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THE DESIGN OF A PUSH-PULL SWITCH MODE POWER SUPPLY FOR AC DRIVES

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ABSTRACT

In this study, the step up DC-DC converter that works in push-pull connection has been taken up. The voltage-drops that occured since input voltage was low, have been minimized. The IRF 1310N mosfet which had minimum conduction resistance was preferred when we were selecting semiconductor elements used for voltage-drop minimization. As the current value switched in converter was high, the current capacity was increased by connecting the mosfets as parallel. At the same time the voltage-drop occured on semiconductor element, was minimized.

Switching mode power supply at 28 kHz is capable of providing 380 Watt . It gives the input power from a standard 12 volt battery, and its output voltage is 311V. As the realized converter has the closed loop feature, output voltage remains stable within determined power limits. When the converter is used as pre-circuit of the inverter, it simplifies the control cards highly as the inverter input voltage remains constant. Converter design shows that theoretical analyses and experimental results are in accordance with.

1. INTRODUCTION

The most significant problems in DC-AC converters used in these days are regulation and response time [2]. These converters are generally fed by serialized battery groups and the batteries are discharged depending on the density of the power. When the batteries are discharged, terminal voltages decrease and output voltage of the DC-AC converter changes. In 20kHz switching frequency, there are approximately 200 PWM data for any voltage value in the inverters of which output voltage is real sinus. With the change of input voltage, inverter will require new 200 PWM data in order to regulate output voltage to its previous value. Considering that the battery voltage is allowed to change 20% in total, this shows that the output voltage will change 44 volts for a voltage value of 220 volts. This means that 8800 PWM data are saved in one volt resolution and this reveals that a very complicated control card is required.

However, the response time of the control card is also very important. If we consider existing microprocessors, the achievement of this work with high speed will become difficult and microprocessors will substituted by DSPs, besides requiring an extra cost.

Using a closed loop switching mode power supply that will compensate the voltage changes in battery, in order to avoid these disadvantages is one of the various alternatives. By means of the switching mode power supply, unproductive big clumsy iron transformers that will be used in 50Hz will be eliminated. Regulation will also be better and, ripple quantity on DC bus and filter elements will highly diminish according to switching frequency [1]. Thus, there will be no need for the complicated control cards to be used in the inverter. That input voltage is in a low value like 12 volts causing many disadvantages. For this reason, the most appropriate system for this work is push-pull converter connection that mid-point transformer is used. In this converter, as one semiconductor element is used for every half-wave, voltage-drop is 50% less than it is in full bridge converters.

2. PUSH-PULL CONVERTER CONNECTION

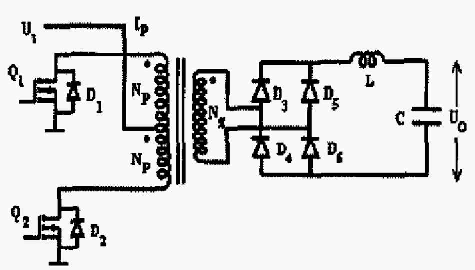

Fig. 1. Converter principle scheme

Power is converted with a very high efficiency by using different circuits. In regard of cost, dimension and performance, mid-point push-pull configuration is the most appropriate circuit [6]. The push-pull converter power circuit which consists of two semiconductors is shown in Fig. 1.

The source terminals of power semiconductors are grounded in the converter. Thus, the drive circuit is simplified and it is not needed the insulated voltage

