GENERAL DESCRIPTION

The PT4213A is a primary side regulated constant current controller, which is designed for LED lighting applications. The device regulates output current without the secondary feedback loop.

The device integrates an oscillator, a current sense circuit, CC control circuit, a pre-driver and a complete set of protection circuits to protect against all fault conditions including output open/short circuit, line under-voltage, and over temperature protection. The PT4213A is available in SOT23-6 package.

FEATURES

- CC Without Secondary Feedback
- Inductance Compensation
- Low Startup Current (<10µA)
- Adjustable Primary Side Current limit
- FB Over Voltage Protection
- VCC/FB UVLO
- Feedback Loop Open Circuit Protection
- Over Temperature Protection
- RoHS compliant

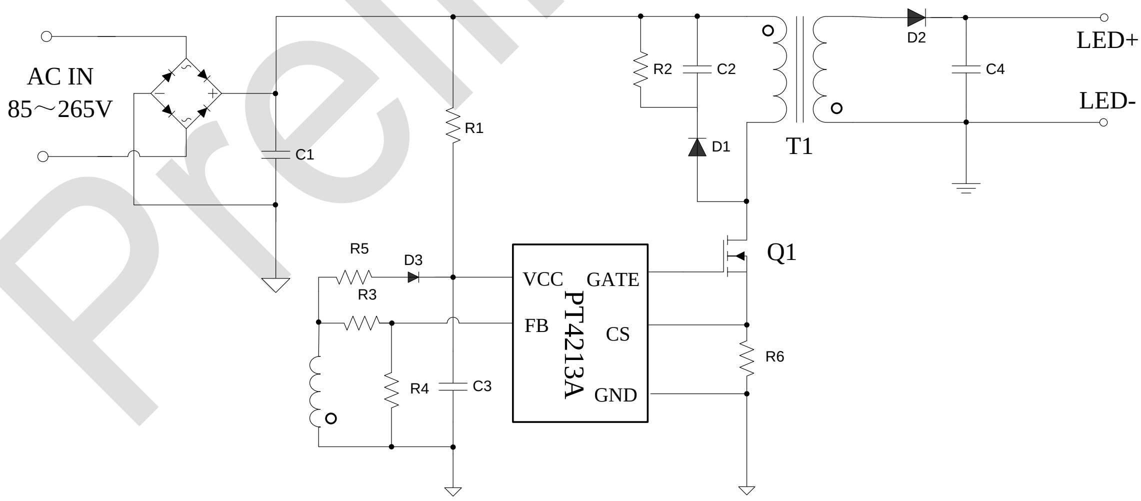
APPLICATIONS

- Off-line High Brightness General LED Lighting
- Integrated LED Lamps such as GU10, E27, PAR20/20/38 LED Lamps

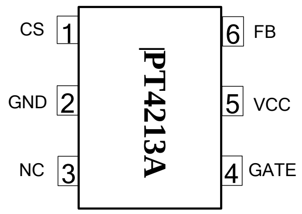
ORDERING INFORMATION

PACKAGE	TEMPERATURE RANGE	ORDERING PART NUMBER	TRANSPORT MEDIA	MARKING
SOT23-6	-40℃ to 85℃	PT4213AE23F	Tape and Reel 3000 units	4213

TYPICAL APPLICATIONS CIRCUIT



PIN ASSIGNMENT



SOT23-6

PIN DESCRIPTIONS

PIN No.	PIN NAMES	DESCRIPTION
1	CS	Primary Side Current Sense Input
2	GND	Ground
3	NC	Not connected
4	GATE	Drive output
5	VCC	Power supply, the device is supplied by an auxiliary winding.
6	FB	Auxiliary Winding Voltage Sense Input

ABSOLUTE MAXIMUM RATINGS (note1)

SYM	PARAMETER	VALUE	UNIT
VCC	VCC pin Voltage	-0.3~30	V
V_{FB}	FB pin Input Voltage	-0.3~5	V
V_{CS}	CS pin Voltage	-0.3~5	V
V_{GATE}	GATE pin output voltage	-0.3~15	V
Topt	Operating Junction Temp. Range	-40 to 150	℃
Tstg	Storage Temp. Range	-55 to 150	℃
ESD	HBM	2000	V
RΘJA	SOT23-6	250	℃/W

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. Recommended Operating Range indicates conditions for which the device is functional, but do not guarantee specific performance limits. Electrical Characteristics state DC and AC electrical specifications under particular test conditions which guarantee specific performance limits. This assumes that the device is within the Operating Range. Specifications are not guaranteed for parameters where no limit is given, however, the typical value is a good indication of device performance.

