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1 Introduction

ZLED7001 is a controller IC designed for building DC-to-DC converters with only a few external components. Operating as a buck converter, it is especially suitable for highly efficient driving of LED loads. When using an external switching transistor with the ZLED7001, the supply voltage and LED power can be selected in a very wide range. However, some basic considerations must be taken into account when driving LEDs from a line voltage as high as 110 VAC or 230 VAC.

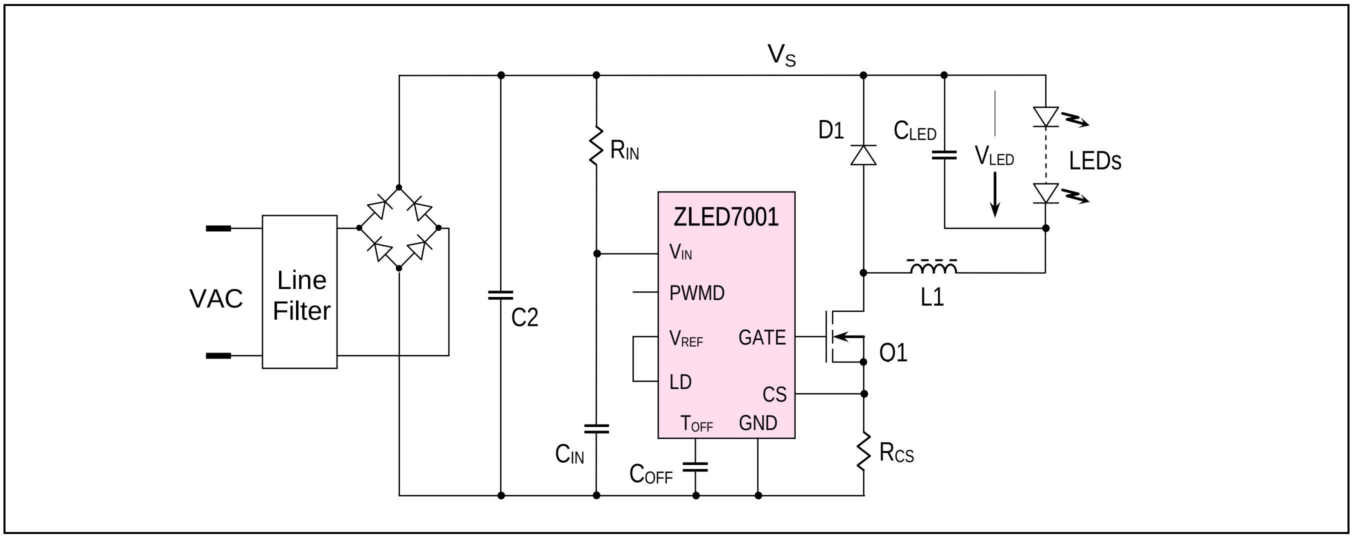
1.1. LED Requirements

LEDs should not be driven from a constant voltage, in which case the LED current and power would be primarily undefined, resulting in poor control of brightness. LEDs require either a constant current or constant power so that the voltage across the LED can settle to the level required by the LED, which varies over temperature and with each device. The ZLED7001 can derive the required constant current drive for the LEDs from a variable voltage (e.g., rectified AC line voltage).

1.2. Buck Converter Principle of Operation

Figure 1.1 shows the basic application circuitry using the ZLED7001 in a buck converter circuit with current control. This is the simplest way to control LEDs from a high supply voltage with high efficiency. When the switching transistor Q1 is turned on, current flows through the LED chain and the inductor L1, increasing linearly over time. Note that due to resistive losses, the increase in current is not exactly linear, but the effect is not significant and therefore not included in the following calculations. When Q1 is switched off, the inductance drives current in the same direction, flowing across free wheel diode D1 and circulating in the free wheel loop while current decays linearly over time.

Figure 1.1 Basic Buck Converter LED Driver Circuit using the ZLED7001



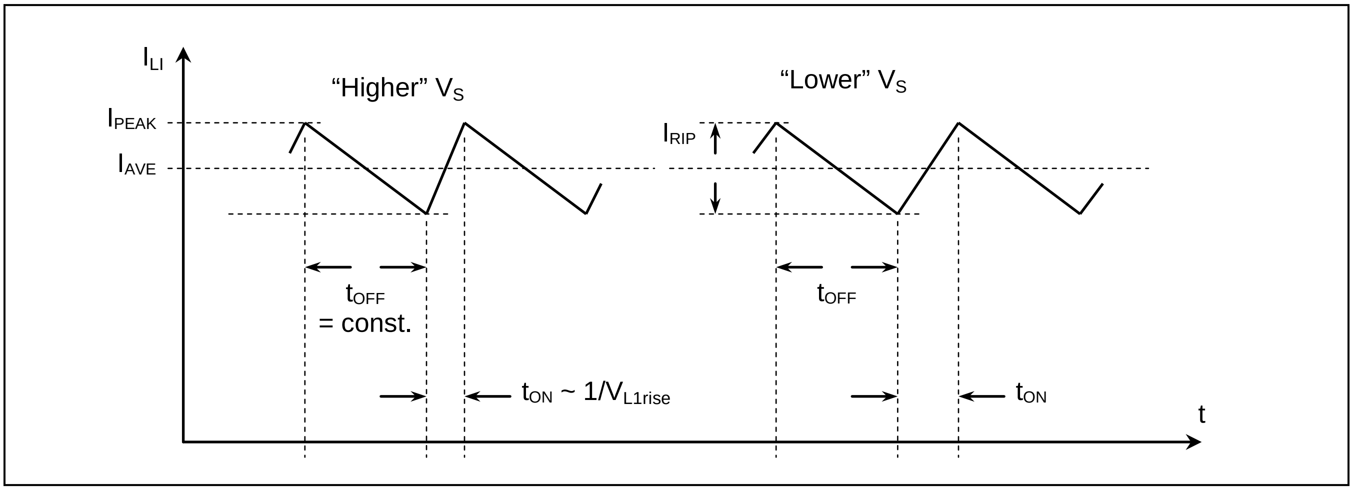
The rate of current change dI/dt is proportional to the voltage V_L across the inductor divided by the inductance L as shown in equation (1):