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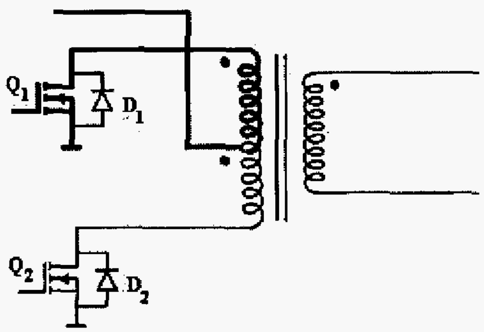
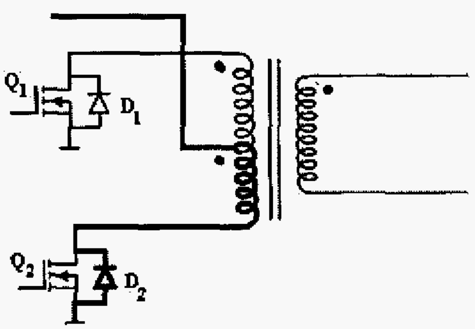
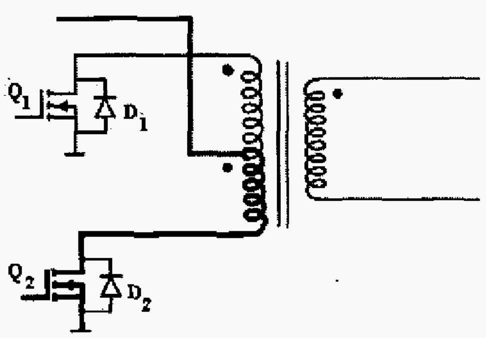
source for the drive. Despite stating the advantages in the introduction part, the major disadvantage of this converter is U_{DS} which is the voltage occured between drain-source of the semiconductor and should be at least double of the source voltage. (U_{DS} >2 U_{supply}). This leak voltage is tried to be kept in certain limits by means of snubber circuits. Another problem is that the mosfets get into conduction simultaneously. In order to avoid this problem, an appropriate dead time value has been determined considering the features of the drive circuit and the semiconductor [1].



a) Q1 mosfet is in conduction



b) D2 diode is in conduction



c) Q2 mosfet is in conduction

Fig. 2. Operating condition of push-pull converter

The voltage controlled converter shown in Fig. 2, gets into conduction between t1>t>t2 by means of PWM signal which is applied to Q1 mosfet gate. During the conduction, the primary current Ip that rises in value, will flow through the battery centre-tapped transformer coil and mosfet. Meanwhile, drain-source area of the mosfet will be exposed to the Vsat voltage that is conduction voltage drop. However, the mosfet in the other branch is being exposed to doubled battery voltage that is feeding voltage. At the same time, the current flowing in secondary circuit charges the shock coil in secondary circuit. When the PWM signal which is applied to Q1 mosfet gate stops, t2<t<t3, a new working area, is entered. In this part, D2 diode existing in mosfet own structure, will assist in transferring the energy accumulated on transformer coil to the source. The energy is not transferred in high speed, causing extreme voltages on mosfet. This may also cause leakage condition on the related element. The shock coil which is charged during the conduction transmits a current decreasing in secondary circuit while mosfets are off-states. Ripple Quantity in the circuit changes according to the dependence of the size of the shock coil. Generally, the shock coil current ripple is calculated as 15%.

In order to form the other half-wave of the switching frequency, Q2 mosfet existing in the other branch is kept in conduction with PWM signal applied to its gate during t3<t<t4 period. The current in the secondary part begins to increase again. The rectifying diodes existing in secondary circuit convert the ac signal in full - wave and then they rectify it in very small ripple waves by means of a filter consists of L-C one. Thus, the needed DC bus is obtained. The related waveforms are shown in Fig. 3.

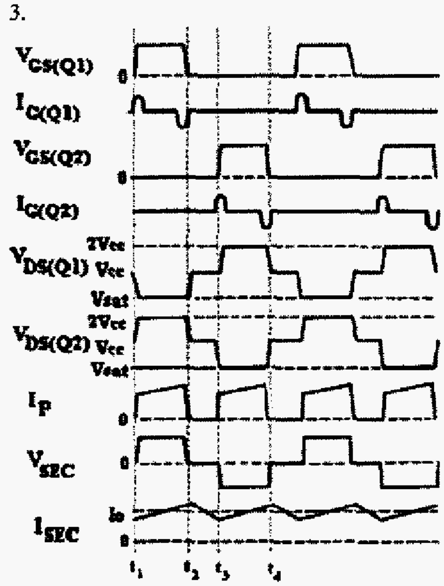


Fig. 3. Switching mode power supply waveforms

3. THE DESIGNING PHASE

When switching mode power supply was designed, a high frequency of 28kHz was preferred considering the cost, dimension and performance triple. Thus, while dimension of the core was diminishing, the efficiency was raised to a very high value of 90%.

