A 500W 100 kHz Resonant Converter Using HEXFETs®

(HEXFET is a trademark of International Rectifier)

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Summary

This application note describes the implementation of a 100 kHz resonant converter using International Rectifier's HEXFET power MOSFETs as the switching elements. The design procedure is discussed and the completed converter schematic is provided, including suitable feedback, drive and protection circuitry.

Introduction

Sine wave converters offer a number of advantages over square wave converters. The absence of high di/dt and dv/dt makes filtering and EMI control easier. Switching losses are lower since current is zero at the instant of switching. Rectifier recovery losses are lower since di/dt is low. Overall component stress is lower resulting in improved unit reliability.

The maximum usable frequency of switching converters is determined to a large extent by magnetic considerations. Obtaining the required performance is easier in the case of a resonant converter as the transformer only sees a sine wave at the fundamental frequency. Square wave converters have to handle harmonics as well as the fundamental and therefore require a higher bandwidth transformer in order to avoid unacceptable loss or distortion. Resonant converters can thus obtain the benefits of high frequency operation without sacrificing cost, efficiency, reliability or electromagnetic compatability.

HEXFET Advantages

Bipolar transistors and SCRs suffer from storage time phenomena resulting in long turn-off times. This directly limits the maximum switching frequency in bridge-type resonant topologies. Drive circuitry required by these devices at higher frequencies with these devices is generally uneconomic.

Power MOSFETs, on the other hand, are majority carrier devices and exhibit no storage time phenomena. Drive requirements are relatively simple even at high operating frequencies. Thus resonant converters operating at hundreds of kiloHertz are quite feasible with HEXFETs.

Converter Details

The converter parameters are as follows:

Output Voltage Vo = 5Vdc

Output Current Io = 100 Amps

Input Voltage Vin = 310 Vdc (+/- 10%)

The input voltage chosen is that typically derived from full-wave rectification of a 220V ac supply.

Converter Topology

The circuit, shown in Figure 1, is based on the half-bridge resonant converter with a full-wave rectifier and LC filter on the secondary side. An analysis of this circuit, along with guidelines for its design can be found in Reference 1. This type of converter is designated a parallel resonant converter since the load is placed in parallel with a resonant circuit element.

A full-bridge version of the circuit with twice the power handling capability is shown in Figure 2.

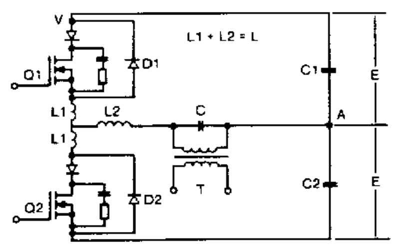


Figure 1. Half-bridge converter typology.

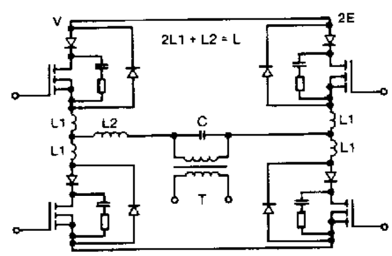


Figure 2. Full-bridge converter typology.

Circuit Operation

The resonant circuit consists of inductance L (L = L1 + L2) and capacitance C. C1 and C2 are sufficiently large so the potential at point A remains substantially constant. The load is connected across the capacitor via a transformer which provides isolation and a step-up or step-down capability. The equivalent inductance referred to the primary is large with respect to the resonant inductance, and the leakage inductance and interwinding capacitance are low. This means that the basic operation of the resonant circuit is not affected by these elements.

The operation of the converter on no-load is as follows. Suppose the switching frequency is made equal to half the circuit resonant frequency. Initial inductor currents and capacitor voltages are zero.

Q1 turns on. Current flows in the resonant circuit with a sinusoidal waveshape. C is charged to a peak value approaching 2E since the Q of the circuit is high. Q1 is held on until the current reverses at which instant it is turned off. The current is then carried by D1 back to the supply. The current will eventually fall to zero leaving C fully discharged. Q2 is turned on and a similar cycle occurs with the current direction reversed. Thus a full sinusoidal-type voltage will appear across C. Figure 3 shows these waveforms.

