CHAPTER 8

Lighting

Fluorescent Lamp Control

8.1.1 Efficient Fluorescent Lighting using Electronic Ballasts

This section provides a general background to fluorescent lamps and their control requirements, with emphasis placed on high frequency electronic ballasts and their advantages over conventional 50/60Hz "magnetic" ballasts. Simplified examples of popular electronic ballast topologies suitable for low cost / economy applications are introduced.

The fluorescent lamp.

A fluorescent tube is a low pressure mercury vapour discharge lamp containing an inert gas consisting of argon or krypton at low pressure (below 1 atmosphere) plus a small measured dose of mercury. There is a filament at each end which, when hot, emit electrons to sustain the discharge when the lamp is operating. The mercury vapour discharge produces ultraviolet light which is converted to visible light by the phosphors coating the inside of the glass tube. The glass blocks the exit of the ultraviolet radiation but allows the visible radiation through. See Fig. 1.

Fluorescent tubes exist in many shapes and sizes. Apart from the many compact types that have appeared on the market in recent years as energy efficient replacements for incandescent lamps, the traditional linear tubes range from 150mm 4W up to the very high output 2400mm 215W.

Modern fluorescent tubes incorporating the latest triphosphor technology (i.e. red, green and blue phosphors similar to those used in modern high brightness television picture tubes) possess efficacies of around 80 lumens per lamp Watt compared with 68 lumens per lamp Watt for the older most efficient "white" fluorescent tubes and around 12 lumens per Watt for an incandescent bulb. Moreover, the triphosphor lamps reveal colour and skin tones more accurately than the standard "white" lamps, which suffer from a deficiency in output at the red end of the spectrum. This results in a greenish hue and a suppression of red colours from anything illuminated by them.

The elimination of the traditional causes of criticism for fluorescent lighting means that this form of lighting is becoming more acceptable in wider applications than ever before. Adjustment of the ratios of the three phosphors can create colour appearances from a very warm, intimate, incandescent equivalent colour temperature of 2700K through the cool, clean, businesslike 4000K to the very cool daylight colour temperature of 6500K, all with high efficacies and good colour rendering properties. Before the availability of triphosphors, these qualities have always been mutually exclusive. You could either have high efficacy and poor colour rendering or poor efficacy and good colour rendering, but not both.

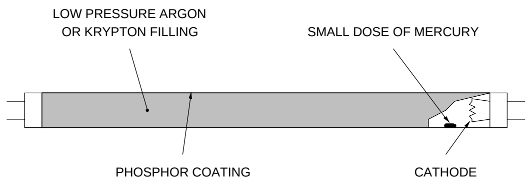


Fig. 1. A fluorescent tube.

A non-operating fluorescent tube will appear as an open circuit, since there is no electrical connection from one end to the other. In order to "strike the arc", a high voltage must be applied across the lamp in order to ionise the gas within. This will instantly "cold start" the lamp and shorten its life by sputtering electron-emitting material from its cathodes.

However, if the cathodes (heaters) are first preheated to generate a space charge of electrons at each end of the lamp, the strike voltage is considerably reduced and lamp life will not be unduly compromised by the start-up.

As soon as arc current flows, the lamp’s electrical impedance will drop. It now exhibits a negative impedance characteristic, where an increase in current is accompanied by a reduction in lamp voltage. There must therefore be a current limiting device in circuit to prevent the rapid onset of runaway and destruction of the lamp.

The lamp running current should ideally be sinusoidal to minimise the radiation of electromagnetic interference from the lamp and its supply wires. Sinusoidal lamp current also maximises lamp life. A peak current approaching twice the RMS current will prematurely deplete the electron emitting material from the lamp cathodes. (For a sinewave the peak value is only 1.414 times the RMS value.)

There should also be no D.C. component to the lamp current; that is, the positive and negative half cycles should be of equal duration. If this is not the case, the resulting partial rectification will result in premature depletion of the electron emitter from one of the lamp cathodes.

The ballast.

The requirements of a fluorescent lamp ballast are to:
(a) Preheat the cathodes to induce electron emission.
(b) Provide the starting voltage to initiate the discharge.
(c) Limit the running current to the correct value.

There are several types of mains frequency "magnetic" ballast available. By far the most common circuit for 230V mains supplies has traditionally been the switchstart ballast (see Fig. 2), where lamp ballasting is provided by the choke. Other circuits include, in order of popularity, the semi-resonant circuit and the quickstart circuit.

The switchstart circuit has been widely adopted because of its simplicity, low cost and improved efficiency when compared with the alternative options mentioned above. Another reason is that the 230V mains voltage is sufficiently higher than the tube running voltage to allow the use of the simple series impedance ballast in almost all cases. Where this is not possible, for example in most 120V supplied circuits, the lamp is controlled by a quickstart circuit incorporating voltage step-up.

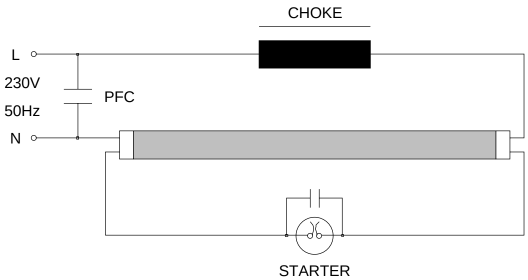


Fig. 2. Conventional switchstart "magnetic" ballast circuit.

Switchstart ballast operation.

When the voltage is applied to the circuit, the lamp does not operate at first, so the full mains voltage appears across the starter via the choke and lamp cathodes.

The starter consists of bi metallic contacts sealed within a small discharge bulb with an inert gas filling such as argon or neon. The mains voltage causes a glow discharge within the starter which heats up the bi metallic contacts, causing them to close. This completes the circuit and allows preheat current to flow through the choke and both cathodes.

Since the glow discharge within the starter has now ceased, the bi metallic contacts cool down and open. Because the inductance of the choke tries to maintain current flow, the voltage across the lamp rises rapidly and strikes the lamp. If it does not, the starter's contacts close again and the cycle repeats.

Once the lamp has started, the choke controls its current and voltage to the correct levels. The lamp running current is enough to keep the cathodes (heaters) hot and emitting electrons without the need for separate heater supplies, which would otherwise be wasteful of energy. Since the lamp's running voltage is much lower than the mains voltage, there is now not enough voltage to cause a glow discharge in the starter, so it remains open circuit.

