# A Multiple Output, Off-Line Switching Power Supply Using HEXFETs®

*(HEXFET is the trademark for International Rectifier Power MOSFETs)*

**By PETER WOOD**

### Introduction

The recent introduction of HEXFET power MOSFETs has provided a significant improvement in the reliability of switching power supplies. Today, with more and more semiconductor manufacturers entering the power MOSFET arena, the initial objections of high cost and poor availability have largely disappeared.

While it is still true that large chip MOSFETs have not yet reached price parity with equivalent bipolar transistors, the inherent performance advantages of the HEXFET over bipolars makes their use very attractive, as shown in Table 1.

### The Power Supply

A large segment of the switching power supply market is applied to computers of varying sizes and types. All these machines operate from "standard" DC voltages, and with this in mind, the power supply to be described provides these voltages. A medium-sized computer would require up to 250 Watts of processed power with the following specs:

(1) +5V DC, 10 to 20A with a total ripple and regulation of ±50mV, overvoltage protected.
(2) +12V DC and -12V DC, 0-1A with a total ripple and regulation envelope of ±100mV, and a common return.
(3) +26V DC, 1-3A with a total ripple and regulation envelope of ±1V and a common return.
(4) Input voltage range 95-130V AC and 190-260V AC at 48-420 Hz.

Overcurrent protection would be required on all outputs, and the power supply should have a minimum efficiency of 75% at full load. Other requirements such as VDE compliance and EMI attenuation would also be required by most users, but these have been extensively covered in other literature and are considered outside the scale of this application note.

### Circuit Approach

Several "ground rules" were adopted to assure low cost, high efficiency, and the best possible reliability, and these are listed below:

(1) In order to minimize the voltage stress on as much of the circuit as possible, the current-driven configuration was chosen.
(2) In order to maintain the best regulation of the unsensed output voltages, the power switch was operated at a constant 50% duty cycle.
(3) In order to reduce the number (and cost) of magnetic components, no output filter chokes were used, and the single primary inductor performed the dual functions of pulse train integrator for the front end buck regulator and current source for the power switch.
(4) 50 kHz operation was chosen as the best compromise between conventional magnetic designs and minimum size (cost) magnetic components.
(5) 115/230V AC operation was obtained by using a voltage doubler

**Table 1.**

| Parameter | Bipolar | HEXFET |
|---|---|---|
| Switching Performance | Temperature Dependent | Temperature Independent |
| Switching Speed | 100-500nSec | 20-100nSec |
| Minority Carrier Storage | 1-2$\mu$Sec | None |
| Peak Current Rating | Limited by Gain | Not Gain Limited |
| SOA | $I_{S/B}$ Limited | Power Limited Only |
| Drive | Current | Voltage |
| Drive Power | Up to 5W | None* |
| Reverse Bias | $I_{B2}$ Drive is required | None Required |