Let f_r be the circuit resonant frequency and f_o the frequency of the output waveform. Both devices are turned on at that frequency, 180 degrees out of phase with each other. The on-time of the devices is then determined by the characteristics of the resonant circuit.

The switching frequency may be varied to control the output waveform. The theoretical maximum for zero current switching operation is equal to the resonant frequency of the circuit. In practice the switching frequency is held below the circuit resonant frequency to allow a deadband between turn-off and turn-on of opposite devices.

Increasing f_r/f_o results in a reduction of the RMS output voltage and an increase in the distortion, to the point where the output no longer resembles a sinusoid. The basic circuit operation is the same with a dead time between conduction cycles.

Reducing f_r/f_o causes the magnitude of the output to increase and the waveshape to approach a true sinusoid. In this instance diode currents are not zero at the switching instants. For example, current is commutated from the lower diode to the upper HEXFET when it turns-on. That HEXFET then charges the resonant capacitance from its initial negative voltage to a peak value greater than 2E. Current then reverses passing through the upper diode discharging C. The lower HEXFET will then be turned on while current is still flowing and the voltage on C is positive. The cycle repeats.

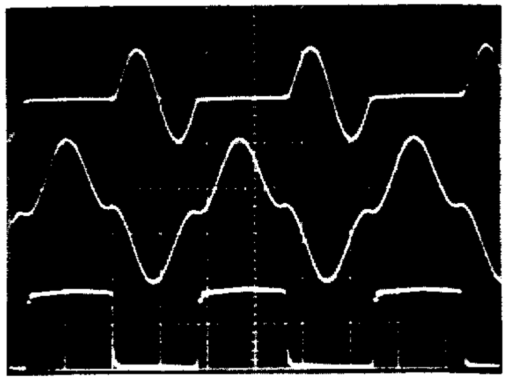


Figure 3. Converter waveforms — no load.

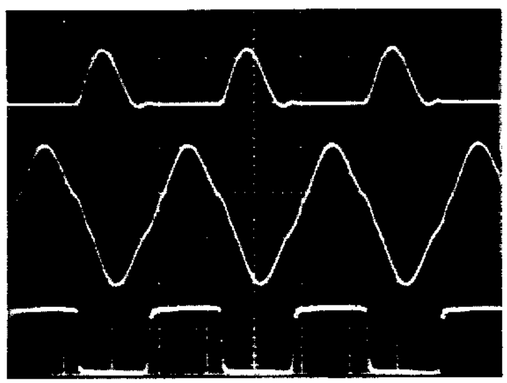


Figure 4. Converter waveforms — loaded.

At some critical value of f_r/f_o the output waveform will have minimum distortion. Reducing f_r/f_o still further results in a large increase in the output peak to peak value and a small increase in distortion. In the limiting case of $f_o/f_r = 1$ the peak voltage across C will equal the product of E and the circuit Q.

The operation under load is similar to the no-load case. The current pulse through the HEXFETs broadens slightly while that through the diode falls in width and peak value. The Q of the resonant circuit is reduced and the output peak-to-peak voltage will be reduced for a given switching frequency. If the load is reactive the resonant frequency will be altered, changing f_r/f_o and affecting the output accordingly. Figure 4 illustrates waveforms for a converter under load. The switching frequency is as close to the resonant frequency as is practicable so that maximum power might be delivered.

The feedback loop may be closed and the switching frequency controlled in order to regulate the output voltage. Output control is achieved by varying the switching frequency and therefore this converter is not suitable for applications requiring a constant frequency sinusoidal output. Various control strategies may be adopted. If a sinusoid with low distortion is required f_r/f_o can be varied between 1.1 and approximately 1.35. If a constant RMS output is required f_r/f_o may be varied upwards from 1.1 at full load with minimum dc input voltage.

A dc output may be obtained by rectifying and filtering the transformer secondary voltage. The parallel resonant converter should feed a current load, as opposed to a voltage load, in order that the resonant tank operation is not affected. An LC filter circuit is therefore required in the secondary stage.

