

Application Note AN4143

LCD Backlight Inverter Drive IC (FAN7310)

1. Description

Design goals for a CCFL inverter used in a notebook computer or other portable applications include small size, high efficiency, and low cost. The FAN7310 provides the necessary circuit blocks to implement a highly efficient CCFL backlight power supply in a small footprint 20-SSOP package. The FAN7310 typically consumes less than 4mA of operating current, improving overall system efficiency. External parts count is minimized and system cost is reduced by the integration of features as a feedback-controlled PWM driver stage, soft start, open lamp regulation, and protection, UVLO, and a synchronization circuit with an external signal. The FAN7310 includes an internal shunt regulator, allowing it to operate with an input voltage from 5 to 24V. It supports analog and burst dimming modes of operation.

The FAN7310 provides open lamp regulation and protection. Open lamp regulation protects the transformer from over-voltage during start up or when an open lamp occurs. The transformer voltage is regulated by reducing duty cycle when an over-voltage is detected. Open Lamp protection can be used to shut down an IC when an open lamp condition continues for more than a specific time.

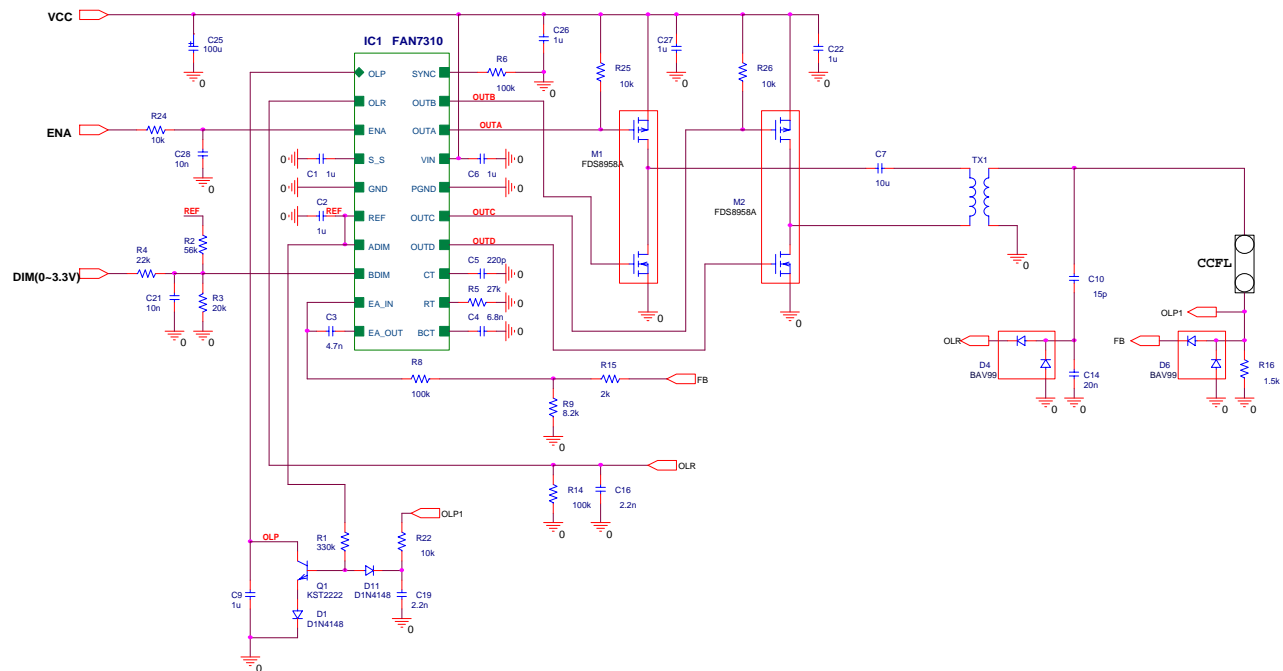


Figure 1-1. Application Circuit

2. Block Diagram and Basic Operation

2-1. Block Diagram

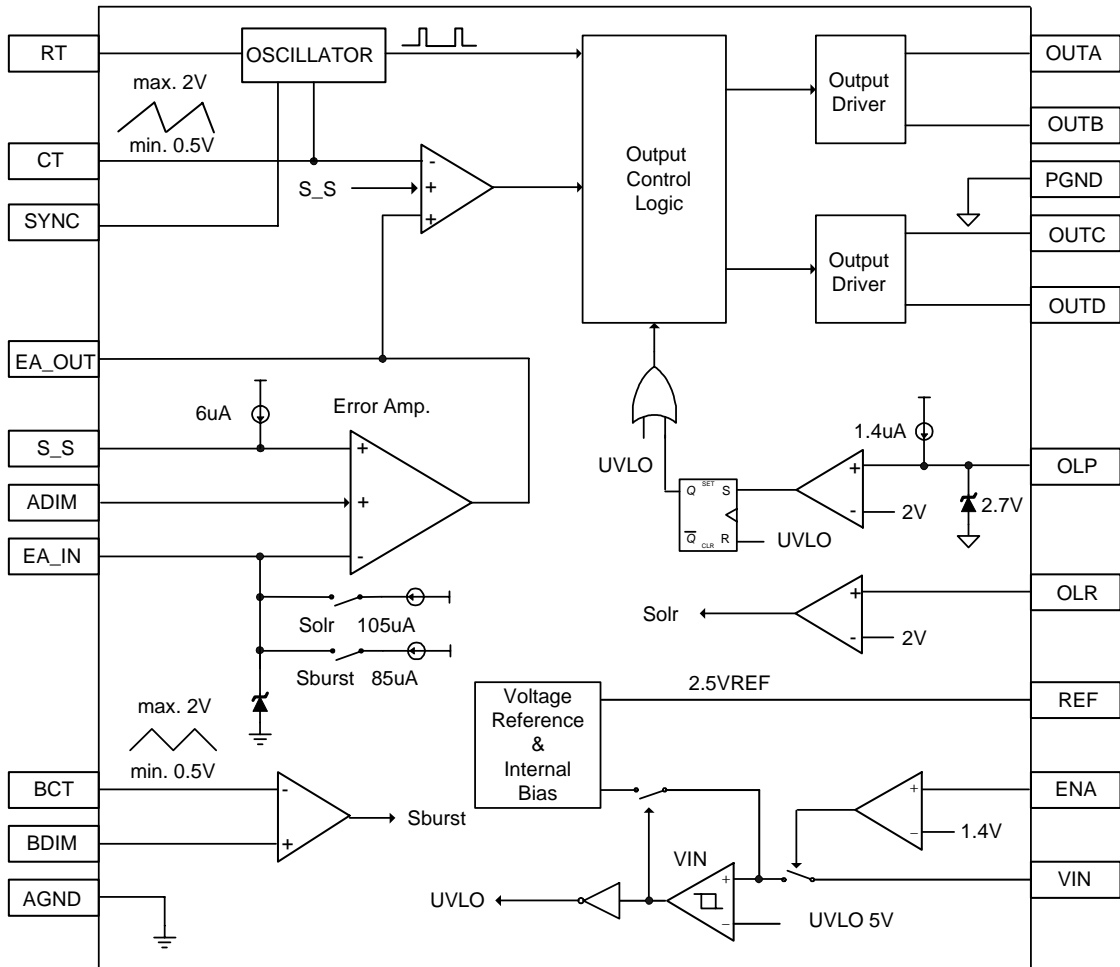


Figure 2-1. Block Diagram

2.2 Under Voltage Lockout (UVLO)

The under voltage lockout(UVLO) circuit guarantees stable operation of the IC's control circuit by stopping and starting it as a function of the VIN value. The UVLO circuit turns on the control circuit when VIN exceeds 5V. When VIN is lower than 5V, the IC's standby current is less than 200uA.

