AN937
APPLICATION NOTE

DESIGNING WITH L4971, 1.5A HIGH EFFICIENCY DC-DC CONVERTER

1 INTRODUCTION

The L4971 is a 1.5A monolithic dc-dc converter, step- down , operating at fix frequency continuous mode. It is realised in BCD60 II technology, and it is available in two plastic packages, DIP8 and SO16L.

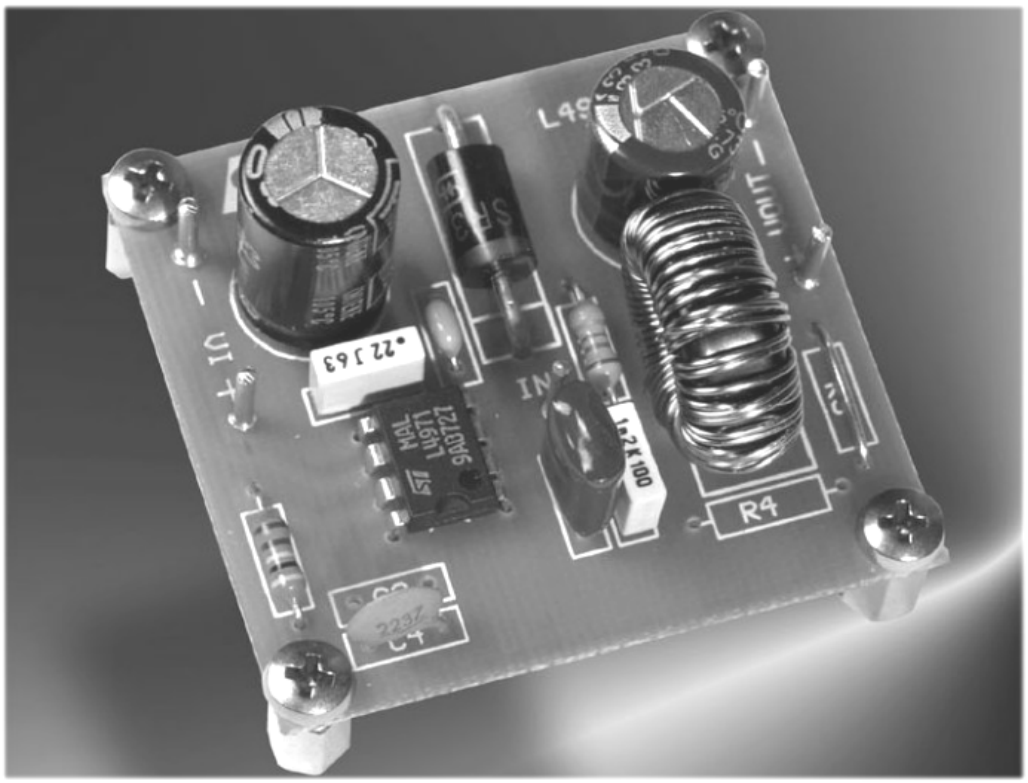
One direct fixed output voltage at 3.3V ±1% is available, adjustable for higher output voltage values, till 40V, by an external voltage divider.

The operating input supply voltage ranges from 8V to 55V, while the absolute value, with no load, is 60V. New internal design solutions and superior technology performance allow to generate a device with improved efficiency in all the operating conditions and with reduced EMI due to an innovative internal driving circuit, and reduced external component counts.

While internal limiting current and thermal shutdown are today considered standard protection functions, mandatory for a safe load supply, oscillator with voltage feedforward improves line regulation and overall control loop.

Soft-start avoids output overvoltages at turn-on, while, shorting this pin to ground, the device is completely disabled, going into zero consumption state.

Figure 1. .



2 DEVICE DESCRIPTION

For a better understanding of the device and it's working principle, a short description of the main building blocks is given here below, with packaging options and complete block diagram. Figure 2 show s the two packaging options, with the pin function assignments.

Figure 2. Pins Connections .

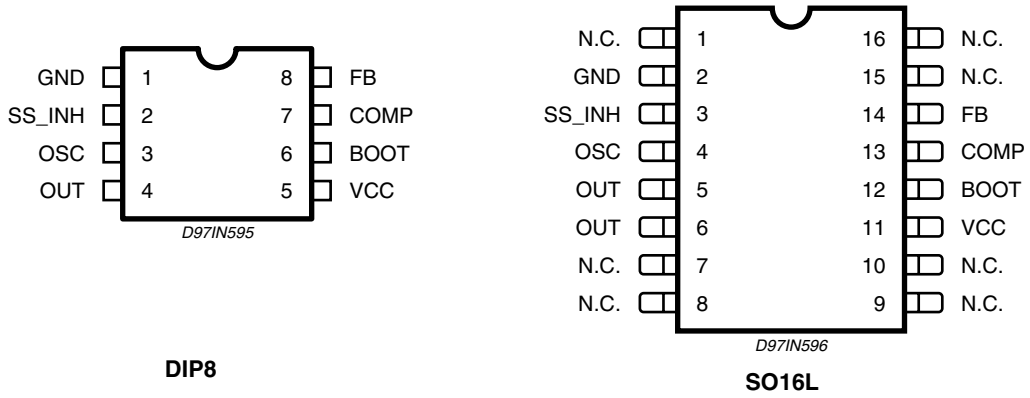
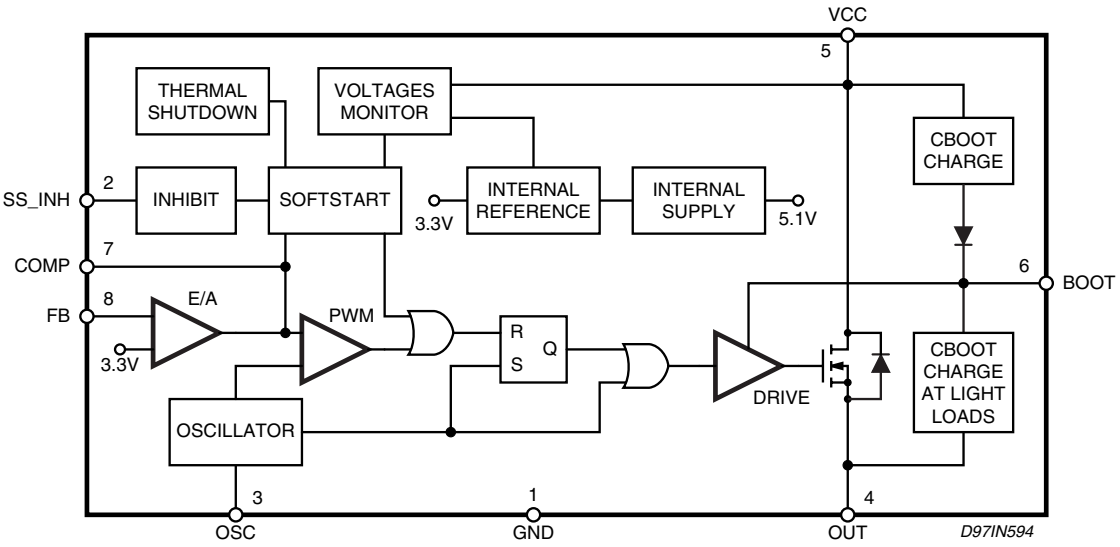


Figure 3. Block Diagram.