RECOMMENDED OPERATING RANGE

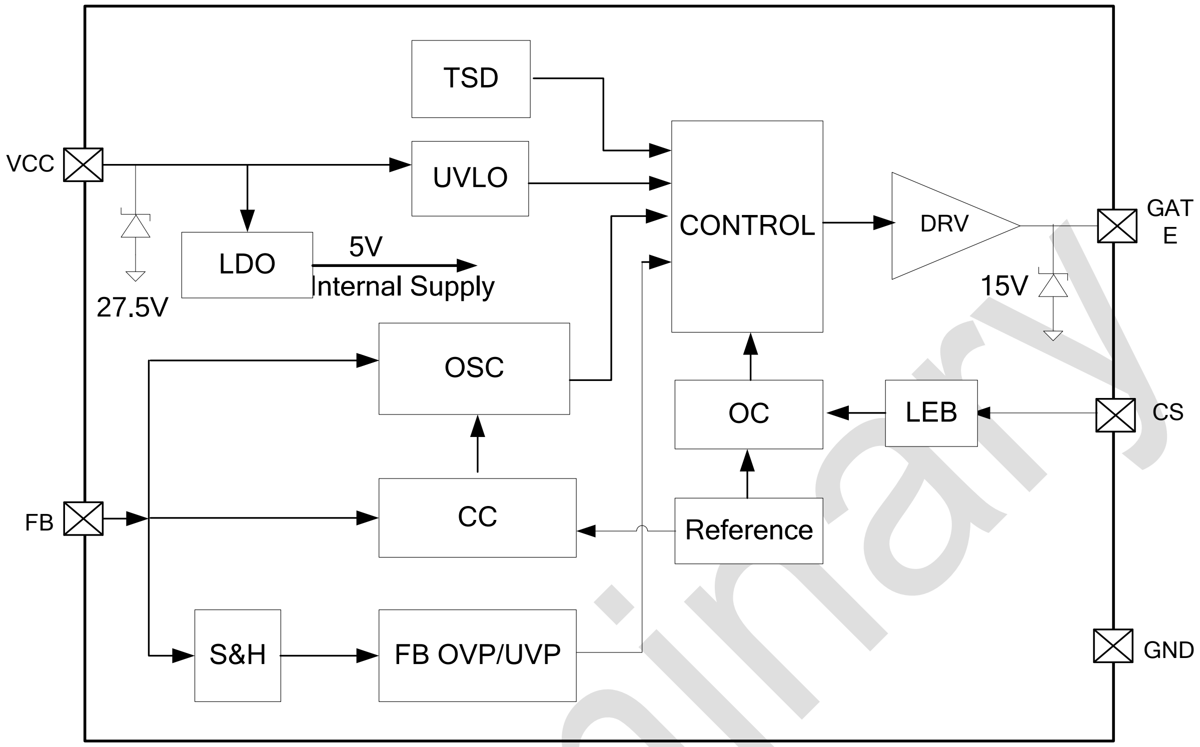
SYM	PARAMETER	VALUE	UNIT
VCC	VCC pin Operating Voltage	9.5~26	V
T_A	Operating Ambient Temperature	-20~85	℃

ELECTRICAL CHARACTERISTICS

(T_A=25℃, VCC=16V，f_{sw}=65kHz unless specified otherwise)

SYMBOL	PARAMETER	TEST CONDITION	MIN	TYP	MAX	UNIT
SUPPLY VOLTAGE (VCC)						
I_{START}	Start up current	VCC=12.0V		1	10	μA
V_{VCC_ON}	VCC turn on threshold	VCC Rising	12.5	14.0	15.5	V
V_{VCC_OFF}	VCC minimum operating voltage	VCC Falling	7.5	8.5	9.5	V
V_{VCC_Clamp}	VCC Clamp Voltage	Icc=10mA		33		V
V_{VCC_OVP}	VCC Over Voltage Protection Threshold		26	27.5	29	V
I_{VCCQ}	VCC Supply Current	No switching		350	700	μA
FEED BACK VOLTAGE SENSE PIN (FB)						
V_{FBMAX}	FB Over Voltage Protection		2.4	2.5	2.6	V
V_{FBMIN}	FB Minimum Voltage			0.8		V
I_{FB_OPEN}	FB Open Loop Current			-85		μA
CURRENT SENSE INPUT PIN (CS)						
VCS	Current Limit Threshold	Ifb=0	490	500	510	mV
T_{LEB}	Current sense Leading Edge Blanking Time			250		ns
DRIVE OUTPUT (GATE)						
Dmax	Maximum driving pulse duty cycle			65		%
Tr	Rising time	Cgate=1nF		220		ns
Tf	Falling time	Cgate=1nF		120		ns
V_{GS_MAX}	Gate clamp voltage			15		V
PROTECTION						
T_{SD}	Thermal Shut Down Threshold			150		℃