2.3 ENA

Applying voltage higher than 2V to ENA pin enables the operation of the IC. Applying voltage lower than 0.7V to ENA pin disables the operation of the inverter.

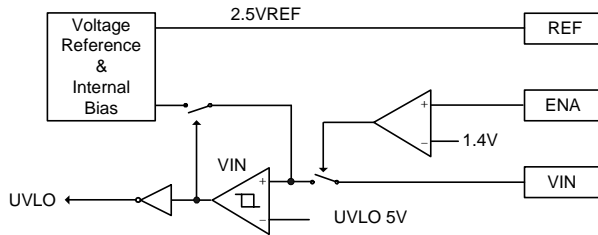


Figure 2-2. Under Voltage Lockout and ENA Circuits

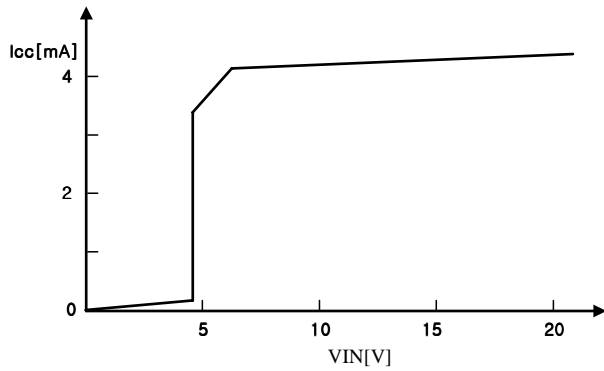


Figure 2-3. Start Voltage & Operating Current

2.4 Soft Start

The soft start function is provided that S_S pin is connected through a capacitor to GND. A soft start circuit ensures a gradual increase in the input and output power. The capacitor connected to S_S pin determines the rate of rise of the duty ratio. It is charged by a current source of 6uA.

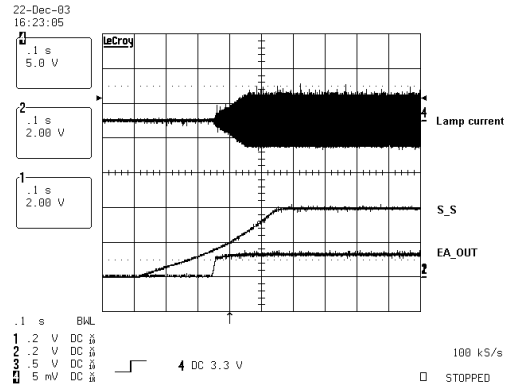
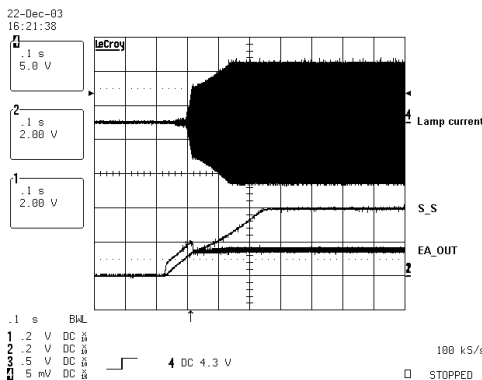


Figure 2-4. Soft Start During Initial Operation

2.5 Oscillator

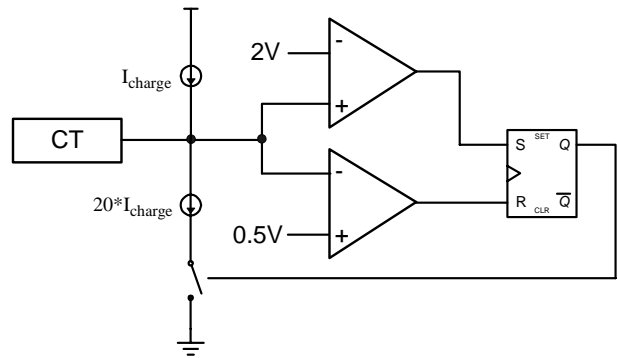
1) Main Oscillator

Timing capacitor CT is charged by the reference current source, formed by the timing resistor RT whose voltage is regulated at 1.25V. The sawtooth waveform of the Main Oscillator circuit charges up to 2V. Then the capacitor begins discharging down to 0.5V. The capacitor starts charging again and a new switching cycle begins.

$$I_{charge} = \frac{3}{4} \frac{1.25}{R_T} \tag{2.1}$$

The main frequency can be programmed by adjusting the values of RT and CT. The main frequency can be calculated as below.

$$f_{op} = \frac{19}{32 R_T C_T} \tag{2.2}$$



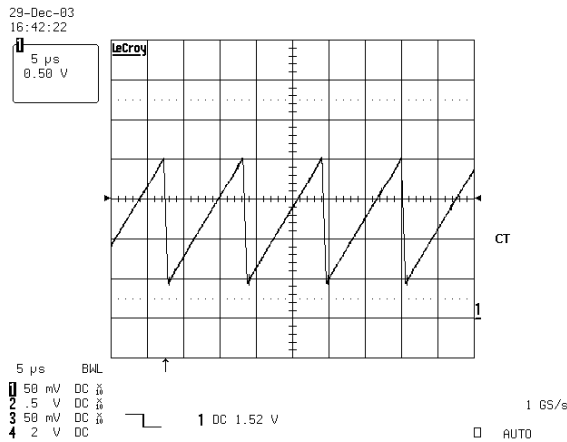


Figure 2.5 Main Oscillator Circuit & Waveform

2) Burst dimming oscillator

Burst dimming timing capacitor BCT is charged by the reference current source, formed by the timing resistor RT whose voltage is regulated at 1.25V. The sawtooth waveform charges up to 2V. Once reached, the capacitor begins discharging down to 0.5V, and then starts charging again and a new switching cycle begins.

$$I_{charge} = \frac{3}{32} \frac{1.25}{R_T} \tag{2.3}$$

The burst dimming frequency can be programmed by adjusting the values of RT and BCT. The burst dimming frequency can be calculated as below.

$$f_{burst} = \frac{3.75}{64 R_T BCT} \tag{2.4}$$

The burst dimming frequency should be greater than 120Hz to avoid visible flicker. To compare the input of BDIM pin with the 0.5~2V triangular wave of burst oscillator makes the PWM pulse for burst dimming. The PWM pulse controls EA_OUT voltage by summing 85uA into EA_IN pin. Figure 2.6 shows burst dimming oscillator circuit and waveform.