3 POWER SUPPLY, UVLO AND VOLTAGE REFERENCE.

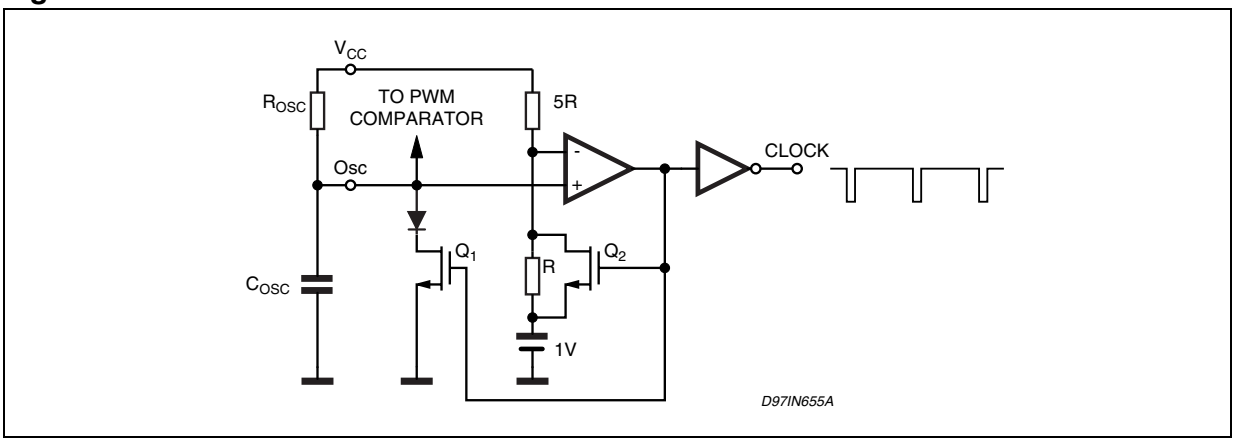
The device is provided with an internal stabilised power supply (of about 12V typ.) that powers the analog and digital control blocks and the bootstrap section.

From this preregulator, a 3.3V reference voltage ±2%, is internally available.

3.1 Oscillator and voltage feedforward

Just one pin is necessary to implement the oscillator function, with inherent voltage feedforward.

Figure 4. Oscillator Internal Circuit



A resistor Rosc and a capacitor Cosc connected as shown in fig. 4, allow the setting of the desired switching frequency in agreement with the below formula:

$$F_{SW} = \frac{1}{(R_{oasc} \cdot C_{osc}) In\left(\frac{6}{5}\right) + 100 \cdot C_{osc}}$$

Where F_{sw} is in kHz, R_{osc} in KΩ and C_{osc} in nF.

The oscillator capacitor, C_{osc}, is discharged by an internal mos transistor of 100Ω of R_{dson} (Q1) and during this period the internal threshold is setted at 1V by a second mos, Q2 . When the oscillator voltage capacitor reaches the 1V threshold the output comparator turn-off the mos Q1 and turn-on the mos Q2, restarting the C_{osc} charging.

The oscillator block, shown in fig. 4, generates a sawtooth wave signal that sets the switching frequency of the system.

This signal, compared with the output of the error amplifier, generates the PWM signal that will modulate the conduction time of the power output stage.

The way the oscillator has been integrated,does not require additional external components to benefit of the voltage feedforward function.

The oscillator peak-to-valley voltage is proportional to the supply voltage, and the voltage feedforward is operative from 8V to 55V of input supply.

$$\Delta V_{osc} = \frac{V_{CC} - 1}{6}$$

Also the ΔV/Δt of the sawtooth is directly proportional to the supply voltage. As Vcc increases, the Ton time of the power transistor decreases in such a way to provide to the chocke, and finally also the load, the product Volt. sec constant.

Fig 6 show how the ducty cycle varies as a result of the change on the ΔV/Δt of the sawtooth with the Vcc. The output of the error amplifier doesn't change to maintain the output voltage constant and in regulation. With this function on board, the output response time is greately reduced in presence of an abrupt change on the supply voltage, and the output ripple voltage at the mains frequency is greately reduced too.

Figure 5. Device switching frequency vs Rosc and Cosc.

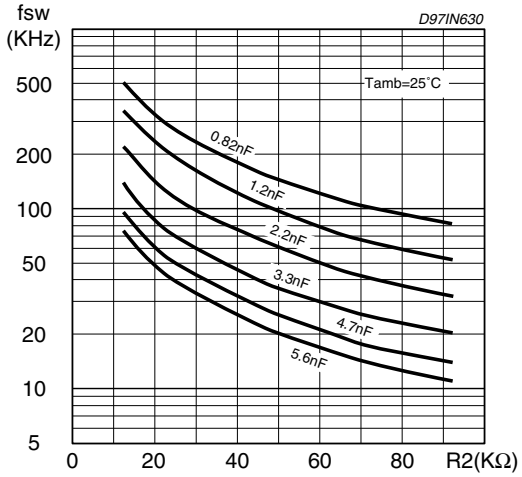


Figure 6. Voltage Feedforward Function.

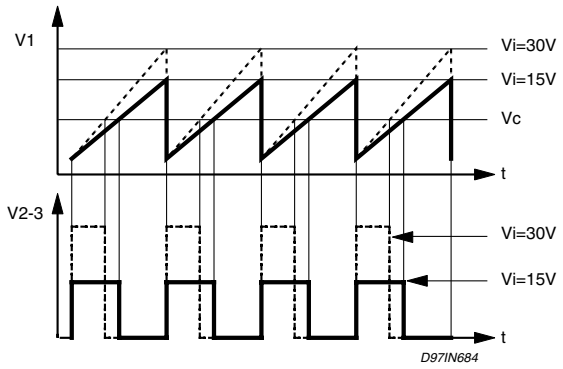
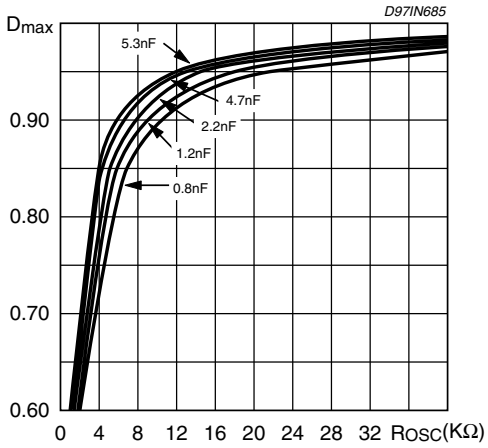


Figure 7. Maximum Duty Cycle vs. Rosc and Cosc as parameter.



In fact, the slope of the ramp is modulated by the input ripple voltage, generally present in the order of some tents of Volt, for both off-line and dc-dc converters using mains transformers. The charge and discharge time is approximable to:

$$Tch = R_{osc} \cdot C_{osc} \cdot \ln\left(\frac{6}{5}\right)$$

$$Tdis = 100 \cdot C_{osc}$$

The maximum duty cycle is a function of Tch, Tdis and an internal delay and is represented by the equation:

$$Dmax = \frac{R_{osc} \cdot C_{osc} \cdot \ln\left(\frac{6}{5}\right) - 80 \cdot 10^{-9}}{R_{osc} \cdot C_{osc} \cdot \ln\left(\frac{6}{5}\right) + 100 \cdot C_{osc}}$$

and is rapresented in figure 7:

3.2 Current protection

The L4971 has two current limit levels, pulse by pulse and hiccup modes.

Increasing the output current till the pulse by pulse limiting current treshold (Ith1 typ. value of 2.5A) the controller reduces the on-time till the value of T_B = 300ns that is a blanking time in which the current limit protection does not trigger. This minimun time is necessary to avoid undesirable intervention of the protection due to the spike current generated during the recovery time of the freewheeling diode.

In this condition, because of this fixed balnking time, the output current is given by:

$$I_{max} = \frac{[V_{CC} \cdot T_B \cdot F_{SW} - V_f \cdot (1 - T_B \cdot F_{SW})]}{[R_O + (R_D + R_L)(1 - T_B \cdot F_{SW}) + (R_{dson} + R_L) T_B \cdot F_{SW}]}$$

In fig 8, the pulse by pulse protection is sufficient to limit the current.