**Drive current is required to charge and discharge $C_{iss}$ but not to maintain drain current.*

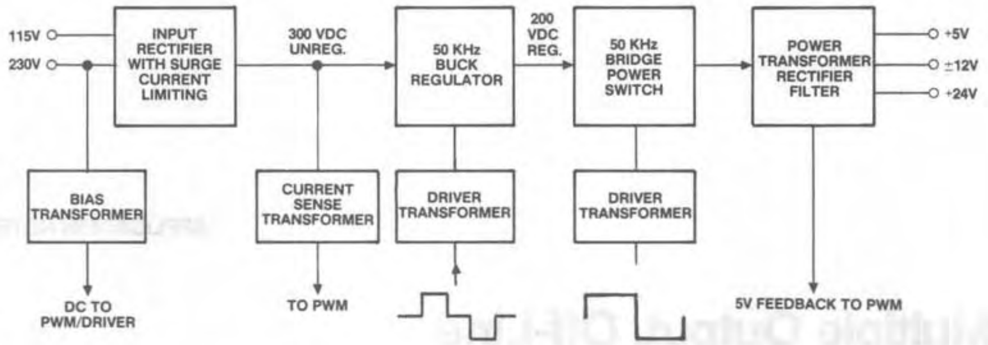



**Figure 1.** Block Diagram of Power Flow.

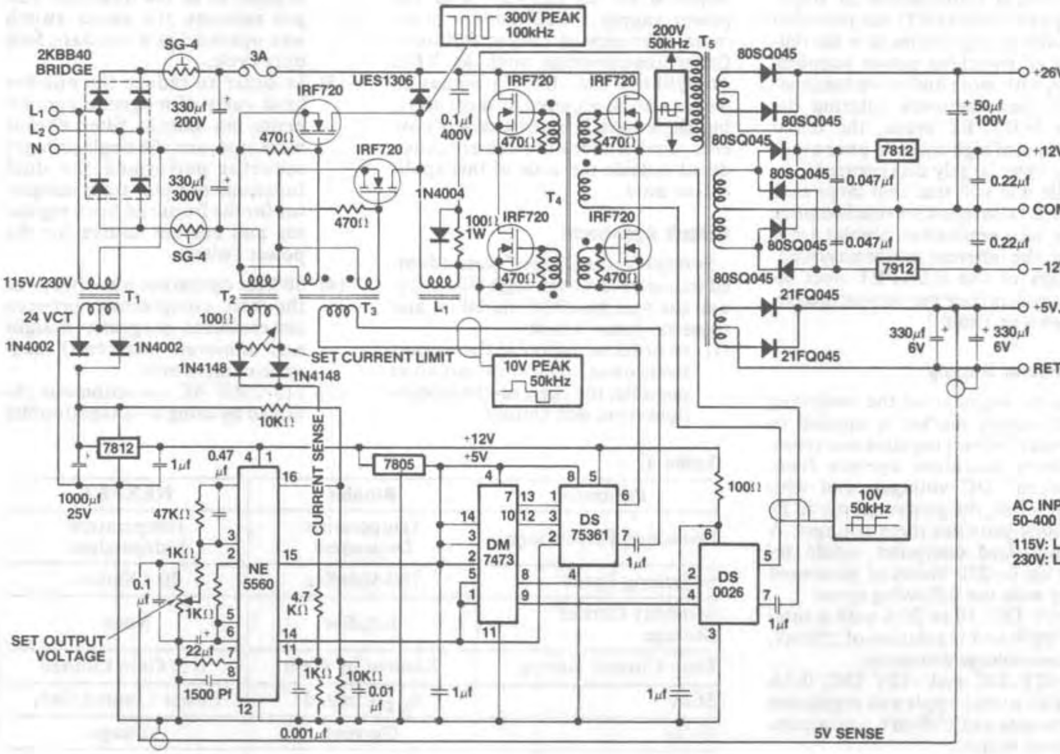



**Figure 1A.** 50kHz, 230W HEXFET Power Supply Schematic.

to supply a nominal 300V DC bus in the 115V input mode and a conventional bridge rectifier for 230V input.

(6) Control loop stability and response was enhanced by the use of a PWM (Pulse Width Modulator) with feed forward capability so that line voltage variations were regulated independently from the closed loop control.

(7) 45V Schottky rectifiers were adequate to supply all outputs since all input line conditions are regulated at the front end buck regulator.

(8) The approach is applicable for power levels from 50W to 2kW using available semiconductors.

(9) No power-wasting snubber circuits were used.

10) Soft starting, under and overvoltage protection, and current limiting were all provided by the PWM.

A block diagram representing the power flow of this approach is shown in Figure 1.

### Input Rectifier

Dual input voltage capability is achieved by the use of a diode bridge, with split reservoir capacitors, as shown in Figure 2.

In the low input voltage (115V AC) mode, AC power is applied between one of the input Lines $L_1$ or $L_2$ and the neutral N. The reservoir capacitors $C_1$ and $C_2$ are each charged to the peak voltage of the AC waveform (approximately 150V DC with 110V AC input), and since they are in series, the total unregulated voltage is around 300V DC.

When a 230V AC input is used, $L_1$ and $L_2$ become the input terminals, and the rectifier functions as a conventional bridge circuit again yielding a 300V DC bus across $C_1$ and $C_2$.

In-rush surge currents are lowered somewhat by the use of surge limiting thermistors, which, because of their negative temperature coefficient, minimize their dissipation under steady-state conditions.