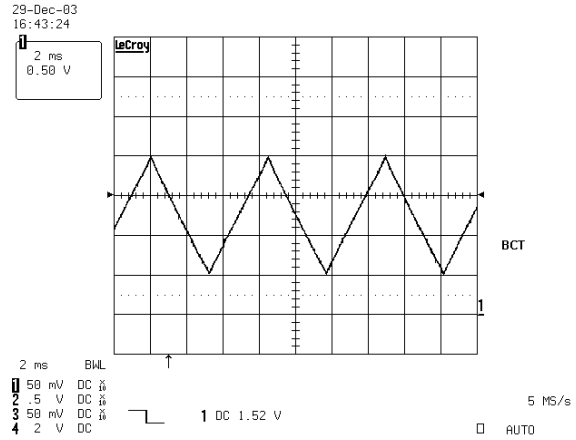
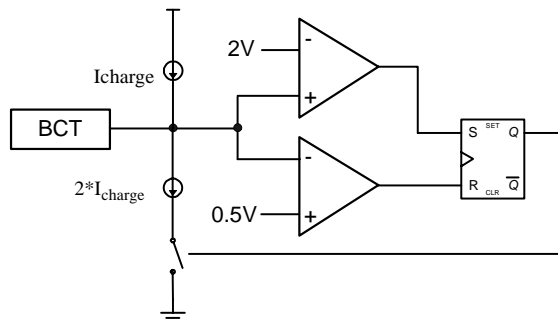


Figure 2.6 Burst Oscillator Circuit & Waveform

2.6 Analog Dimming

For analog dimming the lamp intensity is controlled with the ADIM signal. 2.5V on ADIM commands full brightness. Analog dimming waveforms are shown in Figures 2.7 and 2.8.

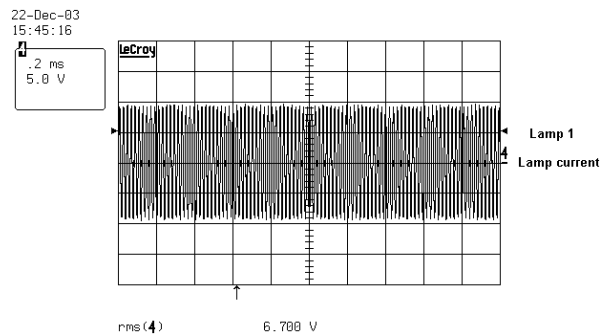


Figure 2.7 Analog Dimming at Maximum

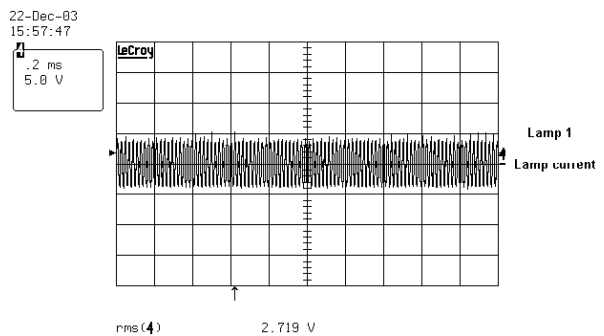


Figure 2.8 Analog Dimming at Minimum

Setting lamp current sensing resistors

1) Positive Polarity Analog Dimming

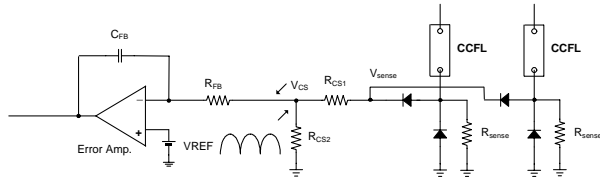


Figure 2.9 Calculating value of the analog dimming circuit parameter

Lamp current is sensed at R_{sense} and the sensed voltage is divided by R_{CS1} and R_{CS2} and is averaged at Error Amp. by R_{FB} and C_{FB} .

$$R_{sense_eq} = \frac{1 + \frac{V_{DF}}{I_{Lamp} \cdot (R_{CS1} + R_{CS2})}}{R} \approx R_{sense} // (R_{CS1} + R_{CS2}),$$

V_{DF} is diode forward voltage

$$V_{sense} = \left(\frac{1}{\pi} \int_0^{\pi} \sqrt{2} \cdot I_{Lamp} \cdot R_{sense_eq} \cdot \sin\theta \cdot d\theta \right) - V_{DF}$$

$$= \frac{2}{\pi} \cdot \sqrt{2} \cdot I_{Lamp} \cdot R_{sense_eq} - V_{DF} \quad (1)$$

$$V_{cs} = V_{sense} \cdot \frac{R_{CS2}}{R_{CS1} + R_{CS2}} = \left(\frac{2}{\pi} \cdot \sqrt{2} \cdot I_{Lamp} \cdot R_{sense} - V_{DF} \right) \cdot \frac{R_{CS2}}{R_{CS1} + R_{CS2}} \quad (2.5)$$

Equation (2.5) assumes that the error amplifier loop is closed. The relationship between V_{CS} and V_{ref} is given in equation (2.6)

$$V_{ref} = V_{CS} = V_{sense} \cdot \frac{R_{CS2}}{R_{CS1} + R_{CS2}}$$

$$= \left(\frac{2}{\pi} \cdot \sqrt{2} \cdot I_{Lamp} \cdot R_{sense} - V_{DF} \right) \cdot \frac{R_{CS2}}{R_{CS1} + R_{CS2}} \quad (2.6)$$

$$\frac{R_{CS1}}{R_{CS2}} = \frac{V_{sense}}{V_{ref}} - 1 \quad (2.7)$$

For example, suppose below.

$$V_{ref} = 2.5V, I_{Lamp} = 6.5mA, R_{sense} = 1K\Omega, R_{CS1} = 10K\Omega$$

From these values, we can get the approximate value of R_{CS2} . To get more precise value of R_{CS2} , you need iterative calculation. First, use R_{sense} to calculate R_{CS2} , because we don't know the R_{sense_eq} value. After the first calculation, we can get the value of R_{sense_eq} . Then use R_{sense_eq} to calculate R_{CS2} value. Calculate iteratively until the previous R_{sense_eq} value becomes almost equal to the current R_{sense_eq} value.

The data to input		
The calculated data		
V_{REF}	2.5	
I_{Lamp}	6.5	
R_{sense}	1	K ohm
R_{sense_eff}	0.95	K ohm
Diode drop voltage	0.3	V
V_{sense}	5.259453252	V
R_{CS1}/R_{CS2}	1.103781301	
R_{CS1}	10	K ohm
R_{CS2}	9.059765727	K ohm

$R_{sense_effective} = 0.950148969 \text{ K ohm} = R_{sense} // (R_{CS1} + R_{CS2})$

2) Negative Polarity Analog Dimming

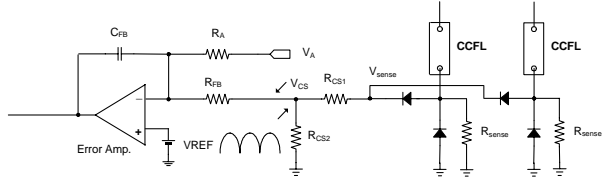


Figure. 2.10 Calculating value of the analog dimming inverting circuit parameter

Lamp current is sensed at R_{sense} and the sensed voltage is divided by R_{CS1} and R_{CS2} and is averaged at Error Amp. by R_{FB} and C_{FB} .

$$R_{sense_eq} = \frac{1 + \frac{V_{DF}}{I_{Lamp} \cdot (R_{CS1} + R_{CS2})}}{R} \approx R_{sense} // (R_{CS1} + R_{CS2}),$$