SIMPLIFIED BLOCK DIAGRAM



OPERATION DESCRIPTION

The PT4213AA consists of an oscillator, feedback circuit, over-temperature protection, frequency jittering, current limit circuit, leading-edge blanking, and constant current control circuitry. The switching frequency is modulated to regulate the output current to provide a constant current characteristic. It senses and regulates output current from primary side of transformer and is ideal for high precise, high reliability and cost effective LED lighting application.

THEORY OF OPERATION

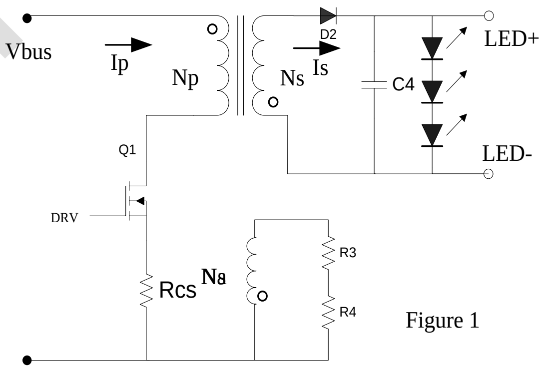


Figure 1

Figure 1 illustrates a simplified flyback converter. When the switch Q1 turns on, the voltage across the primary winding is V_{bus}. Assuming the voltage dropped across Q1 is zero, the current in Q1 ramps up linearly at a rate of V_{bus}/L_p. When the current in Q1 reaches a predefined value of I_{pk_pri} the controller forces Q1 turns off.

During the Q1 on-time, the rectifying diode D2 is reverse biased and the load current Io is supplied by the secondary capacitor C4. When Q1 turns off, D2 conducts and the stored energy is delivered to the output.

The PT4213AA is designed to operate in discontinuous conduction mode (DCM), in DCM the energy stored in primary winding is transported to the secondary winding completely during each cycle. The output current is determined by R_{cs} and primary/secondary turns ratio of the transformer. Assuming the transformer primary/secondary turns ratio is N_{ps}. The output current is given by following expression:

$$I_{out} = \frac{0.1125}{R_{cs}} * N_{ps}$$

START UP

Once the AC voltage is applied to the application circuit, the V_{bus} charges V_{CC} pin up through the start up resistor R1, when the voltage on V_{CC} pin reaches its V_{CC_ON} threshold the controller starts to deliver the driving pulses to power switch Q1 and V_{CC} is powered by auxiliary winding. Thanks to the very small start up current, a large start up resistor R1 could be used in the start up circuit to minimize power loss. For the applications with general 90-264Vac input range, a 1/8W resistor between 0.5Mohm and 3Mohm and a 4.7uF/50V capacitor C3 compose a simple and reliable start up circuit.

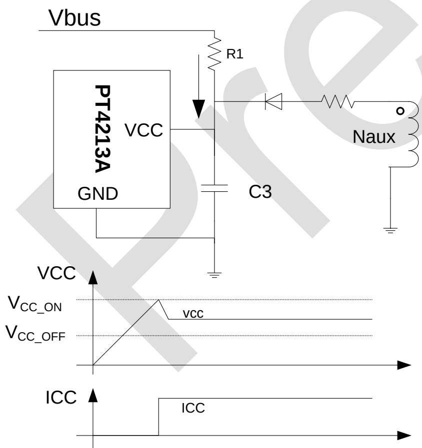


Figure 2

CONSTANT CURRENT (CC) OPERATION

The PT4213A regulates output current from the primary side control. The switching frequency is adjusted as the feedback pin voltage increases to provide a constant output current regulation.

PRIMARY INDUCTANCE COMPENSATION

With the inductance compensation function output current is free of primary inductance variation. If the primary magnetizing inductance is either too high or low the converter will automatically adjust the Oscillation frequency to maintain the constant output current.

CURRENT SENSE AND LEB

The current of the power MOSFET is transferred to voltage signal through a resistor connected between source terminal and GND and feeds to CS input. At each switching cycle when the voltage of CS input excess the internal threshold the driving signal is terminated after a short delay. The relationship between the CS threshold and the primary peak current passing through power MOSFET Q1 follows below expression: $I_{pk\_pri}=V_{cs}/R_{cs}$; I_{pk_pri} is the peak current through power MOSFET, V_{CS} is the voltage threshold of pin1 of PT4213A and R_{cs} represents sensing resistor.