Converter Design Procedure

To facilitate the adaptation of this demonstration circuit to suit individual needs, the procedure followed in the design of the converter is summarized here.

1. Choose the switching frequency (f_o), for full load and minimum input.

   f_o = 100 kHz maximum.

2. Let $f_r/f_o = 1.1$ at full load and minimum input.

   The resonant frequency (f_r) is then fixed at 110 kHz.

3. Choose Lm/L where Lm is the primary equivalent inductance of the transformer.

   Lm should be chosen large enough such that the resonant frequency is not affected. Low values of Lm/L imply a low cost transformer but at the expense of increased circuit voltages.

   Let Lm/L = 50

4. Choose the minimum value of $R/\sqrt{L/C}$.

   Low values of $R/\sqrt{L/C}$ reduce circulating current and thus component cost. However, the maximum capacitor voltage ($f_r/f_o = 1.1$) decreases rapidly as $R/\sqrt{L/C}$ falls below 5.

   Let $R/\sqrt{L/C}$ minimum = 1.7.

5. Calculate R. (Full load value.) R is the effective load resistance seen by the resonant capacitor. It may be estimated as follows:

   $$R = \frac{V_s n^2}{I_s}$$

   where n is the turns ratio of the transformer. V_s and I_S are the average voltage and current on the secondary, respectively.

   $$I_s = \frac{P_o}{V_o} = 100\text{A}$$

   V_s = 6.0V approximately. This takes account of the rectifier drop and other secondary losses.

The turns ratio of the transformer is chosen to be 32. This gives R equal to approximately 60.

6. Calculate $\sqrt{L/C}$.

   $R/\sqrt{L/C} = 1.7 \quad \sqrt{L/C} = 35.$

7. Calculate the component values using the value of $\sqrt{L/C}$ and f_r.

   a) $$C = \frac{10^2}{2\pi f_r \sqrt{L/C}}\ \mu\text{F}.$$

   $$\frac{10^6}{2\pi \times 110 \times 10^3 \times 35} = 0.041\,\mu\text{F}.$$

   Eighteen 2.2 nF Low Loss film capacitors were connected in parallel.

   b) $$L = \frac{\sqrt{L/C} \times 10^6}{2\pi f_r}\ \mu\text{H}.$$

   $$\frac{35 \times 10^6}{2\pi \times 110 \times 10^3} = 50.6\,\mu\text{H}.$$

   The resonant inductance L consists of 2 components L1 and L2 as shown in Figure 1. Reducing L1 decreases the nominal voltage stress on the HEXFETs. However, L1 is necessary to reduce the magnitude of the diode recovery current spike. Also it provides some protection in the event of simultaneous HEXFET conduction. L = L1 + L2.

   L1 was chosen to be 46 μH and L2 4 μH. The winding details are supplied in the appendix.

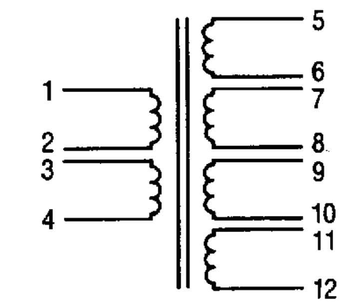
   c) $$Lm = \frac{Lm}{L} \times L\,\mu\text{H}.$$

   $$Lm = 50 \times 50.2 \times 10^{-6} = 2.51\,\text{mH}.$$

   The transformer winding details are given in Figure 5. The final value of primary equivalent inductance was 3 mH.

T1 — DRIVER TRANSFORMER

CORE — MAGNETICS, INC. 41605 TC — W



(1-2) (3-4) (5-6) (7-8) (9-10) (11-12)
WIND 20 TURNS, 6 IN HAND NO. 32.

T2 — T3 — T4 — CURRENT TRANSFORMER

CORE — MAGNETICS, INC. 41605 TC — W

(1-2) WIND 100 TURNS NO. 32

L1 — RESONANT INDUCTANCE 4 μH

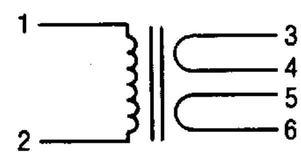
WIND 15 TURNS COPPER AWG 15 WITHOUT CORE ON COIL FORMER ϕ 25mm.