The power factor correction (PFC) capacitor draws leading current from the mains to compensate for the lagging current drawn by the lamp circuit.

Why electronic ballasts?

Electronic ballasts have been available for well over a decade. Recent leaps in performance, coupled with ever increasing energy costs, the increased awareness of the advantages they offer, the increasing environmental awareness of the consumer, and the increased acceptability of the new fluorescent light sources in existing and new applications, have seen an upsurge in electronic ballast use since the beginning of the 1990's.

Replacing the most efficient low loss mains frequency switchstart ballast with an electronic ballast leads to reduced energy consumption and improved performance. The reasons for this are detailed below.

Increased light output.

If the operating frequency is increased from 50Hz to above the audible limit of 20kHz, fluorescent lamps can produce around 10% more light for the same input power (see Fig. 3). Alternatively, the input power can be reduced for the same light output.

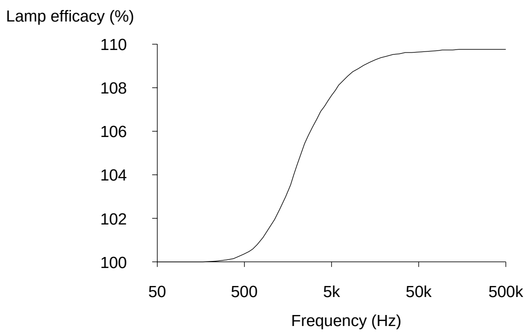


Fig. 3. Typical fluorescent lamp efficacy.

Flicker eliminated.

A fluorescent lamp operating at 50/60Hz will extinguish twice every cycle as the mains sinewave passes through zero. This produces 100/120Hz flicker which is noticeable or irritating to some people. It will also produce the well-known and potentially dangerous stroboscopic effects on rotating machinery.

If the lamp is operated at high frequency, however, it produces continuous light. This is because the time constant and hence the response time of the discharge is too slow for the lamp to have a chance to extinguish during each cycle.

The output waveform of an electronic ballast will usually be slightly modulated by 100/120Hz "ripple". Provided this is kept to a reasonable level by filtering within the ballast, the drawbacks associated with 100/120Hz flicker are eliminated.

Audible noise eliminated.

Since electronic ballasts operate above the audible range, they do not suffer from the audible noise problems that can occur with mains frequency magnetic ballasts. The familiar buzzing noise is caused by vibrations in the laminations and coil of the choke. This can then excite vibrations in the steel body of the fitting which effectively amplifies the original noise.

Lower ballast power.

An electronic ballast will consume less power and therefore dissipate less heat than a mains frequency magnetic ballast. For example, for two 1500mm 58W energy-saving lamps, the typical ballast power dissipations might be 13W per ballast for two 50Hz magnetic ballasts compared with 9W for a single electronic ballast driving two lamps.

The energy-saving benefits of electronic ballasts have made it possible to obtain the same light output from fluorescent lamps as would be obtained using a conventional 50/60Hz magnetic ballast, for a total circuit power (i.e. lamp and ballast) that is actually less than the rated lamp power alone. This is due to two reasons.

Firstly, the lamp can be underrun at high frequency for the same light output. Secondly, the power consumed by the ballast can be so low that the total circuit power is still less than the rated power printed on the lamp. Because of this, energy cost reductions of 20 - 25% are achievable.

Extended lamp life.

An electronic ballast which "soft starts" the lamp (i.e. provides preheat to the cathodes before applying a controlled starting pulse) will dislodge a minimum quantity of material from the cathodes during starting. This will give longer lamp life when compared to the uncontrolled impulses to which the lamp is subjected in a switchstart circuit.

Versatile lamp control.

Electronic ballasts are available which permit lamp dimming. This gives substantial energy savings in situations where the lights are linked to an automatic control system which detects ambient light levels and adjusts lamp output to maintain a constant level of illumination. Lights may also be programmed to dim during intervals when areas are not in use, for example during lunch breaks.

Electronic ballasts can incorporate feedback to detect the operating conditions of the lamp(s) so that failed lamps can be switched off to avoid annoying flicker and possible ballast damage. They can also incorporate regulation, whereby a constant light output is maintained over a range of input voltages. Operation can be either from AC or DC supplies for emergency lighting applications.

Compact and light weight.

Owing to the high frequency of operation, the magnetic components in an electronic ballast are compact and lightweight with cores of ferrite material, whereas at mains frequency the ballast choke must be larger and heavier with bulkier copper windings and a core of laminated steel.

The shape and geometry of a mains frequency choke is determined by magnetic efficiency requirements, whereas the circuitry within an electronic ballast can be arranged to produce a very slim final package. This permits new levels of slimness and compactness for the final ballast.

Electronic ballast topologies.

The typical building blocks of an electronic ballast are shown in Fig. 4.

An increasing number of electronic ballasts are employing active power factor correction in the form of a boost converter between the rectifier and DC filter stages. (Figure 5 shows a simplified boost converter arrangement.) This obliges the ballast to draw current over most of each mains half cycle instead of the usual current spike that a rectifier / DC filter would demand at each peak of the voltage waveform. This reduces the harmonic content of the current and improves the power factor. It will also reduce the size of the electromagnetic interference (EMI) filter required, since filtering is now required at the higher harmonic frequencies of the boost converter switching frequency instead of at the mains frequency and harmonics of it.

Electronic ballasts take many forms. The simplest and most economical form might consist of a free-running self-oscillating circuit using bipolar transistors. This would be an open loop circuit (i.e. no feedback to detect lamp operating conditions).

More expensive options might contain a controlled oscillator in a closed loop circuit using MOSFETs. Here, features could include regulation for varying AC and DC supply voltages, adjustable lamp brightness, soft starting and a mechanism to detect and shut down failed lamps.