### 50kHz Buck Regulator

The regulation of output voltages against line, load, and temperature effects is performed by the buck regulator stage of the power supply. Pulse width modulated drive signals are impressed on $Q_1$ and $Q_2$, which conduct on alternate half cycles of the drive waveforms. At first glance it may appear that the switching function could be performed by a single transistor — and indeed it could, were it not for the fact that the duty cycle ratio needs to be 0 to almost 100%, which poses a severe driver transformer design problem. In addition to the wide duty cycle capability, the two-transistor design affords two separate heat paths for power dissipation. This results in smaller HEXFET switches with a resulting lower cost for the devices themselves plus a much simpler drive circuit.

*NOTE:* It can be less expensive to specify two smaller chip HEXFETs than one of twice the active area.

Notice also that the switches are placed in the negative bus so that capacitively coupled switching spikes do not appear in the drive circuits or current transformer.

The pulse train appearing at the inductor $L_1$ is at 100kHz and is commutated by diode $CR_1$. No DC filter capacitor is used across the output of the buck regulator, since it is desired to present a high output impedance to the power switching inverter circuit which follows.

### 50kHz Bridge Power Switch

In a current-fed converter, the output switch must have the following properties:

(1) DC bus current flows continuously (ideally).

(2) DC bus voltage must be collapsible at constant current.

(3) DC bus voltage must be transient clamped so that high impedance conditions due to leakage inductance of the power transformer do not cause bus overvoltages.

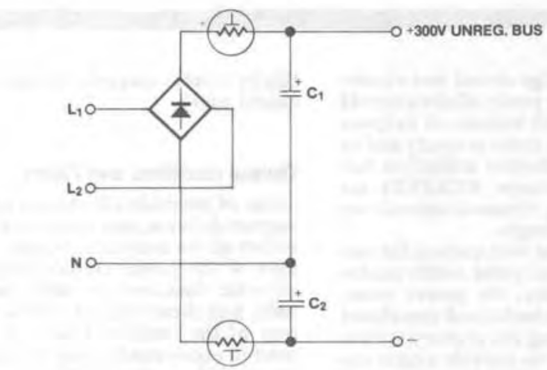



**Figure 2.** Input Rectifier

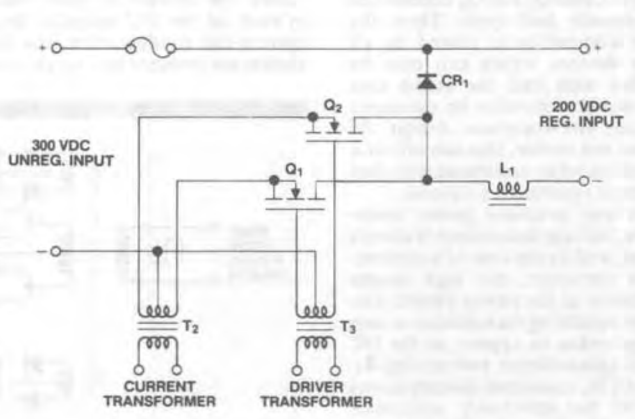



**Figure 3.** Buck Regulator

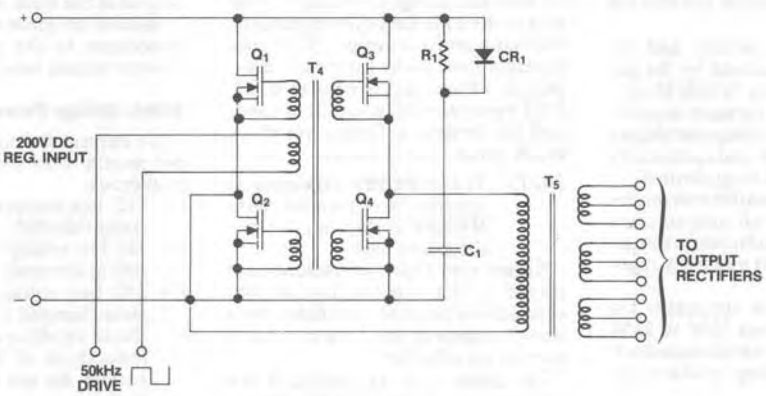



**Figure 4.** Power Switch

The full bridge circuit was chosen (even though a push-pull circuit could have been used) because of its lower switch voltage stress property and its superior transformer utilization factor. Also, because HEXFETs are voltage-driven, the transformer driver is extremely simple.

