AN-995A

Electronic Ballasts Using the Cost-Saving IR215X Drivers

Introduction

Electronic ballast circuits have recently undergone a revolution in sophistication from the early bipolar designs of ten years ago. This has been brought about partly by the advent of power MOSFET switches with their inherent advantages in efficiency, but mainly by incentives and utility rebate programs sponsored by domestic and foreign governments. New IEC requirements have also spurred the design of high power factor ballasts and are starting to impose further restrictions on crest factor, ballast factor and life expectancy (see IEC 555 Standard.)

Until power semiconductors allowed for today's innovations in ballast design, coil and core fluorescent ballasts were manufactured in large quanitities by a few key suppliers.

Now there are hundreds of electronics companies that are "in the ballast business" and more are joining their ranks all the time.

Most electronic ballasts use two power switches in a totem pole (half-bridge) topology and the tube circuits consist of L-C series resonant circuits with the lamp(s) across one of the reactances. Figure 1 shows this basic topology.

In this circuit the switches are power MOSFETs driven to conduct alternately by windings on a current transformer. The primary of this transformer is driven by the current in the lamp circuit and operates at the resonant frequency of L-C.

Unfortunately, the circuit is not self starting and must be pulsed by the DIAC connected to the gate of the lower MOSFET.

After the initial turn-on of the lower switch, oscillation sustains and a high frequency square wave (30 - 80 kHz) excites the L-C resonant circuit. The sinusoidal voltage across C is magnified by the Q at resonance and develops sufficient amplitude to strike the lamp, which then provides flicker-free illumination.

This basic circuit has been the standard for electronic ballasts for many years, but has the following inherent shortcomings:

1) not self starting
2) poor switch times
3) labor intensive torroidal current transformer
4) not amenable to dimming
5) expensive to manufacture in large quantity.

Next Generation Ballast

These criticisms have all been resolved in the new, cost-saving International Rectifier IR215X Control IC series.

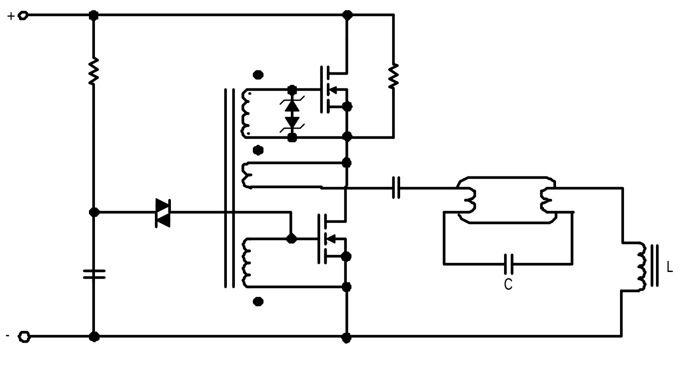


Figure 1. Electronic ballast using transformer drive

International Rectifier Control ICs are monolithic power integrated circuits capable of driving low-side and high-side MOSFETs or IGBTs from logic level, ground referenced inputs. They provide offset voltage capabilities up to 600 VDC and, unlike driver transformers, can provide super-clean waveforms of any duty cycle 0 - 99%.

The IR215X series is a recent addition to the Control IC family and, in addition to the above features, these devices have a front end similar in function to the CMOS 555 timer IC.

These drivers provide the designer with self-oscillating or synchronized oscillation functions merely with the addition of external R_T and C_T components (Figure 2). They also have internal circuitry which provides a nominal 1.2 µs dead time between outputs and alternating high side and low side outputs for driving half-bridge power switches.

When used in the self oscillating mode the frequency of oscillation is given by:

$$f = \frac{1}{1.4 \times (R_T + 75\Omega) \times C_T} \qquad (1)$$

The three available self-oscillating drivers are IR2151, IR2152 and IR2155.

IR2155 has larger output buffers that switch a 1000 pF capacitive load with t_r = 80 ns and t_f = 40 ns. It has micro power start-up and 150 ohm RT source.

IR2151 has t_r and t_f of 100 ns and 50 ns and functions similarly to IR2155.

IR2152 is identical to IR2151 but with phase inversion from R_T to L_O.

IR2151 and 2152 have 75 ohm R_T source (Equation 1.)

These drivers are intended to be supplied from the rectified AC input voltage and for that reason they are designed for minimum quiescent current and have a 15V internal shunt regulator so that a single dropping resistor can be used from the DC rectified bus voltage.