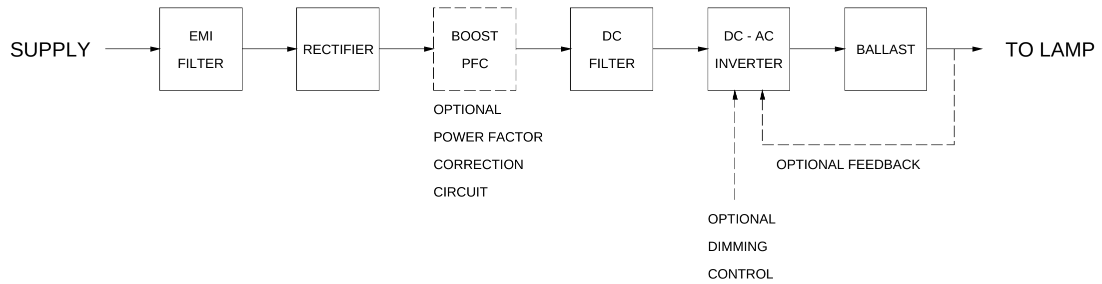


Fig. 4. Electronic ballast block diagram.

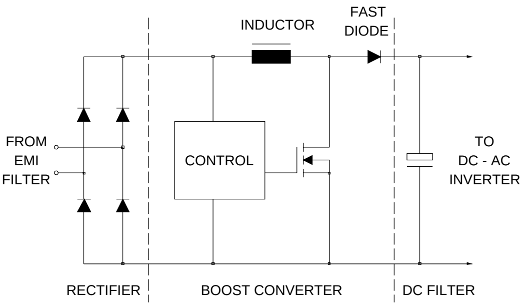


Fig. 5. A boost converter for active power factor correction.

Blocking oscillator.

The most basic form of electronic ballast uses a blocking oscillator as shown in Fig. 6. Its use is restricted mainly to low voltage DC, low power ballasts as used in handlamps, leisure lighting and emergency lighting, where operation is only for short periods. This is because the lamp has a severely limited life when it is driven by a spiky waveform, rich in harmonics, such as that produced by this circuit. This topology might typically be used to operate tubes of 4W to 13W ratings only because of the excessive voltage and current stresses and switching losses that would be experienced by the transistor in higher power mains voltage versions.

Voltage step-up to drive the lamp from the low voltage supply is achieved by the turns ratio of the transformer primary and secondary, while oscillation is maintained by the positive feedback supplied by the auxiliary winding connected to the transistor's base. The values of R, C, transformer primary inductance L_{PRI} and the transistor parameters set the oscillation frequency and the mark / space ratio of the waveform (which should be 1:1 for the reason given in the first section).

No separate ballast inductor is required, since the only energy delivered to the lamp during the transistor's OFF time is what was stored in L_{PRI} during the preceding ON time. The transistor remains OFF and will not turn ON again until all the stored energy has been delivered to the load. Lamp power is therefore controlled by the amount of energy stored in the L_{PRI} during each ON period.

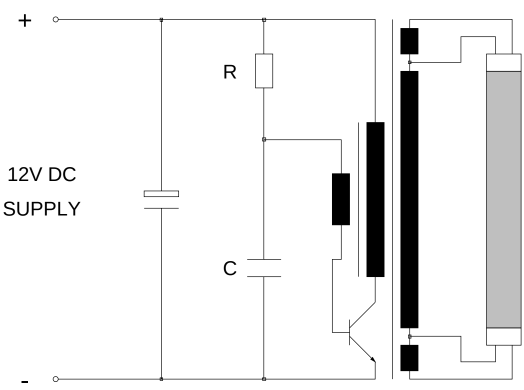


Fig. 6. Basic low voltage ballast using a blocking oscillator.

Unlike the blocking oscillator, mains powered electronic ballasts usually use two switching power transistors in a push pull or half bridge configuration. This can either be a self oscillating or a driven oscillator circuit. The driven oscillator option permits easier lamp control and dimming. The self oscillating option has cost advantages where the benefits of high frequency lighting are required without the necessity for lamp dimming.

The push pull inverter.

A push pull circuit can appear as a voltage fed inverter with series resonant load or a current fed inverter with parallel resonant load. In both cases a centre tapped transformer is required.

Voltage fed push pull inverter.

Figure 7 shows a simplified circuit. This example provides isolation of the output from the mains supply with a separate secondary winding.

In the voltage fed arrangement, the D.C. rail voltage is fed straight to the centre tap. Both ends of the winding are connected to zero volts via transistors, which are alternately switched on during operation. The alternate passage of current in opposite directions through each half of the primary winding induces a square wave voltage across the secondary.

Since the full D.C. rail voltage appears across half the primary winding at a time, twice this voltage will appear across the whole primary winding. This means that during each transistor's "off" period, it will experience a maximum theoretical V_{CE} of 2 x D.C. rail voltage.

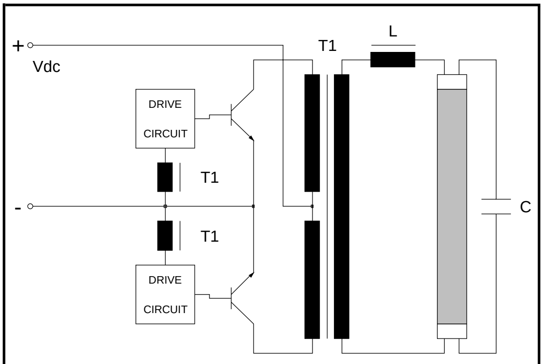


Fig. 7. Voltage fed push pull ballast with isolated output.

When power is first applied, the secondary voltage should not be high enough to cold start the lamp, which should remain in the high impedance state. The only current flowing will be through the series resonant combination of L & C, and both lamp cathodes. This preheat current will be enough to initiate electron emission from the cathodes which will in turn lower the lamp striking voltage to a point where the voltage across the capacitor can then start the lamp (usually within a second).

After starting, the lamp voltage will drop and the current will be limited and filtered by L. C will help to filter out residual harmonic frequencies and its current will fall to negligible proportions at the fundamental operating frequency. The resulting lamp current will closely resemble a sinewave.

The transistor base drives are derived from auxiliary windings on the transformer which provide the necessary positive feedback. An advantage with this transformer-based arrangement is the isolation it provides between the lamp and the mains supply.

Current fed parallel resonant push pull inverter.

The main difference with this circuit over the previous one is that the D.C. rail voltage is fed to the transformer centre tap via an inductor which acts as a current source. A capacitor C across the transformer primary forms a parallel resonant load in combination with the primary winding inductance (see Fig. 8). Instead of a square wave as in the voltage fed circuit, a full wave rectified sinewave appears at the centre tap whose theoretical peak amplitude is $\pi/2$ x V_{DC}. Twice this amplitude appears across the whole winding for the same reason as in the voltage fed push pull circuit. Therefore the maximum theoretical $V_{CE} = \pi$ x V_{DC}.

