherwise increase the power output. An increase in input voltage, results in an increased current flow from the current limit pin, which ultimately reduces the power output.

D105, C105 and R109 form the bias power supply for the FPS. D103 is needed if the voltage used for the regulation is higher than the desired power supply level for the FPS.

VI. Design Guidelines

There are design tools for the dimensioning a flyback design on the internet homepage of Fairchild Semiconductor: www.fairchildsemi.com. These tools simplify the calculation of the transformer as well as the snubber network and components on the secondary side. If there is a wide range of output voltages, Vcc should be selected in such a way that its value is in the middle of this range. If the value of Vcc is too high for the supply of the FPS, the Zener diode D103 will be required. Otherwise, an additional auxiliary winding is necessary.

When using the design tools from Fairchild Semiconductor, the input voltage, output voltage and output current must be specified. As a rule of thumb use 2-3 µF/W for the DC link capacitor. The next step is to choose a value for voltage V_{RO}, the reflected output voltage. The sum of V_{RO}, the maximum input voltage and the snubber voltage should not exceed 80-90% of the rated voltage for the switching element.

The next steps are to insert the switching frequency, the adjusted current limit and the ripple factor into the calculation sheet. In DCM, the ripple factor is set to 1.

For the transformer design, it is useful in some applications to use a larger core than needed to keep the current stress on the FPS and on the secondary side lower. For the individual windings the value for the current density is considered to be 5 A/mm2. After that the number of parallel wires and their diameters can be determined easily.

Select output diodes with a voltage rating equal to 1.5 to 2 times that calculated for voltage stress. The output capacitors should have a very low impedance (low ESR) to reduce output voltage spikes.

For the snubber calculation assume 2.5% of primary inductance as the value for the leakage inductance. Set the snubber capacitor voltage to 1.5 V_{RO}.

The feedback signal in this application, as already described, is generated by the current source of the FPS.

Output current is adjusted by resistors R102, R105, and R108.

The value of the peak current limit to be used for a particular output current is found as follows. The constant current operation occurs at the transition between the constant output voltage mode and constant output power mode. This occurs when:

$$V_O I_O = \frac{1}{2} L_P I_{PEAK}^{\;2} f_{SW} \tag{3}$$

where V_O is the output voltage limit required by the system design and I_O is the output current.

An initial value for R102 may be calculated by solving (3) for I_{PEAK}, then substituting this result for I_{LIMIT} in (1). This result for R102 must then be increased to compensate for the currents drawn by R105 and R108.

In practice, R102, R105 and R108 are determined by experimentation. R105 adjusts the current difference during minimum and maximum output voltage. The effect of the line input voltage is adjusted with R108. Table A-2 in the Appendix shows the incremental approach used to determine these values.

As a rule of thumb R102 is in the range of 1 to 5 kilohm, R105 in the 10 to 50kohm and R108 in the 100 to 200 kilohm range.

The output voltage limit is set by the Zener diode voltage (D106):

$$\mathbf{V_Z + V_{BE} + V_F = \frac{n_{S3}}{n_{S2}}(V_O + V_F)} \tag{4}$$

where V_Z is the Zener diode voltage, V_{BE} is the operating base-emitter voltage of Q101, V_F is the forward conducting voltage of D107 and D201, n_{S2} is the number of turns on the secondary output and n_{S3} is the number of turns on the auxiliary output.

VII. Test Results

The results presented here are for an application to drive high power OSRAM LED's having a forward voltage of around 3V driven at a current of 700mA. These LED's are very bright. Do not look at the LED's when they are switched on. The output voltage of the current source can vary between 12V and 22V depending on how many LED's are connected in series.

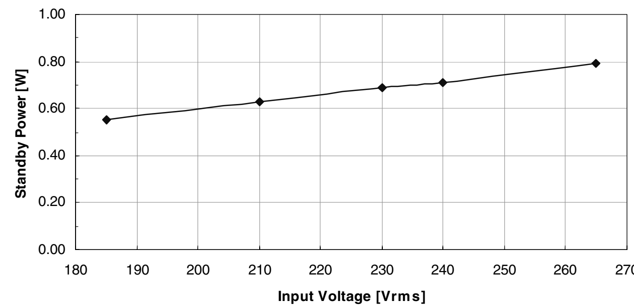

Figure VII-1: Standby power versus input voltage

Figure VII-1 shows the standby power versus input voltage for no load.

