A new magnetron driving method using a phase-shifted active clamp forward converter for sulfur plasma tube applications

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Abstract
This paper presents a new method for driving the magnetron for a sulfur plasma tube using the two-switch forward converter structure with a phase-shifting active clamp (1000±40 W, 285 mA, 4 kV). Depending on the output voltage level required, the magnetron driver circuit is boosted and insulated and also has high gain. The use of the two-switch forward converter reduces the transistor voltage stress. In addition, applying the clamp structure balances the magnetization current. Besides, by controlling the phase shift of the clamp transistor, while maintaining the magnetic current balance of the power transformer, the duty cycle of the main transistors can be increased. This allows more voltage level transfer to the secondary winding of the transformer. Therefore, with a fixed transformer core, more voltage transmission is possible in comparison with a conventional forward converter. Moreover, using the PFC converter improves the power factor and stabilizes the DC-link voltage. The magnetron driver circuit provides a maximum power of 1 kW with an average power of 125–250 W by adjusting the converter's active time under minimum loss conditions. The magnetron driver circuit provided is validated by the simulation and experimental results.

1 | INTRODUCTION

The plasma sulfur tube is a plasma light bulb that is produced by microwave [1]. A magnetron tube is a microwave generator used to supply the energy needed to produce plasma from sulfur gas. The magnetron is a non-linear load whose resistance depends on the beam voltage. Therefore, in Magnetron Driver Circuit (MDC) design, isolated topology is used to supply the magnetron power supply. This prevents damage to the switching elements [2]. The proposed topology block diagram is presented in Figure 1.

The early MDCs used low-frequency converters, so their volume, weight, and losses were high [3, 4]. In [5, 6], the MDC Class E converter has been used. The advantage of this method is its simplicity. High voltage transistor stress can be considered as the disadvantage of this method. In [2, 7], the half-bridge inverter structure for MDC has been used. The disadvantage of this method is the high power transformer losses. In [8, 9], the full bridge inverter structure is used. The high number of power switches can be regarded as the disadvantage of this method. In this study, a new method for magnetron initiation for sulfur plasma tube is presented. The two-switch forward converter structures with active clamp and phase shift are used in the MDC design. The main idea of this research is presented in the control circuit. This is achieved by controlling the phase shift of the clamp switch while balancing the magnetization current of the transformer.

Magnetron specifications are given in Section 2, MDC performance in Section 3, Control Block Diagram in Section 4 and Simulation and Laboratory Test Results in Section 5.

2 | MAGNETRON TUBE SPECIFICATIONS

In a magnetron with the intersection of the DC electric field and the magnetic field in a very high vacuum, microwave is produced [10, 11]. The characteristic magnetron curve is shown in Figure 2. Magnetron has three working areas—non-anonymous, oscillating and discharging. When the magnetron beam voltage...
(V_{Mb}) is below the oscillation threshold voltage (V_{ot} = 3800 V), the beam current is negligible and no microwave is produced. Whenever the magnetron beam voltage exceeds V_{ot}, the magnetron enters the oscillating area and produces microwave. Now, if the voltage exceeds 4 kV, the magnetron enters the discharge area and is ready for re-operation. At this point, the output capacitors are discharged and the output voltage drops rapidly [12, 13]. Equations (1) and (2) show the magnetron circuit relationships.

$$
\begin{align*}
R_m &= 1 \, \text{M} \quad V_{Mb} < 3.8 \, \text{kV} \\
R_m &\approx 700 \, \Omega \quad 3.8 \, \text{kV} \leq V_{Mb} < 4 \, \text{kV} \\
R_m &\approx 0 \quad V_{Mb} > 4 \, \text{kV}
\end{align*}
$$

$$
i_{Lm} = V_{ot} \cdot i_o + R_m \cdot i_o^2
$$

In Equations (1) and (2), R_m is the magnetron resistance, V_{ot} is the oscillation threshold voltage, P_o is the magnetron power and i_o is the magnetron current.

3 | MDC DESIGN

The input power of the converter is 220 V, 50 Hz and the maximum voltage required is 4 kV. So, the MDC gain should be high. Also, taking the magnetron specifications into account, the isolated structure is essential for MDC [14, 15]. Since the magnetron receives maximum power from the grid in a short time, the use of PFC compensates for the adverse load condition. The PFC also stabilizes the DC link at 350 V. The PFC converter operates at a switching frequency of 100 KHz and its efficiency is more than 95%. V_{dc} is used for Filament and High Voltage. Figure 3 presents the proposed MDC.