A spike is inevitable on the sensed signal on R_{cs} at the instance when the power MOSFET is turned on due to the recovery time of the secondary rectifier and the snubber circuit. The LEB has been implemented in PT4213A, during the LEB time the current sense comparator is disabled so the switching signal can not be terminated by the turn-on spike on the sensed signal so the external RC filter can be removed.

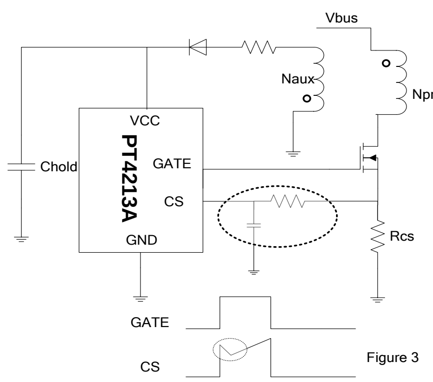


Figure 3

CURRENT LIMIT COMPENSATION

The current limit circuit senses the current in the power MOSFET from the resistor connecting between the source of MOSFET and GND. The current sense resistor converts the current in the power MOSFET to voltage signal and then feeds to CS pin. When the voltage on CS pin exceeds the internal threshold V_{CS} which is defied as 500mV, the power MOSFET is turned off for the remainder of that cycle. Excellent regulation performance is achieved with C.R.PowTech proprietary line regulation control technique.

SHORT CIRCUIT PROTECTION

In the event of a fault condition such as an output short condition the PT4213A enters into an appropriate protection mode as described below. In the event the feedback pin voltage falls below 0.8V during the discharged period of the inductance of the primary winding, the converter enters into short circuit protection mode after the feedback pin sampling delay for the duration in excess of ~30 ms, wherein the device is disabled. V_{CC} will then drops due to internal power consumption. When V_{CC} drops below the V_{VCC_OFF} turn-off threshold, the PT4213A will be totally shut down and the start up sequence will kick in and V_{CC} is charging up again. The device is alternately enables and disables until the fault condition is removed.

OPEN CIRCUIT PROTECTION

In the event of a fault condition of the output open circuit the PT4213A enters into an appropriate protection mode as described below. In the event the feedback pin voltage is over 2.5 V during the discharged period of the L_p (the inductance of the primary winding of the main transformer), the switching frequency is decreased and peak current is also decreased to 50%. If the over voltage condition duration time excess 8 consecutive cycles the controller is disabled.

FREQUENCY JITTERING

The frequency jittering is implemented in the PT4213A. The oscillation frequency is modulated so that the tone energy is spread out. The spread spectrum minimizes the conduction band EMI and therefore reduces system design challenge.

OVER TEMPERATURE PROTECTION

The thermal shutdown circuitry senses the die temperature. When the die temperature is above 150°C the device is disabled and remains disable until the die temperature falls by 20°C.

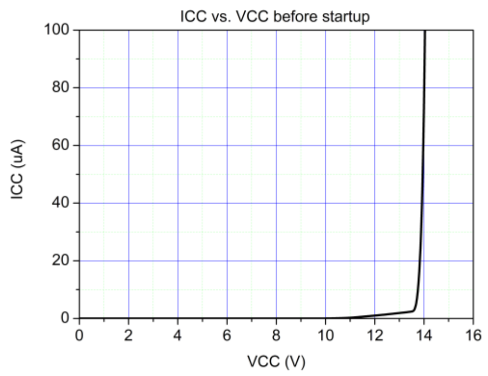
VCC OVER VOLTAGE PROTECTION

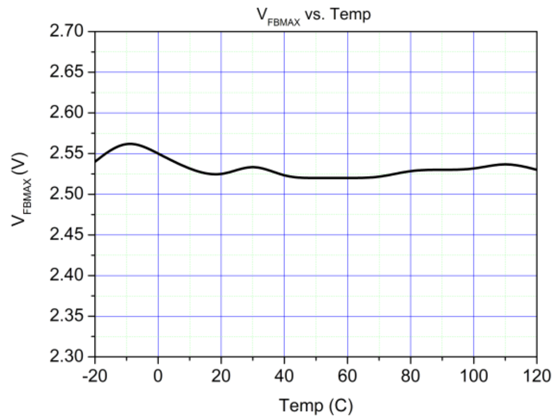
V_{CC} over voltage protection is designed to protect the device from damage of over voltage. When the voltage on V_{CC} reaches the OVP threshold the PT4213A stops delivering PWM signal to the power MOSFET, voltage on V_{CC} begins to drop due to the internal power consumption, the PT4213A will recover from OVP status when voltage on V_{CC} drops below the OVP release threshold.