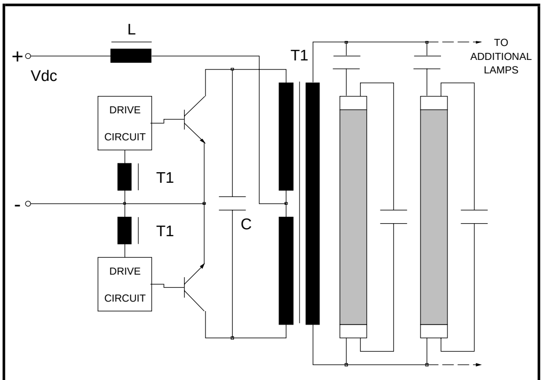


Fig. 8. Current fed parallel resonant push pull ballast with isolated output.

Since each successive half sine produces current flow in opposite directions through the two half windings, a sinewave is produced across the whole winding whose peak to peak amplitude is 2π x V_{DC}.

The additional cost of the inductor might be regarded as a disadvantage. However, the beauty of current fed parallel resonant circuits, of which this is one example, is that they naturally produce a sinusoidal output, so selection of the ballast components for their harmonic filtering properties is no longer so important. This allows the use of a series ballast capacitor instead of the series L normally required.

Another benefit with this type of circuit is its ability to continue normal operation with varying or open circuit loads. This permits independent operation of parallel-connected lamps across the secondary, each with its own ballast capacitor, where failure of one or more lamps will not affect the operation of the remaining lamps. This is unlike series-connected lamps, where the failure of one tube will disable all the tubes on that ballast.

Sinusoidal output topologies are very popular in the self oscillating low cost ballast market because of these advantages and the circuit simplicity.

The half bridge inverter.

The half bridge topology contains two npn transistors connected in series across the D.C. rail with the load connected to their mid point. The half bridge is so called because the return path for the load current is provided by two series-connected capacitors across the D.C. rail. (A full bridge circuit would have transistors in these positions also, but this arrangement is rarely used in electronic ballasts for fluorescent lamps. Although the required voltage rating of the transistors would be halved, this would not compensate for the increased cost of four power transistors instead of two, and the extra complication of controlling the timing of the switching of all four transistors.)

The two capacitors, which have a very low reactance and are essentially a short circuit at the ballast operating frequency, create a mid-point A.C. reference between the D.C. rails. This blocks the D.C. offset equal to half the rail voltage that would be applied to the lamp if the return path were merely taken to one of the rails.

Current fed parallel resonant half bridge inverter.

Figure 9 shows the simplified circuit. Transformer isolation is provided, and the sinusoidal output permits the use of ballast capacitors as for the current fed push pull topology. The series inductance L in each power supply line acts as the current source.

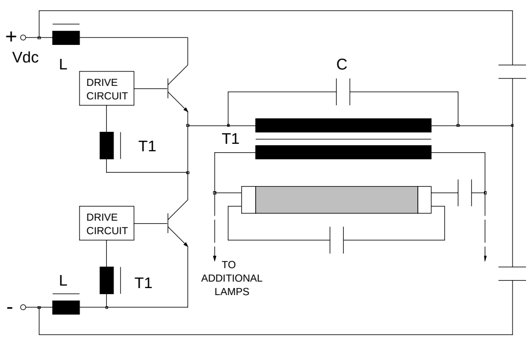


Fig. 9. Current fed parallel resonant half bridge ballast with isolated output.

As each transistor conducts in turn, the current fed resonant load causes alternate polarity half sinewaves with peak voltages of $\pi/2 \times V_{DC}$ to appear at one end of the transformer primary. Each half sine appears across the non-conducting transistor. Therefore the maximum theoretical $V_{CE} = \pi/2 \times V_{DC}$.

The sum of these half sines produces a full sinewave with a peak to peak amplitude of $\pi \times V_{DC}$. However, as the return current flows to the A.C. half rail created by the half bridge capacitors, only half this voltage appears across the primary, resulting in a peak to peak primary voltage of $\pi/2 \times V_{DC}$.

Voltage fed half bridge inverter.

See Fig. 10. This circuit does not employ a transformer so output isolation is not provided. Feedback to drive the transistors is now supplied from two auxiliary windings on the current transformer CT1 in the lamp current path.

As this is a voltage fed circuit whose output is not naturally sinusoidal, lamp starting, ballasting and waveform shaping are provided by the series L and parallel C as for the voltage fed push pull circuit.

In the voltage fed half bridge circuit, since the transistors are "firmly anchored" to the supply rails without any current source series inductance, they will experience a maximum theoretical V_{CE} equal to the D.C. rail voltage.

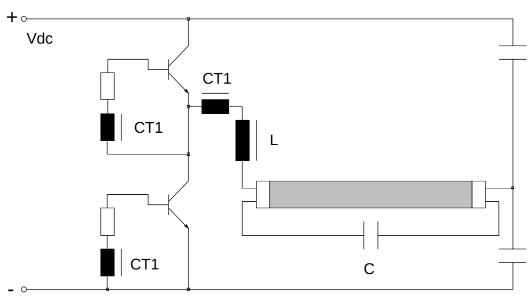


Fig. 10. Voltage fed half bridge ballast.

Variation on the voltage fed half bridge circuit.

A variation on this circuit is shown in Fig. 11, where the two half bridge capacitors are replaced by the single D.C. blocking capacitor C2. This enables the load to be returned to the positive D.C. rail.

The circuit operates as follows:
On initial power-up, before the lamp has struck, C1, L and C2 form a series resonant circuit. C2 is larger than C1 so it looks like a short circuit compared to C1. C1 therefore dominates and dictates the resonant frequency in

combination with L. A high voltage is developed across C1 at resonance which starts the tube. At this point the tube voltage across C1 collapses and C2 then takes over in dictating a lower running frequency in combination with L.

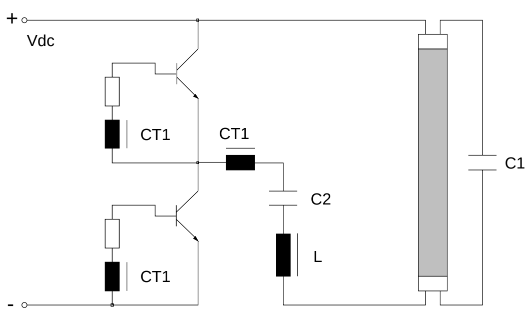


Fig. 11. Variation on voltage fed series resonant half bridge circuit.