Note also that with current fed versus voltage fed pulse width modulated converters, the power transformer always looks back into closed switches feeding the primary, so it is not necessary to provide commutation diodes across each power switch.

The bridge circuit delivers an AC output of the same magnitude as the DC input bus (approx.) and is driven so that $Q_1$ and $Q_3$ conduct on one-half cycle, and $Q_2$ and $Q_4$ conduct on the opposite half cycle. Thus, the power dissipation is shared by all power devices which can now be specified with half the active area than would otherwise be necessary with any two-transistor design. As pointed out earlier, this can afford a cost saving when compared with two devices of equivalent chip area.

With any practical power transformer, leakage inductance is always present, and in the case of a current-driven converter, the high source impedance of the power switch during the switching transistions causes voltage spikes to appear on the DC bus. A spike clipper comprising $R_1$, $C_1$ and $CR_1$ connected directly across the DC bus effectively attenuates these spikes due to its low transient impedance. The power dissipated in $R_1$ is directly proportional to the magnitude of the leakage inductance of $T_5$, so it is important to minimize this by careful magnetic design (discussed later).

## Output Rectifiers and Filters

One of the main advantages of the current-driven square wave converter is that all the secondary output voltages of the power transformer are accurate functions of turns ratios only, and therefore, the reverse ratings of the rectifier diodes do not need to accommodate any AC input line considerations at all.

In practical terms, this means that 45 Volt Schottky rectifiers can be used for all output voltages of this power supply, as indicated in Figure 5.

Note the absence of filter chokes in each of the DC outputs. In the current-fed configuration, the filter chokes are replaced by a single inductor on the primary side of the power switch (see Buck Regulator).

Because of the very coarse turns resolution of the power transformer (approximately 5.6 Volts per turn), it is necessary to regulate the 12 Volt outputs with 3 terminal DC regulators. However, since the transformer output voltages are independent of AC input line variations, the overhead voltages of the 12 Volt regulators are small, and efficiency is not seriously compromised.

The +26 Volt output is derived from the +12 Volt regulator input voltage (approximately 16.2V DC) added to another 10.6V DC rectified output, making a total voltage of around 26.8V DC nominal. Note that Schottky rectifiers are used here also. Since the bulk of the output power is in the 5 Volt output, a feedback voltage for regulation is taken from this output.

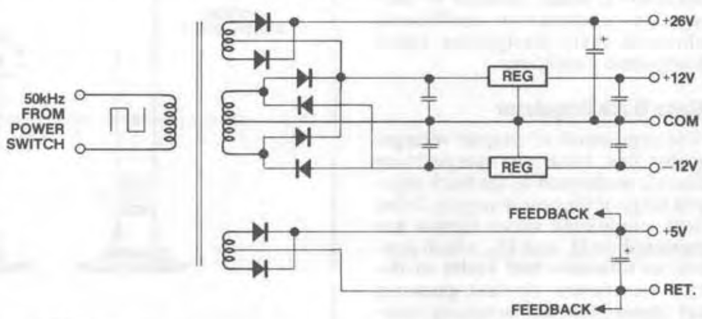



**Figure 5.** Output Circuitry

## PWM and Drivers

Figure 6 is a block diagram of the Signetics NE5560 pulse width modulator control circuit.

In addition to the usual analog-to-pulse width functions, this control circuit has a feed forward function which allows a constant volt-seconds output as a very desirable characteristic, since it reduces the gain requirements of the closed loop control functions and hence makes stabilization of that control loop much simpler. In this power supply, the feed forward control voltage is derived from the DC bias supply obtained from a small mains frequency transformer (see Figure 1).

The output pulse train from the PWM is at 100kHz and is used to trigger a dual flip-flop (DM7473) in order to generate the required 50kHz drive waveforms.