L2 — RESONANT INDUCTANCE

WIND 56 TURNS COPPER AWG 15 WITHOUT CORE ON COIL FORMER ϕ 40mm.

T5 — POWER TRANSFORMER

CORE — SIEMENS B66335 — GG000 — X127 (E55)



(1-2) WIND 32 TURNS BIFILAR NO. 18.

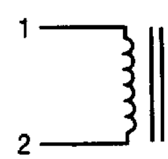
(3-4) SINGLE TURN FORMED BY COPPER STRIP 0.5 x 20mm.

(5-6) SINGLE TURN FORMED BY COPPER STRIP 0.5 x 20mm.

PRIMARY AND SECONDARY WINDING MUST BE INTERLACED IN ORDER TO HAVE LOW LEAKAGE INDUCTANCE.

L — SMOOTHING INDUCTANCE

CORE — SIEMENS B66335 — G0500 — X127 (E55)



(1-2) WIND 4 TURNS FORMED BY COPPER STRIP 1 x 25mm.

Figure 5. Transformer and inductor design details.

8. Device selections.

   It is necessary to estimate the current load and voltage stress on the circuit devices in order that a suitable selection be made. Computer modelling greatly assists resonant converter design in this respect as long hand solutions are likely to be complex. Reference 1 provides some guidelines for the half-bridge circuit.

9. Choice of HEXFET

   The peak voltage experienced by the HEXFET is given by:

$$2E + \frac{L_2\,(V_{cmax} - E)}{L_1 + L_2}$$

   This is approximately 350V maximum.

   However the reverse recovery of the freewheeling diode in parallel with the HEXFET will cause a large voltage spike to appear across the HEXFET when current is commutated to the opposite device. An RC snubber was placed across the HEXFET to limit the spike and prevent ringing. 500V HEXFETs were chosen to allow a suitable safety margin in voltage rating.

   The HEXFETs must be rated to carry the worst case current waveform. A HEXFET can operate at a junction temperature of 150 degrees C but the designer may wish to operate the device at a lower temperature. The final device choice will depend on package constraints, heatsink size and other economic considerations. An IRF450 device was chosen. This has a BVdss rating of 500V and is rated to carry 8A continuously at a case temperature of 100 degrees C.

10. Choice of diodes.

   The diode should have a peak reverse blocking capability equal to the voltage rating of the HEXFET. It should be thermally rated according to the maximum current conditions which will occur when the converter load current is a minimum. The diode should be of a fast recovery type as current is commutated from one diode to an opposite HEXFET. The diode should have a soft recovery characteristic to reduce voltage spikes and EMI.

   HEXFETs incorporate an intergral drain-source diode (Reference 2). While this is a fast diode by traditional standards, its recovery time is long compared with the switching speeds that HEXFETs are capable of.

   For this reason a 16FL60S02 diode was placed in series with the HEXFET to isolate the body-drain diode. Another 16FL60S02 was placed across the diode-HEXFET pair to act as free-wheeling diode. This is a fast recovery rectifier (trr = 200 nS) with a soft recovery characteristic. IF(avg) = 16 A at 100 degrees C case temperature and V_{RRM} = 600 V.

Circuit Details

Figure 6 details the power section of the completed converter. The dc input rail is obtained through rectification of the mains utility supply. A voltage doubler or full bridge rectifier configuration is used depending on whether the line supply is American or European. If a well-filtered dc rail were available the reservoir capacitors would only need to be ten to twenty times the value of the resonant capacitance value.

An International Rectifier 201CNQ045 center tap Schottky module was chosen for secondary rectificaiton. This has an average current capability of 200A and a VFM of 0.67V at IFM = 100A.