This circuit is the one most commonly used in the electronically ballasted compact fluorescent lamps and it lends itself to driven as well as self oscillating circuits.

Summary.

The circuit examples presented in this Publication all use bipolar transistors, mainly for cost advantage reasons, especially where high voltage devices up to 1000V rating and above are required. Ballast manufacturers have perfected many good, reliable designs using such devices in circuits based on the simplified topologies shown.

Popular topologies for low cost electronic ballasts have proved to be the current fed parallel resonant circuits. To summarise the reasons for this, they naturally produce the ideal sinewave output. This permits the use of simple ballast capacitors instead of inductors. The circuits also maintain safe operation with abnormal load conditions. Lamps can be operated in parallel, where the failure of one or more lamp will not disable the remaining lamps.

The current fed topologies require higher voltage transistors than the voltage fed topologies. For example, for the current fed half bridge topology, allowing for safety margins of around 400V for voltage spikes at start-up and 110% mains voltage, a 120V ballast would require transistors with typical voltage ratings of at least 700V. The ratings for 230V mains would typically be at least 950V, and for 277V mains typical voltage ratings of at least 1100V would be required.

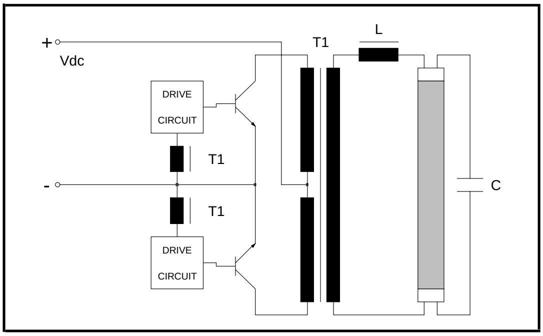
The ratings for a current fed push pull topology would be 1000V, 1500V and 1700V respectively.

8.1.2 Electronic Ballasts - Philips Transistor Selection Guide

Section 8.1.1 provides an introduction to fluorescent lamps and the circuits required to operate them for maximum life and efficiency. Several simplified electronic ballast topologies are introduced.

This section lists those topologies with the theoretical voltage demands they place on the transistors, together with a selection table of suitable Philips transistors.

a) Voltage fed push pull inverter.



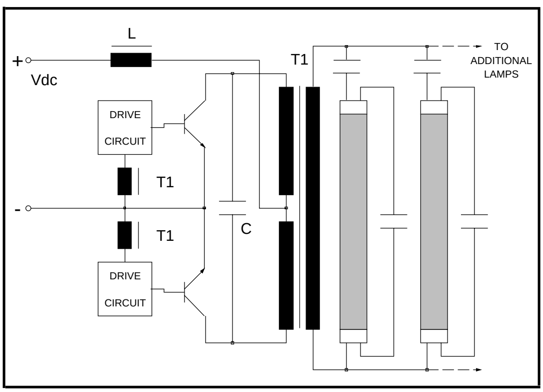
The D.C. rail voltage appears at the transformer centre tap.

Therefore $V_{c.t.} = V_{DC}$.

Half of the transformer's primary winding is energised with the full D.C. rail voltage at any one time. Therefore twice this voltage will appear across the whole winding (autotransformer effect). This voltage appears across each transistor in turn when it is non-conducting. So, during stable circuit operation and neglecting unforeseen voltage spikes:

$$V_{CE(max)} = 2 \times V_{DC}.$$

b) Current fed push pull inverter.



The transformer centre tap is no longer connected directly to the D.C. rail. The voltage developed across the series inductor L as each transistor conducts results in a positive half sinewave at the centre tap whose average voltage is equal to the D.C. rail voltage. A half sine instead of a rectangular pulse is produced because of the resonant nature of the load.

Therefore $V_{c.t.(ave)} = V_{DC}$.

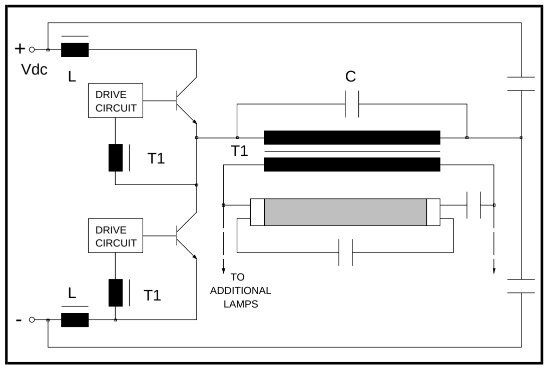
The peak value of this waveform can be shown by integration to be $\pi/2$ x its average value.

Therefore $V_{c.t.(pk)} = \pi/2 \times V_{c.t.(ave)} = \pi/2 \times V_{DC}$.

Each successive half sine is conducted through alternate halves of the primary, so twice this amplitude appears across the full primary. This gives a peak voltage of twice the peak centre tap voltage appearing across the non-conducting transistor (as for the voltage fed push pull circuit), so:

$$V_{CE(pk)} = \pi \times V_{DC}.$$

c) Current fed half bridge inverter.



The transformer primary is driven from one end by the collector-emitter junction point of the two transistors. If this were a voltage fed circuit without any series L, the primary would be alternately connected to the positive and negative rails by the alternate transistor switching to produce a square wave with a peak to peak amplitude of V_{DC}. However, because this is a current fed resonant circuit, the conduction of each transistor will produce a half sine whose average voltage is equal to the D.C. rail voltage.

Therefore $V_{(ave)} = V_{DC}$.

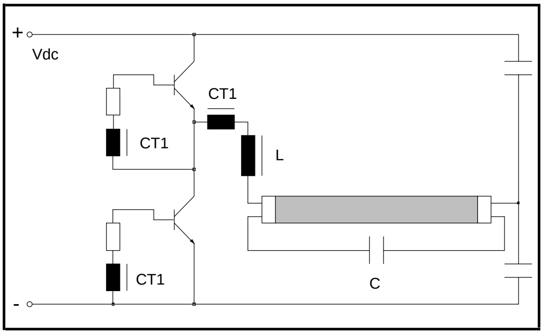
By integrating it can be shown that the half sine will have a peak amplitude of $\pi/2$ x its average value.