Two different drive waveforms are required for the buck regulator and the power switch, respectively. Pulse width information in the form of a 50kHz AC waveform is provided by a dual NAND gate with power output stage (DM75361). The inputs for this gate are the PWM signal from the NE5560 and the square wave output from one of the DM7473 flip-flops. The other drive waveform for the power switch is a 50% square wave, which is derived from the other DM7473 output driving a DS0026 driver which provides the necessary low impedance, fast-switching waveforms.

When driving power HEXFETs through driver transformers, it is important to preserve the accuracy of the drive waveforms, because MOSFETs can easily be damaged by excessive gate voltage spikes due to poor driver transformer design or drivers with uncontrolled high impedance outputs. The drivers in the power supply (DS75361 and DS0026) both have totem pole output stages which always present a low impedance condition to the drive transformers. Also, the DC supply to these drivers is regulated at 12 Volts so that under high or low AC input line conditions, the drive amplitudes are constant.

## Magnetic Component Design

The choice of operating frequency is in large measure dependent on the types and complexities of magnetic components. Unlike sine wave transformers, switching power supplies demand wide bandwidth designs capable of supporting not only the fundamental switching frequencies but also the fast wavefronts associated with efficient power transfer.

The circuit isolation obtainable through transformers applies not only for DC conditions, but more importantly for switching conditions where capacitive coupling between windings or even within a winding can cause unwanted ringing or common mode spikes.

If switching frequencies are too high, leakage inductances cause inefficient operation because of the dissipation in snubber circuits required to control voltage spikes. Conversely, if switching frequencies are too low, magnetic components become larger, and the increased winding capacitances add to the common mode problem, not to mention the additional cost of the magnetics themselves. For the above reasons, it was decided that 50kHz would be a satisfactory compromise between size, ease of winding, available cores and cost.

## Current Transformer T2

Current from the DC unregulated bus is sensed by a current transformer (Figure 8) in series with the buck regulator switches. The primary winding for this toroidal transformer is shown as a center-tapped winding (see Figure 3), but in practice, this winding is made by passing the drain connection leads once