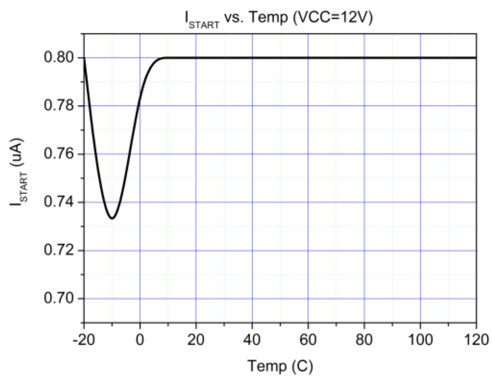
GATE OUTPUT

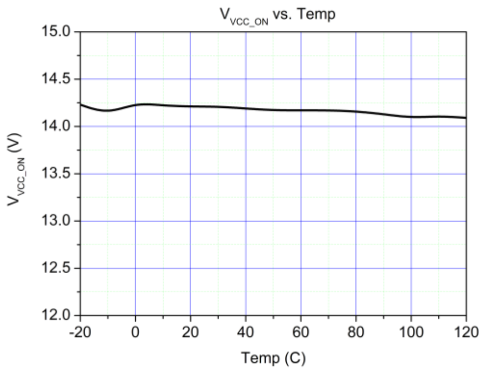
The output drives the GATE of the power MOSFET. The optimized totem-pole type driver offers a good tradeoff between driving ability and EMI. Additionally the output high level is clamped to 15V by an internal clamp so that power MOSFET transistor can be protected against undesirable gate over voltage. A resistor between GATE and GND initials the gate voltage to zero at the off state.

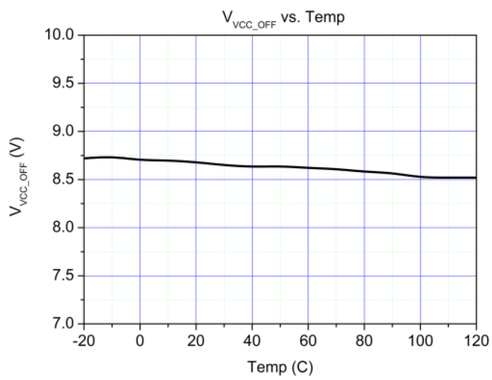
TYPICAL PERFORMANCE CHARACTERISTICS

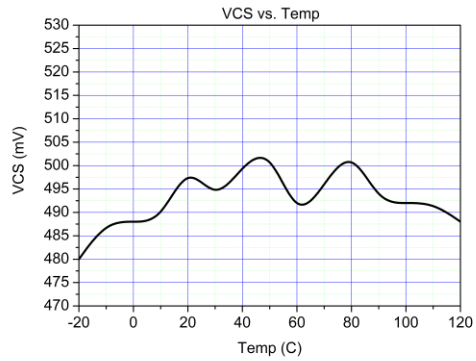












DESIGN Example and Notes:

A design example is given to illustrate how to design an Off-line AC/DC LED driver step by step based on PT4213A. Refer to figure 4 for the application circuit.

1 Determine input and output

The design spec is given in Table 1. It is intended to drive 5pcs 1W white LED in series for general lighting

Table 1

Parameters	Symbol	Limits
Input Voltage	Vin	90-264Vac
Frequency	f_{line}	47-64Hz
Output Voltage	Vout	16V
Max Vout	$Vout_{max}$	17.5V
Min Vout	$Vout_{min}$	15V
Output Current	Iout	320mA

2 Set the switching frequency

The maximum operating frequency is limited by

PT4213A sampling delay time on FB pin during the flyback period. To sample the voltage on FB pin correctly the secondary side discharge duration time must be more than 3.5uS, and the maximum limited switching frequency for PT4213A is 128 kHz. In practice, the suitable switching frequency for this design is set to 65kHz considering the EMI, efficiency and the size of the main transformer.