Therefore $V_{(pk)} = \pi/2 \times V_{(ave)} = \pi/2 \times V_{DC}$.

This voltage appears across the non-conducting transistor, so:

$$V_{CE(pk)} = \pi/2 \times V_{DC}.$$

d) Voltage fed half bridge inverter.



As the transistors are now connected directly to the D.C. rails, their alternate switching will switch the transformer primary between the D.C. rails only.

Therefore $V_{(max)} = V_{DC}$.

As this voltage appears across the non-conducting transistor:

$$V_{CE(max)} = V_{DC}.$$

Transistor selection guide.

This guide lists suitable transistors with maximum recommended output powers for the different topologies. It assumes that the ballast's D.C. rail is obtained from rectified and smoothed A.C. mains. If boost power factor correction is included which boosts the D.C. rail voltage to around 400V irrespective of mains voltage, the suggested transistors for 277V mains should be selected.

	TOPOLOGY:	a) V. fed P.P.		b) C. fed P.P.		c) C. fed H.B.		d) V. fed H.B.	
A.C. SUPPLY:	120V	BUW84/85	35W	BUX87P	13W	BUW84/85	25W	BUW84/85	15W
		BUX84/85	35W	BUX85	55W	BUX84/85	25W	BUX84/85	15W
		BUT211	90W	BUT11A	140W	BUT211	70W	BUT211	40W
		BUT18A	110W	BUT18A	170W	BUT18A	80W	BUT18A	55W
		BUT12A	140W	BUT12A	230W	BUT12A	110W	BUT12A	70W
		BUW12A	140W	BUW12A	230W	BUW12A	110W	BUW12A	70W
	230V	BUX87P	15W	BU1706A	230W	BUX87P	13W	BUW84/85	30W
		BUW85	70W	BU1706AX	230W	BUW85	55W	BUX84/85	30W
		BUX85	70W	BU508A	360W	BUX85	55W	BUT211	80W
		BUT11A	170W			BUT11A	140W	BUT18A	100W
		BUT18A	210W			BUT18A	160W	BUT12A	140W
		BUT12A	280W			BUT12A	220W	BUW12A	140W
		BUW12A	280W			BUW12A	220W		
	277V & most boosted designs	BU1706A	170W	BU1706A	260W	BU1706A	130W	BUW84/85	40W
		BU1706AX	170W	BU1706AX	260W	BU1706AX	130W	BUX84/85	40W
		BU508A	280W			BU508A	220W	BUT211	100W
								BUT18A	125W
								BUT12A	170W
								BUW12A	170W

8.1.3 An Electronic Ballast - Base Drive Optimisation

This section investigates the transistor base drive circuit in a current fed half bridge ballast. (Fig. 1 shows the simplified circuit.) The effect on switching waveforms of progressing from a simple base drive circuit to the optimised solution will be shown.

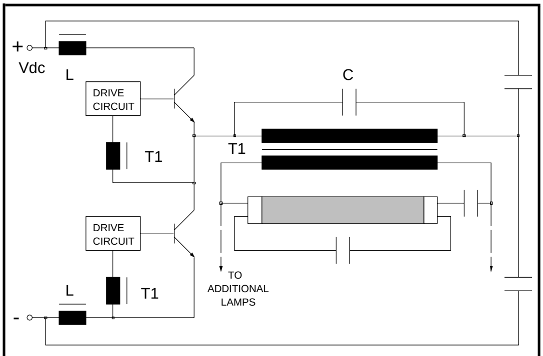


Fig. 1. Current fed half bridge ballast.

Base drive requirements.

1. Each transistor must not be overdriven and oversaturated when conducting otherwise excessive base power dissipation will result. The time will also be increased in bringing the transistor out of saturation during turn-off, leading to increased switching losses.

2. The transistor must not be underdriven because this will result in excessive collector-to-emitter voltage (V_{CE}) during conduction, leading to excessive ON-state losses or inability to sustain oscillation. However, because the transistor is unsaturated, there will be less charge to extract from the base, resulting in a shorter storage time and faster turn-off.

3. Reliable and correct circuit operation should be maintained for all expected transistor gains, maximum and minimum load, maximum and minimum supply voltage and all component tolerances.

Base drive optimisation.

The transformer's auxiliary windings which provide base drive might contain just one or two turns each. In order to provide rapid transistor turn-off, their peak loaded output voltage would need to be such that the transistor 'sees' a turn-off voltage of around minus 5V. An approximation to this drive voltage could be arrived at empirically by increasing the number of auxiliary turns one by one. Any final voltage adjustment, if necessary, can be achieved by varying the base drive components.

Simple base drive.

In order to meet the requirements of non-saturation and rapid turn-off, the simplest base drive might consist of a resistor to limit the positive base current and a Schottky diode in parallel with it to discharge the base as quickly as possible. See Fig. 2.

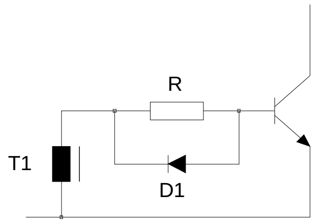
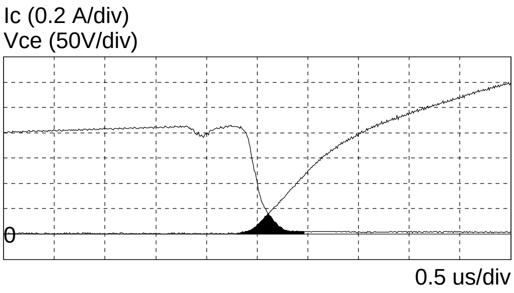


Fig. 2. Simple base drive.

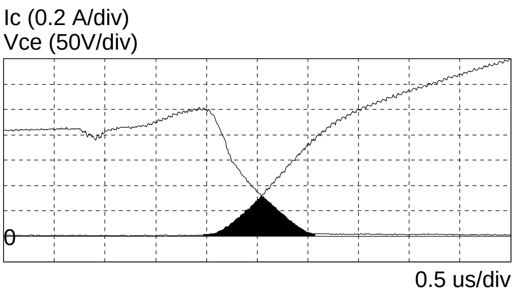
A Schottky diode is specified for its fast switching and low forward voltage drop to best meet the rapid turn-off requirements. A 1A 40V device such as the BYV10-40 is ideally suited.