V_{DF} is diode forward voltage

$$V_{sense} = \left(\frac{1}{\pi} \int_0^{\pi} \sqrt{2} \cdot I_{Lamp} \cdot R_{sense_eq} \cdot \sin\theta \cdot d\theta \right) - V_{DF}$$

$$= \frac{2}{\pi} \cdot \sqrt{2} \cdot I_{Lamp} \cdot R_{sense_eq} - V_{DF} \quad (1)$$

$$V_{cs} = V_{sense} \cdot \frac{R_{CS2}}{R_{CS1} + R_{CS2}} = \left(\frac{2}{\pi} \cdot \sqrt{2} \cdot I_{Lamp} \cdot R_{sense} - V_{DF} \right) \cdot \frac{R_{CS2}}{R_{CS1} + R_{CS2}} \quad (2.8)$$

Equation (2.8) assumes the error amplifier loop is closed. The relationship between V_{CS} and V_A (dimming control voltage) is given in equation (2.9)

$$V_{ref} = \frac{V_A \cdot R_{FB} + V_{CS} \cdot R_A}{R_{FB} + R_A} \quad (2.9)$$

You can program the relationship between dimming control voltage and lamp current can be easily programmed for the application. For example, suppose:

$$V_{Amin.} = 0, I_{Lamp,max} = 7mA \quad (2.10)$$

$$V_{Amax.} = 3.3, I_{Lamp,min} = 3mA \quad (2.11)$$

$$I_{Lamp,min} = \alpha \cdot I_{Lamp,max} \quad (2.12)$$

Substituting for V_A and V_{CS} in equation (2.9) from equation (2.10) results in

$$V_{ref} = \frac{V_{CSmax} \cdot R_A}{R_{FB} + R_A} \quad (2.13)$$

Substituting for V_A and V_{CS} in equation (2.9) from equation (2.11) results in

$$V_{ref} = \frac{V_{Amax} \cdot R_{FB} + V_{CSmin} \cdot R_A}{R_{FB} + R_A} = \frac{V_{Amax} \cdot R_{FB} + \alpha \cdot V_{CSmax} \cdot R_A}{R_{FB} + R_A} \quad (2.14)$$

Multiplying equation (2.13) by $R_{FB}+R_A$ gives

$$V_{ref} \cdot R_{FB} + V_{ref} \cdot R_A = V_{CSmax} \cdot R_A \quad (2.15)$$

Multiplying equation (2.14) by $R_{FB}+R_A$ gives

$$V_{ref} \cdot R_{FB} + V_{ref} \cdot R_A = V_{Amax} \cdot R_{FB} + \alpha \cdot V_{CSmax} \cdot R_A \quad (2.16)$$

$$(V_{ref} - V_{Amax}) \cdot R_{FB} + V_{ref} \cdot R_A = \alpha \cdot V_{CSmax} \cdot R_A \quad (2.17)$$

Multiplying equation (2.15) by α gives

$$\alpha \cdot V_{ref} \cdot R_{FB} + \alpha \cdot V_{ref} \cdot R_A = \alpha \cdot V_{CSmax} \cdot R_A \quad (2.18)$$

Subtracting equation (2.17) from equation (2.18) gives

$$(V_{Amax} + V_{ref}(\alpha - 1)) \cdot R_{FB} + V_{ref}(\alpha - 1) \cdot R_A = 0 \quad (2.19)$$

$$(V_{Amax} + V_{ref}(\alpha - 1)) \cdot R_{FB} = V_{ref}(1 - \alpha) \cdot R_A \quad (2.20)$$

Equation (2.20) can be rewritten as

$$R_A = \frac{(V_{Amax} + V_{ref}(\alpha - 1)) \cdot R_{FB}}{V_{ref}(1 - \alpha)} = \beta \cdot R_{FB} \quad (2.21)$$

R_A is calculated by selecting R_{FB} and solving equation (2.21). Substituting for R_A in equation (2.13) from equation (2.21) and rewriting gives

$$\begin{aligned} V_{CSmax} &= \frac{V_{ref} \cdot (R_{FB} + R_A)}{R_A} = \frac{V_{ref} \cdot (1 + \beta) \cdot R_{FB}}{\beta \cdot R_{FB}} \\ &= \frac{V_{ref} \cdot (1 + \beta)}{\beta} = V_{ref} \cdot \left(1 + \frac{1}{\beta}\right) \\ &= V_{sense} \cdot \frac{R_{CS2}}{R_{CS1} + R_{CS2}} \\ &= \left(\frac{2}{\pi} \cdot \sqrt{2} \cdot I_{Lamp} \cdot R_{sense} - V_{DF}\right) \cdot \frac{R_{CS2}}{R_{CS1} + R_{CS2}} \quad (2.22) \\ \frac{R_{CS1}}{R_{CS2}} &= \frac{V_{sense}}{V_{ref} \cdot \left(1 + \frac{1}{\beta}\right)} - 1 \quad (2.23) \end{aligned}$$

Forexample, suppose below.

$$\begin{aligned} V_{ref} &= 2.5V, I_{Lampmax} = 6.7mA, I_{Lampmin} = 4mA, \\ R_{FB} &= 100K\Omega, R_{sense} = 1.5K\Omega, R_{CS1} = 10K\Omega \end{aligned}$$

From these values, we can obtain the value of R_{CS2} . To get more precise value of R_{CS2} , you need iterative calculation.

First, use R_{sense} to calculate R_{CS2} , because we don't know R_{sense_eq} value. After first calculation, we can get R_{sense_eq} . Calculate R_{CS2} value using R_{sense_eq} . Calculate iteratively until the previous R_{sense_eq} value becomes almost equal to the current R_{sense_eq} value.

The data to input		The calculated data	
Min.	-	V_{REF}	0
Typ.	2.5	V_A	3.2
Max.	-	I_{lamp}	6.7
		α	0.597014925
		$I_{lamp_max}/I_{lamp_min}$	
R_{FB}	100	K ohm	
R_A	217.6296296	K ohm	
β	2.176296296		
V_{CSmax}	3.64874064	K ohm	
R_{sense}	1.5	K ohm	
R_{sense_eff}	1.3861		
Diode drop voltage	0.3	V	
Avg. max. V_{sense}	8.061120587	V	
R_{CS1}/R_{CS2}	1.209288459		
R_{CS1}	10	K ohm	
R_{CS2}	8.269325588	K ohm	
		$R_{sense_effective}$	1.386187316
		$=R_{sense}/(R_{CS1}+R_{CS2})$	

2.7 Burst Dimming

Lamp intensity is controlled with the signal BDIM. 0V on BDIM commands full brightness. The duty cycle of the burst dimming comparator determines the lamp brightness as a percent of the rated lamp current. Burst dimming is implemented by summing 85uA into the feedback node to turn the lamp. If there is sufficient voltage for lamp to strike, the feedback loop controls the lamp at rated current using a fixed current sense resistor. When the voltage of EA_IN is brought higher than V_{ref} , EA_OUT becomes low and the MOSFET stops switching. Then the resonant tank voltage decays until the lamp extinguishes. C_{FB} is reduced, if possible, to speed up the lamp re-strike. Burst dimming waveforms are shown in Figures 2.11, 2.12, and 2.13.