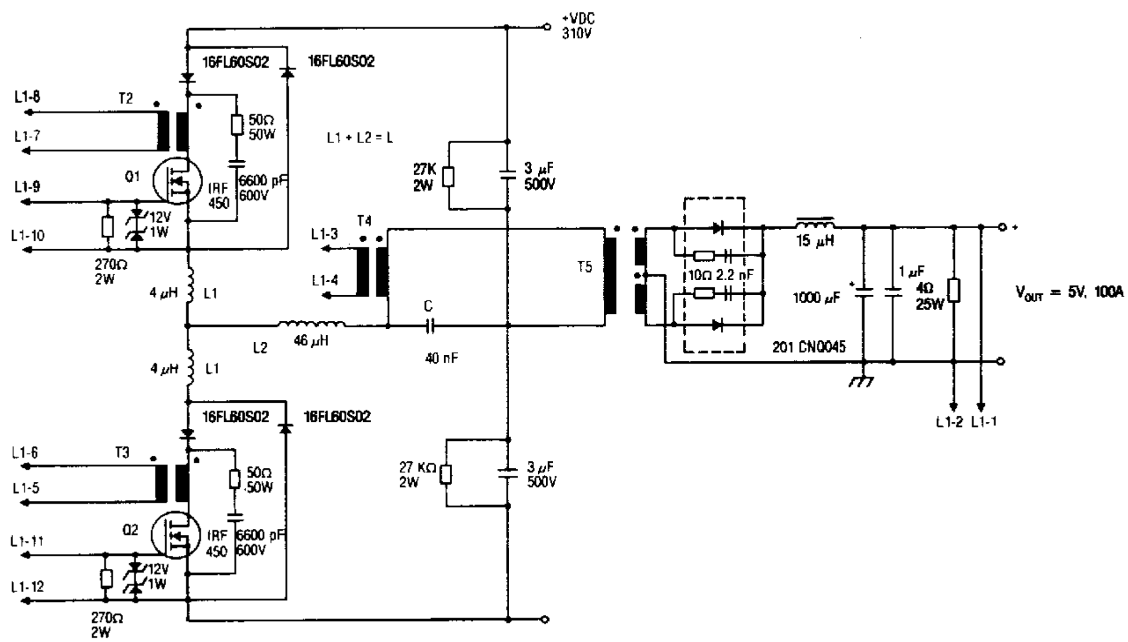


Figure 6. Power converter schematic.