Referring again to Figure 2, note the synchronizing capability of the driver. The two back-to-back diodes in series with the lamp circuit are effectively a zero crossing detector for the lamp current. Before the lamp strikes, the resonant circuit consists of L, C1 and C2 all in series.

C1 is a DC blocking capacitor with a low reactance, so that the resonant circuit is effectively L and C2. The voltage across C2 is magnified by the Q factor of L and C2 at resonance and strikes the lamp.

After the lamp strikes, C_2 is effectively shorted by the lamp voltage drop and the frequency of the resonant circuit now depends upon L and C1.

This causes a shift to a lower resonant frequency during normal operation, again synchronized by sensing the zero crossing of the AC current and using the resultant voltage to control the driver oscillator.

In addition to the driver quiescent current, there are two other components of DC supply current that are a function of the actual application circuit:

1) current due to charging the input capacitance of the power switches

2) current due to charging and discharging the junction isolation capacitance of the International Rectifier gate driver.

Both components of current are charge-related and therefore follow the rules:

$$Q = CV \qquad (2)$$

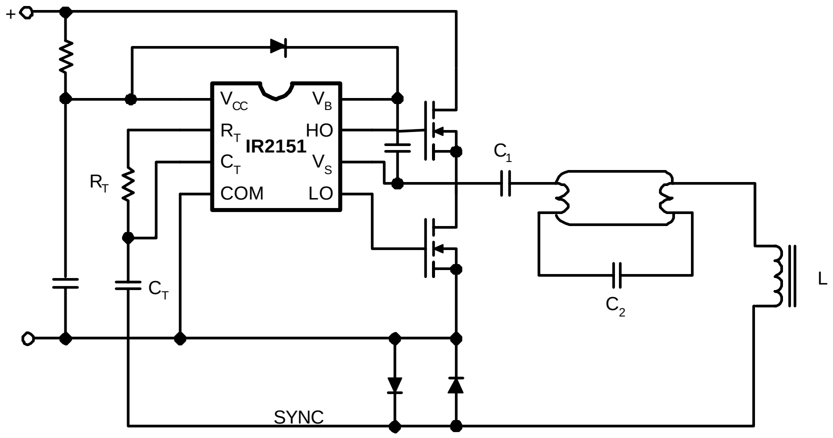


Figure 2. Electronic ballast using IR2151 driver

It can readily be seen, therefore, that to charge and discharge the power switch input capacitances, the required charge is a product of the gate drive voltage and the actual input capacitances and the input power required is directly proportional to the product of charge and frequency and voltage squared:

$$\text{Power} = \frac{QV^2}{2} \times f \qquad (3)$$

The above relationships suggest the following considerations when designing an actual ballast circuit:

1) select the lowest operating frequency consistent with minimizing inductor size;

2) select the smallest die size for the power switches consistent with low conduction losses (this reduces the charge requirements);

3) DC bus voltage is usually specified, but if there is a choice, use the lowest voltage.

NOTE: Charge is not a function of switching speed. The charge transferred is the same for 10 ns or 10 µs switch times.

Let us now consider some practical ballast circuits which are possible with the self-oscillating drivers. By far the most popular fluorescent fixture is the so-called 'Double 40' type which uses two standard T12 or T8 lamps in a common reflector.

Two suggested ballast circuits are shown in figures 3 and 4. One is a low power factor circuit, and the other uses a novel diode/capacitor configuration to achieve a power factor > 0.95.

The low power factor circuit shown in figure 3 accepts 115 VAC or 230 VAC 50/60/400 Hz inputs to produce a nominal DC bus of 320 VDC. Since the input rectifiers conduct only near the peaks of the AC input voltage, the input power factor is approximately 0.6 lagging with a non-sinusoidal current waveform. This type of rectifier is not recommended for anything other than an evaluation circuit or low power compact fluorescents and indeed may become unacceptable as harmonic currents in power distribution systems are further reduced by power quality regulations.

Note that the International Rectifier IR2151 Control IC operates directly off the DC bus through a dropping resistor and oscillates at around 45 kHz in compliance with the following relationship:

$$f = \frac{1}{1.4 \times (R_T + 75\Omega) \times C_T}$$

Power for the high side switch gate drive comes from a bootstrap capacitor of 0.1 µF which is charged to approximately 14V whenever V_S (lead 6) is pulled low during the low side power switch conduction. The bootstrap diode 11DF4 blocks the DC bus voltage when the high side switch conducts. A fast recovery diode (<100 ns) is required to ensure that the bootstrap capacitor is not partially discharged as the diode recovers and blocks the high voltage bus.