T_{ON} =32,84µs (max)
T_{OFF} =2.85µs (min) (1)

Duty cycle gets the following value,

$$d_{\max} = \frac{T_{ON}(\max)}{T(peryot)} = \frac{32,84}{35,7} = \%92 \quad (2).$$

Thus, enough dead time is provided and mosfets into conduction simultaneously are prevented. Considering that the output power is 380 watt and efficiency is n=90% being a value of,

$$P_{IN} = \frac{311 \cdot 1,22}{0,90} = 422 Watt \quad (3)$$

is received from the input. The current flowing in primary part can be calculated as follows;

Average primary current (I_{dc}),

$$(I_{dc}) = \frac{P_{IN}}{V_{IN}} = \frac{422}{10} = 42,2A \quad (4)$$

If duty cycle is considered, the current through the mosfet will be as follows;

$$I_P = \frac{I_{dc}}{d(\max)} = \frac{42,2}{0,92} = 45,87A \quad (5).$$

In order to switch this current in the converter, the IRF 1310N, a semiconductor which has a very low transmission resistance (R_{ds}=0,036Ω) and switching losses are very lower than the other mosfets, has been used. For the switching primary circuit, 3 pieces of the mosfets which have 42A current transfer capacity, have been parallel connected. At the same time, optimization in regard of the voltage-drop, has been acquired.

$V_{ds}(on) = R_{ds}(on) * I_P = (0,036/3) * 45.87 = 0,557V$. Considering that the battery voltage is 12 Volts, 4,6% of the voltage drops on mosfet. Power consumption on body of the mosfets is,

$P_{dc} = (I_P)^2 * R_{ds}(\max) * \text{duty} / 2$
$= (45,87)^2 * 0,012 * 0,92/2 = 11,61$ Watts (6)

As the transformer used, the core which could provide the required power in the desired switching frequency, has been determined by EC70 in the related tables.

$V_{pri}(\min) = V_{in}(\min) - V_{ds} = 11 - 0,557 = 10,4$ V
$V_{sec}(\min) = V_{out}(\max) + V_{diyot} + V_L$
$= 311 + 1,3 + 0,1 = 312,4$ V (7)

Converting the rate of the transformer has been calculated from the voltages as follows;

$N = V_{pri}(\min) \cdot \text{duty}(\max) / V_{sec}(\min) = N_{pri} / N_{sec}$

$$N = \frac{312,4}{10,4} = 31 \div 1 \quad (8).$$

In order to minimize the transmission area, primary winding number will be taken as 3,5 tours and secondary winding number (31x3.5=110) will increase. In order to prove this condition, the ferrit core which is calculated on the table is able to transfer the desired power in the following equation.

$$A_c(\min) = \frac{V_{pri}(\min) \cdot duty(\max) \cdot 10^4}{2 \cdot Frequency \cdot N_{pri} \cdot \Delta B(Tesla)} (cm^2) \quad (9)$$

If the related values are placed in equation 9,

$$A_c(\min) = \frac{10 \cdot 0,92 \cdot 10^4}{2 \cdot 28000 \cdot 3,5 \cdot 0,28} = 1,66cm^2 \quad (10).$$

As dimensions of the area where the total magnetic flux flows, are 2,79cm^2 >1,66cm^2, the usage of the ferrit core is appropriate. The most significant problem experienced in magnetic design in high frequency is decreasing eddy current losses and skin effects by optimizing cable diameter. The current flowing through the copper cable, is influential in different depths depending on the frequency. This depth is calculated as;

$d_{pen} = 7,5 / (\text{frequency}^{0,5})$ (cm) $= 7,5 / (28000^{0,5})$

$d_{pen} = 0,04482$ cm (11)
where is a cm per 100 °C in copper conductor. In this case, the diameter of the copper conductor to be used in primary coil should be higher than this value. Practically, the current density of the copper is 450A/cm2 and effective current value is calculated as follows;

$$I_{Prms} = I_P \cdot \sqrt{\left(\frac{d(\max)}{2}\right)} = 31A \quad (12).$$

Therefore, from the current density value, the primary coil area is obtained as 0,068 cm^2. Considering that the influence depth, using a copper foil of 0,05 x 3 cm in the primary part will be convenient. If the secondary circuit is calculated in same methods, it appears that it will be made of five copper wires of 0.4mm diameter.