$$\frac{dI}{dt} = \frac{V_{L1}}{L1} \quad (1)$$

If dI/dt is considered as constant, the current change ΔI during a given time t equals

$$\Delta I = \frac{dI}{dt} * t \qquad (2)$$

Figure 1.2 Converter Current Wave Shapes



In a situation of constant average current, as shown in Figure 1.2, both rising and falling current changes ΔI_{RISE} and ΔI_{FALL}, defined as the current ripple I_{RIP}, must have the same amplitude. During t_{ON} †and t_{OFF}, the respective voltages across L1 are

$$\text{For dI/dt rising: } V_{L1rise} = V_S - V_{LED} \qquad (3)$$

$$\text{For dI/dt falling: } V_{L1fall} = V_{LED} + V_{D1} \qquad (4)$$

where V_S is the momentary supply voltage, V_{LED} is the total LED forward voltage and V_{D1} is the forward voltage of the free wheel diode D1.

Considering $\Delta I_{RISE} = \Delta I_{FALL}$ as described above, I_{RIP} can be calculated as

$$\Delta I_{RISE} = \Delta I_{FALL} = I_{RIP} \Rightarrow (V_S - V_{LED}) * t_{ON} = (V_{LED} + V_{D1}) * t_{OFF} \qquad (5)$$

† See section 1.4 regarding minimum t_{ON}.

Equation (5) can be modified to

$$V_S * t_{ON} = V_{LED} * (t_{ON} + t_{OFF}) + V_{D1} * t_{OFF}$$

$$V_S * t_{ON} + V_{D1} * t_{ON} = V_{LED} * (t_{ON} + t_{OFF}) + V_{D1} * (t_{ON} + t_{OFF})$$

$$\frac{t_{ON}}{t_{ON} + t_{OFF}} = \frac{t_{ON}}{t_P} = d = \frac{V_{LED} + V_{D1}}{V_S + V_{D1}} \qquad (6)$$

with the switching period $t_P = 1/f_{SW}$, switching frequency f_{SW} and duty cycle d.

If the forward voltage of D1 can be neglected, the equation simplifies to

$$\frac{V_{LED}}{V_S} \approx d \qquad (7)$$

This basic relationship for the buck converter shows that the ratio of output voltage over supply voltage is equal to the switching duty cycle. This has a severe impact on circuit dimensioning.

1.3. Buck Converter Output Current Control

Equations (6) and (7) demonstrate that the duty cycle d determines the output voltage as a fraction of the supply voltage V_S. As previously discussed, LEDs should be driven with constant current or power. Consequently the circuit design must allow the duty cycle to automatically "adapt"; e.g., to the desired LED current. The ZLED7001 offers this feature based on peak current switching and constant off-time.

Although current is only monitored during the switching transistor's on-time, the constant off-time approach enables good control of average load current. If the forward voltages of the LEDs and the free wheel diode are considered constant, with the free wheel voltage $V_{FW} = V_{LED} + V_{D1}$, the current ripple can be calculated as

$$I_{RIP} = t_{OFF} * \frac{V_{FW}}{L1} \qquad (8)$$

The average LED current is therefore

$$I_{LED} = I_{PEAK} - \frac{I_{RIP}}{2} \qquad (9)$$

I_{PEAK} is the current level at which the transistor must be switched off in order to achieve the desired average current. This current is sensed as a voltage drop across R_{CS} (see Figure 1.1) and compared to an internal threshold of 240 mV (typical) in the ZLED7001. The on-time t_{ON} of Q1 adapts automatically according to the momentary DC voltage and the voltage across the LEDs.

These calculations for maintaining a constant current independent of the supply are based on the assumption that the forward voltage remains constant. In reality this is not the case; for instance, the forward voltage decreases with temperature. The impact of this effect can be estimated: For example, for a typical current ripple of 30% of the average current, $I_{PEAK} = 1.15 * I_{AVE}$. For this example, assume that the forward voltage decreases by 10%, which results in a 10% reduction of I_{RIP} during the constant t_{OFF}; i.e., I_{RIP} is now 27% of I_{AVE}. Since I_{PEAK} does not change, I_{AVE} increases from $I_{PEAK}/1.15$ to $I_{PEAK}/1.135$; in other words by 1.3%–an insignificant effect of the change in forward voltage on the average current.

1.4. Limitations Caused by Switching Transients

Every time the switching transistor Q1 is turned on, it must discharge the parasitic capacitance on its drain node to ground. This is basically its own drain capacitance, the free wheel diode's reverse capacitance and the parasitic capacitance of the inductor. The transient current is only limited by the speed of the gate voltage being turned on and the drain current capability of the saturated transistor. It is normally much higher than the operating current through the inductor. This current causes a significant voltage drop across R_{CS}, higher than the switch-off threshold. Therefore, after turning the transistor on, it is necessary to introduce a "blanking" time t_{BLANK}, during which the internal comparator is disconnected from the current sense voltage.