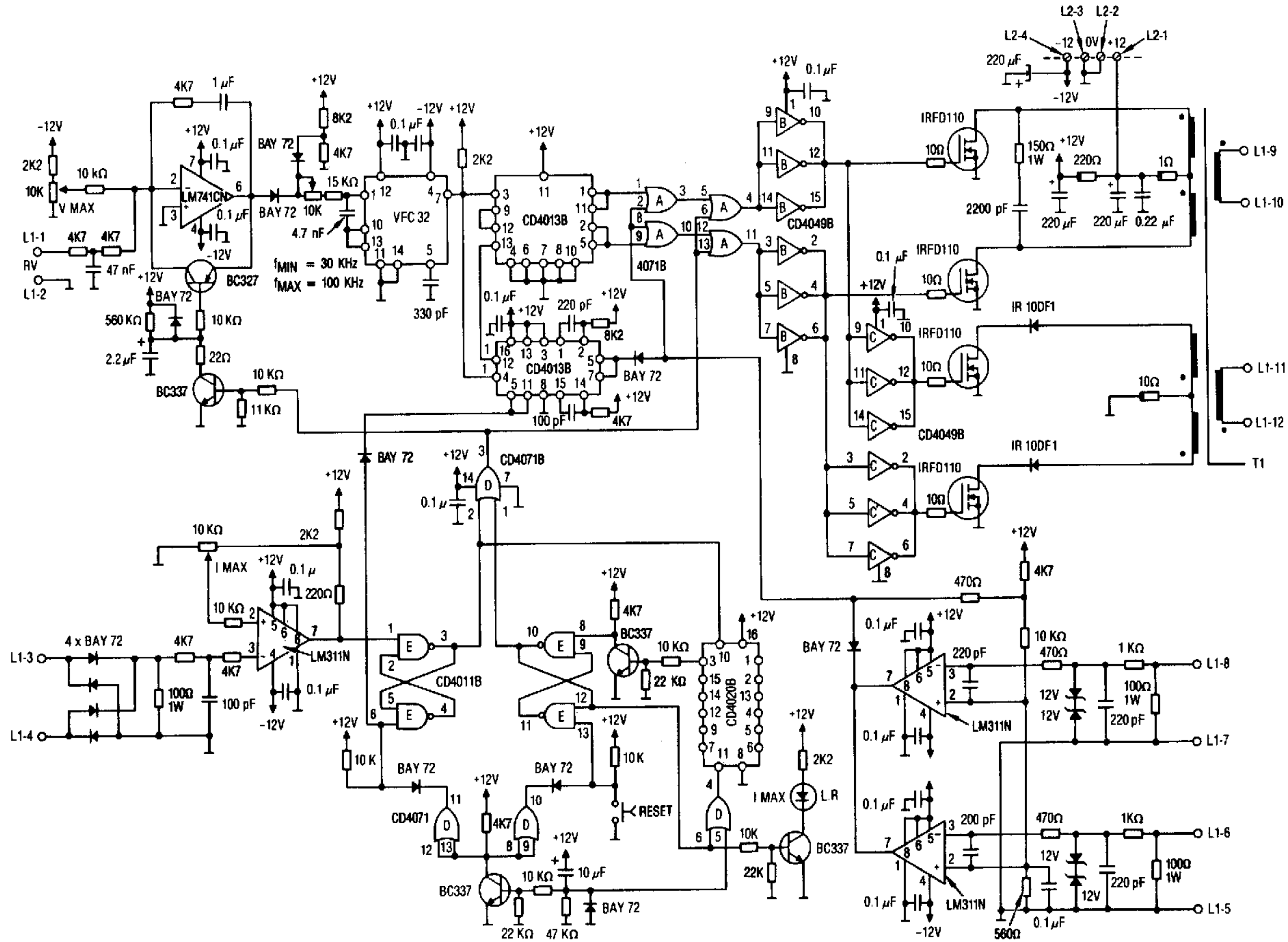


Figure 7. Control circuit schematic.

NOTES: 1. BAY 72 = 1N4148
2. 4K7 = 4.7 KΩ

The output filter values are chosen to provide the required output ripple. The filter capacitor consists of 10 x 100 μF electrolytic capacitors and 10 x 0.1 μF capacitors all in parallel. This arrangement was necessary to ensure acceptable high frequency performance. The smoothing inductance winding specification is included in figure 5. A minimum load is necessary to ensure continuous inductor current.

Current transformers T2 and T3 detect the HEXFET currents. These signals are required for controlling the on-time of the devices. The gates of the HEXFETs are protected by back-to-back zener diodes located as close as possible to the HEXFETs. 270 Ohm resistors are also placed across the gate-source terminals to reduce ringing and to reduce the susceptibility of the gate curcuit to noise and drain voltage spikes coupled to the gate by drain-gate capacitance.

Figure 7 gives details of the control circuit. An explanatory block diagram is given in Figure 8.

The VFC 32 is a voltage-to-frequency converter (Burr-Brown). The digital output is open collector and the pulse repetition rate is proportional to the analog signal from the error amplifier. The CD4013B is a dual D-type flip flop. The output to the OR gates are complimentary and toggle for every positive going transistion on the input from the VFC32. This effectively enables the output power HEXFETs alternately.

The CD4098B is a dual monostable multivibrator. Both monostables are connected in the non-retriggerable mode. The leading edge of the waveform from the VFC32 at pin 12 causes the output at pin 5 to go low for 2 μS. The output of gate D at pin 3 remains low except when an overload condition has been detected and thus the 2 μS pulse acts as an initial turn-on pulse for one of the HEXFETs.

The gate input capacitance of the HEXFET is charged up via one leg of a centre-tapped pulse transformer driven in a push-pull mode. A second centre-tapped winding maintains a low impedance across the gate-source terminals of each device when both devices are in the off-state.

Feedback from pulse transformers T2 or T3 hold the HEXFET on as long as device current remains positive. When the resonant current passes through zero the HEXFET is turned-off.