In fig 9 the pulse by pulse protection is no more effective to limit the current due to the minimun Ton fixed by the blanking time TB, and the hiccup protection intervenes because the output peak current reachs the relative threshold. At the pulse by pulse intervention (point A) the output voltage drops because of the Ton reduction, and the current is almost constant. Going versus the short circuit condition, the current is only limited by the series resistances RD and RL (see relation above) and could reach the hiccup treshold (point B), set 20% higher than the pulse by pulse. Once the hiccup limiting current is operating, in output short circuit condition, the delivered average output current decreases dramatically at very low values (point C).

Figure 8. Output Characteristics

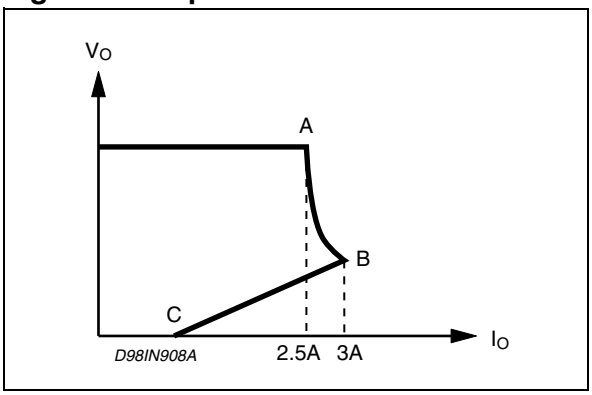


Figure 9. Output Characteristics

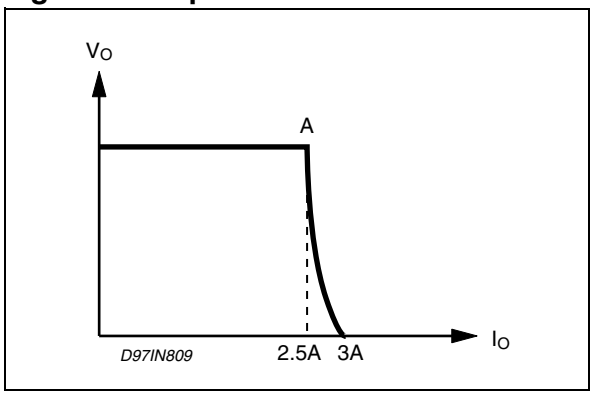


Figure 10. Current Limiting Internal Schematic Circuit.

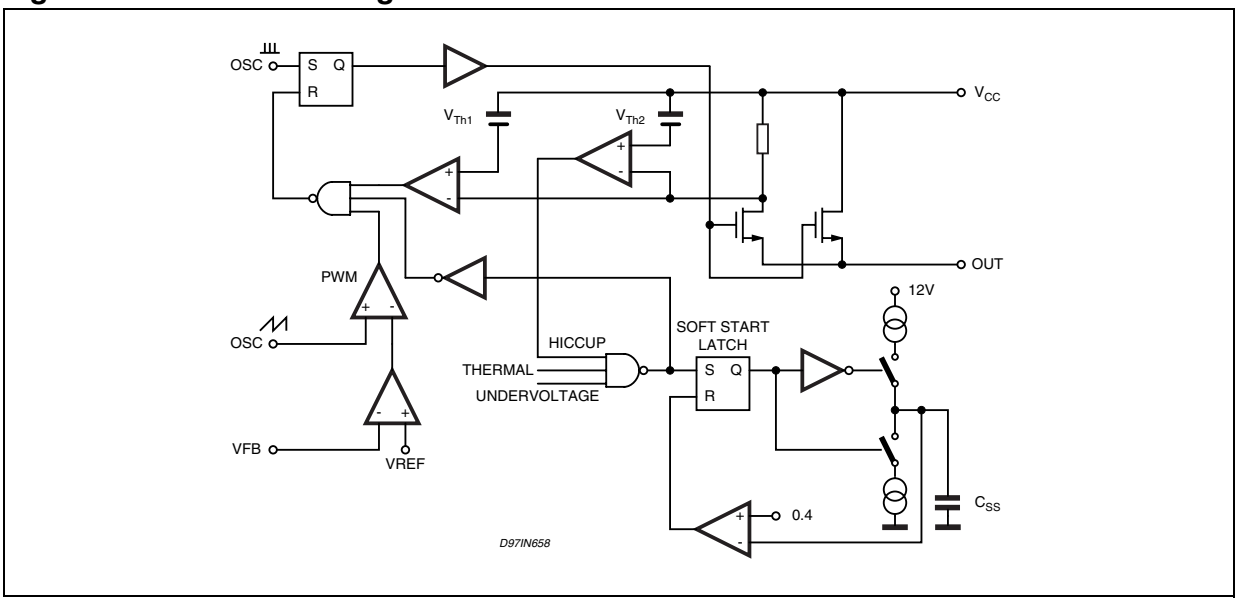


Figure 11. Output current and soft start voltage

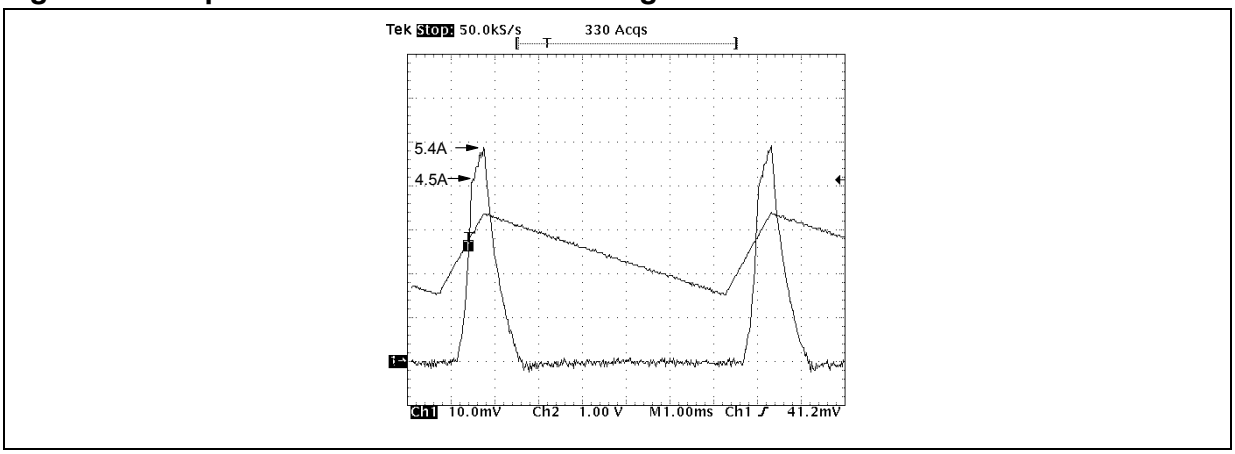


Fig. 10 shows the internal current limiting circuitry. Vth1 is the pulse by pulse while Vth2 is the hiccup threshold.

The sense resistor is in series with a small mos realised as a partition of the main DMOS.

The Vth2 comparator (20% higher than Vth1) sets the soft start latch, initialising the discharge of the soft start capacitor with a constant current (about 22mA). Reaching about 0.4V, the valley comparator resets the soft start latch, restarting a new recharge cycle.

Fig. 11 Shows the typical waveforms of the current in the output inductor and the soft start voltage (pin 2).

During the recharging of the soft start capacitor, the Ton increases gradually and, if the short circuit is still present, when Ton>TB and the output peak current reachs the threshold, the hiccup protection intervenes again. So, the value of the soft start capacitor must not be too high (in this case the Ton increases slowly thus taking much time to reach the TB value) to avoid that during the soft start slope, the current exceeds the limit before the protection activation.