For this design example, the switching frequency f_{sw} and operating period T_{sw} is set as below:

$$f_{sw} = 65kHz$$

$$T_{sw} = 1/f_{sw} = 15.4\mu s$$

3 The maximum duty cycle

For PT4213A, the discharged time t_{dis} of the inductance of the primary winding is fixed to be 45% of the operating cycle. Keep in mind the operation of the LED driver using PT4213A should be DCM in the full input voltage range in order to regulate the LED current precisely. So:

$$T_{sw} = t_{on} + t_{dis} + t_{dead} \qquad (1)$$

t_{on}: turn on time of the switch MOSFET Q1

t_{dis}: discharge time of L_p

t_{dead}: dead time of the flyback DCM mode

$$t_{dis} = 0.45 * T_{sw} \qquad (2)$$

$$t_{dead\_\min} = 1.5 * 2\pi\sqrt{L_p * C_{ds}} \qquad (3)$$

For this design example, the minimum t_{dead_min} can be estimated to be approximately 20% of the switch cycle as a start point and then adjusted after the LED driver is made.

$$t_{dead\_\min} = 1.5 * 2\pi\sqrt{L_p * C_{ds}} = 20\% * T_{sw} = 3.08\mu s$$

$$t_{dis} = 45\% * T_{sw} = 6.93\mu s$$

$$t_{on\_\max} = T_{sw} - t_{dis} - t_{dead\_\min}$$

$$D_{\max} = t_{on\_\max} / T_{sw} = 35\%$$

4 Turns Ratio

The maximum allowable N_{ps} is determined by the voltage stress on power switch Q1, the minimum operating voltage $V_{in\_dc\_min}$ and primary side inductance L_p. The minimum allowable N_{ps} is limited by the maximum allowable reverse voltage on the secondary rectifying diode D3. So the N_{ps} is the trade-off among the Drain-Source breakdown voltage of the power switch Q1, the breakdown voltage of the secondary rectifying diode D3, the primary side inductance L_p, the primary peak current I_{pk_pri}, and the minimum operating input voltage $V_{in\_dc\_min}$. The smaller of N_{ps}, the smaller of $V_{in\_dc\_min}$ and less capacitance of the bulk capacitor is required. But it is led to higher reverse voltage for the secondary rectifying diode D3. In practice, the recommended N_{ps} is less than 5.0 for general applications with 85VAC-265VAC input.

$$V_{in\_dc\_\min} * t_{on\_\max} = N_{ps} * V_{out} * t_{dis} \qquad (4)$$

From equation (4), the below equation is given

$$N_{ps} = V_{in\_dc\_\min} * t_{on\_\max} / V_{out} * t_{dis} \qquad (5)$$

For this design example, the minimum operating input voltage is estimated to be approximately 60Vdc as a start point.

$$N_{ps} = V_{in\_dc\_\min} * t_{on\_\max} / V_{out} * t_{dis} = 60 * 5.39 / 16 * 6.93 = 2.92$$

5 Calculate the current sense resistor

Once N_{ps} is determined, the current sense resistor R_{cs} can be calculated from following expression:

$$R_{cs} = \frac{0.1125}{I_{out}} * N_{ps} \qquad (6)$$

$$I_{pk\_pri} = \frac{V_{cs}}{R_{cs}} = \frac{0.5V}{R_{cs}} \qquad (7)$$

For this design example,

$$R_4 = R_{cs} = \frac{0.1125}{I_{out}} * N_{ps} = \frac{0.1125 * 2.92}{0.32} = 1.0\Omega$$

$$I_{pk\_pri} = \frac{V_{cs}}{R_{cs}} = \frac{0.5V}{R_{cs}} = \frac{0.5V}{1.0\Omega} = 0.5A$$

6 Calculate the primary inductance

Primary inductance L_p is determined by the output power, the switch frequency f_{sw}, the primary peak current I_{pk_pri} and the flyback converter efficiency η. The relationship between primary energy and the secondary energy is given by:

$$\frac{1}{2}L_p * I_{pk-pri}{}^2 * F_{sw} * \eta = V_{out} * I_{out} \quad (8)$$

From the equation (8), the primary inductance L_p of the main transformer can be derived:

$$L_p = \frac{2*V_{out}*I_{out}}{I_{pk-pri}{}^2 * F_{sw} * \eta} \quad (9)$$

For this design example,

$$L_p \le \frac{2*V_{out}*I_{out}}{I_{pk-pri}{}^2 * F_{sw} * \eta} = \frac{2*16*0.32}{0.5^2 * 65*10^3 * 0.90} = 0.7mH$$

For this design example, L_p is chosen to be 660µH.

7 Calculate the primary winding

In order to keep the transformer from saturation, the maximum flux density must not be exceeded. So the minimum turns of the primary winding must meet:

$$N_p = \frac{L_p * I_{pk\_Pri}}{A_e * \Delta B_{max}} \quad (10)$$

Where ΔB_{max} is the maximum allowed flux density and Ae is the core effective area.

For this design example, EE16 core is selected. From the EE16 transformer core datasheet, Ae is 19.2, ΔB_{max} is chosen to be 2500 Gauss. So:

$$N_p = \frac{L_p * I_{pk\_pri}}{A_e * \Delta B_{max}} = \frac{0.66*10^{-3}*0.5}{19.2*10^{-6}*0.25} = 68.75$$

69 turns is chosen for this deign example.