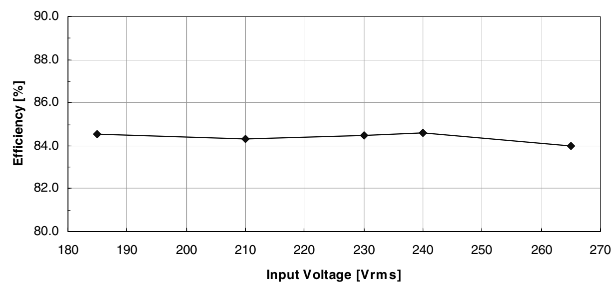

Figure VII-2: Efficiency versus input voltage

The constant current source was driven at the maximum power and the efficiency was measured for a range of input voltages.

For measuring the V/I characteristics, the output voltage was varied and the output current was measured. These tests were performed using an input voltage of 230Vrms.

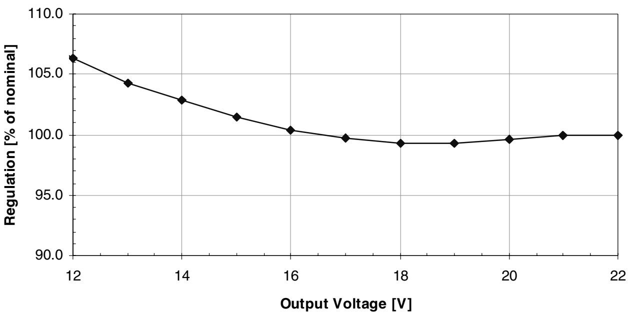

Figure VII-3: Regulation versus load (output voltage)

The constant current plot shows that the current level stays within 5% of the nominal except at very light loads, showing excellent performance.

In comparison with Figure II-1, we show the constant current characteristic of the test circuit. Figure VII-4 shows the plot when R105 and R108 have the values specified in the schematic of Figure A-2. R102 is not mounted, so the internal current limit of the FPS is used for this purpose. The voltage limit between 23V and 25V dominates the curve.

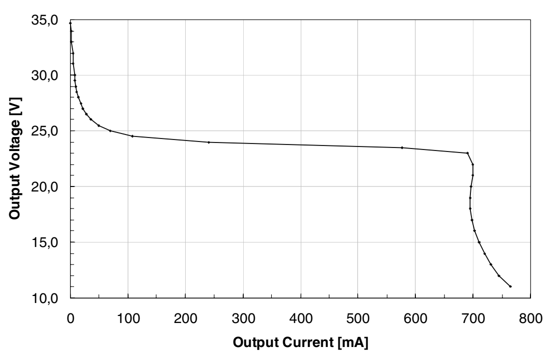

Figure VII-4: V/I characteristic for primary side regulated constant current source

By adjusting the resistors, a foldback characteristic can be obtained. Here R102 was set to 2 kilohm, which sets the maximum current at just over 200mA, and R105 set to 15 kilohm, which is a low resistor value forcing the foldback. R108 was not mounted.

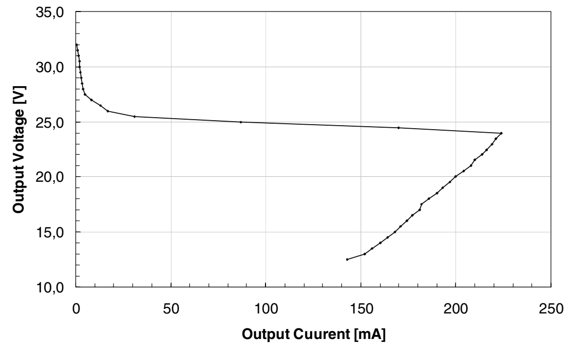

Figure VII-5: Foldback characteristic

Finally the EMI plot was measured at maximum load and at nominal input current. The results show that there is no problem to meet EN55011/22 Class B EMI limits.