The following diagrams of Fig12 and Fig13 show the maximum allowed soft-start capacitor as a function of the input voltage, inductor value and switching frequency. A minimun value of the

soft start capacitance is necessary to guarantee, in short circuit condition, the functionality of the limiting current circuitry.
Infact, with a capacitor too small, the frequency of the current peaks (see figure 11) is high and the mean current value in short circuit increases.
Example: for a maximum input voltage of 55V at 100kHz, with an inductor of 260mH, it is possible to use a soft start capacitor lower than 470mF . With such a value, the soft start time (see fig. 16) of about 10ms for an output voltage of 5V

Figure 12. Maximum soft start capacitance with f_{sw} = 100kHz

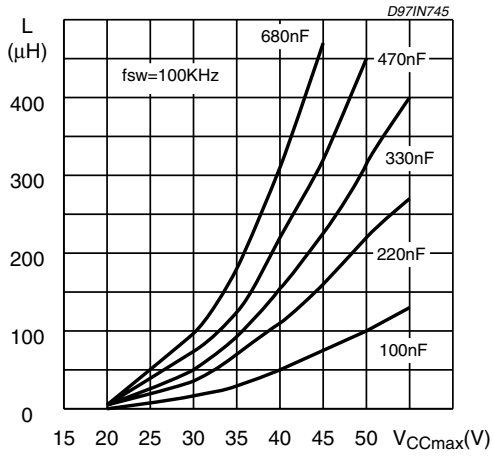
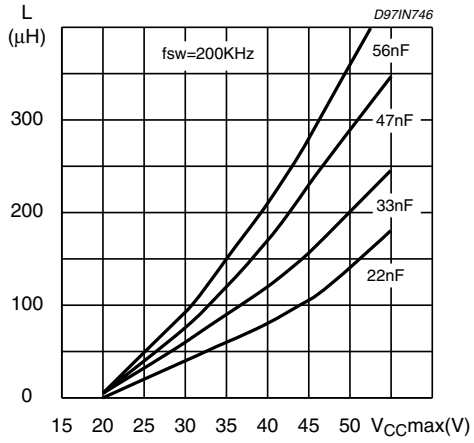


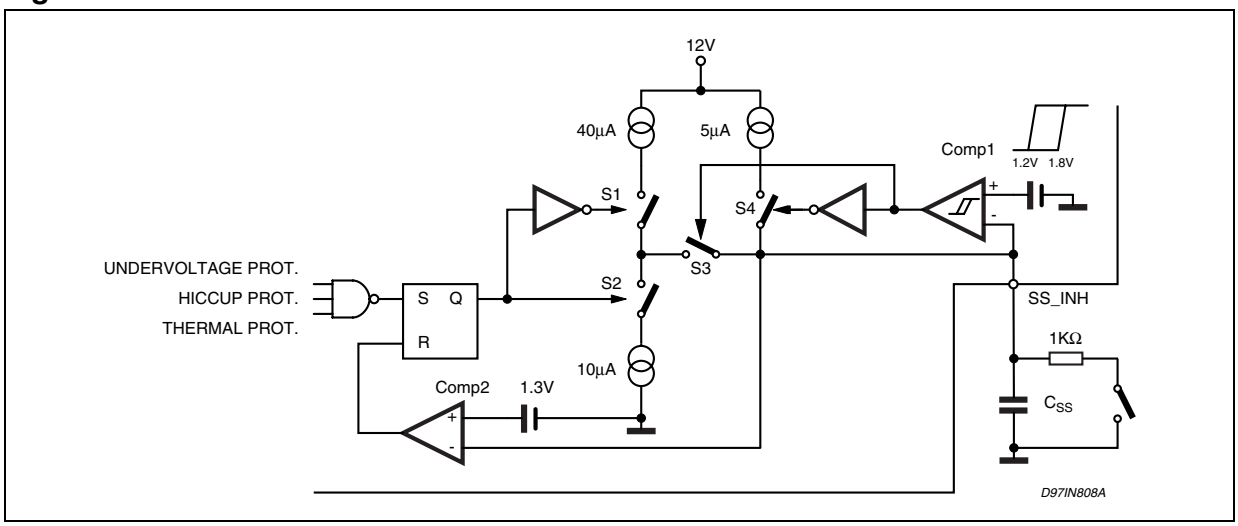
Figure 13. Maximum soft start capacitance with f_{sw} = 200kHz



3.3 Soft Start and Inhibit Functions

The soft start and the inhibit functions are realised using one pin only, pin2. Soft-start is requested to inizialise all internal functions with a correct start-up of the system without overstressing the power stage, avoiding the intervention of the current protection, and having an output voltage rising smoothly without output overshoots.
At Vcc Turn-on or having had an intervention of inhibit function, an initial 5µA internal current generator starts to charge the soft-start capacitor, from 0V to about 1.8V. From this hysteretic threshold, a 40µA current generator is activated, putting in off state the previous generator.

Figure 14. Soft-Start and inhibit functions Internal Circuit



At this point, the output PWM starts, initiating the rising phase of the output voltage.
The soft-start capacitor is quickly discharged in case of:

- Thermal protection intervention
- Hiccup limiting current condition
- Supply voltage lower than UVLO off threshold.

The soft-start and inhibit schematic diagram is shown in fig 14.
At device turn-on, the soft-start capacitor has no charge, with 0V at its terminals.
From 0V to 1.8V, switch S3 is opened and S4 is closed.
Soft-start capacitor is charged with 5μA.
At 1.8V, comp1 change the output status, opening S4 and closing S3, and the device starts to generate the PWM signal, rising smothly the output voltage.
Till this moment, S2 is opened, S1 closed.
By closing S3, the soft-start capacitor is charged with 40μA reaching its saturation voltage.
This procedure is repeated at each Vcc turn-on.

