

APPLICATION NOTES

Resonant Fluorescent Lamp Converter Provides Efficient and Compact Solution

Mark Jordan
Senior Design Engineer

John A. O'Connor
Applications Engineer

Unitrode Integrated Circuits
Merrimack, NH 03054

Abstract - This paper describes a zero voltage switched (ZVS) resonant converter for driving cold cathode fluorescent lamps. Primarily intended for liquid crystal display (LCD) back-lighting, the circuit features minimal component count and size. A specially designed integrated circuit provides all control functions for a current fed push-pull ZVS converter, and also contains an auxiliary pulse width modulated (PWM) controller to develop a programmable supply voltage for the LCD. Analysis and simulation of the converter, and a complete circuit schematic are presented. The analysis and simulation results are validated by experimental circuit waveforms and critical performance parameters.

Introduction

The proliferation of laptop and notebook computers places an ever increasing demand on display technology. High resolution and contrast are required to run today's graphics based programs, increasing the conflict between display performance, size and efficiency. The LCD with cold cathode fluorescent back lighting best satisfies this design requirement, however the lamp and its high voltage AC supply still remain the major contributor to battery drain.

The cold cathode fluorescent lamp (CCFL) requires 1-2 kV to fire. Sine wave drive is preferred to minimize RF interference and maximize lamp efficiency over time. Converter efficiency and size are extremely critical. These formidable requirements demand a highly efficient conversion topology and maximum circuit integration.

A zero voltage switched resonant topology will maximize efficiency by eliminating losses associated with charging parasitic capacitances to high voltages. This topology can be controlled

using discrete circuitry. The most common implementation is a Royer oscillator modified to provide ZVS operation. While this at first appears to be a good solution, and is commonly used today, it suffers from several limitations.

High voltage DC to AC conversion is only part of the display supply. The average output current must be programmable for lamp intensity control, and the LCD requires a programmable low voltage supply for contrast adjustment. This additional circuitry, implemented discretely or with multiple ICs results in a large number of components, significantly impacting size and reliability. Synchronization is also preferred to eliminate beat frequency effects such as lamp intensity modulation, further complicating the design. Minimizing circuit complexity and bulk are best achieved through integration.

Cold Cathode Lamp Characteristics

The CCFL presents a highly nonlinear load to the converter as illustrated in fig. 1. Initially when the lamp is cold (inoperative for some finite time), the voltage to fire the lamp is typically more than three times higher than the sustaining voltage. The lamp characterized in fig. 1 fires at 1600V and exhibits an average sustaining voltage (V_F) of 300V. Notice that the lamp initially exhibits a positive resistance and then transitions to a negative resistance above 1mA. These characteristics dictate a high output impedance (current source) drive to suppress the negative load resistance's effect and limit current during initial lamp firing. Since the ZVS converter has a low output impedance, an additional "lossless" series impedance such as a coupling capacitor must be added.

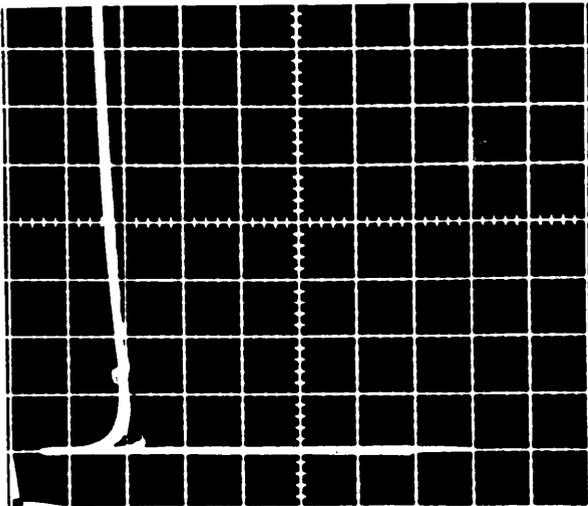


Fig. 1 Cold Cathode Lamp Current as a function of Voltage
Vertical: 2mA/div. Horizontal: 200V/div.