8 Calculate the secondary winding

The turns of the secondary winding can be calculated:

$$N_s = \frac{N_p}{N_{ps}} \quad (11)$$

For this design example,

$$N_s = \frac{N_p}{N_{ps}} = \frac{69}{2.92} = 23.63$$

23 turns is chosen for this design example.

9 Calculate the auxiliary winding

The turns of the auxiliary winding can be calculated:

$$N_a = \frac{V_{cc}}{V_{out}} * N_s \quad (12)$$

In practice Vcc is set to be 12V.

For this design example,

$$N_a = \frac{V_{cc}}{V_{out}} * N_s = \frac{12}{16} * 23 = 17.25$$

17 turns is chosen for this design example.

10 Determine feedback resistors Rfb_up and Rfb_dn

For better regulation the current flowing out from FB pin of PT4213A at 220VAC input voltage is better to be approximate 1mA, so the R_{fb_up} is given by:

$$R_{fb\_up} = \frac{\sqrt{2}*220V}{1mA} * \frac{N_a}{N_p} \quad (13)$$

R_{fb_up} unit is $k\Omega$.

For this design example,

$$R_5 = R_{fb\_up} = \frac{\sqrt{2}*220V}{1mA} * \frac{17}{69} = 76.6k\Omega$$

R5 is chosen to be $75k\Omega$ for R_5 for this design

The low side feedback resistor is selected to set the output OVP. R_{fb_dn} should be selected so that when output voltage reaches OVP, the voltage on FB pin reaches V_{FBMAX} during transformer reset time. The relationship between V_{ovp} and V_{FBMAX} is:

$$V_{FBMAX} = \frac{R_{fb\_dn}}{R_{fb\_up} + R_{fb\_dn}} \frac{N_a}{N_s}(V_{ovp} + V_d) \quad (14)$$

Re-arrange above expression R_{fb_dn} can be derived as:

$$R_{fb\_dn} = \frac{V_{FBMAX} * R_{fb\_up}}{\frac{N_a}{N_s}(V_{ovp} + V_d) - V_{FBMAX}} \quad (15)$$

V_d: forward voltage drop of the secondary side rectifier diode, V_{FBMAX} is maximum FB pin operating voltage which has a typical value of 2.5V.

For this design example,

$$R_6 = R_{fb\_dn} = \frac{2.5*75}{\frac{17}{23}(20+0.5) - 2.5} = 14.82k\Omega \cdot$$

R6 is chosen to be 15 $k\Omega$; output OVP is set to be 20V for this design example.

11 Select the secondary and auxiliary rectifying diode

Maximum reverse voltage on secondary and auxiliary rectifying diode is:

$$V_{sec\_diode} \ge \left(\frac{\sqrt{2}*265*N_s}{N_p} + V_{out}\right) \qquad (15)$$

$$V_{aux\_diode} \ge \left(\frac{\sqrt{2}*265*N_a}{N_p} + V_{cc}\right) \qquad (16)$$

Peak current flowing in secondary rectifying diode is:

$$I_{pk\_sec} = \frac{0.5*N_{ps}}{R_{cs}} \qquad (17)$$

The secondary rectifying diode should be selected with the break down voltage higher than V_{sec_diode} and average forward current should be selected based on the output current and the secondary peak current. The auxiliary rectifying diode should be selected with the break down voltage higher than V_{aux_diode}.

12 PCB Layout

The circuit shown in Figure 4 is configured as a primary-side regulated off-line flyback power supply utilizing PT4213A with 320mA CC output for driving 5pcs 1W LED in series.

AC input power is rectified by BD1. The rectified DC is filtered by the bulk capacitors C1 and C2. Inductor L1, C1 and C2 form a pi (π) filter, which attenuates conducted differential-mode EMI noise.

The secondary side of the transformer is rectified by D3, an ultra fast recovery diode and filtered by C5.

The feedback resistors (R5 and R6) were selected using standard 1% resistor values for closed LED current. The feedback resistors should be placed directly at the FB pin of the PT4213A device to minimize noise.

R1, R2 and C4 forms the start up circuit; the auxiliary winding is powered PT4213A through R7 and D2. C4 should be place as close as possible to the VCC and GND pins.

R4 is the primary side current sense resistor, should be placed close to CS pin of PT4213A.