Figure 15. Timing Diagram in Inhibit, overcurrent and turn off condition

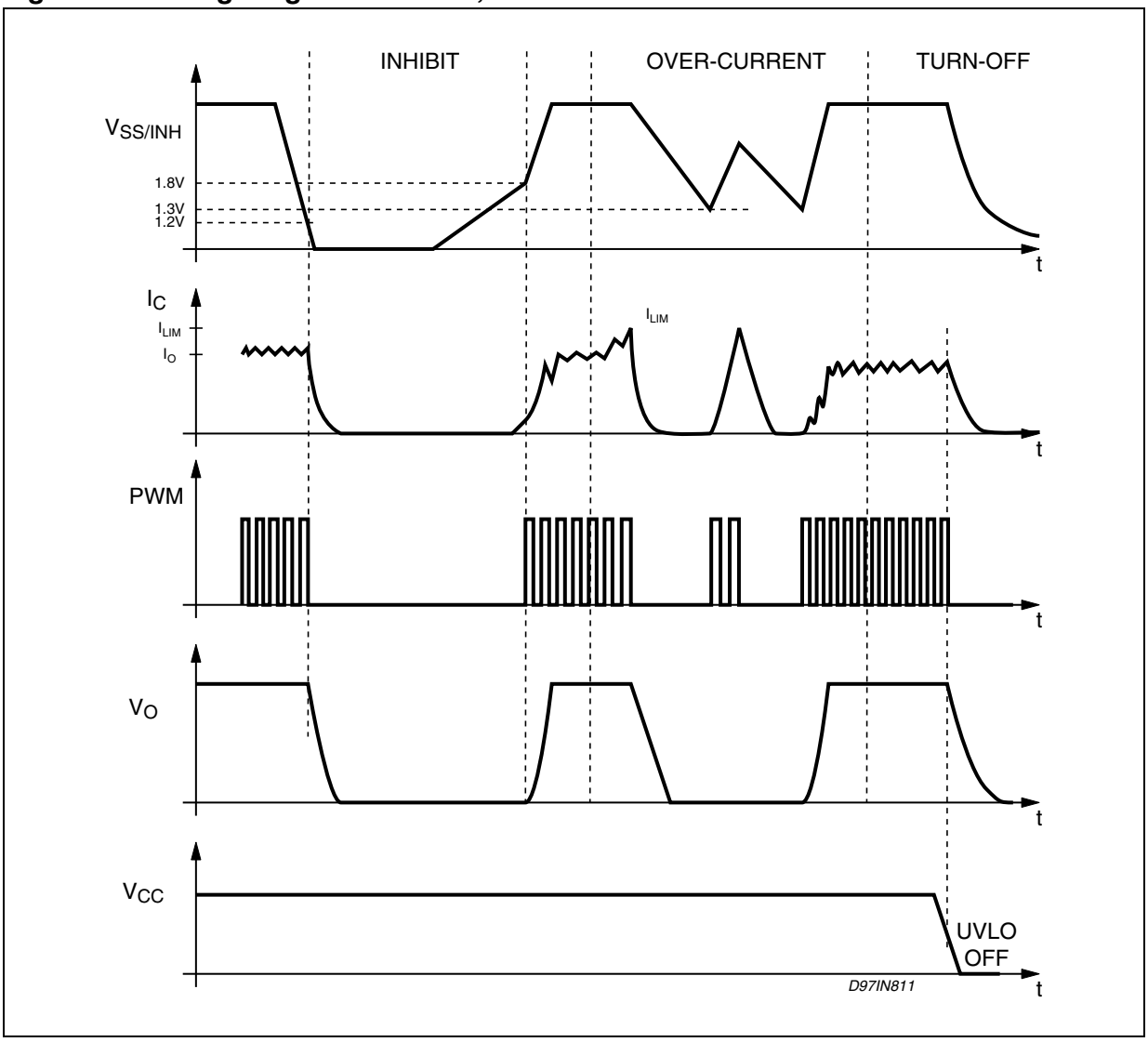
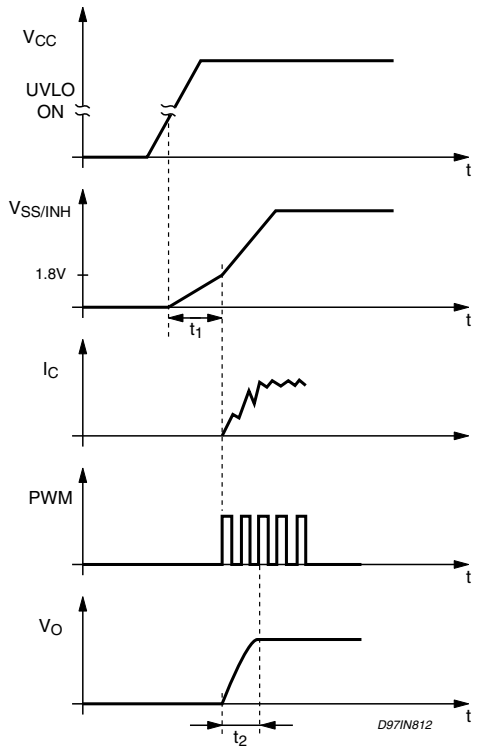


Figure 16. Figure 11b. Start up sequence.



Turning Vcc off, the soft-start capacitor is discharged with a constant 10μA (S2 closed, S3 closed, S1 and S4 open), from the moment when Vcc is crossing the UVLO off threshold.

The final discharge value is 1.2V. In case of the Css is discharged using an external grounded element when the voltage at Css reaches the threshold of 1.3V Comp2 resets the flip flop, S1 is closed, S2 is opened and the 40mA current generator is activated.

The external switch, sinking some mA, discharges the soft-start under the 1.2V Comp1 threshold, opening S3 and closing S4. At this point the device is in disable, sourcing only 5μA through pin 2. When the external grounding element is removed, the device restarts charging the soft start capacitance, initially, with 5μA till the voltage reaches the 1.8V threshold and Comp1 connects the 40μA charging current generator. In case of thermal shutdown or overcurrent protection intervention the power is turned off and the flip flop turns off S2 and turns on S1. The soft-start is discharged till the voltage reaches the 1.3V threshold, and Comp2 resets the flip flop. S1 is closed, S2 is opened and the soft-start capacitance is charged again.

Fig 15 shows the systems signals during Inhibit, overcurrent and Vcc turn off.

t1 and t2 can be calculated by the following equations:

$$t1 = 0.36 \cdot Css; \qquad t2 = \frac{V_O}{Ich \cdot 6 \cdot D_{max}} \cdot C_{ss}$$

where Dmax is 0.95, Css is in μF and Ich is in μA . Soft-start time (t2) versus output voltage and Css is shown in Fig17.

Thanks to the voltage feedforward, the start-up time (t2) is not affected by the input voltage.

Fig18 shows the output voltage start-up using different soft-start capacitance values:

It is mandatory a minimum capacitor value of 22nF. The pin 2 cannot be left open.

Figure 17. Soft start time (t2) vs Vo and Css

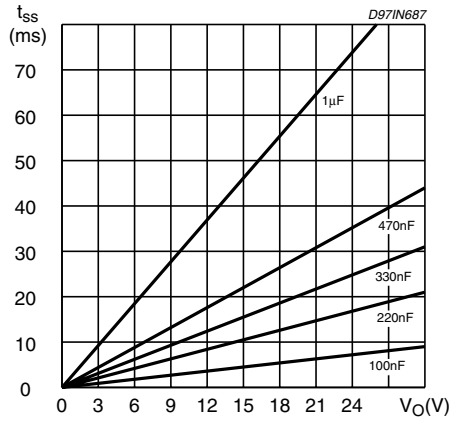
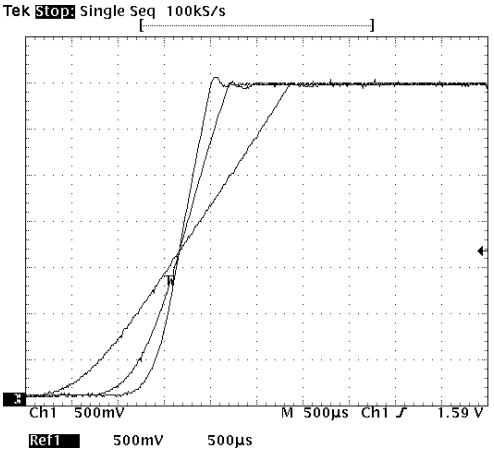


Figure 18. Output rising voltage with Css 56nF, 100nF, 220nF.-



3.4 Feedback disconnection

In case of feedback disconnection, the duty cycle increases versus the max allowed value bringing the output voltage close to the input supply. This condition could destroy the load.
To avoid this dangerous condition, the device is forcing a little current (1.4µA typical) out of the pin 8 (E/A Feedback).
If the feedback is disconnected, open loop, and the impedance at pin 8 is higher than 3.5MΩ, the voltage at this pin goes higher than the internal reference voltage located on the non-inverting error amplifier input, and turns-off the power device.

3.5 Zero load

In normal operation, the output regulation is also guaranteed because the bootstrap capacitor is recharged, cycle by cycle, by means of the energy flowing into the choke.
Under light load conditions, this topology tends to operate in burst mode, with random repetition rate of the bursts. An internal new function makes this device capable of keeping the output voltage in full regulation with 1mA of load current only.
Between 1mA and 500µA, the output is kept in regulation up to 8% above the nominal value.
Here the circuitry providing the control :

1) a comparator located on the bootstrap section is sensing the bootstrap voltage; when this is lower than 5V, the internal power VDMOS is forced ON for one cycle and OFF for the next..
2) during this operation mode, i.e. 500µA of load current, the E/A control is lost. To avoid output overvoltages, a comparator with one input connected to pin 8, and the second input connected to a threshold 8% higher that nominal output, turns OFF the internal power device the output is reaching that threshold.
   When the output current, or rather, the current flowing into the choke, is lower than 500µA, that is also the consumption of the bootstrap section, the output voltage starts to increase, approaching the supply voltage.

3.6 Output Overvoltage Protection (OVP)

The output overvoltage protection, OVP, is realized by using an internal comparator , which input is connected to pin 8, the feedback, that turns-off the power stage when the OVP threshold is reached. This threshold is typically 8% higher than the feedback voltage.

When a voltage divider is requested for adjusting the output voltage, the OVP intervention will be setted at:

$$V_{OVP} = 1.08 \cdot V_{fB} \cdot \frac{(Ra + Rb)}{Rb}$$

where Ra is the resistor connected to the output.