To facilitate analysis, the equivalent CCFL circuit shown in figure 2 is used. V_{FL} is the **average** lamp sustaining voltage over the operating range. The lamp impedance (R_{FL}) is a complex function but can be considered a fixed negative resistance at the sustaining voltage. Stray lamp and interconnect capacitance are lumped together as C_{FL} .

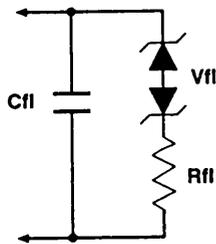


Fig. 2 CCFL equivalent circuit

ZVS Resonant Converter Topology

The current fed push-pull converter shown in fig. 3 is driven at its resonant frequency to provide ZVS operation. The push-pull output MOSFETS (Q1 & Q2) are alternately driven at 50% duty cycle. Commutation occurs as V1 and V2 resonate through zero thereby insuring zero voltage switching. This virtually eliminates switching losses associated with charging MOSFET output and stray capacitance, and reduces gate drive losses by minimizing gate charge.

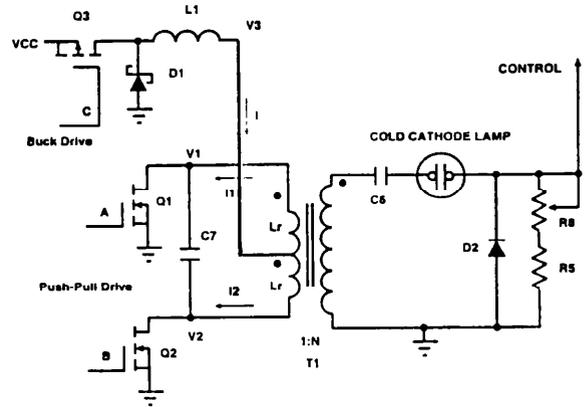


Fig. 3 Current Fed Push-Pull ZVS Resonant Converter

Current is supplied to the push-pull stage by a buck regulator (Q3). The control circuitry forces the average voltage across the current sense resistor ($R8+R5$) and rectifier (D2) to equal a reference voltage. Adjusted R8 varies the current and the lamp's brightness. The non-linearity introduced by D2 is insignificant since RS is adjusted for a particular brightness with no concern of the actual current level.

Winding inductance, L_r , and C_r , the combined effective capacitance of C7 and the reflected secondary capacitances make up the resonant tank. The secondary side of the transformer exhibits a symmetrical sine wave voltage varying from about 300V to 1500V peak. Capacitor C6 provides ballasting and insures that the converter is only subjected to positive impedance loads.

Waveform Analysis

Simulated converter voltage and current waveforms are shown in fig. 4. At time t_0 , the primary current ($I1$ & $I2$) has reached its peak value. The push-pull drain voltages (V1 & V2) have resonated to zero. The primary voltage (V3) has also resonated to zero, and through the control circuitry commutated Q1 off and Q2 on. The energy stored in L_r is also at its peak. This energy is transferred from L_r to the effective resonant capacitance (C_r) during time t_0 to t_1 , causing C_r 's voltage to sinusoidally increase.

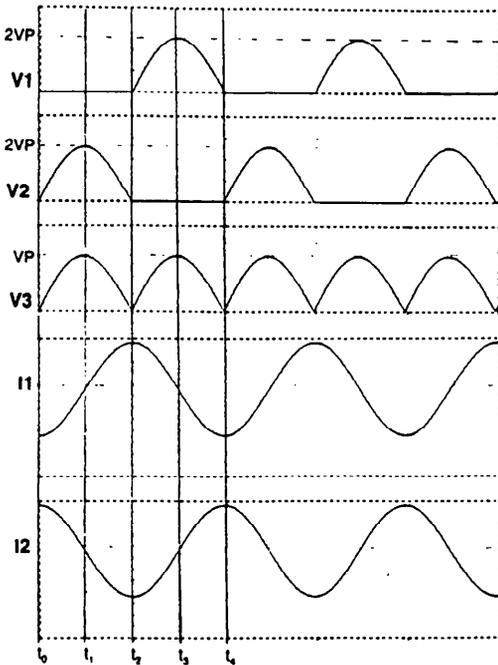


Fig. 4 Converter Voltage and Current Waveforms

At time t_1 , all of the inductive energy in L_R has transferred to C_R , resulting in zero current through L_R and maximum voltage across C_R . From time t_1 to t_2 , the energy transfers from C_R back to L_R , decreasing C_R 's voltage while L_R 's current increases.