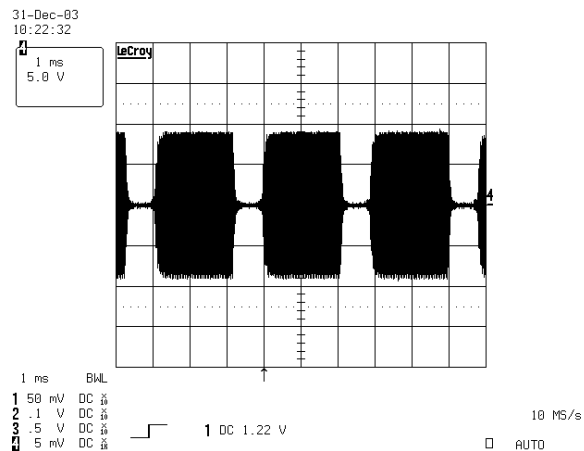


Figure 2.11 Burst Dimming at 75%

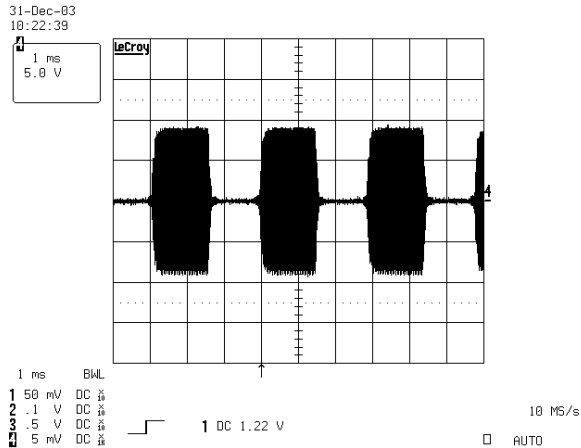


Figure 2.12 Burst Dimming at 50%

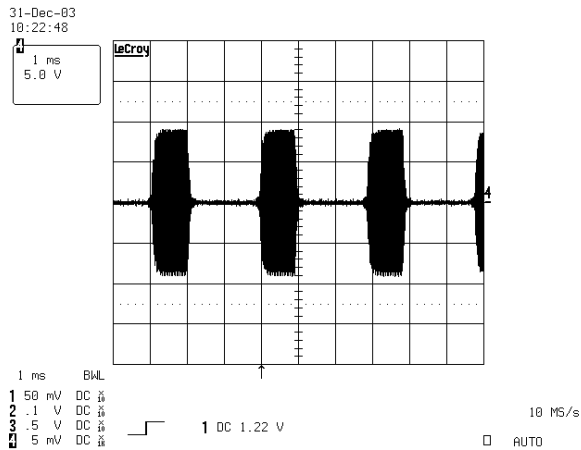


Figure 2.13 Burst Dimming at 25%

2.8 Open Lamp Regulation & Open Lamp Protection

You must suspend power stage operation if an open lamp occurs, because the power stage has high gain. When a voltage higher than 2V is applied to the OLR (Open Lamp Regulation) pin, the part enters the regulation mode and controls EA_OUT voltage to limit the lamp voltage by adding 105uA into the feedback node. The OLP (Open Lamp Protection) capacitor which is connected to the OLP pin, is charged by the 1.4uA internal current source. After reaching 2V, the IC shuts down where all the output is high. Using the approximation $I=C\Delta V/\Delta T$, where $I=1.4\mu A$, $\Delta V=2V$, we have the following approximate relationship between the OLP capacitor and the time ΔT before the IC shuts down. $\Delta T(s)=1.43C(\mu F)$

1) How to operate OLP

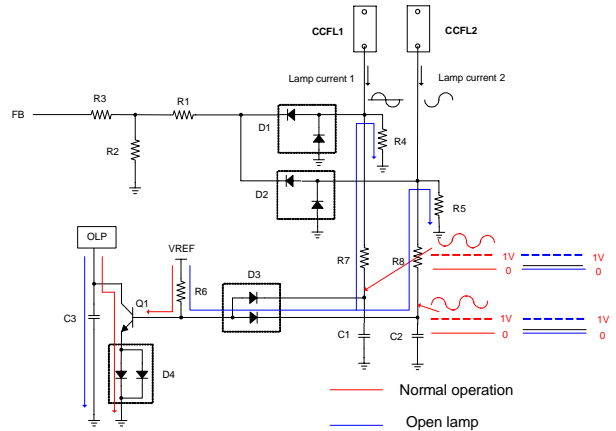


Figure 2.14 Operating OLP

In normal operation, the voltage of D3's cathode is over 1V and D3 is turned off. Q1 is on and OLP remains low. When open lamp occurs, the voltage of D3's cathode is under 1V and either of D3 is turned on. Then Q1 is turned off-and OLP start charging by internal current source 1.4uA. If OLP reaches 2V, IC is shut down.

OLP has internal current source of 1.4uA. The base current of Q1 should be more than 1.4uA/hfe. R6 is determined from this condition. R4, R5, R7 and R8 are determined so that Q1 and D4 are turned off in open lamp condition. C1 and C2 are determined so that the voltage of D8's cathode is over 1V in normal operation.

2) How to operate OLR

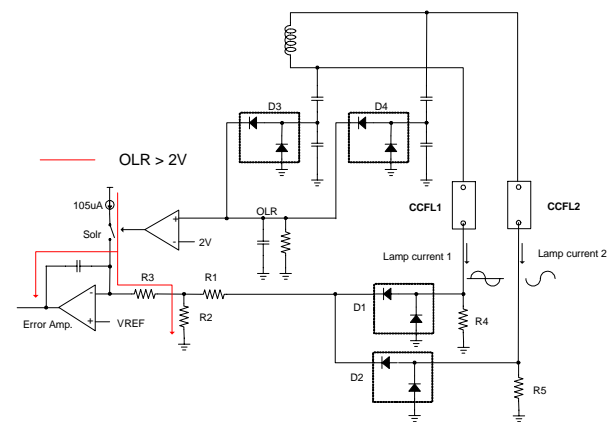


Figure 2.15 Operating OLR

OLR>2V \uparrow \rightarrow Solr on duty \uparrow \rightarrow EA_IN \uparrow \rightarrow EA_OUT \downarrow \rightarrow Switching duty \downarrow \rightarrow Lamp voltage \downarrow
 \rightarrow OLR>2V \downarrow \rightarrow Solr on duty \downarrow \rightarrow EA_IN \rightarrow EA_OUT \downarrow \rightarrow Switching duty \uparrow \rightarrow Lamp voltage \uparrow .

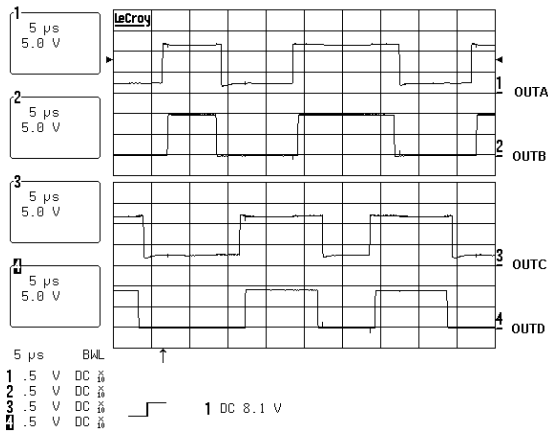


Figure 2.17 New Phase Shift Control Waveforms

2.10 SYNC

The SYNC pin is used for frequency synchronization. Used as an output, SYNC provides a clock signal. As an input, this pin provides a synchronization point. Multiple FAN7310s, each with their own local oscillator frequency, may be connected together by the SYNC pin to synchronize to the fastest oscillator. You can also use to synchronize the FAN7310 to an external clock, provided the frequency of the external signal is higher than the frequency of the local oscillator. An external resistor to GND should be used for stable operation of the SYNC pin is 1V.