3.7 Power Stage

The power stage is realized by a N-channel D-mos transistor with a Vdss in excess of 60V and typ Rdson of 290mOhm (measured at the device pins).
Minimising the Rdson, means also minimise the conduction losses.
But also the switching losses have to be taken into consideration. mainly for the two following reasons:

a) they are affecting the system efficiency and the device power dissipation
b) because they generate EMI.

3.8 TURN - ON

At turn-on of the power element, or better, the rise time of the current(di/dt) at turn-on is the most critical parameter to compromise.
At a first approach, it looks that faster is the rise time and lower are the turn-on losses.
It's not completely true.
There is a limit, and it's introduce by the recovery time of the recirculation diode.
Above this limit, about 100A/usec, only disadvantages are obtained:
1- turn-on overcurrent is decreasing efficiency and system reliability
2- big EMI encrease.
The L4971 has been developed with a special focus on this dynamic area.
An innovative and proprietary gate driver, with two different timings, has been introduced.
When the diode reverse voltage is reaching about 3V, the gate is sourced with low current (see fig 19) to assure the complete recovery of the diode without generating unwanted extra peak currents and noise. After this threshold, the gate drive current is quickly increased, producing a fast rise time till the peak current, so maintaining the efficiency very high.

Figure 19. Turn on and Turn off (pin 2, 3)

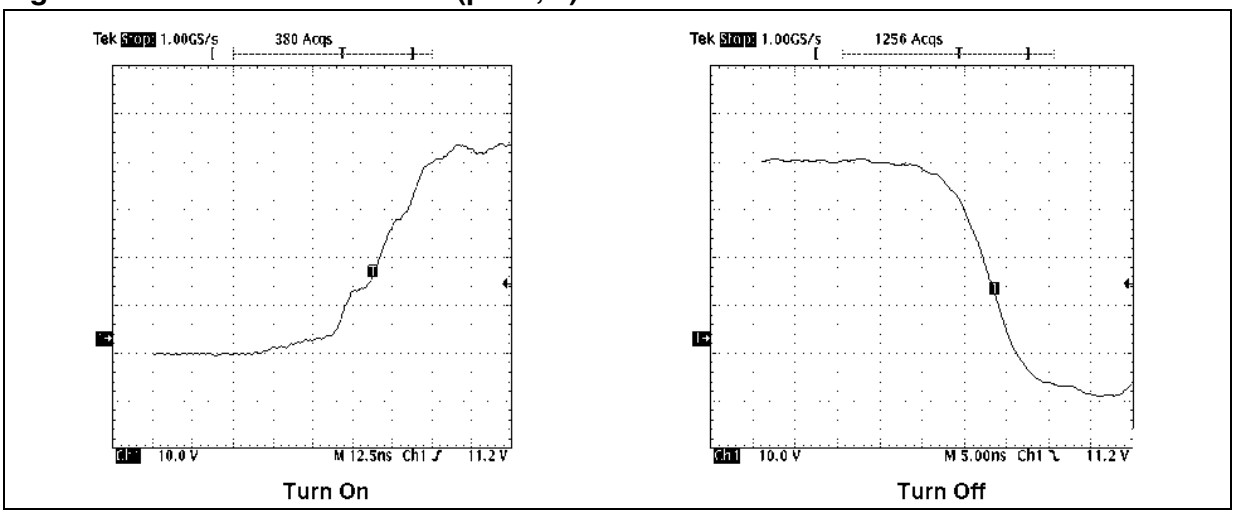
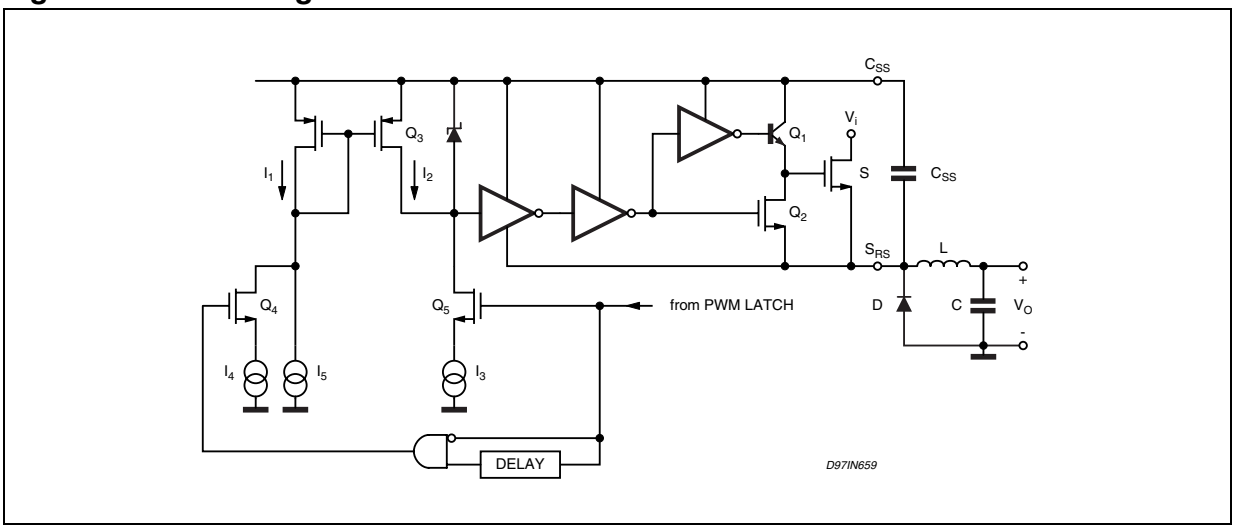


Figure 20. Power stage internal circuit



3.9 TURN - OFF

The turn-off behavior, is shown at Fig. 19.

Fig. 20 shows the details of the internal power stage and driver, where at Q2 is demanded the turn-off of the power swicth, S.

4 TYPICAL APPLICATION.

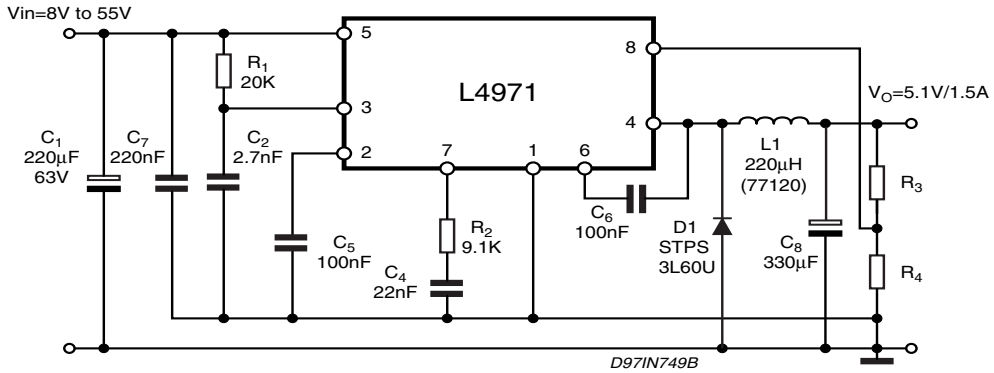
Fig. 21 shows the typical application circuit, where the input supply voltage, Vcc, can range from 8 to 55V operating, and the output voltage adjustable from 3.3V to 40V.

The selected components, and in particular input and output capacitors, are able to sustain the device voltage ratings, and the corresponding RMS currents.