The resonant current through L_R at time t_2 is equal and opposite to its value at t_0 . The reflected load current flows during the MOSFET on time, and is observed as a slight current amplitude asymmetry. The voltages at V1, V2, and V3 have resonated back to zero, causing the control circuitry to commutate Q2 off and Q1 on. The cycle continues symmetrically during the t_2 through t_4 interval, producing fully sinusoidal voltage and current waveforms.

Simplified Converter Model

The converter model shown in fig. 5, which is valid for one half cycle simplifies analysis by reflecting all impedances to the primary and eliminating the transformer. The differential voltage developed across the push-pull stage primary (V1-V2) exhibits twice the voltage excursion as the center-tap (V3). This reflects C7 to V3 through the turns ratio squared, resulting in $4(C7)$ at V3. The secondary winding

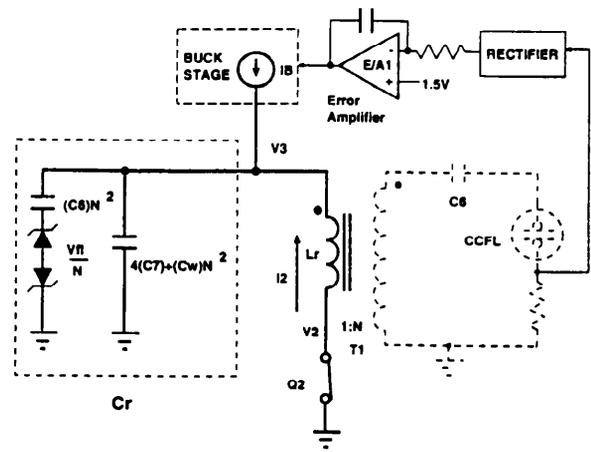


Fig. 5 Simplified converter model

capacitance is also reflected by the square of the turns ratio (n). Reflected winding capacitance is usually significant due to the high turns ratios typically employed. The buck stage operates in continuous current mode and is synchronized to the push-pull stage.

Lamp current is proportional to lamp intensity, and is used as the feedback variable. Buck current (I_b) is the response variable, which in turn regulates the average push-pull primary voltage. The coupling capacitor's high impedance transforms the secondary voltage to lamp current.

Control Equations

Variable Summary:

- C_R = Effective resonant tank capacitance
- C_W = Secondary interwinding capacitance
- F_L = Average lamp voltage
- I_B = Average Buck output current
- L_R = Primary Winding Inductance
- n = Transformer turns ratio
- Z_{sec} = Secondary impedance

Fig. 6 shows the buck output stage and forced output voltage waveform. The output voltage is a rectified sine wave, corresponding to the synchronous, resonant push-pull stage input

voltage. The inductor output configuration exhibits high impedance at the resonant frequency and averages the output voltage throughout the cycle. The buck output voltage as a function of time is:

$$V_{out}(t) = V_P \sin(\omega t)$$

Where the angular frequency is:

$$\omega = 2\pi f = \frac{2\pi}{2t_1} = \frac{\pi}{t_1}$$

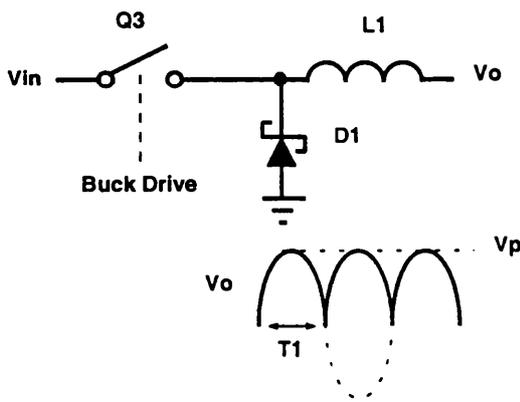


Fig. 6 Buck converter stage

The volts-second product across the inductor must be zero during steady state. Setting the on and off volt-second products equal and integrating gives the buck's transfer function:

$$V_{LR}(t_{on}) = -V_{LR}(t_{off})$$

$$\int_0^{t_{on}} V_{LR} dt = -\int_{t_{on}}^{t_1} V_{LR} dt$$

$$V_P = \frac{\pi}{2} V_i D \tag{1}$$

This transfer function is identical to the familiar DC output buck transfer function, with the $\pi/2$ term accounting for peak versus average output voltage. As with the DC buck, primary voltage varies linearly with duty-cycle.