If we consider that the shock coil used in the secondary of the circuit arranges current ripple value of the secondary current, this allows alernating in 15% of the original current, this is calculated as follows;

$$L = \frac{V_{sec}}{4 \cdot f \cdot \Delta I_{sec}} = \frac{311}{4 \cdot 28000 \cdot \%15 \cdot 1,22} = 15,17mH$$
(13).

Converter diodes used in the output are ultrafast diodes which have 35ns delay, since the frequency is high. If these diodes are not fast enough, extreme voltage occures on mosfets in primary circuit. Finally, C is found by filtering the obtained wave per 100 mV ripple voltage.

$$C = \frac{I_{p-p}}{8 \cdot f \cdot dV} = \frac{1{,}22 \cdot 0{,}15 \cdot 2}{8 \cdot 28000 \cdot 0{,}1} = 17\mu F \qquad (14)$$

4. PARALLEL CONNECTED MOSFETS FOR HIGHER OUTPUT POWER

Development of uninterrupted power supply; If increasing of requested power value and reliability with excessively load currents are considered, parallel connection operation has become an attractive alternative since the current capacity of the existing semiconductors is exceeded. All semiconductor elements can be paralleled in order to meet higher currents. Their own capacity and the voltage-drops in parallel branchs are the same and the currents flowing through the branchs are equal [5].

However, as some of the parameters are not equal, the current values switched in paralleled mosfets may differentiate. Generally, the parameters producing the unbalance condition of the current are threshold voltage, gain and conduction resistance. Conduction voltage also includes thermal compensation [3].

5. MODEL ANALYSIS

Parallel operation is simulated with power circuit which consists of voltage source, inductance and semiconductors in parallel shown in Fig. 4. There are leakage capacities between gate-drain and gate-source of mosfets and inductances occuring on drain and source terminals.

It is accepted that there is a parameter conflict in the mosfets used for simulating the worst situation. N pieces of parallel circuits are separated into two branchs: On the first branch, there is a semiconductor of which parameters are conflicting, on the other branch, there is N-1 semiconductor of which parameters are the same.

For the mosfet model used for the analysis, equations stating the working levels shown in Fig. 5, are as follows;

Active area:

$$i_D = GF(V_{GS} - V_T)^2, \; V_{DS} \geq V_{GS} - V_T \geq 0 \qquad (15)$$

Ohmic area:

$$i_D = GF \cdot V_{DS}\left[2(V_{GS} - V_T) - V_{DS}\right]$$
$$V_{GS} - V_T \geq V_{DS} \geq 0 \qquad (16)$$

where;

i_D = the current flows from drain to source, V_{GS} = the voltage between gate-source, V_T = VGS threshold voltage, V_{DS} = the voltage between drain-source, GF = semiconductor gain coefficient [4].