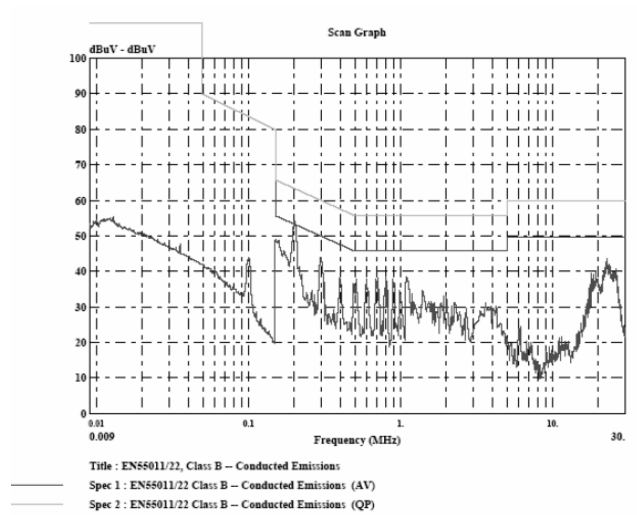

Figure VII-6: EMI plot for test circuit

VIII. CONCLUSION

In comparison with other isolated flyback constant current output power supplies, the reviewed circuit using the FPS has fewer components offering a lower cost system solution. It offers an ideal solution for driving high intensity LED’s.

REFERENCES

[1] Fairchild Semiconductor Application Note AN4105: Design considerations for Switched Mode Power Supplies Using A Fairchild Power Switch (FPS) in a Flyback.

[2] Fairchild Semiconductor Application Note AN4137: Design Guidelines for Off-line Flyback Converters Using Fairchild Power Switch (FPS)

[3] Fairchild Semiconductor Application Note AN4141: Troubleshooting and Design Tips for Fairchild Power Switch (FPS) Flyback Applications

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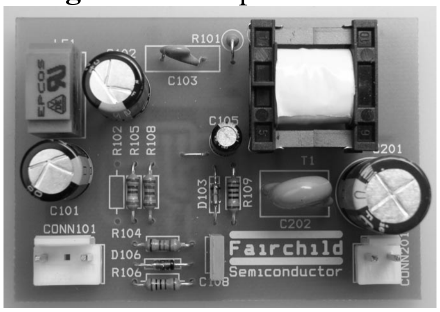
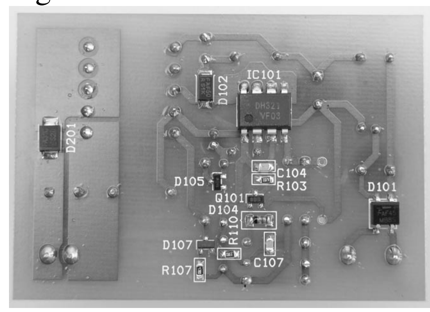
APPENDIX

Figure A-1: Top and bottom views of primary side regulated constant current source





Top view **Bottom view**

Dimensions: 65.5mm x 47.0mm x 25.0 mm (Lx B x H)