The peak primary voltage is also related to peak lamp current by:

$$V_P = \frac{(I_{FL(peak)})(Z_{sec}) + V_{FL}}{n} \tag{2}$$

Setting (1) and (2) equal and solving for $I_{FL(avg)}$ expresses lamp current as a function of duty-cycle:

$$I_{FL(avg)} = \frac{DV_i n - \frac{2 V_{FL}}{\pi}}{Z_{sec}} \tag{3}$$

As expected from fig. 5, the lamp sustaining voltage, V_{FL} introduces a nonlinearity.

Buck output current is related to lamp current by equating input and output powers. The input power is:

$$\begin{aligned} P_{input} &= \frac{1}{t} \int P dt \\ &= \frac{1}{\pi/2} \int_0^{\pi/2} I V_P \sin(t) dt \end{aligned}$$

$$P_{input} = \frac{2 I_B V_P}{\pi}$$

The power to the load is:

$$P_{out} = V_{FL} I_{FL} (avg)$$

For analytical purpose, 100% power transfer is assumed:

$$P_{out} = P_{input}$$

$$V_{FL} I_{FL(avg)} = \frac{2 I_B V_P}{\pi}$$

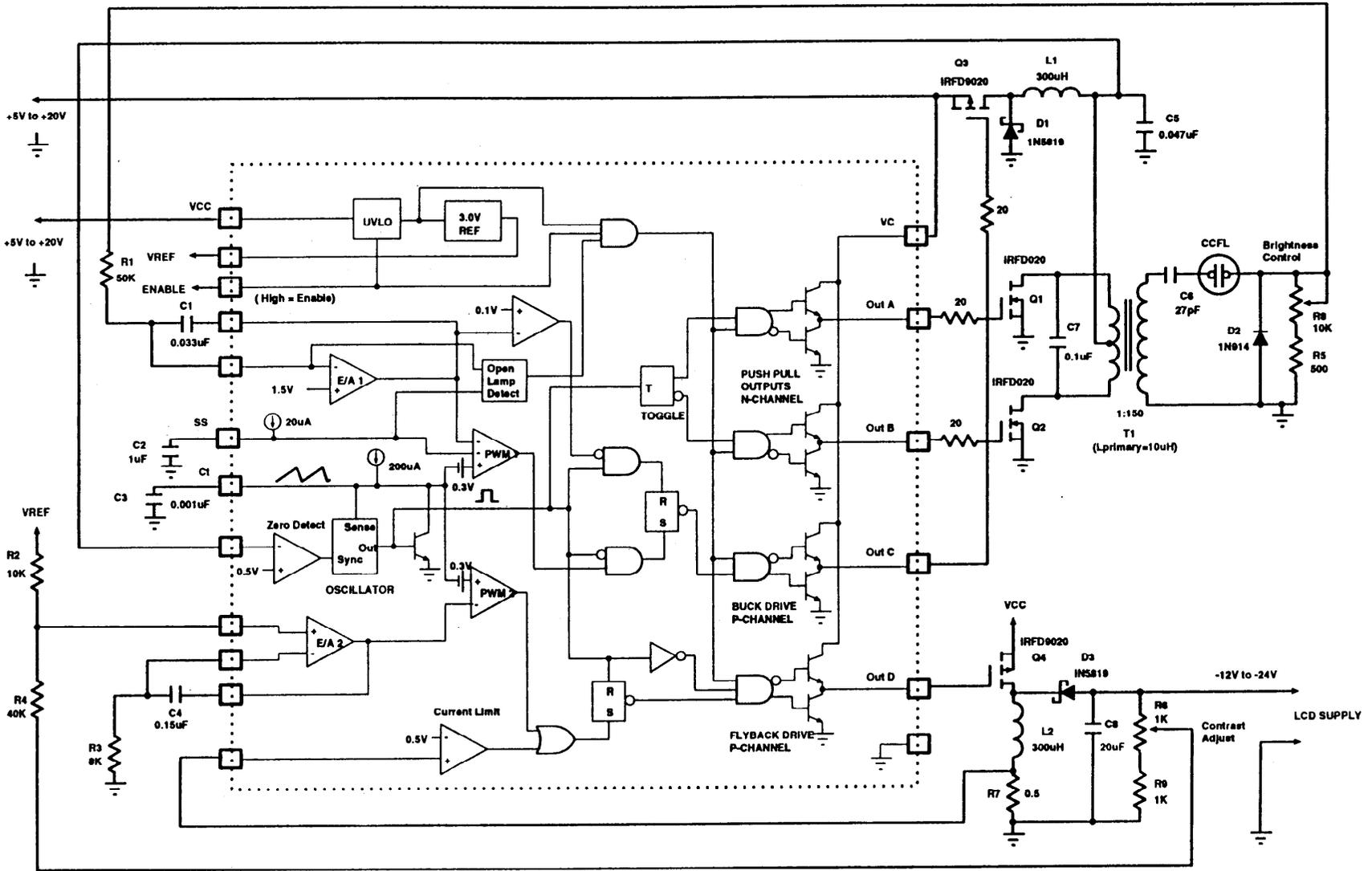


Fig. 7 UC3871 Application circuit

3-374

$$I_B = \frac{\pi I_{FL(avg)} V_{FL}}{2 V_P} \quad (4)$$

Substituting (2) for V_p in (4) gives the buck output current as a function of lamp current:

$$I_B = \frac{n V_{FL}}{Z_{sec} + \frac{V_{FL}}{\pi I_{FL(avg)}}} \quad (5)$$

The resonant frequency is approximately:

$$f_R \approx \frac{1}{2\pi\sqrt{L_R C_R}} \quad (6)$$

The nonlinearity introduced by V_{FL} causes the resonant frequency to vary with load. At very low lamp intensity the secondary voltage barely crests above V_{FL} . The effective resonant capacitance, C_r , is primarily the sum of C_7 and C_W reflected to the primary. As the secondary voltage increase above V_{FL} , the reflected C_6 value adds to the resonant capacitance, decreasing the frequency. The frequency range is approximated by assuming C_6 has negligible effect at minimum lamp intensity, and fully adds to C_r at maximum intensity.