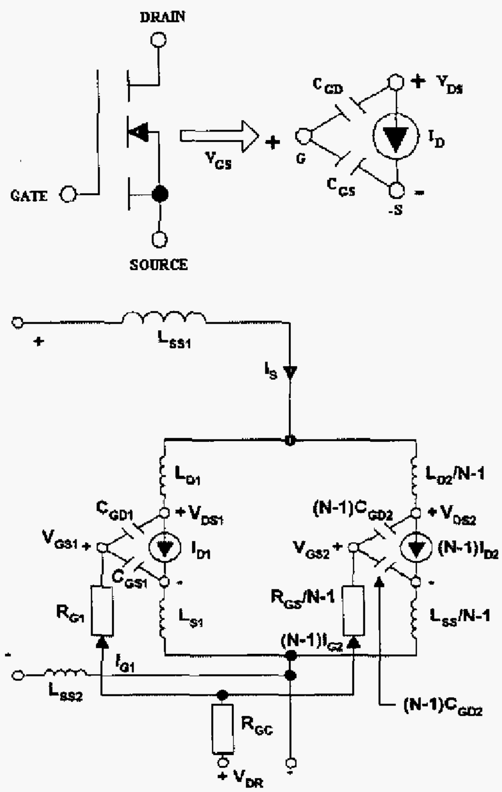


Fig. 4. Equivalent Model of the Mosfet

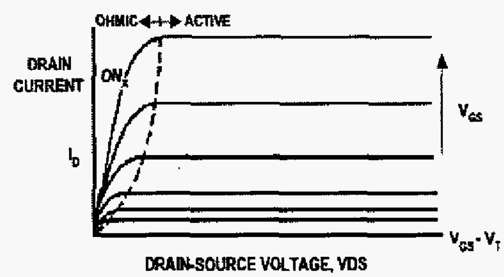


Fig. 5. Activity areas of the Mosfet

6. CONCLUSIONS AND DISCUSSIONS

A push-pull switch mode power supply was designed by paralleling the power with 3-power mosfets that of 42 amperes. The advantage conditions are;

1- Conduction resistance has been reduced from 0,036Ω to 0,018Ω for 25 °C.
2- Voltage-drop on mosfets has been reduced due to decrease of resistance value

3- Current transfer capacity of semiconductor elements has been increased to 126A and the system has become more reliable for the pick currents.

The mosfets which have equal parameters and are produced on the same date have been paralleled in the same branch. Mosfet gates are connected to drive resistances one by one in order to avoid oscillation. A resistance of 1 kΩ has been added between gate-source in order the mosfets not to be influenced by displaced currents. While designing on pcb, the inductance values which distance produces, have been ensured to be in the same level. Paying attention for all these measures, it has been observed that the heat value of the mosfets were the same during the activity and that the currents are switched in accordance with. In this system, efficiency is very high as to the ordinary transformers and the related graphic is shown in Fig. 6. As the losses on ferrit core are very less, they do not cause any heat increase.

In Fig. 7 and 8, the current switched by mosfet and voltage drop occured on it are shown respectively. It is also depicted the scheme of the proposed converter circuit in Fig. 9.

Since paralleled conductors are not used in primary of the transformer, a very high leak inductance has been occured and this inductance has limited the power with 385 watt. In order to enlarge the limit interval, the self value can be raised by using conductors paralleling in primary coil. At the same time, except the drain inductance, some inductances have been occured on pcb, at the active frequency, due to the distance length between transformer and semionductor elements. Therefore, in the drain-source area of the mosfets, very high voltages have been occured. These voltages have been limited in appropriate values by means of snubber circuits, and efficiency has been raised by charging the feeding battery without converting the whole of these excessive voltages to thermal energy.

In this study, there is a voltage feedback provides to remain the voltage constant on the output. Voltage ripple has a value under 30mV by the help of existing filter elements.

A converter with a very high efficiency, has been obtained and parallel working has been verified without any unbalanced operating condition from the measurements in the system.