The ZLED7001 features an internal blanking time of 510 ns (typical), which is sufficient for the current to settle to the value determined by the inductor. However, this also means that 510 ns is the minimum t_{ON} of Q1, since it cannot be switched off before t_{BLANK}, has elapsed. If this limitation for t_{ON} is applied to equation (6), it is evident that as a consequence, the maximum possible switching frequency is limited by the supply voltage and the forward voltage of all diodes. The higher V_S is and the lower V_{FW} is, the lower f_{SWmax} will be. Using equation (7) (neglecting V_{D1}) for a simple estimation and assuming $t_{ON} = t_{BLANK}$, the duty cycle can be calculated as

$$d = \frac{t_{ON}}{t_{ON} + t_{OFF}} = t_{ON} * f_{SW} \approx \frac{V_{LED}}{V_S} \tag{10}$$

$$f_{SW\,max} = \frac{V_{LED\,min}}{V_{S\,max} * t_{BLANK}} \tag{11}$$

1.5. Dimensioning L1

A high switching frequency allows significant advantages in component size and cost. One of the bulkier components in ZLED7001 applications is the inductor L1. Its physical dimensions are mainly determined by the energy it must store and transfer during each individual switching cycle. (Since the magnetizing losses of the ferrite material increase with f_{SW}, the switching frequency is also a factor; however, for simplicity, this is not considered in the following discussion.) During t_{ON}, the inductor's energy increases and during t_{OFF}, it drives the energy into the load. The power that is being provided to the load by the inductor is therefore

$$P_{L1} = P_{OUT} * (1 - d) \tag{12}$$

where Pout is the total power of the LEDs and freewheel diode. This determines the energy the inductor must store in each cycle:

$$E_{L1} = P_{OUT} * \frac{1-d}{f_{SW}} \tag{13}$$

Since d depends only on voltages, not on frequency, the energy is proportional to 1/ f_{SW}.

Note the unexpected result that it is not the current that defines the dimensions of L1, but the load power and switching frequency. Calculating the inductance is relatively simple given the freewheel voltage (total forward voltage for the LEDs and freewheel diode) V_{FW}, off-time t_{OFF} and current ripple I_{RIP}:

$$L1 = t_{OFF} * \frac{V_{FW}}{I_{RIP}} \tag{14}$$

To be precise, V_{FW} should also include the voltage drop across the inductor's DC resistance at the average current I_{LED} if this is not negligible. Current ripple is typically assumed to be 30 to 50% of the average current.

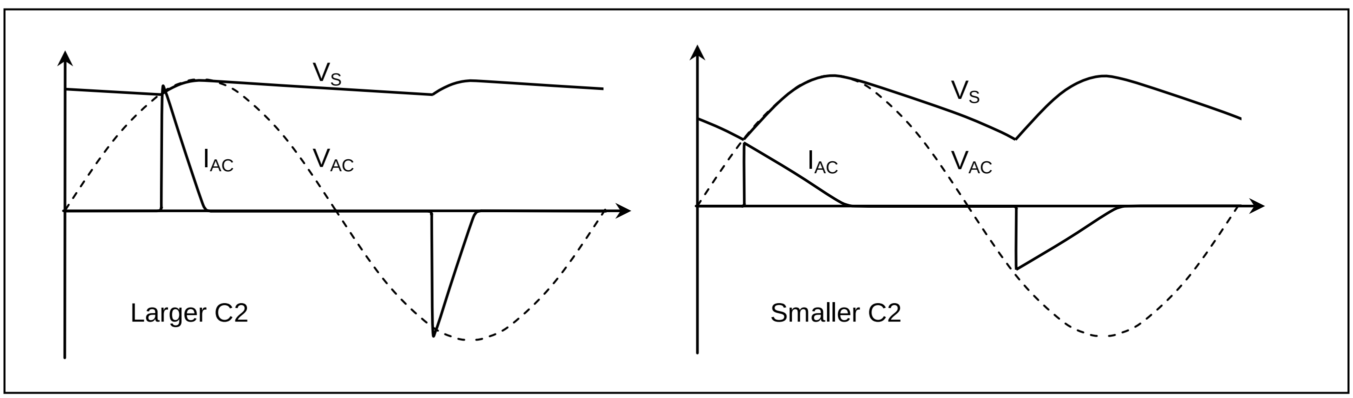
The inductor must not be saturated at the maximum current I_{LED} + ½ I_{RIP}, which is the tripping point of the current comparator in ZLED7001. A margin of 30 to 50% allows handling transient conditions; e.g., during start-up. For proper choice of core material, size and air gap, it is important to know

- Inductance
- Saturation current
- Current ripple
- Switching frequency

1.6. Converter DC Supply

The simplest way to provide a DC supply voltage V_S for the converter is to employ a bridge rectifier and a bypass capacitor C2 on its output (as shown in Figure 1.1). The capacitor charges to the AC peak voltage ($V_{EFF} * \sqrt{2}$) and then maintains the converter's supply for the time the input voltage goes low. Figure 1.3 shows the typical V_S voltage and AC line current shapes for a larger and a smaller bypass capacitor C2. Current from the input is only drawn when the AC amplitude is higher than V_S. Figure 1.3 shows only the stationary line current, when turning the device on, the inrush current is significantly higher.

Figure 1.3 Voltage and Current Wave Shapes



C2 must be large enough to sustain the maximum voltage of the LED chain until the AC amplitude is high enough to recharge it. Conversely, C2 should not be larger than necessary for a number of reasons:

- Minimizing cost and size of the capacitor
- Keeping the inrush current as low as necessary
- Improving the power factor and reducing the harmonic distortion and losses via a longer conduction angle for the line current (discussed in detail in section 1.12)

Since C2 must provide the switching current for the converter, it should be a low ESR type electrolytic or, even better, a ceramic capacitor.

1.7. Bypass Capacitor Size

As can be seen in Figure 1.3, the voltage across C2 does not drop linearly when driving the converter. LED current is constant, but the duty cycle changes with the voltage on C2 and so does the average discharge current. Therefore, it is easier to base calculations for the capacitor size on the load power, which can be considered nearly constant. First, define the worst case operating conditions without parameter degradation. Regarding the size of C2, these are

- minimum supply voltage,
- maximum voltage drop across the LEDs, which corresponds with
- maximum load power (i.e., maximum voltage and current) and
- minimum efficiency η of the converter.