The high frequency output from the half-bridge is a square wave with very fast transition times (approximately 50 ns). In order to avoid excessive radiated noise

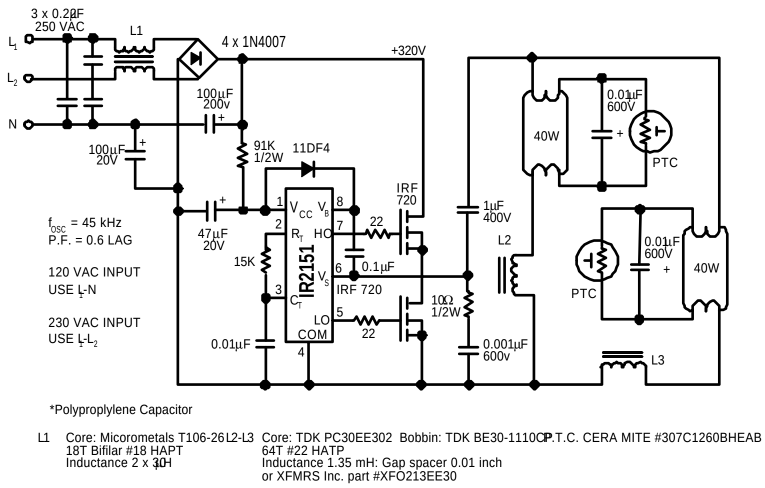


Figure 3. 'Double 40' ballast using IR2151 oscillator/driver

from the fast wave fronts, a 0.5W snubber of 10Ω and 0.001 µF is used to reduce the switch times to approximately 0.5 µs. Note that there is a built-in dead time of 1.2 µs in the IR2151 driver to prevent shoot-through currents in the half-bridge.

The fluorescent lamps are operated in parallel, each with its own L-C resonant circuit. Up to four tube circuits can be driven from a single pair of MOSFETs sized to suit the power level.

The reactance values for the lamp circuit are selected from L-C reactance tables or from the equation for series resonance:

$$f = \frac{1}{2\pi\sqrt{LC}} \qquad (4)$$

The Q of the lamp circuits is rather low because of the need for operation from a fixed frequency which, of course, can vary because of R_T and C_T tolerances. Fluorescent lamps do not normally require very high striking voltages so a Q of 2 or 3 is sufficient. 'Flat Q' curves tend to result from larger inductors and small capacitor ratios where:

$$Q = \frac{2\pi fL}{R} \qquad (5)$$

and R tends to be larger as more turns are used.

Soft-starting with tube filament pre-heating can be cheaply incorporated by using P.T.C. thermistors across each lamp. In this way, the voltage across the lamp gradually increases as the P.T.C. self-heats until finally the striking voltage with hot filaments is reached and the lamp strikes.

High Power Factor

The circuit shown in figure 4 is a passive power factor improvement (no active boost circuit) and is applicable to low power ballasts such as compact fluorescent. It suffers from the disadvantage of low DC rectified output voltage and results in a crest factor of about 2.

Note that a crest factor standard not exceeding 1.7 is recommended by fluorescent lamp manufacturers to realize the maximum life projections of 20,000 hours for these lamps.

$$\text{Crest Factor} = \frac{\text{Peak Current}}{\text{RMS Current}}$$

If the ballast delivers a pure sine wave of voltage and current to the lamp, the crest factor would be $\sqrt{2}$. In an electronic ballast, a DC bus voltage is derived from a mains frequency rectifier and is filtered by means of an electrolytic capacitor. The 2x line frequency ripple voltage on the DC bus gives rise to additional ripple currents in the lamp. Even if the lamp current is sinusoidal (crest factor 1.414) the mains-related ripple adds to the peak current value and causes the crest factor to increase. Referring to the waveforms of figure 5, it is clear that the ripple voltage amplitude is $V_P/2$ which results in a crest factor of approximately 2.

What is needed, therefore, is a power factor correction using active control to minimize current ripple and stabilize the DC bus voltage. Boost regulator correction cir-