4.1 Electrical Specification

Input Voltage Range	8V - 55V
Output Voltage	5.1V ±3% (Line, Load and Thermal)
Output Ripple	20mV
Output Current range	1mA - 1.5A
Max Output Ripple current	15% Iomax
Current limit	2.5A
Switching frequency	100kHz
Target Efficiency	85% @ 1.5A Vi = 505V
	91% @ 0.5A Vi = 12V

Figure 21. Application Circuit



C1=220µF/63V EKE
C2=2.7nF
C5=100nF
C6=100nF
C7=220nF/63V
C8=330µF/35V CG Sanyo
L1=220µH KoolMu 77120 - 65 Turns - 0.5mm
R1=20K
R2=9.1K
D1=STPS3L60U

L4971

V_O(V)	R3(KΩ)	R4(KΩ)
3.3	0	
5.1	2.7	4.7
12	12	4.7
15	16	4.7
18	20	4.7
24	30	4.7

4.2 Input Capacitor

The input capacitor has to be able to support the max input operating voltage of the device and the max rms input current. The input current is squared and the quality of these capacitors has to be very high to minimise its power dissipation generated by the internal ESR, improving the system reliability. Moreover, input capacitors are also affecting the system efficiency.
The max Irms current flowing through the input capacitors is:

$$I_{rms} = I_O \cdot \sqrt{D - \frac{2 \cdot D^2}{\eta} + \frac{D^2}{\eta^2}}$$

where η is the expected system efficiency, D is the duty cycle and Io the output dc current.
This function reaches the maximum value at D = 0.5 and the equivalent rms current is equal to Io/2. The following diagram Fig. 23 is the graphical rappresentation of the above equation, with an estimated efficiency of 85% , at different output currents.
The maximum and minimum duty cycles are:

$$D_{max} = \frac{V_O + V_f}{V_{ccmin} + V_f} = 0.66 \qquad D_{max} = \frac{V_O + V_f}{V_{ccmax} + V_f} = 0.1$$

where Vf is the freewheeling diode forward voltage. This formula is not taking into account the power mos Rdson, considering negligible the inherent voltage drop, respect input and output voltages. At full load,1.5A, and D=0.5, the rms capacitor current to be sustained is of 0.75A.
The selected 220µF/63V Roderstain is able to support this current.

Figure 22. Efficiency vs Output Current

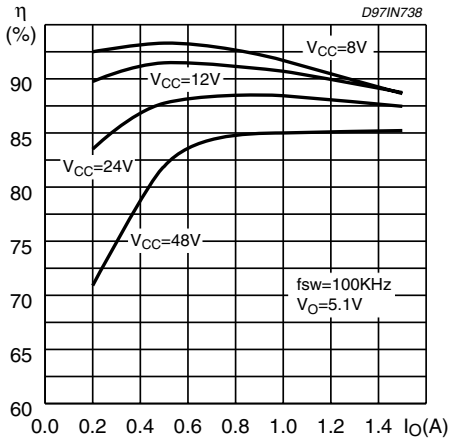
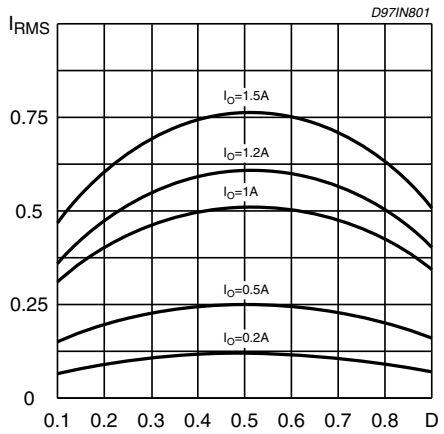


Figure 23. Input Capacitance rms current vs. duty cycle



4.3 Inductor Selection

The inductor ripple current is fixed at 10% of Iomax and is 0.15A, the inductor needed is:

$$L = (V_O + V_f) \cdot \frac{(1 - D_{min})}{\Delta I_O \cdot f_{sw}} = 310\mu H \qquad \text{Eq 1}$$

The $L \cdot I_O^2$ is 0.58 and the size core chose is 77120 (125µ) Magnetics KoolMµ material and are wiring 65Turns. At full load the magnetising force is about 25 Oersted, the inductance value is reduced at about L = 220µH and the ripple current increases at 0.24A (16% Iomax).

It is possible to graficate the Eq 1 as a function of Vo and Vinmax at 100kHz and 200kHz (see Figs. 24 and 25).

These curves are useful to define the inductor value immediately.

Example: With a maximum input voltage of 15V at 100kHz, fixed the curve (1) in Fig24 and with an output voltage of 5V the inductor needed is 220µH.

Figure 24. Inductor needed as a function of maximum input voltage and output voltage at fsw=100kHz

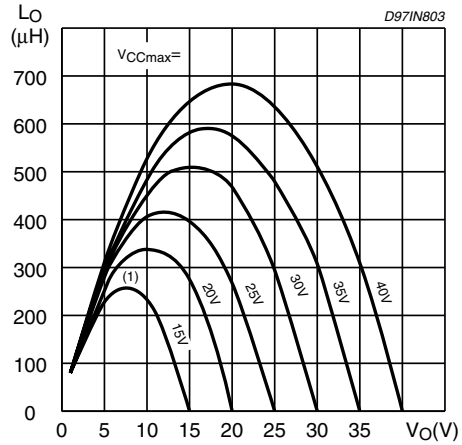
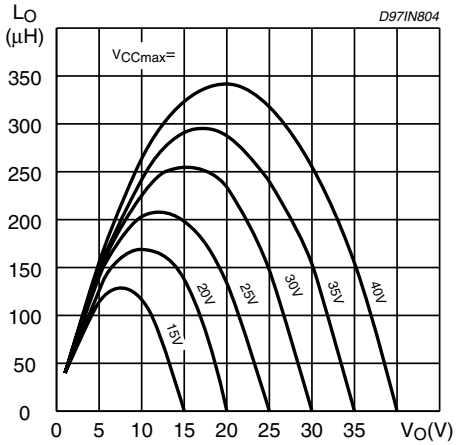


Figure 25. Inductor needed as a function of maximum input voltage and output voltage at fsw=200kHzt



4.4 Core Losses

Core losses are proportional to the magnetic flux swing into the core material. To evaluate the flux swing is used the following formula:

$$\Delta B = \frac{L \cdot \Delta I_O \cdot 10^{-4}}{N_O \cdot A_{le}} = 423 Gauss$$

where Ale is the core cross section [m^2].
The choosen core material family has an empirical equation to calculate the losses:

$$P_L = \Delta B^2 \cdot f_{sw}{}^{1.5} \cdot V_L \cdot 10^2 = 141 mW$$

Where ΔB is in KGauss, fsw in kHz and V_L is the core volume in cm^3. The core increasing temperature is:

$$\Delta T = \left(\frac{P_l}{13.6}\right)^{0.833} = 7°C$$

4.5 Output Capacitors

The selection of Cout is driven by the output ripple voltage required, 1% of Vo. This is defined by the output capacitance ESR and with the maximum ripple current (0.24A) the maximum ESR is:

$$ESR = \frac{\Delta V_O}{\Delta I_L} = \frac{0.051}{0.24} = 212 m\Omega$$

The selected capacitance is 330µF/35V CG Sanyo with ESR = 86mΩ and the ripple voltage is 0.40% of V_o (20mV).
The drop due to a fast load variation of 1A produce an output drop of:

$$ESR \cdot \Delta I_o = 86 mV$$

that is the 1.6% of the output voltage.
Output capacitance has to support a load transient until the inductor current reaches the increased current. The output drop during an output current variation is:

$$\Delta V_O = \frac{(\Delta I_O)^2 \cdot L_O}{2 \cdot C_O \cdot (V_{inmin} \cdot D_{max} - V_O)} \qquad \text{Eq 2}$$

where ΔI_o is the load current load variation 0.5A to 1.5A, D_{max} is the maximum duty cycle, 95%, V_o is 5.1V, L_o is 220µH.
Equation 2, normalized at Vo, is represented in the following diagram, Fig. 26, as a function of the minimum input voltage.
These curves are rapresented for different output capacitor 100µF, 220µF, 2x220µF, 3x330µF.