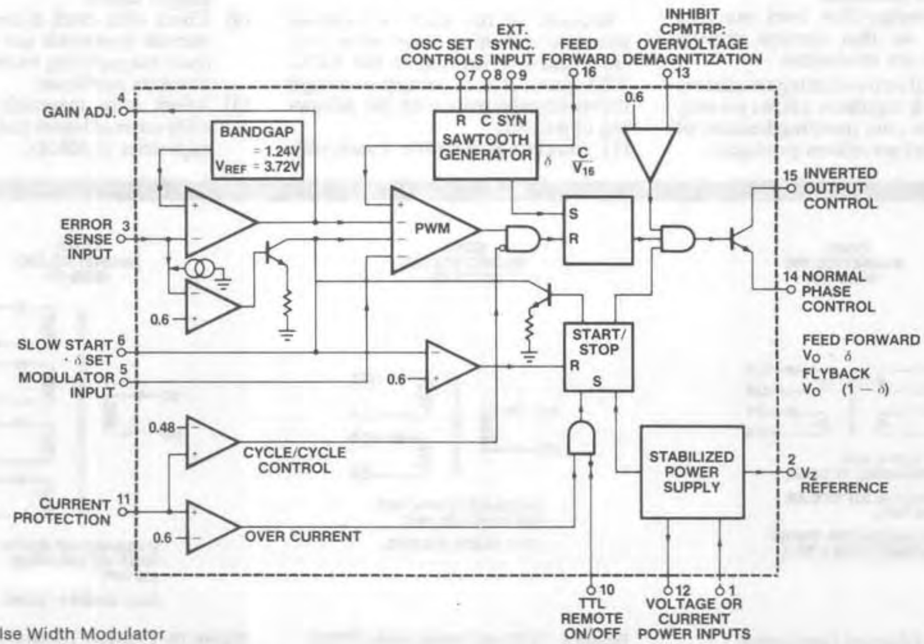



**Figure 6.** Pulse Width Modulator

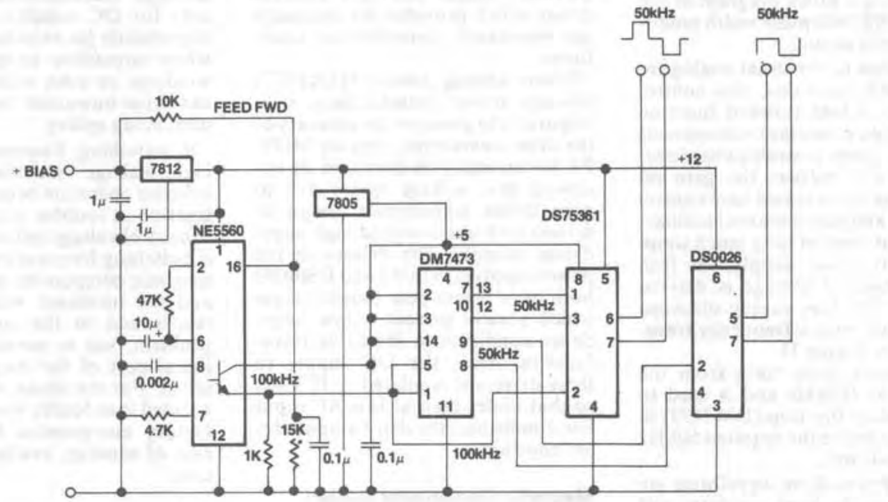



Figure 7. PWM and Driver

through the core only, thus forming a bar primary. Errors in current sensing occur because of variations in magnetizing current at differing flux levels, and for this reason, two design considerations are important:

(1) The core hysteresis loop must be narrow so that magnetizing current is small and low currents can be measured.

(2) The design flux level must be small so that current sensing errors are minimized.

The use of two switching transistors in the buck regulator allows an easy method for core resetting because of the AC flux waveform produced.

The secondary winding is bifilar wound with the winding occupying 360° of the core so that core flux is uniformly distributed and a center tap can be formed. The core is a small tapewound toroid of Square Permalloy 80, which provides the necessary high sensitivity and linearity.

## Driver Transformers T3 and T4

Because of the need to produce accurate switching waveforms into the capacitive loading of the HEXFET gates, it is necessary to design driver transformers with the following objectives:

(1) Maximum possible bandwidth by reducing leakage inductance.

(2) Minimum possible leakage inductance by using:
  (a) unity turns ratios
  (b) toroidal gapless cores
  (c) multifilar windings occupying 360°

(3) Minimum number of turns to reduce winding capacitances and copper losses.

(4) Cores with small diameters and narrow hysteresis curves to reduce magnetizing currents to an absolute minimum.

(5) Select core materials for low eddy current losses and high flux capacities at 50kHz.

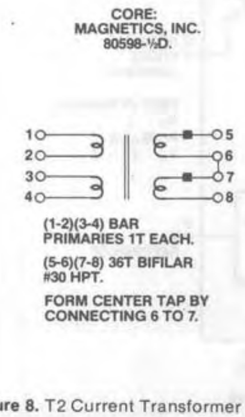



Figure 8. T2 Current Transformer

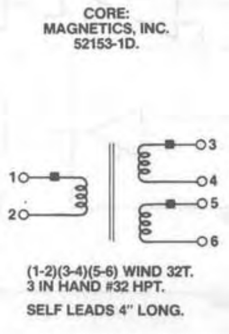



Figure 9. T3 Driver Transformer (PWM)

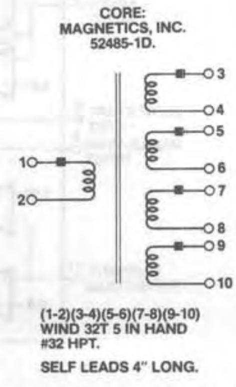



Figure 10. T4 Driver Transformer (Bridge)

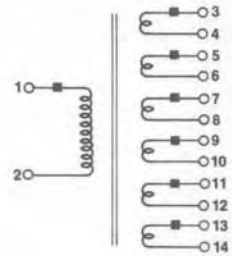
### Power Transformer T5

Because of the very coarse turns resolution due to the high operating frequency, the design of this transformer (Figure 11) is performed in the following sequence:

(1) Determine the secondary voltage for 5V DC output at full load (5 Volts + one diode drop).
(2) Select core size based on a single turn winding at a flux density of around 2Kgauss.
(3) Calculate turns ratios for other secondaries.
(4) Calculate primary turns for 200V DC input to bridge power stage. (This corresponds to the maximum duty cycle condition of the PWM at 95V AC input line.)