The threshold voltage of the SYNC pin is 1V.

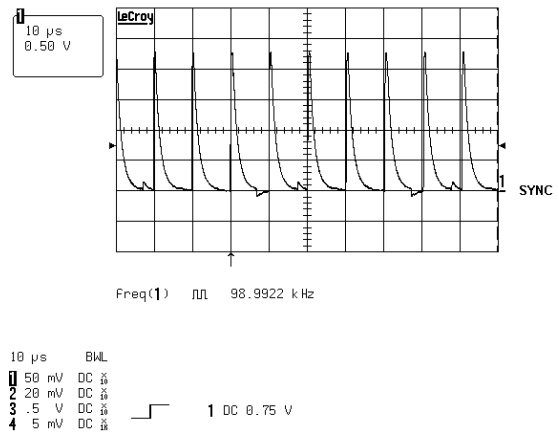


Figure 2.18 SYNC Pin Output Waveform

2.11 PCB Layout Guideline

1. Separate grounding for analog and power portions of circuitry is one of the simplest and most effective methods of noise suppression as shown in Figure 2.19.
2. The traces between drive output and the MOSFET gates should be as short as possible and as wide as possible.
3. The traces of RT, CT and BCT should be kept away from high current components and traces.

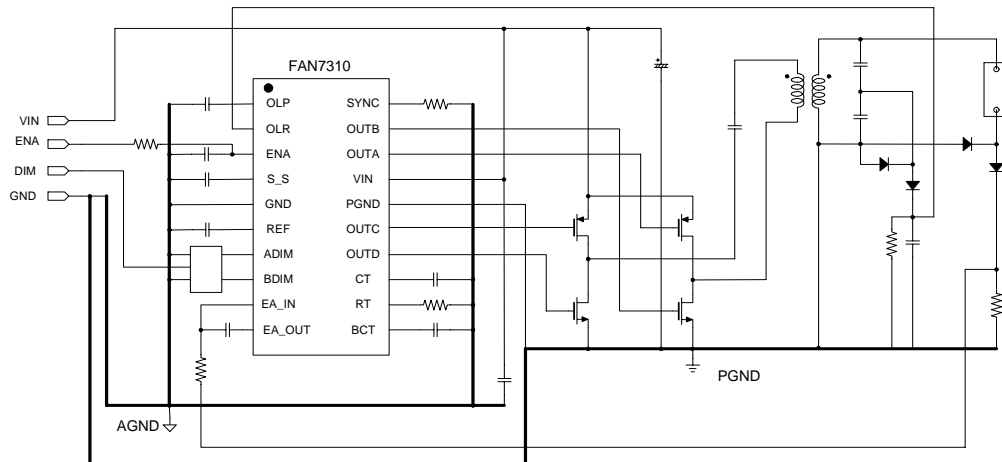


Figure. 2.19

3. Power Stage Design

3.1 Resonant Circuit

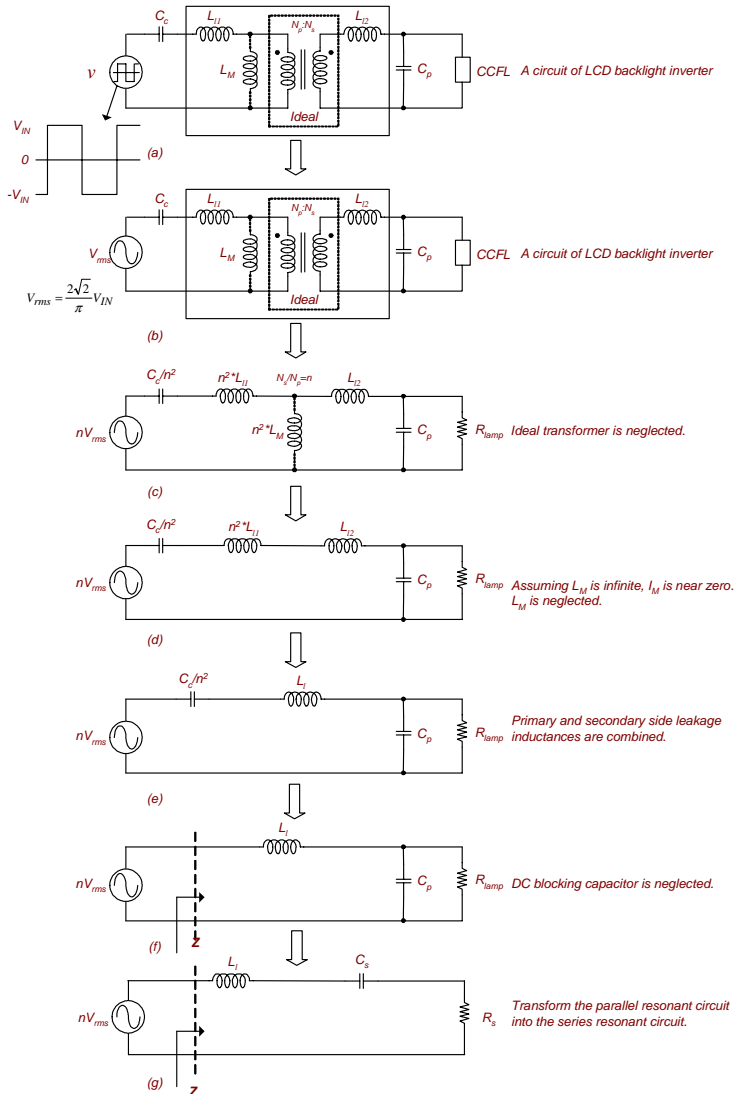


Figure 3.1 Circuit of LCD Backlight Inverter

The resonant circuit in Fig 3.1(f) is a second-order low-pass filter and can be described by the following normalized parameters:

- The corner frequency

$$\omega_o = \frac{1}{\sqrt{L_1 C_p}} \quad (3.1)$$

- The characteristic impedance

$$Z_o = \omega_o L_1 = \frac{1}{\omega_o C_p} = \sqrt{\frac{L_1}{C_p}} \quad (3.2)$$

- The loaded quality factor at the corner frequency f_o

$$Q_L = \omega_o C_p R_{lamp} = \frac{R_{lamp}}{\omega_o L_1} = \frac{R_{lamp}}{Z_o} \quad (3.3)$$

- The resonant frequency that forms the boundary between capacitive and inductive loads

$$\omega_r = \frac{1}{\sqrt{L_1 C_s}} \quad (3.4)$$

- The loaded quality factor at the resonant frequency f_r

$$Q_r = \frac{\omega_r L_1}{R_s} = \frac{1}{\omega_r C_s R_s} \quad (3.5)$$

The input impedance of the resonant circuit in Fig 3.1(f) is

$$Z = j\omega L + \frac{R_{lamp} \cdot \frac{1}{j\omega C_p}}{R_{lamp} + \frac{1}{j\omega C_p}} = \frac{R_{lamp} \left[1 - \left(\frac{\omega}{\omega_o}\right)^2 + j \frac{1}{Q_L} \left(\frac{\omega}{\omega_o}\right) \right]}{1 + j Q_L \left(\frac{\omega}{\omega_o}\right)} \quad (3.6)$$