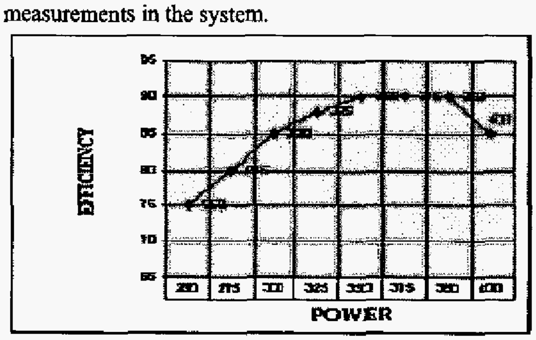


Fig. 6. Efficiency versus power in the converter

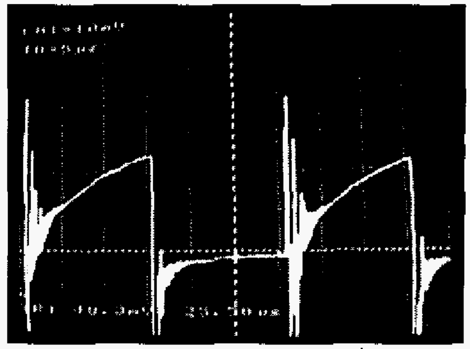


Fig. 7. Primary current versus time in the converter

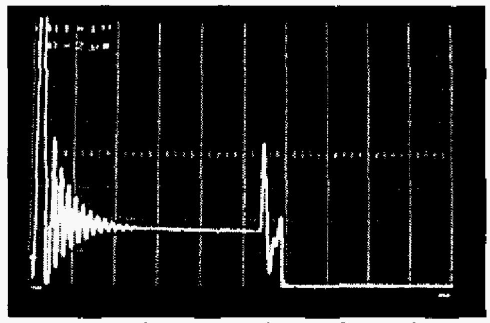


Fig. 8. Voltage versus time on the mosfet

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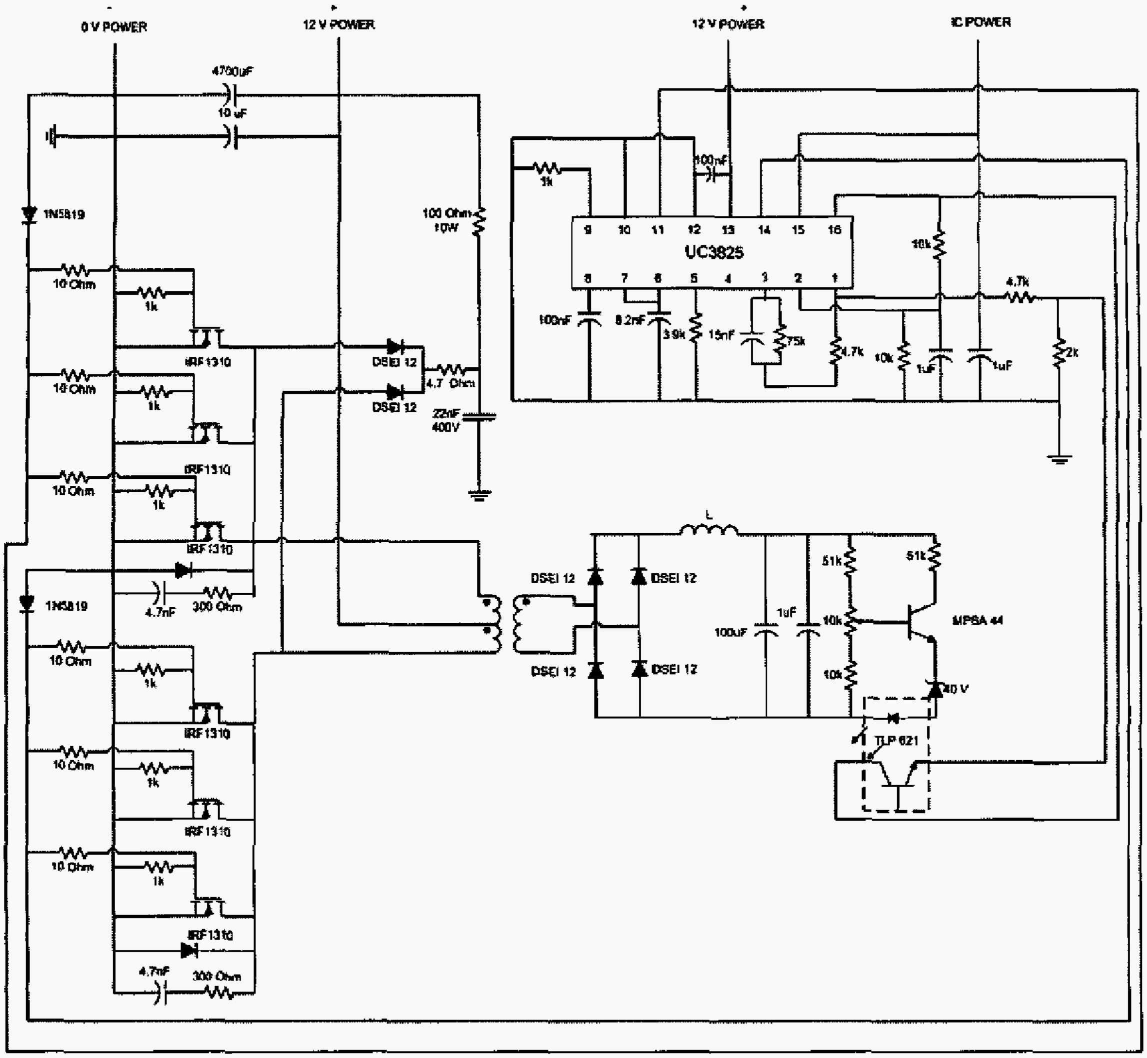

Fig. 9. Scheme of the proposed converter circuit