The peak resonant inductor current is the sum of the reflected load current from (5) and the resonant current:

$$I_{LR(peak)} = \frac{V_P}{Z_{tank}} + I_b \quad (7)$$

The tank impedance is determined by setting the resonant energy storage terms equal:

$$\frac{1}{2} L_R I^2 = \frac{1}{2} C_R V^2$$

Solving for V_R/I_R gives the tank impedance:

$$Z_{tank} = \sqrt{\frac{L_R}{C_R}} \quad (8)$$

Although relatively large currents are circulated through the resonant tank, the switches operate at low current levels. This is a direct result of the continuous resonant topology; the switches only must handle the energy that is removed by the load and lost in parasitics. The peak switch current is:

$$I_{sw(peak)} = I_b \quad (9)$$

The UC3871 A Completely Integrated Solution

Fig. 7 shows a complete application circuit using the UC3871 Synchronous Resonant Fluorescent lamp and LCD driver. The IC provides all drive, control and housekeeping functions to implement CCFL and LCD converters. The buck output voltage (transformer center-tap) provides the zero crossing and synchronization signal. The LCD supply modulator is also synchronized to the resonant tank.

The buck modulator drives a P-channel MOSFET directly, and operates over a 0-100% duty-cycle range. The modulation range includes 100%, allowing operation with minimal headroom. The LCD supply modulator also directly drives a P-channel MOSFET, but it's duty-cycle is limited to 95% to prevent flyback supply foldback.