,where

$$\frac{Z}{Z_o} = \frac{Q_L^2 \left[1 - \left(\frac{\omega}{\omega_o}\right)^2 \right]^2 + \left(\frac{\omega}{\omega_o}\right)^2}{1 + \left(Q_L \frac{\omega}{\omega_o}\right)^2} \quad (3.7)$$

$$\varphi = \arctan \left\{ Q_L \left(\frac{\omega}{\omega_o}\right) \left[\left(\frac{\omega}{\omega_o}\right)^2 + \frac{1}{Q_L^2} - 1 \right] \right\} \quad (3.8)$$

$$R_s = Z \cos \varphi \quad (3.9)$$

$$X_s = Z \sin \varphi \quad (3.10)$$

The resonant frequency f_r is defined as a frequency at which the phase shift is zero. The ratio of the resonant frequency f_r to the corner frequency f_o is

$$\frac{f_r}{f_o} = \sqrt{1 - \frac{1}{Q_L^2}}, \quad \text{for } Q_L \geq 1 \quad (3.11)$$

The loaded quality factors Q_L and Q_r are related by

$$Q_r = \frac{\omega_r L}{R_s} = \frac{1}{\omega_r C_s R_s} = \omega_r C_p R_{lamp} \\ = Q_L \left(\frac{\omega_r}{\omega_o}\right) = \sqrt{Q_L^2 - 1}, \quad \text{for } Q_L \geq 1 \quad (3.12)$$

3.2 Voltage Transfer Function

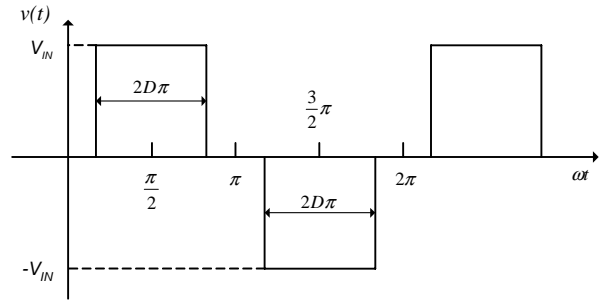


Figure 3.2 Input Voltage of the Resonant Circuit

As shown in Fig 3.2, the input voltage of the resonant circuit v is a square wave of magnitude V_{IN} , given by

$$v = 0, \quad \text{for } 0 < \omega t \leq \left(\frac{1}{2} - D\right)\pi \\ v = V_{IN}, \quad \text{for } \left(\frac{1}{2} - D\right)\pi < \omega t \leq \left(\frac{1}{2} + D\right)\pi \\ v = 0, \quad \text{for } \left(\frac{1}{2} + D\right)\pi < \omega t \leq \left(\frac{3}{2} - D\right)\pi \\ v = -V_{IN}, \quad \text{for } \left(\frac{3}{2} - D\right)\pi < \omega t \leq \left(\frac{3}{2} + D\right)\pi \\ v = 0, \quad \text{for } \left(\frac{3}{2} + D\right)\pi < \omega t \leq 2\pi \quad (3.13)$$

The fundamental component of this voltage is

$$v_{i1} = V_m \sin(\omega t), \quad (3.14)$$

in which the amplitude of v_{i1} can be found from Fourier analysis as

$$V_m = \frac{4}{\pi} V_{IN} \sin D\pi \quad (3.15)$$

Hence, you can obtain the rms value of v_{i1}

$$V_{rms} = \frac{V_m}{\sqrt{2}} = \frac{2\sqrt{2}}{\pi} V_{IN} \sin D\pi \quad (3.16)$$

Which leads to the voltage transfer function from V_{IN} to the fundamental component at the input of the resonant circuit

$$M_{Vs} = \frac{V_{rms}}{V_{IN}} = \frac{2\sqrt{2}}{\pi} \sin D\pi \quad (3.17)$$

According to Fig 3.1(f), the voltage transfer function of the resonant circuit is

$$M_{Vr} = \frac{V_{Ri}}{\sqrt{2} \cdot nV_{rms}} = \frac{\frac{R_{lamp}}{j\omega C_p}}{R_{lamp} + \frac{1}{j\omega C_p}} \cdot \frac{R_{lamp}}{j\omega L + \frac{R_{lamp}}{j\omega C_p} + \frac{1}{j\omega C_p}} \quad (3.18)$$

$$= \frac{1}{1 - \left(\frac{\omega}{\omega_0}\right)^2 + j\frac{1}{Q_L}\left(\frac{\omega}{\omega_0}\right)} = M_{Vr}^{e^{j\phi}} \quad (3.18)$$

where

$$M_{Vr} = \frac{V_{Ri}}{nV_{rms}} = \frac{1}{\sqrt{\left[1 - \left(\frac{\omega}{\omega_0}\right)^2\right]^2 + \frac{1}{Q_L^2}\left(\frac{\omega}{\omega_0}\right)^2}} \quad (3.19)$$

$$\phi = -\arctan \left[\frac{\frac{1}{Q_L}\left(\frac{\omega}{\omega_0}\right)}{1 - \left(\frac{\omega}{\omega_0}\right)^2} \right] \quad (3.20)$$

The maximum value of M_{Vr} is obtained by differentiating the quantity under the square-root sign with respect to f/f_0 and setting the result equal to zero. Hence, the normalized peak frequency is

$$\frac{f_{pk}}{f_0} = 0, \quad \text{for } 0 \leq Q_L \leq \frac{1}{\sqrt{2}} \quad (3.21)$$

$$\frac{f_{pk}}{f_0} = \sqrt{1 - \frac{1}{2Q_L^2}}, \quad \text{for } Q_L \geq \frac{1}{\sqrt{2}}$$

, resulting in the maximum magnitude of the voltage transfer function of the resonant circuit

$$M_{Vr(max)} = 1, \quad \text{for } 0 \leq Q_L \leq \frac{1}{\sqrt{2}}$$

$$M_{Vr(max)} = \frac{Q_L}{\sqrt{1 - \frac{1}{4Q_L^2}}}, \quad \text{for } Q_L \geq \frac{1}{\sqrt{2}} \quad (3.22)$$

The magnitude of the dc-to-ac voltage transfer function of the LCD backlight inverter without losses is obtained from (3.17) and (3.22)

$$M_{VI} = \frac{V_{lamp}}{V_{IN}} = M_{Vs} \cdot (nM_{Vr})$$

$$= \frac{2\sqrt{2} \cdot n \sin D\pi}{\pi \sqrt{\left[1 - \left(\frac{\omega}{\omega_0}\right)^2\right]^2 + \frac{1}{Q_L^2}\left(\frac{\omega}{\omega_0}\right)^2}} \quad (3.23)$$

The maximum magnitude of the dc-to-ac voltage transfer function of the LCD backlight inverter without losses is

$$M_{VI(max)} = \frac{2\sqrt{2} \cdot n}{\pi}, \quad \text{for } 0 \leq Q_L \leq \frac{1}{\sqrt{2}}$$

$$M_{VI(max)} = \frac{2\sqrt{2} \cdot nQ_L}{\pi \sqrt{1 - \frac{1}{4Q_L^2}}}, \quad \text{for } Q_L \geq \frac{1}{\sqrt{2}} \quad (3.24)$$

3.2 Design Procedure

A LCD monitor backlight circuit is presented here to illustrate a design based on the FAN7310. The inverter is designed to drive two CCFLs with the following specifications.