In addition to the individual LED tolerances, the LED forward voltage has a negative temperature coefficient, so the maximum voltage occurs at the lowest operating temperature. The efficiency of the converter must be estimated for the initial design; some recommendations will be given later.

The discharge time of C2 is longer than ¼ and shorter than ½ of the AC period; i.e., 5 ms < t_{DIS} < 10 ms at 50 Hz line frequency and 4.17 ms < t_{DIS} < 8.33 ms at 60 Hz. This might be sufficient information for a simple approach; for a more precise value, calculate the angle at which the line voltage has reached the minimum voltage on C2 (V_{C2MIN}) with some margin for voltage drops across the rectifier, inductor resistance, switching transistor and shunt resistor being considered; e.g., a margin of 3 V in total:

$$\varphi = \arcsin\left(\frac{V_{C2MIN} + 3V}{V_{ACpeak}}\right) \tag{15}$$

where V_{ACpeak} is the amplitude of the minimum AC supply voltage. This leads to the total discharge time t_{DIS} per half cycle of the line voltage:

$$t_{DIS} = \frac{\frac{1}{4} + \frac{\varphi}{360°}}{f_{LINE}} \tag{16}$$

The energy stored in a given capacitor is

$$E_C = \frac{C}{2} * V^2 \tag{17}$$

The energy provided to the converter during the discharge time equals the difference between energy levels at peak voltage and minimum voltage and must be identical to the required power multiplied by discharge time.

$$\frac{C2}{2} * \left(V_{PEAK}{}^2 - V_{C2MIN}{}^2\right) = \frac{P_{LED}}{\eta} * t_{DIS} \Rightarrow$$

$$C2 = 2 * \frac{P_{LED} * t_{DIS}}{\eta * \left(V_{ACpeak}{}^2 - V_{C2MIN}{}^2\right)} \tag{18}$$

This is the minimum required capacitance value; however, especially when using an electrolytic capacitor, the nominal value should be selected higher; e.g., by a factor of 2, since these capacitors have a tendency to dry out over time and also lose some capacitance at low temperatures.

1.8. Open Load

One big advantage of the buck converter over other converters, such as boost or flyback converts, is that it is inherently open-load protected. Since the inductor is connected in series with the load, there is no more current flow in the switching transistor when the load fails open. The remaining magnetic energy of the inductor simply discharges into the output capacitor C_{LED}.

Still there is an important consideration: when the load fails open, the voltage across C_{LED} increases to the peak voltage of the AC supply and remains there. To be precise, there is a short transient condition in which the inductor charges the capacitor to an even higher voltage than the momentary V_S, but this will immediately be discharged (in a damped oscillation) after the inductor's current has reached zero. Even though the nominal supply of the LED chain may be relatively low, the voltage rating of the capacitor must match this fault condition in case an open load occurs.

1.9. Output Capacitor across LEDs

Selecting the correct capacitance for the output capacitor C_{LED}, is challenging. Its purpose is to reduce the current ripple of the LEDs, as well as avoiding EMI radiation if the wiring between the converter and LEDs is relatively long. To simplify calculations, assume that the current ripple created by the buck converter is sinusoidal. To reduce this ripple by a certain factor, include a low pass filter with a cut-off frequency f_R that is the same fraction of f_{SW} by which the current ripple of the LEDs should be reduced. From the LED's data sheet, derive the differential resistance $R_{LEDdiff}$ of a single LED as the tangent to the I(V) characteristic at the operating point. For a 1 W LED, $R_{LEDdiff}$ is typically in the order of 1.5 Ω. The size of C_{LED} can be calculated with equation (19) where n is the number of LEDs connected in series:

$$C_{LED} \geq \frac{1}{2\pi * f_R * n * R_{LEDdiff}} \tag{19}$$

C_{LED} should not be made larger than necessary, not only for cost reasons, but also to avoid excessive current during converter start-up when the capacitor must be charged from 0 V and initially is in effect a momentary short circuit on the converter output. Like C2, C_{LED} should be a low ESR type.

1.10. IC Supply

The supply voltage of ZLED7001 must be in the order of 7 V for operation, while current consumption depends on the external power transistor (Q1) and switching frequency f_{SW}. The ZLED7001 features a shunt regulator, but as long as it does not clamp the input voltage, it consumes less than 640 µA. Driving the gate of Q1 causes an additional average current of

$$I_{GATEave} = Q_{GATE} * f_{SW} \tag{20}$$

with Q_{GATE} being the transistor's gate charge. This current increases proportionally to f_{SW}, regardless of how fast or slow the actual switching transitions are.

Although the total supply current is typically only in the range of 1 to 4 mA, power dissipation is not negligible when deriving the supply linearly from V_S. For example, assuming a V_S of 300 V on average (as in the case of a 230 VAC supply), power dissipation would be in the order of 300 mW/mA.

Although it is not the most efficient method, for some applications, it may be reasonable to use a series resistor for the IC supply, simply because it is the cheapest solution.

Figure 1.4 shows an alternative circuit that causes much lower losses. After the converter has started operating, an "auxiliary converter," composed of a small inductor L2, a diode D_{AUX} and a damping resistor R_{DMP}, provides the necessary supply current. When Q1 is turned on, the current in L2 is the same as the current in L1, but since L2 is very small (typically between 5 and 50 µH, depending on the LED current), the auxiliary converter operates in discontinuous conduction mode (DCM); i.e., the current of L2 drops back to zero in every switching cycle. This makes it quite easy to calculate the required inductance of L2. The energy stored in L2 per cycle is