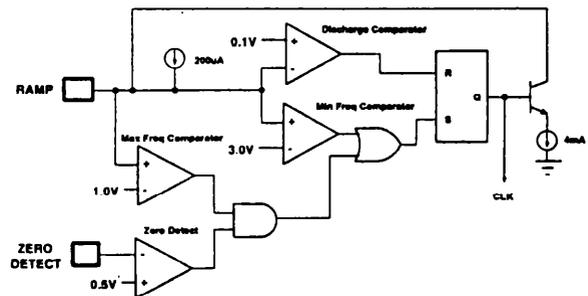


Fig. 8 UC3871 Oscillator Block Diagram

The Oscillator and synchronization circuitry are shown in fig. 8. The oscillator is designed to synchronize over a 3:1 frequency range. In an actual application however, the frequency range is only about 1.5:1. A zero detect comparator senses the primary center-tap voltage, generating

a synchronization pulse when the resonant waveform falls to zero. The actual threshold is 0.5 volts, providing a small amount of anticipation to offset propagation delay.

The synchronization pulse width is the time that the 4mA current sink takes to discharge the timing capacitor to 0.1 volts. This pulse width sets the LCD supply modulator minimum off time, and also limits the minimum linear control range of the buck modulator. The 200 μ A current source charges the capacitor to a maximum of 3 volts. A comparator blanks the zero detect signal until the capacitor voltage exceeds 1 volt, preventing multiple synchronization pulse generation and setting the maximum frequency. If the capacitor voltage reaches 3 volts (a zero detection has not occurred) an internal clock pulse is generated to limit the minimum frequency.

A unique protection feature incorporated in the UC3871 is the Open Lamp Detect circuit. An open lamp interrupts the current feedback loop and causes very high secondary voltage. Operation in this mode will usually breakdown the transformer's insulation, causing permanent damage to the converter. The open lamp detect circuit, shown in fig. 9 senses the lamp current feedback signal at the error amplifiers input, and shuts down the outputs if insufficient signal is present. Soft-start circuitry limits initial turn-on currents and blanks the open lamp detect signal.

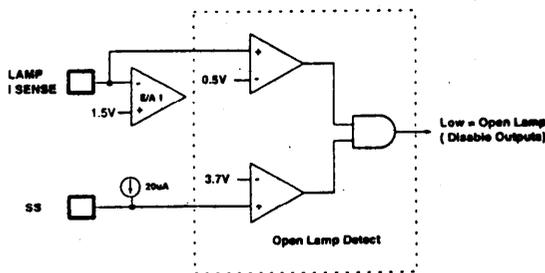


Fig. 9 Open Lamp Detect Circuit

Other features are included to minimize external circuitry requirements. A logic level enable pin shuts down the IC, allowing direct connection to the battery. During shut-down, the IC typically draws less than 100nA. The UC3871, operating from 4.5V to 20V, is compatible with almost all battery voltages used in portable computers. Under-voltage lockout circuitry

disables operation until sufficient supply voltage is available, and a 1% voltage reference insures accurate operation. Both inputs to the LCD supply error amplifier are uncommitted, allowing positive or negative supply loop closure without additional circuitry. The LCD supply modulator also incorporates cycle-by-cycle current limiting for added protection.

Application Circuit Example

The application circuit shown in fig.7 resonates at approximately 50khz. This frequency allow a reasonable compromise between size and efficiency. This relatively low frequency by today's standards. results from high voltage insulation and spacing requirements, and practical limitations in reducing stray and interwinding capacitance. The half wave current sense signal is sensed by Error Amp 1 and averaged by integral compensation. The range of current control is 500 μ A to 10mA.

A flyback converter generates the LCD supply, outputting -12V to -24V to bias monochrome LCDs. Color displays normally require a positive bias voltage. Since this voltage typically must also be stepped up, a coupled inductor flyback is normally used.

Actual circuit waveforms agree with the spice simulated waveforms in fig. 4. Distortion caused by lamp nonlinearity is clearly visible at the operating extremes. At more nominal levels, the waveforms are more ideal, with only a small amount of observable distortion.