REFERENCE CIRCUIT FOR DRIVING 5X1W LED

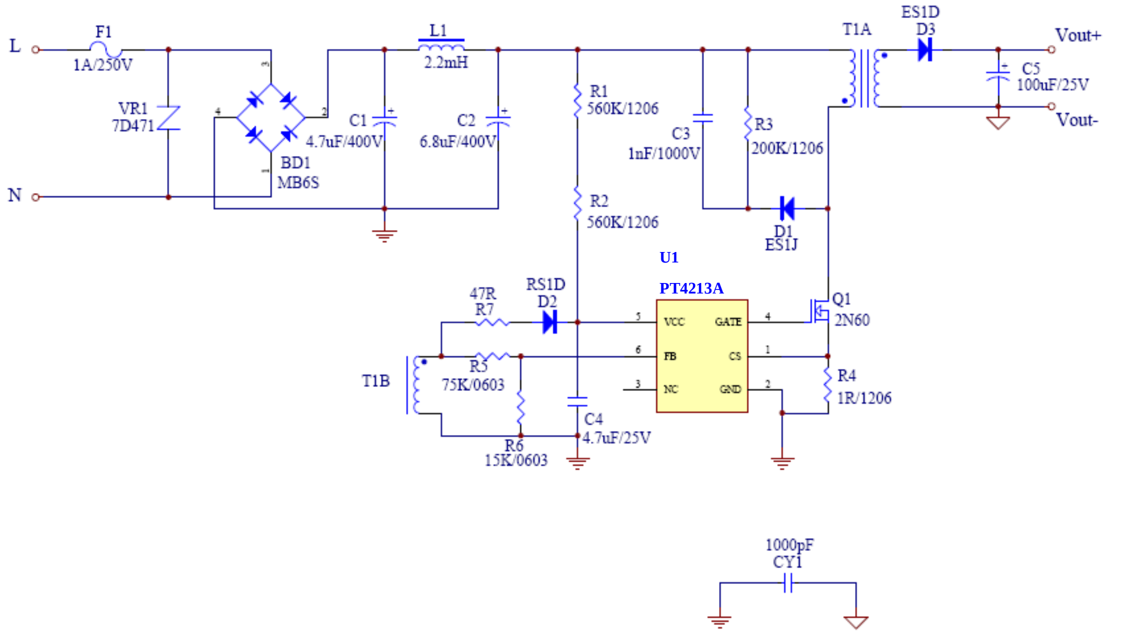
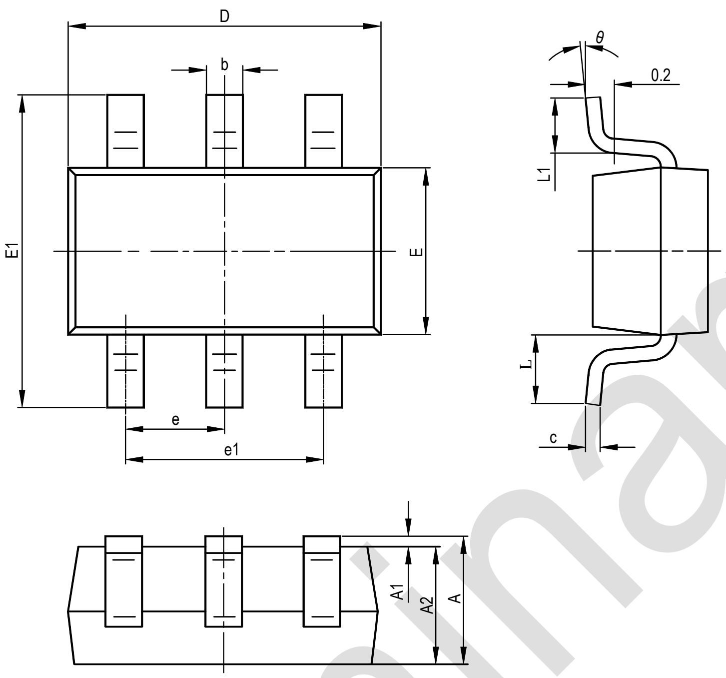


Figure 4

PACKAGE INFORMATION

SOT23-6



SYMBOL	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	1.050	1.250	0.041	0.049
A1	0.000	0.100	0.000	0.004
A2	1.050	1.150	0.041	0.045
b	0.300	0.400	0.012	0.016
c	0.100	0.200	0.004	0.008
D	2.820	3.020	0.111	0.119
E	1.500	1.700	0.059	0.067
E1	2.650	2.950	0.104	0.116
e	0.950TYP		0.037TYP	
e1	1.800	2.000	0.071	0.079
L	0.700REF		0.028REF	
L1	0.300	0.600	0.012	0.024
θ	0°	8°	0°	8°