$$E_{L2} = \frac{L2}{2} * I_{PEAK}{}^2 \tag{21}$$

which leads to the average power

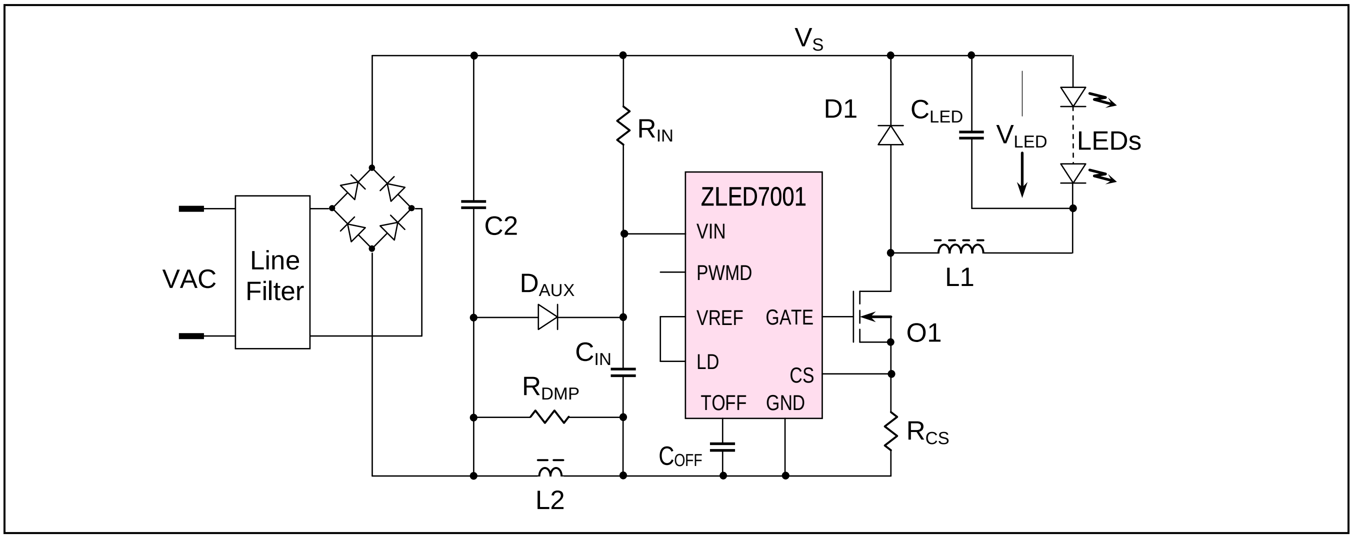
$$P_{L2} = \frac{L2}{2} * I_{PEAK}{}^2 * f_{SW} \tag{22}$$

With the power $I_{IN} * 8$ V (including diode voltage) needed for the IC supply, the minimum inductance for L2 can be calculated as

$$L2_{MIN} = 2 * I_{IN} * \frac{8V}{I_{PEAK}{}^2 * f_{SW}} \tag{23}$$

Although the "inductive discharge" pulse is very short, some losses caused by the damping resistor must be considered, but this additional power will be easily compensated by the start-up circuitry.

Figure 1.4 IC Supply from "Auxiliary Converter"



It is still necessary to power up the ZLED7001 from V_S; e.g., via a series resistor, but this resistor must provide only the ZLED7001's current consumption, and V_S may be considered as the peak voltage at minimum V_{AC}, since C2 is not yet loaded by the LED current. Therefore under the worst case conditions, R_{IN} can be calculated as

$$R_{IN} = \frac{V_{DCpeakMIN}}{640\mu A} \tag{24}$$

1.11. Efficiency Restrictions

High efficiency is an important target for this application design, especially considering that keeping the LED's temperature as low as possible is recommended. Each component of the application circuit dissipates power (including the capacitors), but this discussion will focus on those that contribute most to the total losses:

- Switching transistor Q1
- Series inductor L1
- Free wheel diode D1
- Bridge rectifier diodes D_{B1} to D_{B4}
- ZLED7001 supply resistor R_{IN}

Q1 causes "direct" and "indirect" losses, both linked to the current it must drive. Indirect losses are caused by charging and discharging the gate from the IC supply as shown in equation (20).

Direct losses can be divided into static and dynamic losses. The static power dissipation during the on-state of the transistor is easy to calculate:

$$P_{Q1stat} \approx I_{LED}^{2} * R_{DSon} * d_{AVE} \tag{25}$$

with on-resistance R_{DSon} and average duty cycle d_{AVE}. The "≈" occurs because the drain current I_{DRAIN} during conduction is not constant and P_{DISS} is a square function of I_{DRAIN}, but the error when considering the average current I_{LED} is acceptable.

Assuming that the transient switching current is significantly higher than the operating current, switching losses can be estimated as the total energy of the parasitic capacitance on Q1's drain node, which must be discharged, times the switching frequency. This capacitance is a combination of the inductor's parasitics, Q1's drain-source capacitance C_{DS} and C_{D1}, the reverse capacitance of D1. Semiconductors characteristically have significant non-linear voltage dependence of their capacitances, and especially for the diode, the junction charge accumulated during the conduction phase is more relevant. Typically data sheets for discrete components provide a value for the total charge at a certain switching voltage that does not consider the voltage dependency. However, this is a better approach than assuming a fixed capacitance.

As shown in equation (17), the energy stored in a capacitor is proportional to V^2, while for the semiconductors, the charge is used to estimate switching losses:

$$P_{Q1dyn} \approx \left(\frac{C_{L1} * V_S^{2}}{2} + Q_{DSQ1} * V_S + Q_{D1} * V_S \right) * f_{SW} \tag{26}$$

Equation (26) demonstrates that dynamic losses increase with V_S and f_{SW}, while equation (25) shows that static losses increase with I_{LED}^2 as long as R_{DSon} stays constant, so the normal consequence is to use a transistor with lower R_{DSon} if I_{LED} is higher. Unfortunately, transistors with lower R_{DSon} show higher drain and gate charge (when produced in the same technology), which in turn increases the direct and indirect switching losses.

This leads to an important conclusion regarding the topology of LEDs: when the LED power is given (which is normally the case), a long string of LEDs with high voltage and low current should be used rather than a short string with low voltage and high current to achieve good efficiency.