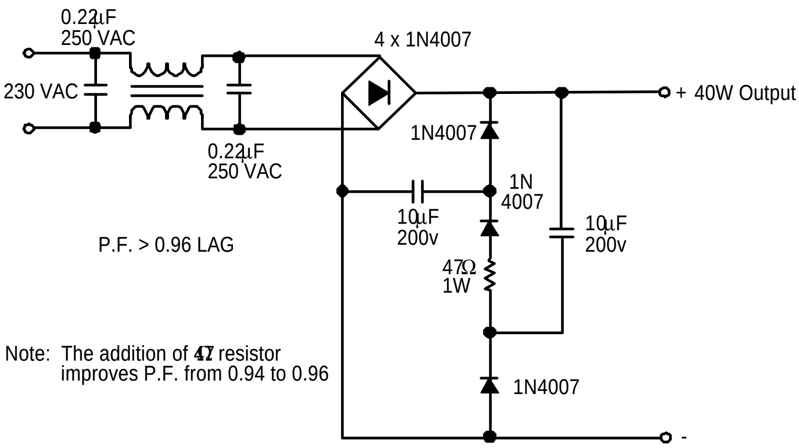


Figure 4. Passive high power factor rectifier/filter

cuits have become popular for off-line power supplies and several semiconductor manufacturers supply control ICs for this topology.

For electronic ballasts, however, the sophistication of these control chips may not be necessary and it is relatively simple to provide power factors exceeding 0.95 by a simple boost topology operating at a fixed 50% duty cycle. Using the IR2151 driver it is also possible to provide dimming merely by changing the duty cycle and, hence, the boost ratio.

Figures 6* and 7 illustrate how this can be accomplished.

Dimming Control

The IR2151 has a 'front end' oscillator circuit akin to the 555 IC and is amenable to the same type of circuitry to control the duty cycle of the output waveforms.

Dimming control to 50% of power input is easily achieved by this control. When R_T (lead 2) switches high the charging path for C_T (lead 3) is through the forward biased diode and the left side of the duty cycle control pot. When C_T charges to two-thirds V_{CC}, R_T switches low and C_T discharges through the right side of the control pot. Until the one-third V_{CC} voltage is reached, the cycle then repeats. Note that although the charge and discharge times of C_T can be varied, the sum of them remains constant and hence the oscillation frequency is also constant. This allows sufficient lamp striking voltage even under dimmed conditions.

In actual operation, the 'on' time of the boost MOSFET is reduced as R_T(CHG) becomes smaller than RT(DISCH). Obviously, if the "on" time of the boost MOSFET is reduced the boost voltage ratio is also reduced proportionately:

$$\text{boost voltage ratio} = V_{IN} \times \frac{1}{1-D} \qquad (6)$$

$$\text{e.g., at 50\% duty cycle: } V_{IN} \times \frac{1}{0.5} = 2V_{IN}$$

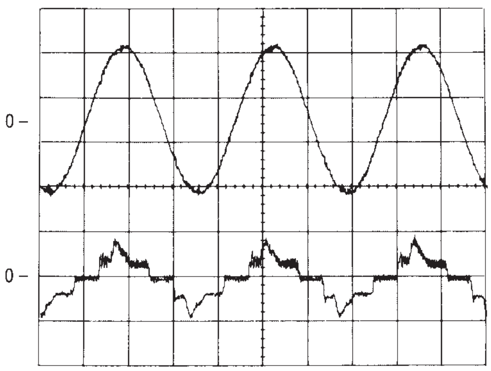
where V_{IN} = instantaneous input voltage and D is the 'on' time ratio of the boost MOSFET.

A variation of this circuit, shown in Figure 8, allows dimming to be controlled remotely by a variable resistor. The circuits of Figures 7 and 8 both suffer from a basic flaw; namely that if the lamps are removed or broken the open circuit DC bus voltage rises until the power MOSFETs avalanche and fail or the filter capacitor overheats and fails due to overvoltage.

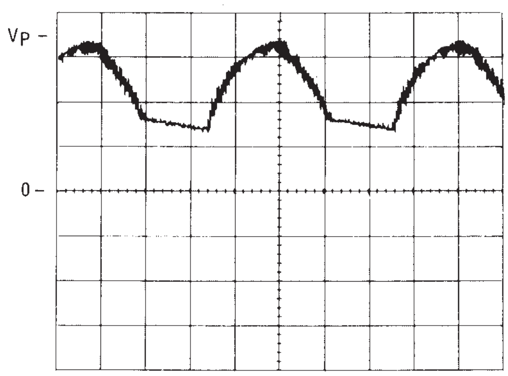
To prevent this, the duty cycle of the boost transistor can be reduced so that the DC bus is regulated to a constant level, as shown in Figure 9.

In operation, the duty cycle of the boost regulator is determined by comparing a fraction of the DC bus voltage with a reference triangle wave appearing on the timing capacitor C_T. The switching levels of the IR2151 Control IC timing circuit occur at one-third V_{CC} and two-thirds V_{CC}. Since V_{CC} is regulated by an internal voltage regulator, the amplitude of the C_T waveform is also regulated.