All of the following waveforms were taken at minimum and maximum lamp intensity to indicate worst case conditions. Nominal measured efficiency was 80%. Further improvement is possible with lower resistance magnetics and lower on resistance MOSFETs. Fig. 10 shows secondary output voltage, fig 11 shows lamp voltage, and fig. 12 shows lamp current. Notice that the lamp voltage is fairly constant with widely varying current. A frequency shift from about 48kHz to 57kHz is also observed over the lamp intensity range. The lamp current exhibits additional harmonics induced by it's nonlinearity. Push-pull MOSFET drain to source voltage is shown in fig. 13, and drain current is shown in fig. 14. The transformer center-tap voltage (buck output) is shown in fig. 15. All waveforms are sinusoidal, exhibiting minimal harmonic content.

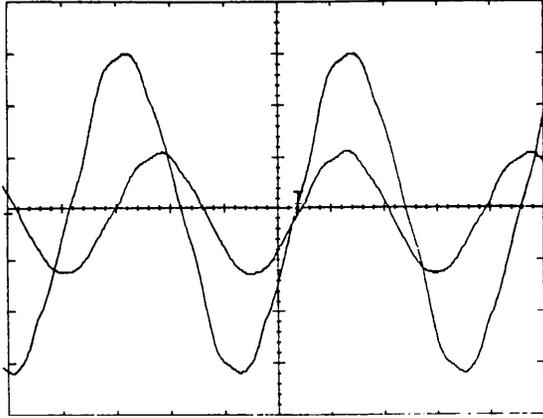


Fig. 10 secondary output voltage
vertical: 500V /div. Horizontal: 5 μ s/div.

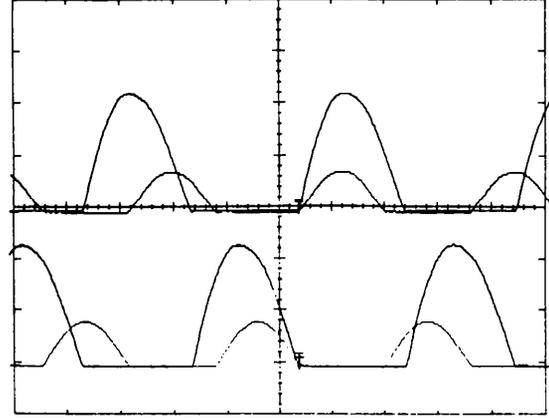


Fig. 13 Push-Pull MOSFET drain to source voltage
vertical: 10V /div. Horizontal: 5 μ s/div.

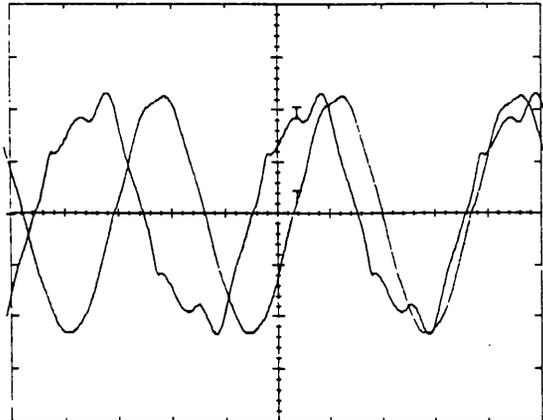


Fig. 11 lamp voltage
Vertical: 200V /div. Horizontal: 5 μ s/div.

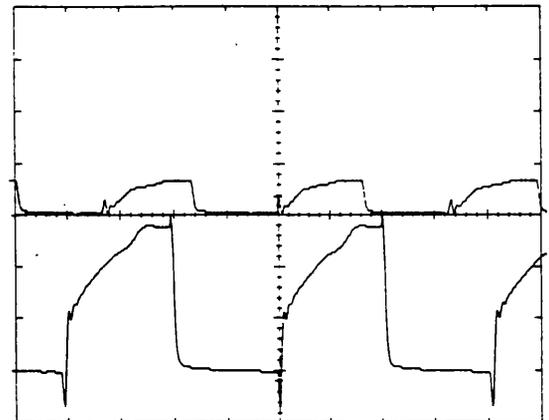


Fig. 14 Push-pull MOSFET drain current
Vertical: 200mA /div. Horizontal: 5 μ s/div.