Panel Model	LM151X2(LG.PHILIPS LCD)
Input Voltage	9 ~ 15V
Striking Voltage	880V _{rms}
Operating Voltage	585V _{rms} (TYP.)
Operating Current	8mA _{rms} (TYP.)
Operating Frequency	50kHz(TYP.)
Rated Power	4.68W/CCFL
Efficiency	85%(TYP.)

1) Select Transformer's Primary Turns

The number of primary turns is determined by Faraday's law. $N_{p,min}$ is fixed by the minimum voltage across the primary and the maximum on-time.

$$N_{p,min} = \frac{V_{IN,min} \cdot \Delta t_{max}}{\Delta B \cdot A_e}$$

, where $N_{p,min}$ = Minimum number of primary turns

$V_{IN,min}$ = Minimum input voltage (Volts)

ΔB = Core magnetic flux density change (Tesla)

Δt_{max} = Maximum overlap on-time of diagonal MOSFET switches (us)

A_e = Core cross-sectional area (mm²)

A transformer used in a full-bridge topology operates in two quadrants of the B-H curve such that the maximum magnetic flux density $B_{max} = 0.5\Delta B$. For most cores, saturation magnetic flux density is about 400mT. Margin considered, determine that the maximum magnetic flux density $B_{max} = 0.5 B_{sat}$. So the maximum magnetic flux density $B_{max} = 200mT$. In an example with a minimum voltage of 9V, operating frequency 50KHz, maximum on-time of diagonal MOSFET switches of 10us and a core cross-sectional area (EPC17, EPC19, EFD1820) of 22mm², the minimum number of primary turns required is

$$N_{p, \min} = \frac{V_{IN, \min} \cdot \Delta t_{\max}}{\Delta B \cdot A_e} = \frac{9 \cdot 10}{400 \cdot 22} \approx 10T_s$$

2) Select Q_L & Operation Frequency and Determine the Turns Ratio

Select a value of 1 for Q_L . Assume that $f_{op}=f_{pk}=50\text{kHz}$ based on the LCD panel specification. From (3.21), the corner frequency is

$$f_o = \frac{f_{pk}}{\sqrt{1 - \frac{1}{2Q_L^2}}} = \frac{50}{\sqrt{1 - \frac{1}{2 \cdot 1.1^2}}} = 70.7(\text{kHz})$$

and from (3.11), the resonant frequency that forms the boundary between capacitive and inductive loads

$$f_r = f_o \cdot \sqrt{1 - \frac{1}{Q_L^2}} = f_{pk} \cdot \frac{\sqrt{1 - \frac{1}{Q_L^2}}}{\sqrt{1 - \frac{1}{2Q_L^2}}} = 0(\text{kHz})$$

So, ZVS (zero-voltage switching) can be achieved at any operating frequency.

For our reference design, the required secondary lamp voltage is 585V, and the minimum voltage is 9V.

Therefore, from (3.23), the minimum number of the turns ratio is

$$n \geq \frac{585}{9} \cdot \frac{\pi \sqrt{\left[1 - \left(\frac{\omega_{op}}{\omega_o}\right)^2\right]^2 + \frac{1}{Q_L^2} \left(\frac{\omega_{op}}{\omega_o}\right)^2}}{2\sqrt{2} \sin D\pi} \approx 62.5T_s$$

$$\therefore M_{VI} = \frac{2\sqrt{2} \cdot n \sin D\pi}{\pi \sqrt{\left[1 - \left(\frac{\omega}{\omega_o}\right)^2\right]^2 + \frac{1}{Q_L^2} \left(\frac{\omega}{\omega_o}\right)^2}} \geq \frac{585}{9}$$

2) Determine The Required Output Capacitance

Using the above specifications, the equivalent resistance of a CCFL is

$$R_{lamp} = \frac{V_{lamp}}{I_{lamp}} = \frac{585}{0.008} \approx 75 \text{ (K}\Omega\text{)}$$

The corner frequency is 70.7KHz. Assume a parasitic capacitance per lamp of 10pF. Each parasitic capacitance is effectively in parallel with each of the output capacitors.

The output capacitor is

$$C_{out} = C_p - C_{para} = \frac{Q_L}{\omega_o R_{lamp}} - 10\text{pF} \approx 21\text{pF}$$

$$\therefore Q_L = \omega_o R_{lamp} C_p$$

Using (3.3), the value of the leakage inductance is

$$L_l = \frac{1}{\omega_o^2 C_p} \approx 164.6(\text{mH})$$

* Considering minimum primary turns, minimum turns ratio and leakage inductance, determine primary turns, turns ratio and the gap of core to get the required leakage inductance. For the sample design, the number of primary turns is $30T_s$ and that of the secondary turns is $2200T_s$. Turns ratio is 66.7.

3) Select the Proper wire Gauges for the Primary and Secondary Transformer Windings.

The approximate primary winding rms current I_p and approximate secondary winding rms current I_s are determined by the following equations.

$$I_p = \frac{\pi}{2\sqrt{2}} \cdot \frac{P_{lamp}}{\eta V_{IN}}$$

$$\therefore P_{lamp} = \eta V_{IN} I_{IN}, I_{IN} = \frac{2\sqrt{2}}{\pi} I_p$$

$$I_s = \sqrt{I_{Lamp}^2 + [2\pi f_{op} C_{out} V_{lamp}]^2}$$

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The values you must know or select initially

Parameter	Description	Typical Value	Units
V_{lamp}	Nominal Lamp Operating Voltage	585	V
I_{lamp}	Nominal Lamp Operating Current	8	mA
f_{op}	Operating Frequency	50	kHz
f_{pk}	Peak frequency	50	kHz
V_{in}	Input voltage	9	V
D	Duty ratio at input voltage	50	%
Q_L	Loaded factor at the corner frequency	1	
C_{para}	Parasitic capacitance	10	pF
A_e	Core cross-sectional area	22	mm ²
B_{sat}	Saturation magnetic flux density	0.4	T
$AL_{leakage}$	AL value of leakage inductance	22	nH/N ²

The values that are calculated

B_{max}	Maximum magnetic flux density	0.2	T
ΔB	Core magnetic flux density change	0.4	T
Δt_{max}	Maximum overlap on-time of diagonal switches	10	us
f_o	Corner frequency	70.71067812	kHz
f_r	Resonant frequency	0	kHz
R_{lamp}	Equivalent resistance of a CCFL	73.125	kΩ
$N_{p,min}$	The minimum number of transformer's primary turns	10	Turns
n_{min}	The minimum number of the turns ratio	62.5	
C_{out}	The output capacitor	20.78	pF
L_l	The leakage inductance of the transformer	164.59	mH
N_p	The number of transformer's primary turns	31	Turns

Select the below values more than minmum turn ratio.

n	The number of the turns ratio	62.5	
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The values that are calculated

N_s	The number of transformrer's secondary turns	1934.1	Turns
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