If the resistor is selected empirically so that the transistor is barely saturating, this simple circuit will work, but only for a given load current, supply voltage, transistor gain and base drive voltage from the transformer auxiliary winding. Altering any of these conditions will either cause underdriving of the transistor and, ultimately, cessation of oscillation, or else the transistor will be overdriven, causing increased collector current fall time and excessive switching losses.

For example, the resistor value was optimised for transistors with low gain limits. Fig. 3 shows the resulting I_C fall at transistor turn-off, while Fig. 4 shows the effect of replacing the transistor with a high gain limit sample. The shaded areas bounded by the I_C and V_{CE} curves represent transistor power dissipation during switching.



Simple base drive.
Fig. 3. Low h_{FE}. I_C fall with V_{CE}.



Simple base drive.
Fig. 4. High h_{FE}. I_C fall with V_{CE}.

Improved circuit.

What is required is a means of providing enough base drive under worst case conditions of maximum load current, minimum supply voltage, minimum transistor gain and minimum base drive voltage, while avoiding excessive saturation in the opposite condition. This can be achieved by diverting excess positive base drive current into the collector path when the transistor is fully turned on. This requirement is partly met by a Baker Clamp arrangement as shown in Fig. 5.

When the transistor is fully conducting, V_{CE} will be at a minimum. This will bring V_C close to V_B so that any excess base drive will then flow through anti saturation diode D2 to the collector. As a first approximation, the single resistor R is divided equally into two and D2 taps its voltage from the mid point. Figs. 6 and 7 show the resulting I_C fall waveforms. Considerably reduced transistor saturation is evident.

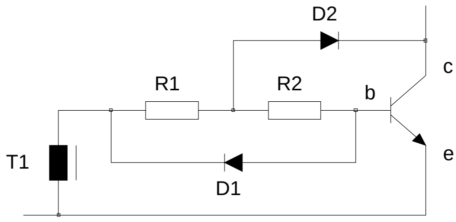
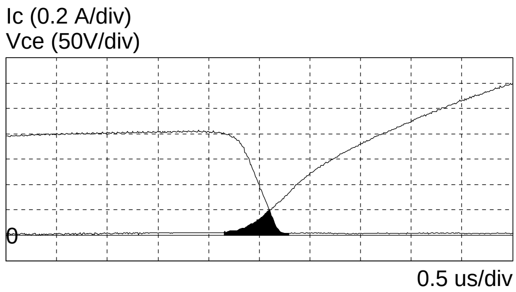
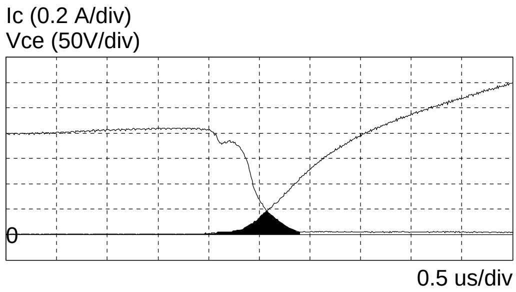


Fig. 5. Improved base drive.



Improved base drive.
Fig. 6. Low h_{FE}. I_C fall with V_{CE}.



Improved base drive.
Fig. 7. High h_{FE}. I_C fall with V_{CE}.

With regard to the base waveforms, where the simple circuit produces more base drive current than is necessary, as shown in Fig. 8, the improved circuit reduces this to that shown in Fig. 9.

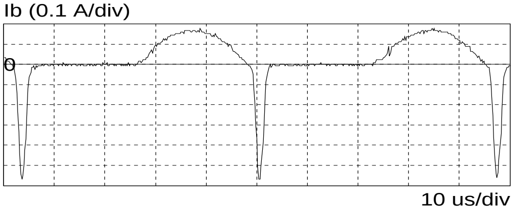


Fig. 8. Simple base drive.
High h_{FE}. I_B.

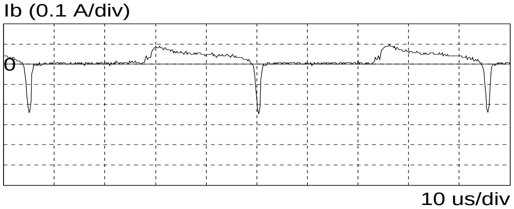
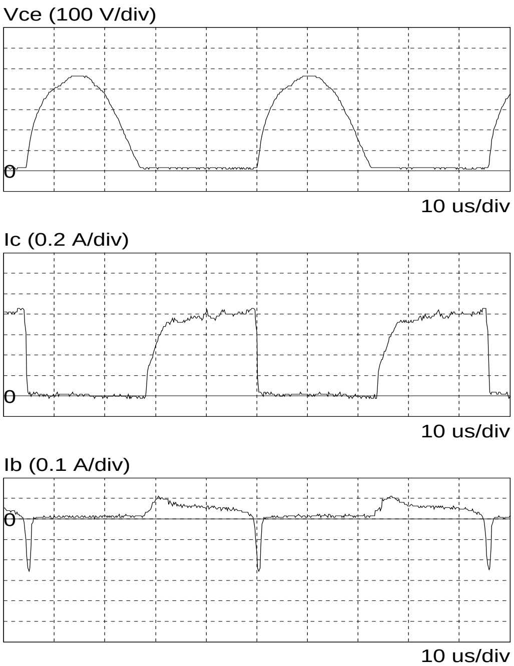


Fig. 9. Improved base drive.
High h_{FE}. I_B.

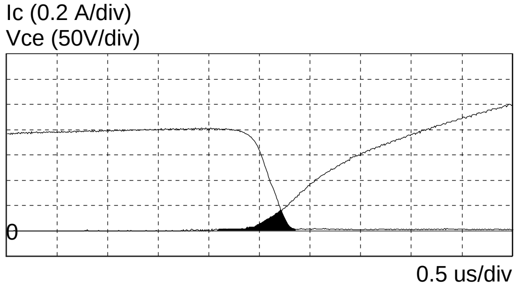
Optimised base drive.