*U.S. Patent No. 5001400 Nilssen, March 1991



AC Voltage and current
PF = 0.96 LAG
200 V/Div., 0.5 A/Div., 5 msec/Div.



DC Bus voltage showing 50% Vp ripple
100V/Div., 2msec/Div.

Figure 5. Waveforms of Figure 4

The LM311 comparator produces a positive output whenever the instantaneous voltage on C_T exceeds a fraction of the DC bus voltage. This output is 'OR'ed with the 50% LO waveform and impedance matched to drive the boost MOSFET by a 2N2222A emitter follower. The DC bus regulation resulting from this technique is 210-225 VDC with an input AC range of 90 VAC to 130 VAC and dimming from 50% to 100% (225 VDC maximum with bulb removed). Dimming is performed by raising the operating frequency to approximately double for a 50% reduction in power output.

Reliable striking of the lamp is always assured at any dimming setting because the circuit is synchronized to the natural resonance of the lamp circuit. Note the back-to-back diodes which form a zero current crossing detector for the lamp current and the connection of C_T to this synchronizing voltage (see also figure 2.)

After the lamp strikes, the synchronization circuit is no longer able to control the frequency which then reverts to whatever is selected by C_T and the variable R_T.

In addition to the popular fluorescent ballast applications, there is a growing interest in High Intensity Discharge (HID) ballasts for outdoor lighting. These too can be simply designed using the International Rectifier IR2151. A 70 watt high-pressure sodium (HPS) ballast is illustrated in Figure 10.

HPS ballasts have some unique requirements not found in fluorescent ballasts. They must:

- not be damaged when operating into open circuit
- supply sufficient energy at 3 - 4 kV to start the lamp
- accomodate large variations in lamp voltage
- not cause arc instability in the lamp
- be matched to lamp characteristics to maximize lamp life

The circuit shown in figure 10 provides an input power factor >0.9 and has DC bus control limiting the voltage to 225 VDC whether or not the lamp is energized. L3 performs two functions:

1) current limiting for the negative resistance characteristics of the lamp
2) a pulse voltage step-up function to strike the HPS lamp.

The 3 kV pulse voltage is derived from a 135V SIDAC which discharges a 1 μF capacitor into the 2 turn winding of L3. The 30:1 step-up ratio of L3 supplies the starting pulse to the lamp. After the lamp strikes, there is insufficient charge voltage on the 1 μF capacitor in the 2 turn winding circuit to avalanche the SIDAC and no fur-

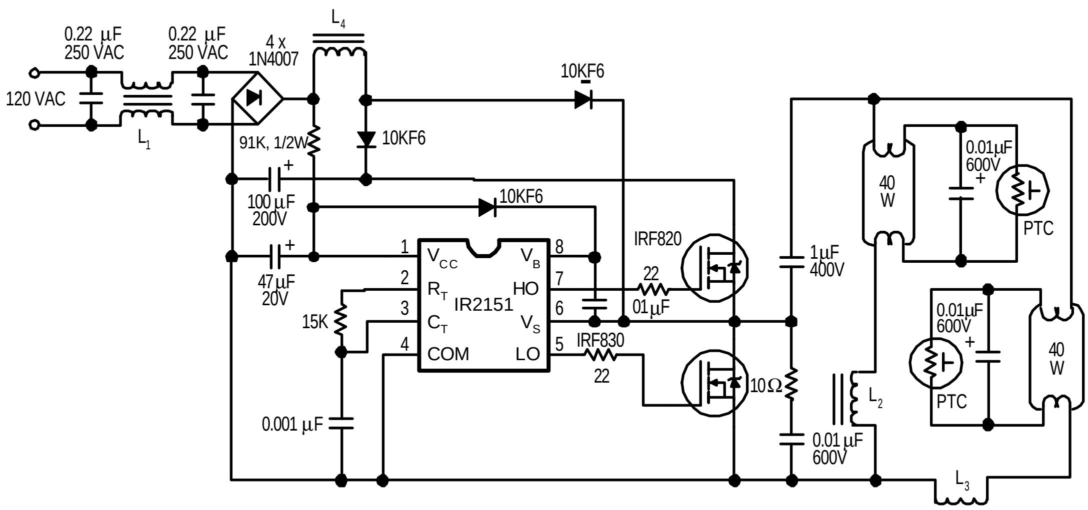


Figure 6. 'Double 40' IR2151 ballast with active power factor correction

ther start pulses are supplied. The hot re-strike time of this ballast is approximately 75 seconds.

The circuits described above have illustrated some of the ways in which the IR2151 Control IC may be used in synchronized and non-synchronized ballasts.