Inductor L1 also generates two types of losses: losses of the winding resistance R_{L1} and magnetizing losses of the ferrite core. The resistive power dissipation is given by

$$P_{L1res} \approx I_{LED}^{2} * R_{L1} \tag{27}$$

The core losses depend on the core material (type of ferrite), magnetic flux variation as a function of current ripple and f_{SW} (directly proportional). Selection of a suitable core for the inductor is described in supplier's data sheets.

The power dissipation of free wheel diode D1 is estimated assuming a constant forward voltage V_{FD1}:

$$P_{D1} = V_{FD1} * I_{LED} * (1 - d_{AVE}) \tag{28}$$

Transient losses caused by junction charge and diode capacitance have already been considered in the switching losses of Q1 (equation (25)). To keep the power dissipation as low as possible, a fast recovery diode with low forward voltage should be selected, considering of course, that it must withstand the maximum peak voltage in the reverse direction.

For simplicity, also assume constant forward voltages for the bridge rectifier diodes, D_{B1} to D_{B4}. Two of the four diodes are always conducting simultaneously, so a voltage drop of $2 * V_{FDB}$ must also be taken into consideration. Estimating the average current is a little more difficult because the supply current varies with duty cycle as a function of momentary DC voltage. For this simple approach, assume the DC voltage as the average between peak voltage and minimum voltage on C2. This leads to an average supply current of

$$I_{S-AVE} \approx \frac{P_{LED}}{\eta} * \frac{2}{V_{PEAK} + V_{C2MIN}} \tag{29}$$

and the rectifier's total power dissipation

$$P_{DB1-DB4} \approx \frac{P_{LED}}{\eta} * \frac{4 * V_{FDB}}{V_{PEAK} + V_{C2MIN}} \tag{30}$$

Actually, it is difficult to prevent this dissipation without implementing an active rectifier. Fortunately the diode losses decrease efficiency by an insignificant amount, in the order of 1%.

Power dissipation of the series supply resistor R_{IN} can be estimated by assuming a linear increase and decrease of V_S (somewhat optimistic):

$$P_{Rin} = \frac{1}{R * T} * \int_0^T \left[V_{PEAK} - \frac{t}{T} * \left(V_{PEAK} - V_{C2MIN} \right) \right]^2 dt$$

$$P_{Rin} = \frac{V_{PEAK} * V_{C2MIN} + \frac{\left(V_{PEAK} - V_{C2MIN} \right)^2}{3}}{R} \tag{31}$$

The alternative solution for providing the operating IC current described in section 1.10 helps improve total efficiency, especially when V_{PEAK} is relatively high; e.g., 325 V with a 230 VAC supply.

1.12. Power Factor

When using a rectifier with a bypass capacitor for the converter's supply, the line current wave shape is far from being sinusoidal, as shown in Figure 1.3. Actually, the conduction angle becomes even smaller, and the current peak higher, when increasing the size of the bypass capacitor. Since the signal is periodic, it can be considered as a superposition of sinusoidal waves, and since it is symmetrical (for positive and negative half waves), it contains only odd harmonics.

These harmonics do not contribute to the real power that supplies the converter, but they create losses in the power supply lines, transformers and back to the generator. The number of alternative lamps using electronic ballasts will increase significantly in the next years, and therefore many electric utilities are advocating for legislation to reduce the harmonic content of line current. As of the end of 2010, there has been discussion of reducing the limit above which power factor correction is required from 75 W to 25 W for all lighting applications.

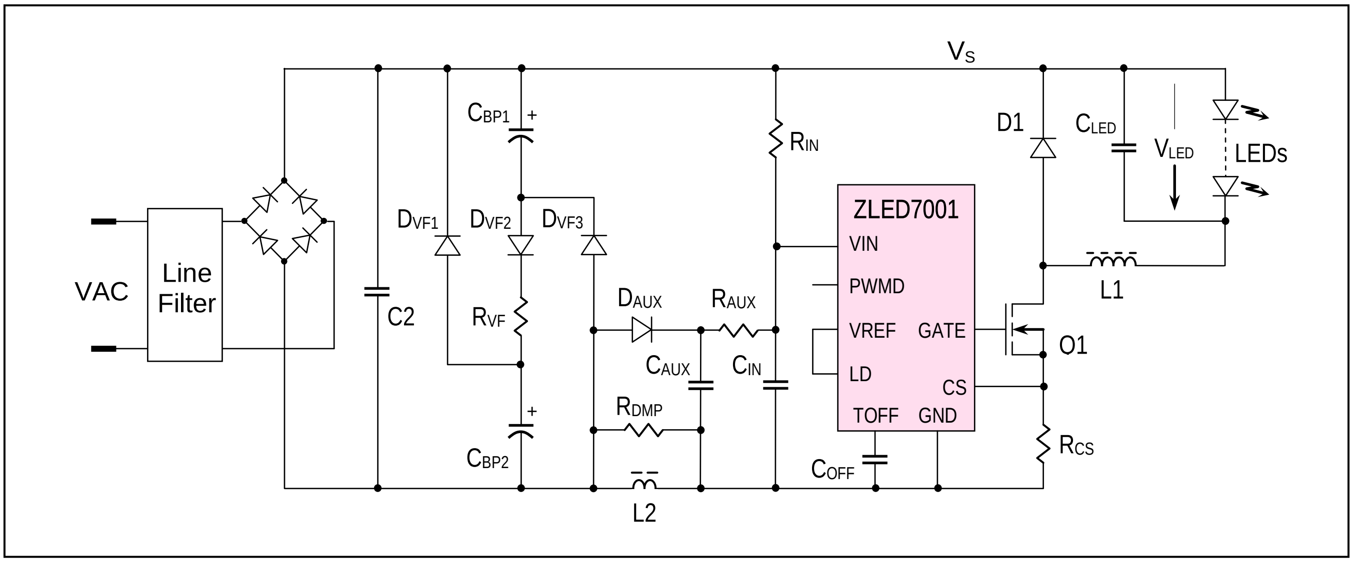
Limits for each individual harmonic are specified in IEC 61000 - 3 - 2, but what is more commonly known is the power factor pf:

$$pf = \frac{\cos\varphi}{\sqrt{1 + \left(\frac{I_2}{I_1} \right)^2 + \left(\frac{I_3}{I_1} \right)^2 + \ldots}}$$

$$pf = \frac{\cos\varphi}{\sqrt{1 + THD^2}} \tag{32}$$

The power factor that can be achieved with the supply generation of Figure 1.1 is typically in the range of 0.5 to 0.65, which is very poor. Even though many LED applications will stay well below 25 W, it makes sense to consider at least a simple improvement of the power factor by including a "valley fill" circuit, as is shown in Figure 1.5.

Figure 1.5 Buck Converter with Valley Fill and Auxiliary Converter



Two bypass capacitors C_{BP1} and C_{BP2} are charged in series connection via D_{VF2} and R_{VF} to the peak of the input voltage, and when the AC input drops below ½ of the peak voltage, they are connected in parallel via diodes D_{VF1} and D_{VF3} and provide the converter's supply. R_{VF} is included to reduce the charging peak of the capacitors. The line current conduction angle improves to > 120°, and even though the current wave shape is still far from being sinusoidal, the reduction of harmonic distortions is quite significant. Figure 1.6 shows a typical waveform and Table 1.1provides a comparison of its first 11 harmonics with the simple circuit of Figure 1.1 (with "large" and "small" bypass capacitors), where cos φ refers to the fundamental wave's phase displacement.

Figure 1.6 Typical Valley Fill Line Current and Comparison of Distortions and Power Factor

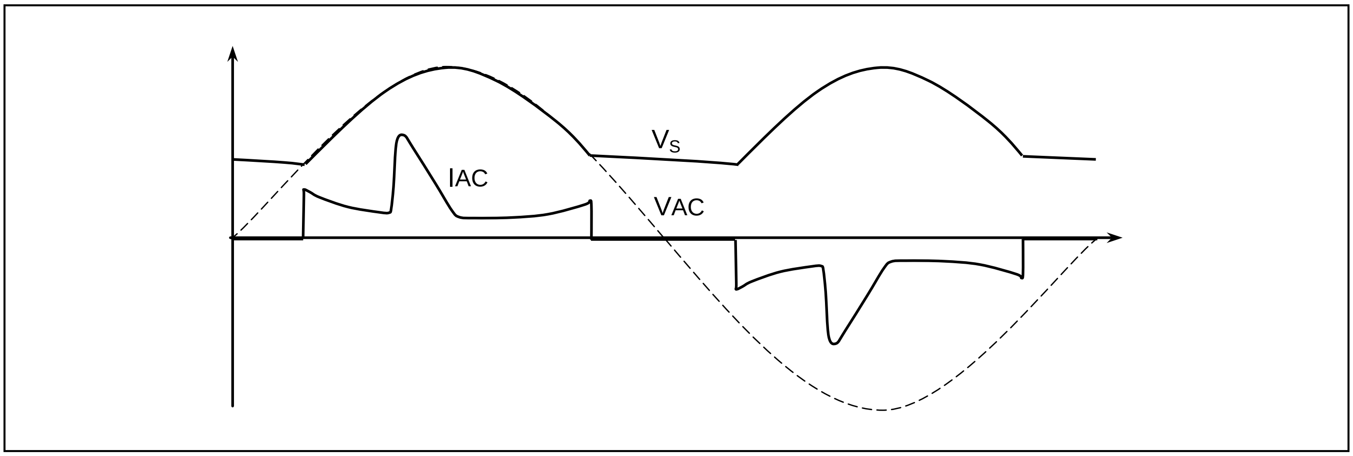


Table 1.1 Comparison of Effects of C2 and the Valley Fill Circuit on Harmonics, cos φ and Power Factor pf

Note: Values for harmonics are relative to the fundamental waveform.

	3^{rd}	5^{th}	7^{th}	9^{th}	11^{th}	cos φ	pf
"Large" C2	95%	84%	71%	56%	41%	0.96	0.51
"Small" C2	71%	35%	22%	20%	9%	0.81	0.62
Valley Fill	22%	25%	32%	25%	24%	0.995	0.86

D_{VF1}, D_{VF2}, D_{VF3} and R_{VF} cause additional power dissipation, but the valley fill approach could still help improve total efficiency by reducing the average DC voltage and thus the transient losses of switching transistor Q1.

