AN954

A Unique Converter Configuration
Provides Step-Up/Down Functions

INTRODUCTION

The use of switching regulators in new portable equipment designs is becoming more pronounced over that of linear regulators. This is primarily due to the need for reductions in size and weight which dictate an ever increasing demand for higher power conversion efficiency from a battery pack. When designing at the board level it sometimes becomes necessary to generate a constant output voltage that is less than that of the battery. The step-down circuit shown in Figure 1A will perform this function efficiently. However, as the battery discharges, its terminal voltage will eventually fall below the desired output, and in order to utilize the remaining battery energy the step-up circuit shown in Figure 1B will be required.

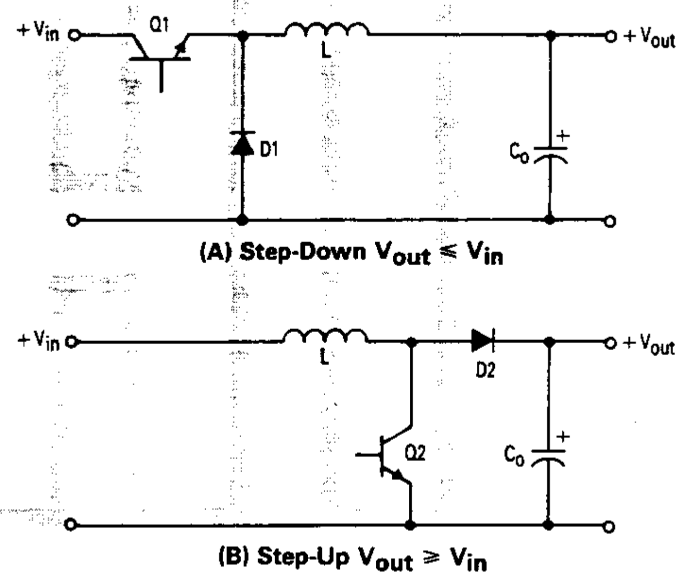


Figure 1. Basic Switching Regulator Configurations

GENERAL APPLICATIONS

By combining circuits A and B a unique step-up/down configuration can be created (Figure 2) which still employs a *simple inductor* for the voltage transformation. Energy is stored in the inductor during the time that transistors Q1 and Q2 are in the 'on' state. Upon turn-off, the energy is transferred to the output filter capacitor and load forward biasing diodes D1 and D2. Note that during t_{on} this circuit is identical to the basic step-up, but during t_{off} the output voltage is derived only from the inductor and is with respect to ground instead of V_{in}. This allows the output voltage to be set to any value, thus it may be less than, equal to, or greater than that of the input. Current limit protection cannot be employed in the basic step-up circuit. If the output is severely overloaded or shorted L or D2 may be destroyed since they form a direct path from V_{in} to V_{out}. The step-up/down configuration allows the control circuit to implement current limiting because Q1 is now in series with V_{out}, as is in the step-down circuit.