Some applications require higher lamp voltages which may be too high for the simple half-bridge topology. By using four power MOSFETs in a full bridge circuit, the output voltage may be doubled without increasing the MOSFET current. A full bridge circuit automatically doubles output power and this topology can be implemented with the IR2151 low-cost master oscillator driving an IR2111 slave circuit. Figure 12 illustrates this topology which is described in the following text.

This ballast is intended to drive two 80W fluorescent lamps such as F96-T12 type. These lamps are operated at the same current as their 48 inch counterparts but require twice the voltage both for striking and normal operation. These slim line lamps have single pin contacts and are designed to be instantly started from suitable ballasts. Since the lamps start with cold electrodes, the ballast must provide in excess of 800V RMS for reliable starting of any lamp at low ambient temperatures.

The circuit shows a full bridge with each leg driven from a separate Control IC. U1 is a self-oscillating driver (IR2151) and U2 is a slave driver (IR2111). The

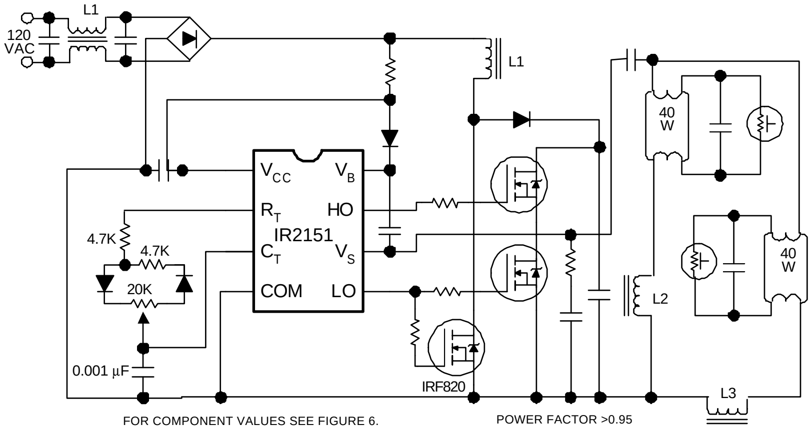


Figure 7. Local dimming by bus voltage control

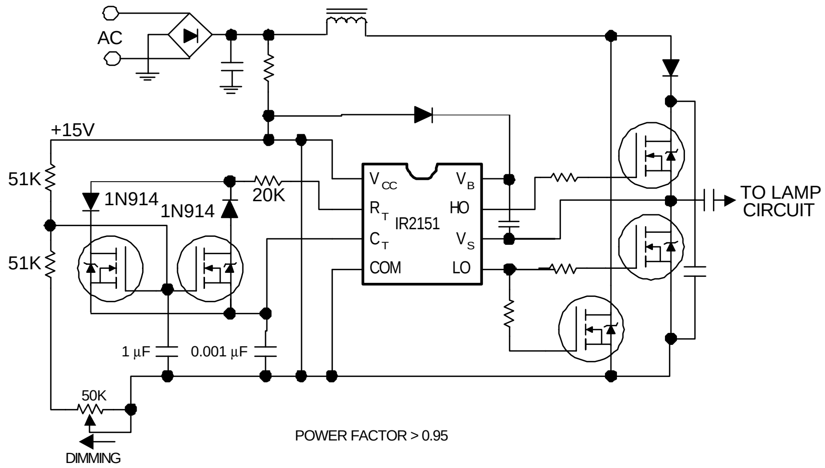


Figure 8. Remote dimming control by variable resistor

operation of the IR2151 is the same as previously described for the ‘Simple Double 40 Fluorescent Ballast.’ The full bridge circuit essentially doubles the available AC output voltage compared with the half-bridge design.

The slave driver U2 is driven from lead 2 of U1 and provides an inversion of its input signal at lead 2 to the LO drive waveform at lead 4. U1 does not have this inversion feature so its LO waveform is in phase with pin 2. When driven in this fashion, it is apparent that Q1 and Q4 conduct together and on the other half cycle Q2 and Q3 conduct together. The resultant output square wave has the same RMS value as the DC bus voltage (400 VDC). The lamp circuits are resonant at the self-oscillating frequency of U1 determined from equation (1).