4.6 Compensation Network

The complete controll loop block diagram is shown in fig. 27. a transfer functions described are:

Figure 26. Output drop vs minimum input voltage

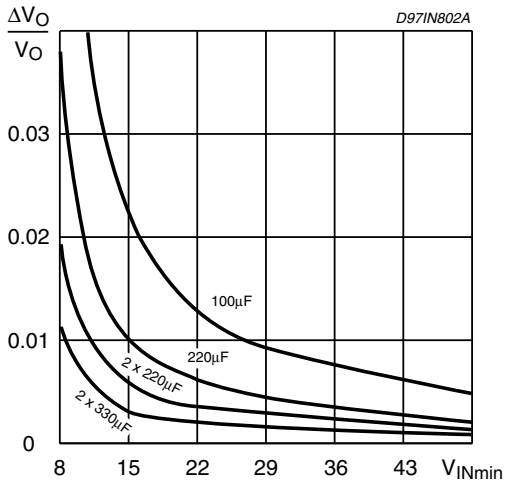


Figure 27. Block diagram compensation loop

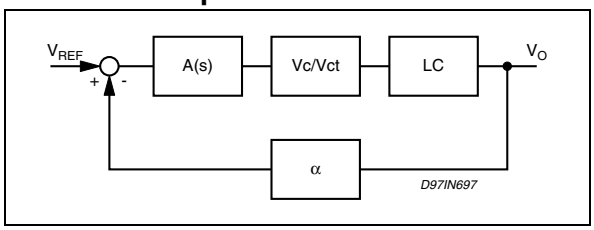


Figure 28. Error Amplifier Compensation Circuit

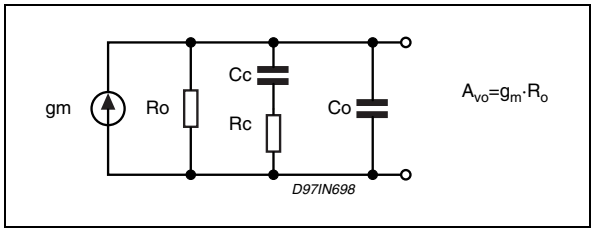
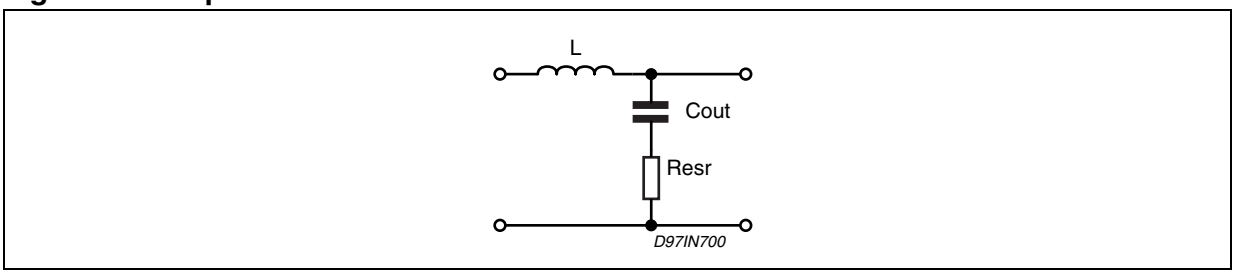


Figure 29. Output Filter



4.7 Error amplifier and compensation block

$$A(s) = \frac{A_{VO} \cdot (1 + s \cdot R_c \cdot C_c)}{s^2 \cdot R_o \cdot C_o \cdot R_c \cdot C_c + s \cdot (R_o \cdot C_c + R_o \cdot C_o + R_c \cdot C_c) + 1}$$

C_o is the parallel between the output and the external capacitance of the Error Amplifier output impedence. Rc and Cc are the compensation values.

4.8 LC Filter

$$A_{o(s)} = \frac{1 + R_{esr} \cdot C_{out} \cdot s}{L \cdot C_{out} \cdot s^2 + R_{esr} \cdot s + 1}$$

4.9 PWM Gain

$$\frac{V_{CC}}{V_{ct}} = \frac{V_{CC} \cdot 6}{V_{CC} - 1} \approx 6$$

where Vct is the peak to peak sawtooth oscillator.

4.10Voltage divider

$$\alpha = \frac{R4}{R3 + R4}$$

The Error Amplifier basic characteristics are:
Ro = 1.2MΩ
Avo = 60dB
Co = 220pF
The poles and zeros value are:

$$F_o = \frac{1}{2 \cdot \pi R_{esr} \cdot C_{out}} = \frac{1}{2 \cdot \pi \cdot 0.086 \cdot 330 \cdot 10^{-6}} = 5.6KHz$$

$$F_p = \frac{1}{2 \cdot \pi \cdot \sqrt{L \cdot C_{out}}} = \frac{1}{2 \cdot \pi \cdot \sqrt{220 \cdot 10^{-6} \cdot 330 \cdot 10^{-6}}} = 590Hz$$

$$F_{ocomp} = \frac{1}{2 \cdot \pi \cdot R_c \cdot C_c} = \frac{1}{2 \cdot \pi \cdot 9.1 \cdot 10^3 \cdot 22 \cdot 10^{-9}} = 795Hz$$

$$F_{p1} = \frac{1}{2 \cdot \pi \cdot R_o \cdot C_c} = \frac{1}{2 \cdot \pi \cdot 1.2 \cdot 10^6 \cdot 22 \cdot 10^{-9}} = 6.02Hz$$

$$F_{p2} = \frac{1}{2 \cdot \pi \cdot R_c \cdot C_o} = \frac{1}{2 \cdot \pi \cdot 9.1 \cdot 10^3 \cdot 220 \cdot 10^{-12}} = 80KHz$$

The compensation is realized choosing the Focomp nearly the frequency of the double pole due to the LC filter.
Using a compensation network with R1= 9.1K, C6 = 22nF and C5 = 220pF obtain the Gain and Phase Bode plot of Figg. 30-31. Is possible to omit C5 because does not influence the system stability but is useful only to reduce the noise. The cut off frequency and a phase margin are:

Fc = 5KHz Phase margin = 21°

Figure 30. Gain Bode open loop plot

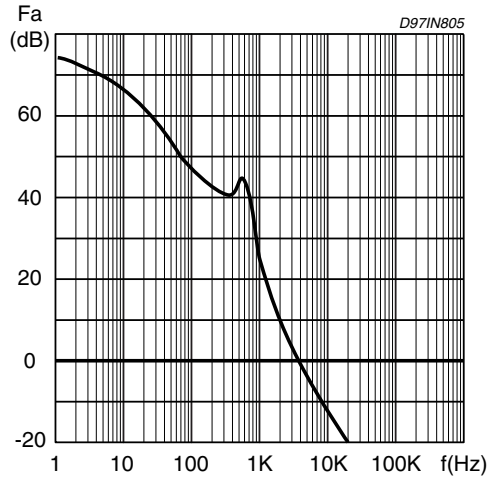
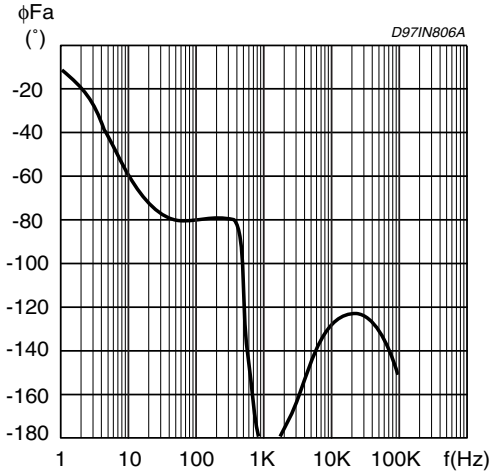


Figure 31. Phase Bode open loop plot



5 REVISION HISTORY

Table 1. Revision History

Date	Revision	Description of Changes
October 2004	13	First Issue in EDOCS
May 2005	14	Updated the Layout look & feel. Changed name of the D1 on the fig. 21

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