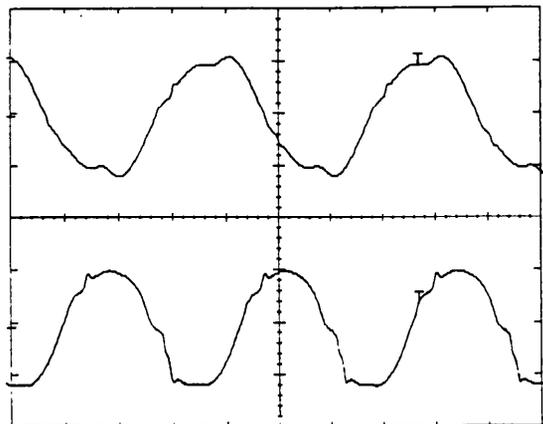


Fig. 12 Lamp current
Vertical: 10mA /div. Horizontal: 5 μ s/div.
500 μ A/div.

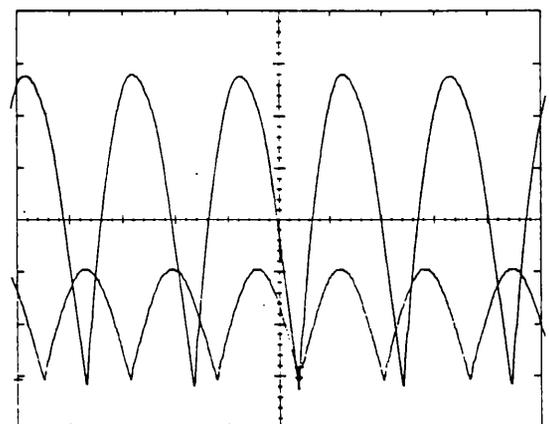


Fig. 15 Buck output voltage
Vertical: 200mA /div. Horizontal: 5 μ s/div.

Summary

The current fed push-pull ZVS converter efficiently develops high voltage, sinusoidal power for driving cold cathode fluorescent lamps. Design

equations have been derived, and verified experimentally, simplifying application circuit design and analysis. The **UC3871** provides a complete solution for high performance back-light and LCD power supplies.

IMPORTANT NOTICE

Texas Instruments and its subsidiaries (TI) reserve the right to make changes to their products or to discontinue any product or service without notice, and advise customers to obtain the latest version of relevant information to verify, before placing orders, that information being relied on is current and complete. All products are sold subject to the terms and conditions of sale supplied at the time of order acknowledgement, including those pertaining to warranty, patent infringement, and limitation of liability.

TI warrants performance of its semiconductor products to the specifications applicable at the time of sale in accordance with TI's standard warranty. Testing and other quality control techniques are utilized to the extent TI deems necessary to support this warranty. Specific testing of all parameters of each device is not necessarily performed, except those mandated by government requirements.

CERTAIN APPLICATIONS USING SEMICONDUCTOR PRODUCTS MAY INVOLVE POTENTIAL RISKS OF DEATH, PERSONAL INJURY, OR SEVERE PROPERTY OR ENVIRONMENTAL DAMAGE ("CRITICAL APPLICATIONS"). TI SEMICONDUCTOR PRODUCTS ARE NOT DESIGNED, AUTHORIZED, OR WARRANTED TO BE SUITABLE FOR USE IN LIFE-SUPPORT DEVICES OR SYSTEMS OR OTHER CRITICAL APPLICATIONS. INCLUSION OF TI PRODUCTS IN SUCH APPLICATIONS IS UNDERSTOOD TO BE FULLY AT THE CUSTOMER'S RISK.

In order to minimize risks associated with the customer's applications, adequate design and operating safeguards must be provided by the customer to minimize inherent or procedural hazards.

TI assumes no liability for applications assistance or customer product design. TI does not warrant or represent that any license, either express or implied, is granted under any patent right, copyright, mask work right, or other intellectual property right of TI covering or relating to any combination, machine, or process in which such semiconductor products or services might be or are used. TI's publication of information regarding any third party's products or services does not constitute TI's approval, warranty or endorsement thereof.