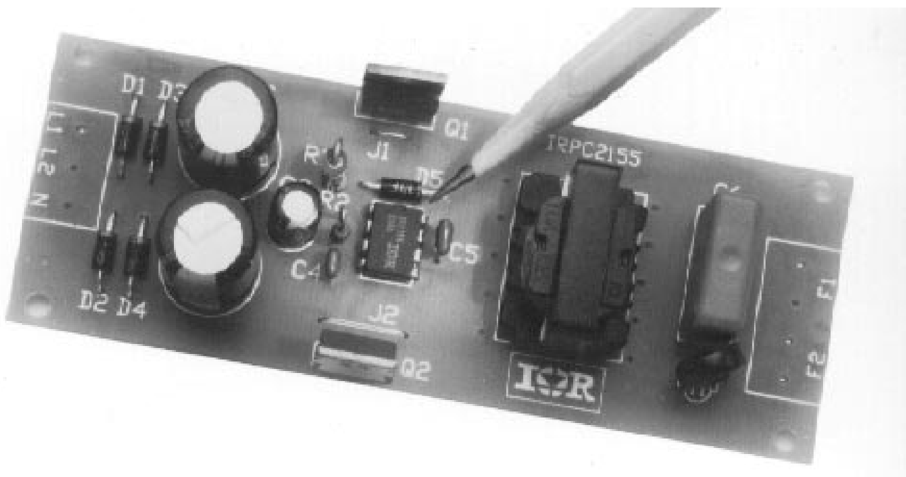
The low-Q lamp circuits have a broad resonance curve so that tolerance buildups of the timing components R1 and C3 do not seriously compromise the available striking voltage for each lamp. Even with a Q of only 2, the RMS lamp striking voltage exceeds 800V — more than sufficient to strike the F96T12 lamps.

Also shown on the schematic is a power factor correction circuit following the AC input rectifier. These circuits use a boost topology to achieve in-phase AC sinusoidal current waveforms with low harmonic content, and are becoming nearly universally required, particularly at higher power levels. Many papers have been presented on the subject and a few semiconductor manufacturers provide control chips and application information on their use.

The ballast circuit will operate with or without a P.F.C. rectifier; the simplest approach being a configuration similar to the ‘Simple Double 40 Ballast’ circuit. If this option is used, the DC bus voltage is around 320 VDC and the values of L2 and L3 should be reduced by 25% to around 1 mH (by increasing the core gap.) The R1 value should also be reduced to provide the slightly higher resonant frequency now required.

Summary

This application note has described a few ballast circuits which are easily implemented with International Rectifier’s IR215X Control IC family. Additional possibilities are limited only by the imagination of the designer.



This PC board (shown actual size) is designed to drive a 13W to 40W fluorescent lamp using the IR2155, IR2151 or IR2152.) Input is 115 or 230 VAC. (Schematic, parts list and board available on request. Ask for Design Tips DT 94-3.)

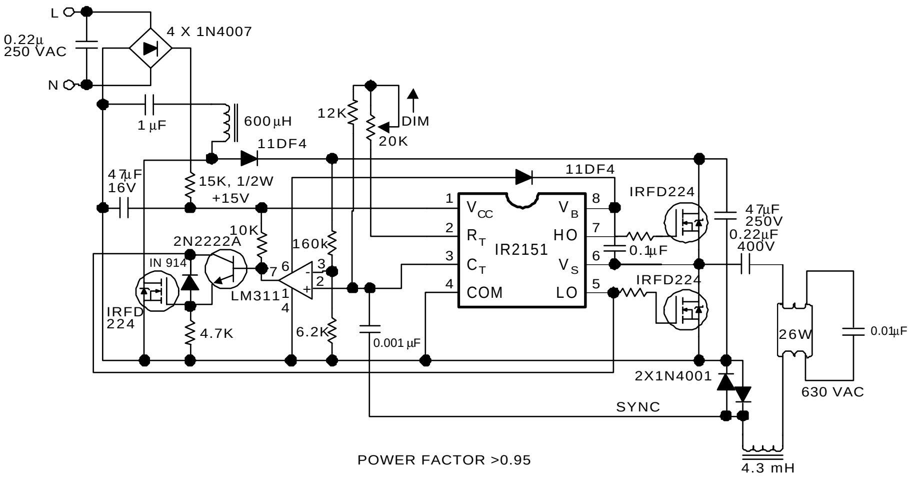
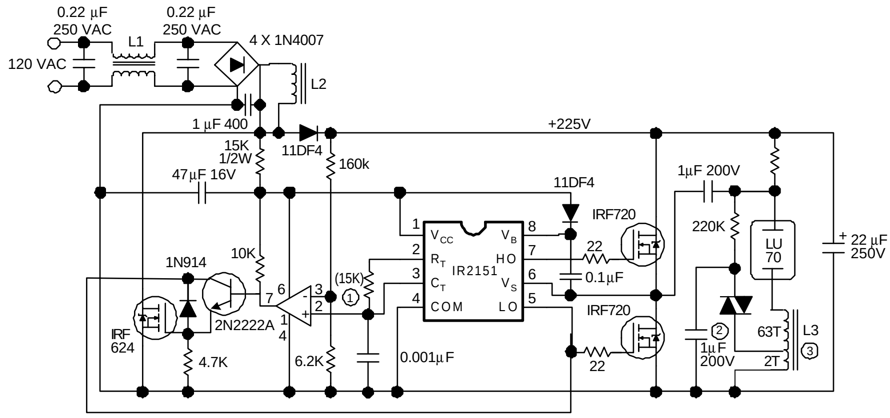