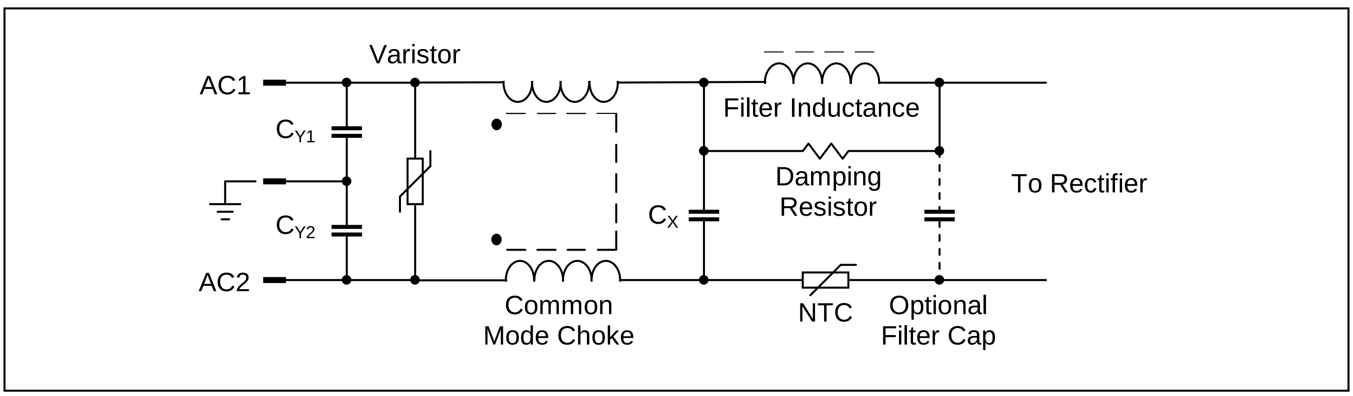
1.13. Line Filter

A line filter is needed to protect the electronics from harsh power supply transients and conversely reduces conducted EMI (electro magnetic interference) caused by the switching converter. An optimal line filter contains common mode and differential mode sections, connected in series. The common mode filter contains two windings on the same magnetic core with identical orientation, but opposite supply current flow direction. It features low impedance for the supply current, but high impedance for common mode line transients. The "X" and "Y" capacitors might be available together with the common mode choke in a single package.

!

Warning: The Y capacitors provide a good HF path to earth potential, but must be quite small (e.g., 2.2 nF) to prevent the risk of potentially dangerous electrical shock to anyone touching the application body if an open earth connection occurs. Since one of the AC supply lines is normally also connected to earth potential, both capacitors must withstand the maximum AC supply voltage. Additionally they must be self-healing in case of over-voltage breakdown for electrical shock protection.

Figure 1.7 Line Filter



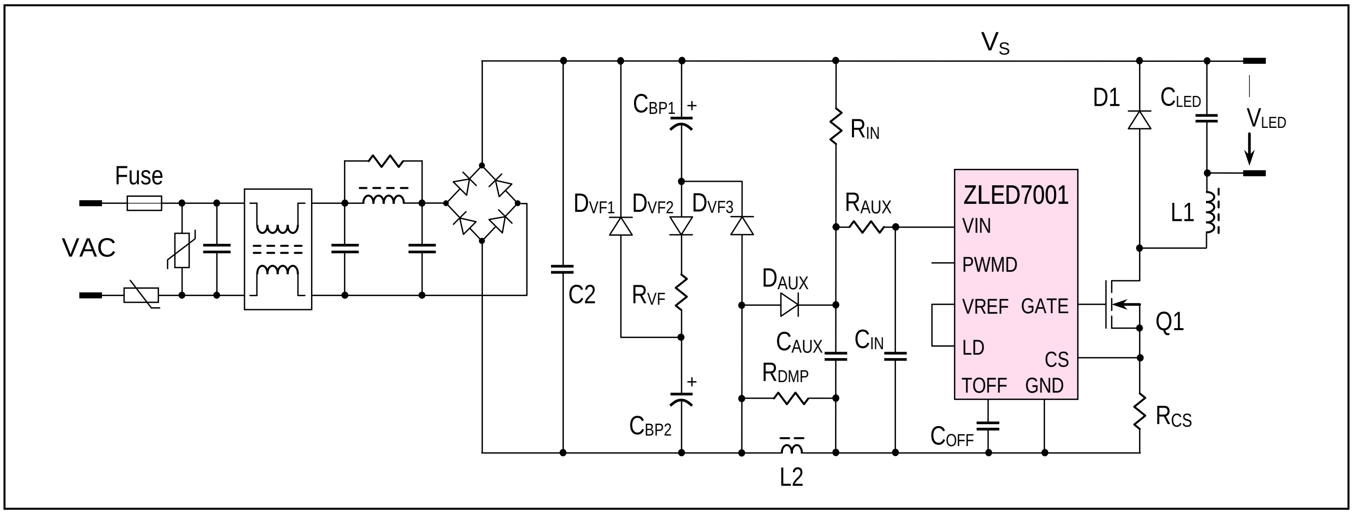
A varistor on the AC supply features over-voltage protection for the electronics, and a series NTC (negative temperature coefficient) resistor reduces inrush current when charging the bypass capacitor(s) after power on. Initially the NTC's resistance is high, and after being powered by the nominal supply current, it heats up and becomes low resistive.

The differential mode filter contains one single or two independent inductances and operates in conjunction with an optional AC capacitor and the bypass capacitor C2. Although the resonance frequency is typically much lower than the converter's switching frequency, sub-harmonic excitation might lead to significant line currents on the filter resonance. Therefore a resistor of a few 100 Ω in parallel with the inductor(s) is recommended to suppress such oscillations.

1.14. Putting it all Together

Figure 1.8 shows a schematic that contains all the necessary components to drive a string of up to 20 LEDs at a current defined by R_{CS} from a 110 or 230 VAC supply. On the input, there is a fuse, a varistor for over-voltage protection, a common mode line filter and a differential line filter.

Figure 1.8 Example Schematic



Note: A similar circuit is used in the ZLED7001 Line Power Kit available at www.ic-board.de (click on ZLED Kits).

The valley fill circuit is an optional assembly, and a single electrolytic bypass capacitor can be substituted for it. R_{IN} generates the initial IC supply voltage, with C_{IN} providing the switching current peaks for the gate of Q1. After the main converter has started operating, the necessary supply current is generated by the auxiliary converter L2, D_{AUX}, C_{AUX}, and R_{AUX} with the damping resistor R_{DMP}. C_{OFF} defines the constant off-time, and R_{CS} sets the current switch-off threshold.

2 Related Documents

Document
ZLED7001 Data Sheet

Visit www.IDT.com/ZLED7001 or contact your nearest sales office for the latest version of these documents.

3 Glossary

Term	Description
NTC	Negative temperature coefficient
Buck Converter	A type of step-down DC-to-DC converter.

4 Document Revision History

Revision	Date	Description
1.0	June 10, 2011	First release of document.
	April 19, 2016	Changed to IDT branding.

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