The current transformer T4 detects output shorts. Pin 3 of gate D goes high. This is reset every second cycle by one of the dual monostables. A CD4020B counter locks out the HEXFET drive after a chosen number of overcurrent detections. A visible warning is set and the circuit must be manually reset.

Figure 9 shows the completed power supply.

Circuit Waveforms

Figure 10 shows the circuit waveforms for three different loading levels.

The switching frequency increases from 60 kHz at no load to 70 kHz at 50A load and 85 kHz at full load of 100A. The switching frequency increases to 100 kHz as the input line falls to its rated minimum. The distortion of the capacitor voltage is significant for the no-load situation but is reduced considerably as f_r/f_o falls.

Conclusion

The advent of HEXFETs has made it practical to operate resonant converter circuits at frequencies beyond the normal operating range of power bipolar transistors resulting in a significant reduction in the size of the magnetic components required. The square safe-operating area of the HEXFET and its high surge-current capability ensure good system reliability and high transient overload capabilities. The simplicity of the HEXFET gate drive circuits reduces the cost of the unit and cuts design effort. In resonant power supplies, as in many other applications, the HEXFET offers important advantages over bipolar transistor in almost every respect. □

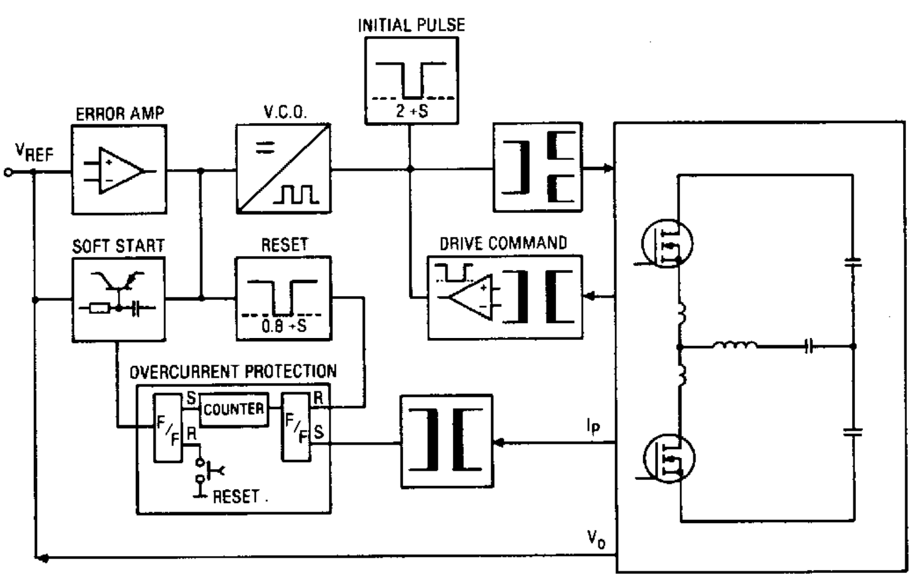


Figure 8. Control circuit block diagram.

References

1. IEEE Transactions on Industry and General Applications. N. Mapham. Volume: IGA-3, No. 2 March/April 1967, pages 176-187.
2. "The HEXFET's Integral Body Diode — Its Characteristics and Limitations". S. Clemente, B. Pelly, B. Smith. International Rectifier Application Note 934B, HEXFET Databook 1985 (HDB-3).
3. "High frequency resonant converter using power MOSFETs". S. Young, G. Castino. Proceedings of the High Frequency Power Supply Conference, Virginia Beach, May 1986. pp 21-35.

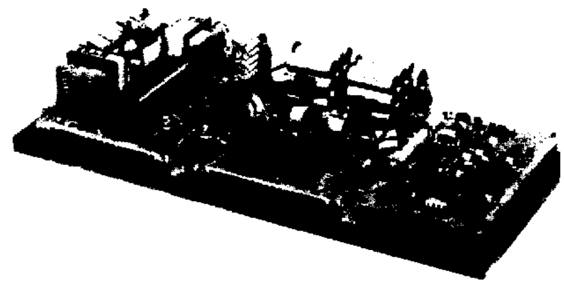


Figure 9. Prototype converter.