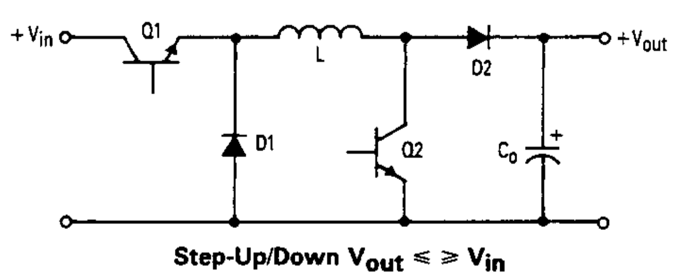
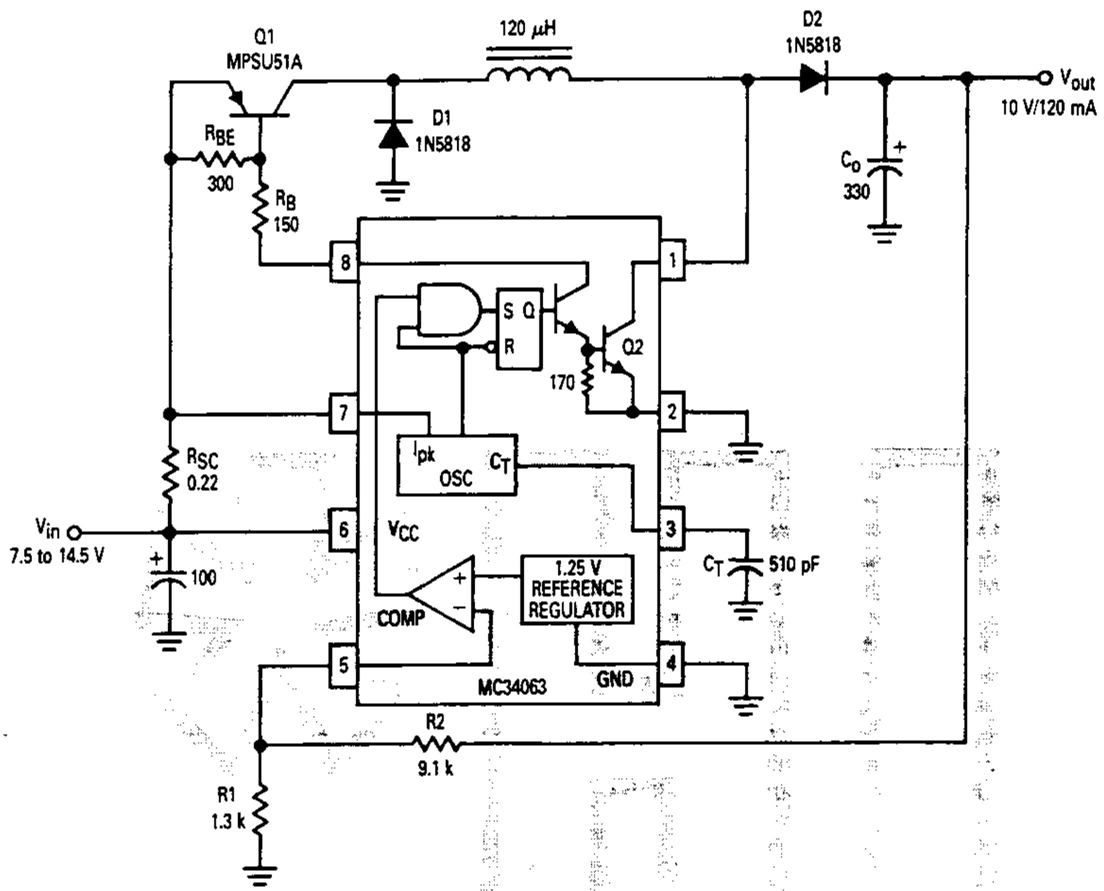


Figure 2. Combined Configuration



Test	Conditions	Results
Line Regulation	V_{in} = 7.5 to 14.5 V, I_{out} = 120 mA	Δ = 22 mV or ± 0.11%
Load Regulation	V_{in} = 12.6 V, I_{out} = 10 to 120 mA	Δ = 3 mV or ± 0.015%
Output Ripple	V_{in} = 12.6 V, I_{out} = 120 mA	95 mV_{p-p}
Short Circuit Current	V_{in} = 12.6 V, R_L = 0.1 Ω	1.54 A
Efficiency	V_{in} = 7.5 to 14.5 V, I_{out} = 120 mA	74%

Figure 3. Step-Up/Down Switching Regulator Design Example

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A complete step-up/down switching regulator design example is shown in Figure 3. In order to implement a minimum component system, the Motorola MC34063 switching regulator subsystem was selected. This device features an on-chip 1.5 A switch transistor that was used to perform the function of Q2. This regulator was designed to operate from a standard 12 V battery pack with the following conditions:

V_{in} = 7.5 to 14.5 V V_{out} = 10 V
f_{min} = 50 kHz I_{out} = 120 mA
$V_{ripple\,(p\text{-}p)}$ = 1% V_{out} or 100 $mV_{p\text{-}p}$

The following design procedure is provided so that the user can select proper component values for his specific converter application.

1) Determine the ratio of switch conduction t_{on} versus diode conduction t_{off} time.

$$\frac{t_{on}}{t_{off}} = \frac{V_{out} + V_{FD1} + V_{FD2}}{V_{in(min)} - V_{satQ1} - V_{satQ2}} = \frac{10 + 0.6 + 0.6}{7.5 - 0.8 - 0.8} = 1.9$$

2) The cycle time of the L-C network is equal to $t_{on\,(max)} + t_{off}$.

$$t_{on\,(max)} + t_{off} = \frac{1}{f_{min}} = \frac{1}{50 \times 10^3} = 20\ \mu s \text{ per cycle}$$

3) Next calculate t_{on} and t_{off} from the ratio of t_{on}/t_{off} in #1 and the sum of $t_{on\,(max)} + t_{off}$ in #2.

$$t_{off} = \frac{t_{on\,(max)} + t_{off}}{\frac{t_{on}}{t_{off}} + 1} = \frac{20 \times 10^{-6}}{1.9 + 1} = 6.9\ \mu s$$

$$t_{on} = 20\ \mu s - 6.9\ \mu s = 13.1\ \mu s$$

4) The maximum on-time is set by selecting a value for C_T.

$$C_T = 4 \times 10^{-5}\, t_{on\,(max)} = 4 \times 10^{-5}\,(13.1 \times 10^{-6}) = 524\ pF$$

Use a standard 510 pF capacitor.