Figure A-2: Circuit diagram

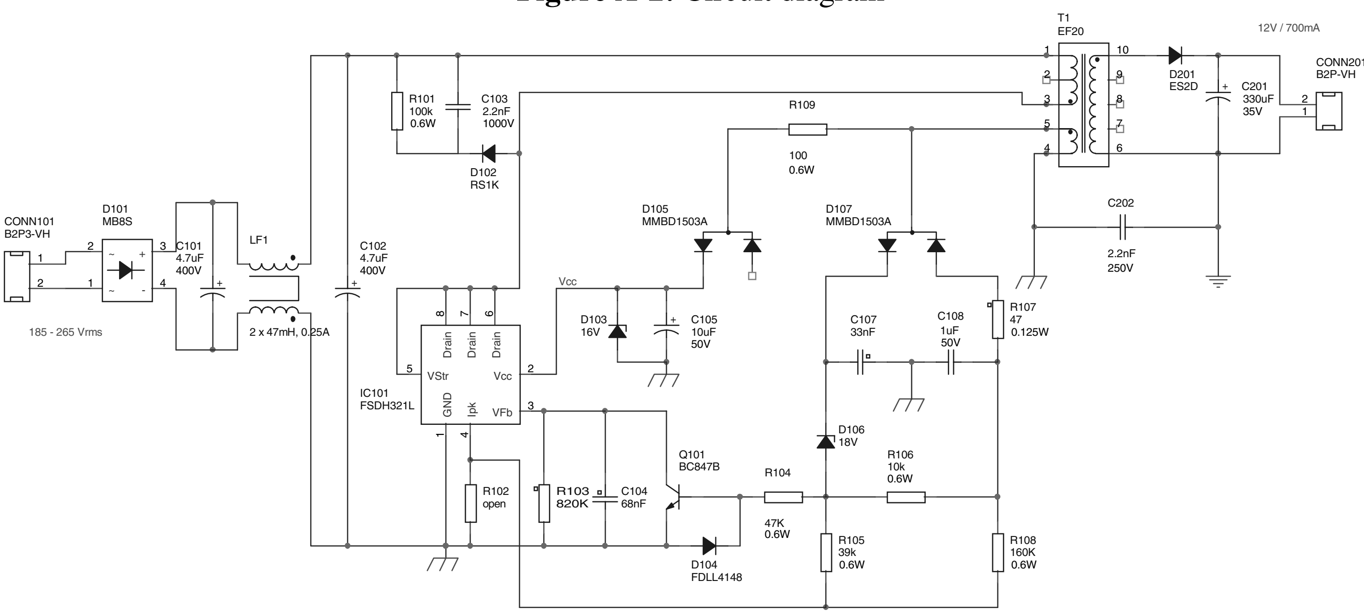


Table A-1: Demo Board Specification

Minimum Input Voltage	185 V_{RMS}
Maximum Input Voltage	265 V_{RMS}
Frequency	50 Hz
Output Voltage and Current	12 V – 22 V / 700 mA constant current

Table A-2: Example of experimental step-by-step adjustment to obtain constant output current

V_{INMIN}	V_{INNOM}	V_{INMAX}	R102	R105	R108	at V_{OUTLOW}	at $V_{OUTHIGH}$
X			Open	62 k	140 k	741 mA	671 mA
	X		Open	62 k	140 k	701 mA	640 mA
		X	Open	62 k	140 k	684 mA	626 mA
X			Open	39 k	150 k	703 mA	708 mA
	X		Open	39 k	150 k	665 mA	674 mA
		X	Open	39 k	150 k	654 mA	664 mA
X			Open	39 k	160 k	730 mA	733 mA
	X		Open	39 k	160 k	703 mA	704 mA
		X	Open	39 k	160 k	693 mA	697 mA

Table A-3: Transformer Specification

Name	Pinning	Layers	Diameter	Turns	Construction	Material
W1a	3 →2	2	1 x 0.18 mm	52	Solenoid	CuLL
W2	7 →6	2	1 x 0.5 mm	19	Solenoid	CuLL
W1b	2 →1	2	1 x 0.18 mm	52	Solenoid	CuLL
W3	5 →4	1	1 x 0.15 mm	15	Spaced	CuLL

*CuLL is copper wire with two thin insulation layers

Core: E20
Material: N27 (Epcos) or equivalent
Bobbin: E20 vertical / 10 pins
Gap in center leg: approximately 0.34 mm for A_L of 135nH/Turns^2

Figure A-3: Transformer Construction

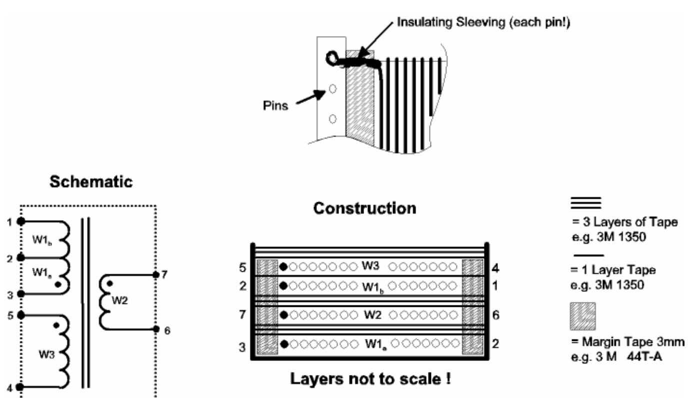


Table A-4: Transformer Characteristics

Parameter	Pins	Specification	Conditions
Primary Inductance	1 → 3	1459 µH +/- 5%	10 kHz, 100 mV, all secondaries open
Leakage inductance	1 → 3	73 µH maximum	10 kHz, 100 mV, all secondaries short