A P.Q. core yields low leakage inductance and simple winding techniques, and for these reasons was chosen for this application. The single turn secondaries are made in the form of U-shaped strips of 0.031 copper insulated with adhesive Mylar tape.

CORE:
TDK #H7C1PQ26/25Z-12.
BOBBIN:
TDK #BPQ26/25-0112DS.
MOUNTING CLIP:
TDK #FPQ26/25A.




| | |
|---|---|
| (1-2) | WIND 32T, 2 LAYERS, 3 IN HAND #30 HPT. INSULATE AYERS WITH 0.02 MYLAR TAPE. TERMINATE TO BOBBIN TERMINALS. |
| (3-4)(5-6)<br>(7-8)(9-10) | WIND 3T BIFILAR #20 HPT. }<br>WIND 2T BIFILAR #20 HPT. } 1 LAYER<br>INSULATE LAYER WITH 0.02 MYLAR TAPE. TERMINATE TO BOBBIN TERMINALS. |
| (11-12) | SINGLE TURN FORMED BY U-SHAPED COPPER STRIP 0.5 X 0.031. INSULATE BOTH SIDES WITH 0.02 ADHESIVE MYLAR TAPE. |
| (13-14) | SINGLE TURN SAME AS (11-12). |

**Figure 11.** Power Transformer T5

### Inductor L1

One of the unique features of this topology is the provision of a single inductor to serve several functions:

- Integration of pre-regulator pulse train to steady state current.
- High impedance current feed to output bridge.
- Spike/Ripple filter for all outputs enabling multiple output voltages to be a function of turns ratios of T5.

The inductance value was chosen to give a di/dt value of 0.5A/$\mu$S: Thus, at low line input (95 Vac), the unregulated bus voltage is approximately 260 Vdc. With a regulated output voltage to the bridge power stage of 200 Vdc:

$$\Delta V = 60V$$

At a pre-regulator switching frequency of 100 KHz:

$$T = 10\ \mu S$$

If D = 0.8

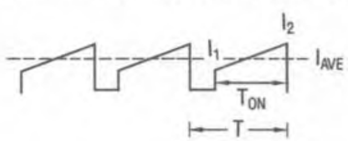
$$T_{ON} = 8\ \mu S$$

Allowing for an incremental current of 2A.

$$L = \frac{V \times T_{ON}}{2A}$$

$$= \frac{60V \times 8\ \mu S}{2} = 240\ \mu H$$

When supplying 230W output, the dc input to the bridge power stage is approx 250W and at 200V this represents an average current of 1.25A.




$$I_{AVE} = I_1 \times D + \frac{I_2 \times D}{2}$$

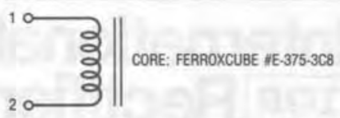
and

$$I_1 + 2A = I_2$$

$$I_{AVE} = 1.25 = 0.8 I_1 + 0.4(I_1 + 2)$$

$$I_1 = 0.375A$$

$$I_2 = 2.375A$$




(1-2) WIND 44T #20 HAPT.
GAP SPACER 0.02" (TOTAL GAP 0.04").

**Figure 12.** Primary Inductor, L1

### Summary

Although the intent of this Application Note was not to provide a "cookbook" recipe for an optimized power supply, the approach taken illustrates the use of small (low cost) HEXFETs to process substantial amounts of power. The current driver configuration minimizes electrical stresses on the power switches and enhances reliability because of the benign operation of all active and passive components in the power train.

An overall efficiency approaching 75% was obtained at full load, and the control loop was stable for all conditions of line and load.

IRF720 HEXFETs were chosen for the original 230 Watt design, but higher power levels could be accomplished merely by suitable sizing of power train components — the waveform generator and drive components would remain the same. □