5) The peak switch current is:

$$I_{pk(switch)} = 2I_{out}\left(\frac{t_{on}}{t_{off}} + 1\right) = 2\,(120 \times 10^{-3})\,(1.9 + 1) = 696\ mA$$

6) A minimum value of inductance can now be calculated since the maximum on-time and peak switch current are known.

$$L_{min} = \left(\frac{V_{in(min)} - V_{satQ1} - V_{satQ2}}{I_{pk(switch)}}\right) t_{on} = \left(\frac{7.5 - 0.8 - 0.8}{696 \times 10^{-3}}\right) 13.1 \times 10^{-6} = 111\ \mu H$$

A 120 μH inductor was selected for $L_{(min)}$.

7) A value for the current limit resistor, R_{SC}, can be determined by using the current limit level of $I_{pk(switch)}$ when V_{in} = 14.5 V.

$$I'_{pk(switch)} = \left(\frac{V_{in} - V_{satQ1} - V_{satQ2}}{L_{(min)}}\right) t_{on(max)} = \left(\frac{14.5 - 0.8 - 0.8}{120 \times 10^{-6}}\right) 13.1 \times 10^{-6} = 1.41\ A$$

$$R_{SC} = \frac{0.33}{I'_{pk(switch)}} = \frac{0.33}{1.41} = 0.23\ \Omega$$

Use a standard 0.22 Ω resistor.

8) A minimum value for an *ideal* output filter capacitor is:

$$C_O \approx \left(\frac{I_{out}}{V_{ripple(p\text{-}p)}}\right) t_{on} \approx \left(\frac{120 \times 10^{-3}}{100 \times 10^{-3}}\right) 13.1 \times 10^{-6} \approx 15.7\ \mu F$$

Ideally this would satisfy the design goal, however, even a solid tantalum capacitor of this value will have a typical ESR (equivalent series resistance) of 0.3 Ω which will contribute an additional 209 mV of ripple. Also there is a ripple component due to the gain of the comparator equal to:

$$V_{ripple(p-p)} = \left(\frac{V_{out}}{V_{Ref}}\right) 1.5 \times 10^{-3}$$

$$= \left(\frac{10}{1.25}\right) 1.5 \times 10^{-3}$$

$$= 12\ mV$$

The ripple components are not in phase, but can be assumed to be for a conservative design. From the above it becomes apparent that ESR is the dominant factor in the selection of an output filter capacitor. A 330 μF with an ESR of 0.12 Ω was selected to satisfy this design example by the following:

$$ESR \approx \frac{V_{ripple(p-p)} - \left(\frac{I_{out}}{C_O}\right) t_{on} - \left(\frac{V_{out}}{V_{Ref}}\right) 1.5 \times 10^{-3}}{I_{pk(switch)}}$$

9) The nominal output voltage is programmed by the R1, R2 resistor divider.

$$R2 = R1 \left(\frac{V_{out}}{V_{Ref}} - 1\right)$$

$$= R1 \left(\frac{10}{1.25} - 1\right)$$

$$= 7\ R1$$

If 1.3 k is chosen for R1, then R2 would be 9.1 k, both being standard resistor values.

10) Transistor Q1 is driven into saturation with a forced gain of approximately 20 at an input voltage of 7.5 V. The required base drive is:

$$I_B = \frac{I_{pk(switch)}}{B_F}$$

$$= \frac{696 \times 10^{-3}}{20}$$

$$= 35\ mA$$

The value for the base-emitter turn-off resistor R_{BE} is determined by:

$$R_{BE} = \frac{10\ B_F}{I_{pk(switch)}}$$

$$= \frac{10\ (20)}{696 \times 10^{-3}}$$

$$= 287\ \Omega$$

A standard 300 Ω resistor was selected.

The additional base current required due to R_{BE} is:

$$I_{RBE} = \frac{V_{BEQ1}}{R_{BE}}$$

$$= \frac{0.8}{300}$$

$$= 3\ mA$$

The base drive resistor for Q1 is equal to:

$$R_B = \frac{V_{in(min)} - V_{sat(driver)} - V_{RSC} - V_{BEQ1}}{I_B + I_{RBE}}$$

$$= \frac{7.5 - 0.8 - 0.15 - 0.8}{(35 + 3) \times 10^{-3}}$$

$$= 151\ \Omega$$

A standard 150 Ω resistor was used.

The circuit performance data shows excellent line and load regulation. There is some loss in conversion efficiency over the basic step-up or step-down circuits due to the added switch transistor and diode 'on' losses. However this unique converter demonstrates that with a simple inductor, a step-up/down converter with current limiting can be constructed. For a more in depth analysis of the MC34063 operation and the derivation of the above design equations, request Motorola Application Note AN920A.

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