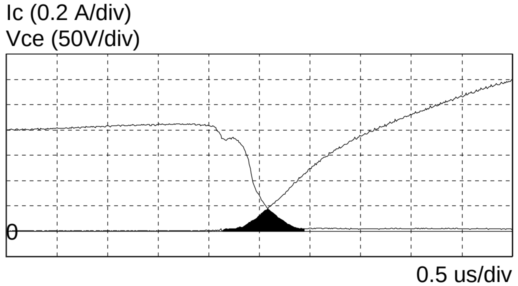
Optimised base drive.
Fig. 10. High h_{FE}. V_{CE}, I_C and I_B.

To ensure correct operation under all conditions, base drive can be optimised by adjusting the ratio of the two resistors to vary the amount of tap-off voltage. With the base resistor divided equally into two, this particular circuit suffered from a lack of base drive at low supply voltage. Too much drive had been diverted away from the base. This was corrected by moving the tap-off point to the right to split the resistor two thirds to one third to reduce the amount of diverted base drive. Referring to Fig. 5, R1 becomes 2/3 x R and R2 becomes 1/3 x R.

Figs. 11 and 12 show the optimised IC fall waveforms. A few cycles of the switching waveforms with optimised base drive are shown in Fig. 10.



Optimised base drive.
Fig. 11. Low h_{FE}. I_C fall with V_{CE}.



Optimised base drive.
Fig. 12. High h_{FE}. I_C fall with V_{CE}.

Startup circuit.

The half bridge circuit as described so far cannot start of its own accord. Both transistors are off and will remain off when power is applied until one of them is artificially turned on to draw current through the transformer primary. This will then induce a voltage in the auxiliary windings which will provide the necessary base drive to maintain self oscillation. Startup is usually achieved using a diac such as the BR100/03. The circuit is shown in Fig. 13.

When power is first applied, oscillator start-up is achieved as follows:

Transistors Q1 and Q2 are initially non conducting. Resistor R4, whose value will be several hundred kilohms, provides a high impedance path between Q2's collector and the positive rail to ensure that Q2 has the full D.C. rail voltage across it prior to start-up.

Capacitor C charges up via R1 until the breakover voltage of the diac D8 is reached. The diac breaks over and dumps the capacitor's charge into the base of Q2 to turn it on. Q2 draws current through the transformer primary. From now on, oscillation is maintained by the voltages induced on the auxiliary base drive windings.

Diode D1 discharges C every time Q2 turns on, thereby preventing the diac's breakover voltage being reached during normal circuit oscillation. This avoids repeated triggering of the diac when it is not required, so preventing oversaturation of Q2. (The length of time for C to charge to the diac's breakover voltage is much longer than the time between ON periods of Q2.)

D4 and D5 provide reverse current protection for Q1 and Q2.

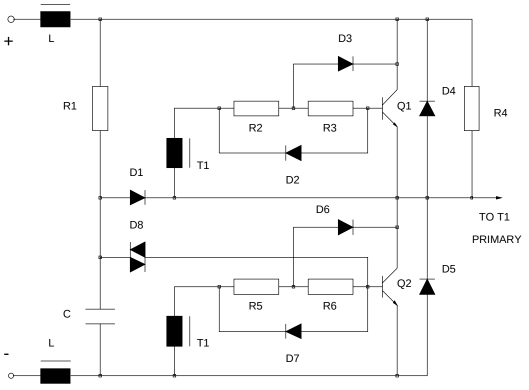


Fig. 13. A classic startup arrangement. (Part of current fed half bridge circuit.)

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Contributing Authors

N.Bennett	D.J.Harper	J.Oosterling
M.Bennion	W.Hettersheid	N.Pichowicz
D.Brown	J.v.d.Hooff	W.B.Rosink
C.Buethker	J.Houldsworth	D.C. de Ruiter
L.Burley	M.J.Humphreys	D.Sharples
G.M.Fry	P.H.Mellor	H.Simons
R.P.Gant	R.Miller	T.Stork
J.Gilliam	H.Misdom	D.Tebb
D.Grant	P.Moody	H.Verhees
N.J.Ham	S.A.Mulder	F.A.Woodworth
C.J.Hammerton	E.B.G. Nijhof	T.van de Wouw

This book was originally prepared by the Power Semiconductor Applications Laboratory, of the Philips Semiconductors product division, Hazel Grove:

M.J.Humphreys	D.Brown	L.Burley
C.J.Hammerton	R.Miller	

It was revised and updated, in 1994, by:

N.J.Ham	C.J.Hammerton	D.Sharples

Preface

This book was prepared by the Power Semiconductor Applications Laboratory of the Philips Semiconductors product division, Hazel Grove. The book is intended as a guide to using power semiconductors both efficiently and reliably in power conversion applications. It is made up of eight main chapters each of which contains a number of application notes aimed at making it easier to select and use power semiconductors.

CHAPTER 1 forms an introduction to power semiconductors concentrating particularly on the two major power transistor technologies, Power MOSFETs and High Voltage Bipolar Transistors.

CHAPTER 2 is devoted to Switched Mode Power Supplies. It begins with a basic description of the most commonly used topologies and discusses the major issues surrounding the use of power semiconductors including rectifiers. Specific design examples are given as well as a look at designing the magnetic components. The end of this chapter describes resonant power supply technology.

CHAPTER 3 describes motion control in terms of ac, dc and stepper motor operation and control. This chapter looks only at transistor controls, phase control using thyristors and triacs is discussed separately in chapter 6.

CHAPTER 4 looks at television and monitor applications. A description of the operation of horizontal deflection circuits is given followed by transistor selection guides for both deflection and power supply applications. Deflection and power supply circuit examples are also given based on circuits designed by the Product Concept and Application Laboratories (Eindhoven).

CHAPTER 5 concentrates on automotive electronics looking in detail at the requirements for the electronic switches taking into consideration the harsh environment in which they must operate.

CHAPTER 6 reviews thyristor and triac applications from the basics of device technology and operation to the simple design rules which should be followed to achieve maximum reliability. Specific examples are given in this chapter for a number of the common applications.

CHAPTER 7 looks at the thermal considerations for power semiconductors in terms of power dissipation and junction temperature limits. Part of this chapter is devoted to worked examples showing how junction temperatures can be calculated to ensure the limits are not exceeded. Heatsink requirements and designs are also discussed in the second half of this chapter.

CHAPTER 8 is an introduction to the use of high voltage bipolar transistors in electronic lighting ballasts. Many of the possible topologies are described.

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