Figure 9. Compact fluorescent with dimming and bus voltage control



NOTE ① Select value for 70 watt lamp power.
② Polypropylene capacitor
③ Adjust gap for L = 400µH

L1 Core: Micrometais #T106-26
Wind: 18T Bilifar #18 AWG HAPT
L = 2 x 30µH

L2 Core: TDK #EE-30Z
Bobbin: TDK #BE-30-1110CP
Wind: 64T #22 AWG HAPT
L = 720 mH with approximately
0.035 inch gap spacer

L3 Core: Phillips EC-35-3C81
Bobbin: Special 3-Slot bobbin (see Figure 11)
Wind: 63T #22 AWG HAPT (21T/slot)
2T #20 AWG HAPT wind over
Low voltage (near ground) slot
Connect 2-turn winding in series
with 63-turn winding.

Figure 10: 70 watt HPS ballast

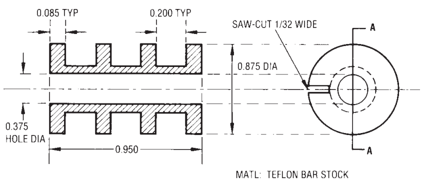


Figure 11: Section A-A view of L3 bobbin

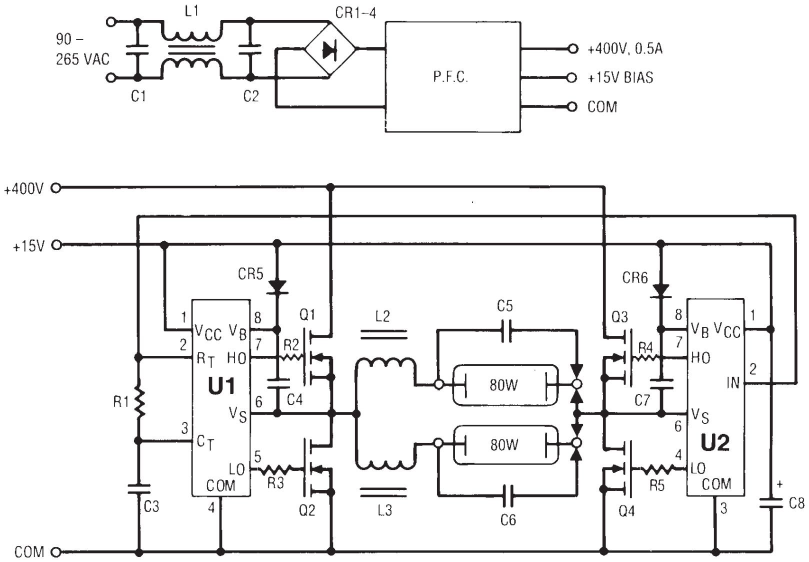


Figure 12. Full bridge 160 watt fluorescent ballast

Component List

U1	IR2151
U2	IR2111
Q1, Q2, Q3, Q4	IRF820
CR1, CR2, CR3, CR4	1N4007
CR5, CR6	10DF6
R1	15K, ¼W
R2, R3, R4, R5	22Ω, ¼W
C1, C2	0.22 µF 250VAC
C3	0.001 µF, 50V
C4, C7	0.1 µF, 50V
C5, C6	0.01 µF, 1600V, Polypropylene
C8	47 µF, 16V, Aluminum Electrolytic
L1	Core: Micrometals # T106-26 Wind: 18T BIFILAR # 18 AWG HAPT L = 2 x 30 µH
L2, L3	Core: TDK # EE-30Z Bobbin: TDK # BE-30-1110CP Wind: 64T # 22 AWG HAPT L = 1.35mH with 0.01 inch Gap Spacer
P.F.C.	Motorola MC34262 Data Sheet Figure 20 Schematic or equivalent from Unitrode, Micro Linear, SGS Thompson, Cherry, etc.