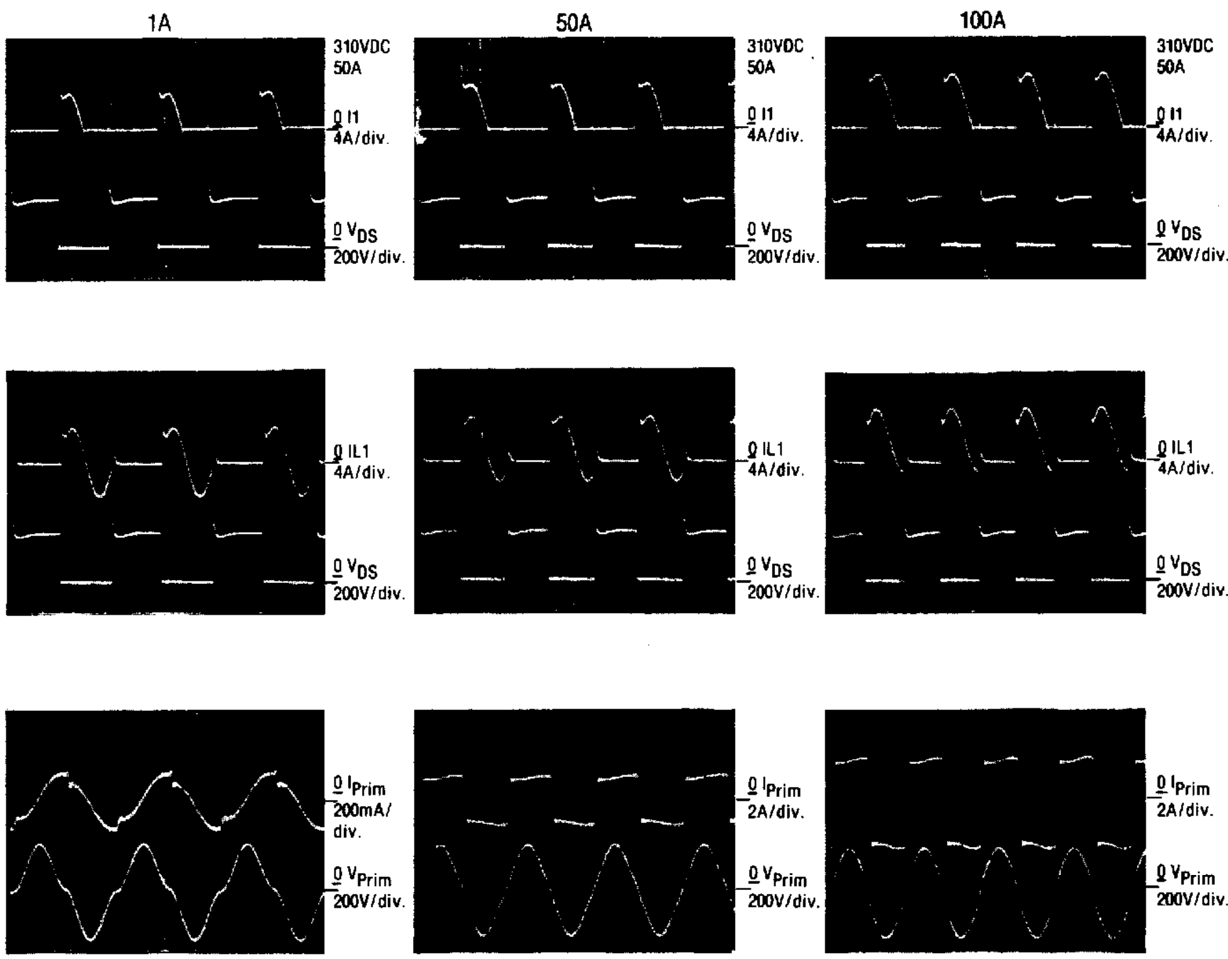


Figure 10. Circuit waveforms for I_h = 1A, 50A and 100A.