In MDC, the switches S_{m1}, S_{m2} are the main transistors, the S_{C} switch is the clamp transistor, the S_{PFC} switch is the PFC transistor, the S_b switch is the filament transistor, the C_{C} is the clamp capacitor, TR_2 is the filament transformer, TR_1 is the power transformer, the ratio of the number of primary to secondary coil rounds N = n_p / n_o, D_{O1} and D_{O2} are output rectifier diodes, C_{O1} and C_{O2} are the output capacitors, R_{L} is the output voltage feedback resistor and R_L is the output current feedback resistor.

3.1 | MDC principal

The function of the converter is analysed in six steps as follows:

**State I** (t_0–t_1): At t_0, the main transistors turn ON on the ZVS condition. The leakage inductance current of the power transformer (i_{Lk}) is positive. The D_{O2} diode charges the C_{O2} capacitor and the D_{O1} diode is OFF. In the primary section, the primary winding current (i_{NP}) causes the i_p current to be positive. In this way, the main transistors pass the positive current i_p in the primary part of the power transformer. Afterwards, it is possible to increase the current and voltage resonance caused by the capacitor equivalent to the output and leakage inductance of the power transformer. Given that the switching frequency is greater than the resonant frequency, this condition continues until the main transistors are switched OFF at time t_1 [Figure 4(a)]. At this stage, i_{L,k} is calculated from Equation (3) and i_{L,m} is expressed by Equation (4).

$$
i_{L,k} = i_{L,k} (t_0) + \frac{V_{OT} - V_{C2}}{L_d} (t - t_0) \quad (3)
$$

where V_{OT} is the transformer output voltage.

$$
i_{L,m} = i_{L,m} (t_0) + \frac{V_d}{L_{m}} (t - t_0) \quad (4)
$$

**State II** (t_1–t_2): At this stage, the main transistors turn OFF, the i_{L,k} current is still positive and the C_{O2} capacitor is charged. In the primary part, i_p is positive and C_{Sm1} and C_{Sm2} conduct the i_p current. The voltage of the main transistors increases from zero to the positive value (V_{dc}/2). At the same time, the V_k voltage goes from (V_{dc} - V_{C1}) to zero and the D_{OC} diode conducts the i_p current. Due to the low capacity of the parasitic capacitors, this step is also very short and the value of i_p can be assumed to be constant [Figure 4(b)]. At this stage, i_{L,K} is expressed by Equation (5).

$$
i_{L,k} = i_{L,k} (t_1) \cdot \cos [\omega_0 (t - t_1)] + \frac{V_{OT} - V_{C2}}{Z_0} \cdot \sin [\omega_0 (t - t_1)]
$$

(5)
where $\omega_0 = 1/\sqrt{C_{O2}L_k}$, $Z_0 = \sqrt{I_k/C_{O2}}$.

**State III** ($t_2$–$t_3$): During this state, $D_k$ turns ON and $i_P$ is positive. The voltage in the primary part of the power transformer is equal to $V_{Cc}$. $i_{Lk}$ is still positive and the $C_{O2}$ capacitor is charged. By charging the $C_{O2}$ capacitor, the $i_{Lk}$ current is reduced to zero. This stage ends when the $i_{Lk}$ current reaches zero. Thus, $i_{Lk}$ is expressed by Equation (6). The $D_{O2}$ diode shuts OFF under normal conditions and the $D_{O1}$ diode turns ON naturally and softly in a negative cycle [Figure 4(c)].

$$i_{Lk} = i_{Lk}(t_2) + \frac{V_{OT} - V_{CO2}}{L_k} (t - t_2) \quad (6)$$

**State IV** ($t_3$–$t_4$): In $t_3$, transistor $S_C$ turns ON. The value of $i_{Lk}$ becomes negative, the $D_{O2}$ diode turns OFF and the $D_{O1}$ diode turns ON. The $C_{O1}$ capacitor is charged and provides resonant current and load current. Also, the $C_{O2}$ capacitor is discharging. Thus, the current of $i_P$ is negative and the transistor $S_C$ conducts the primary winding current of the power transformer. Due to the negative primary voltage, the $i_{Lm}$ current decreases. This condition continues until the $S_C$ transistor turns OFF [Figure 5(a)]. At this state, $i_{Lm}$ is calculated from Equation (7) and $i_{Lk}$ is expressed by Equation (8).

$$i_{Lm} = i_{Lm}(t_3) - \frac{V_{Cc}}{L_{m}} (t - t_3) \quad (7)$$

$$i_{Lk} = -\frac{V_{OT} - V_{CO1}}{L_k} (t - t_3) \quad (8)$$

**State V** ($t_4$–$t_5$): In $t_4$, $S_C$ is OFF and the $C_{SC}$ parasitic capacitor is charged by $-i_P$ current. The voltage of transistor
SC increases from zero to \((-V_{CC})\). At the same time, the Drain-Source voltage of the main transistors decreases from \([V_{dc} - V_{CC})/2\] voltage to zero. The condition of the output diodes and capacitors is maintained. As soon as the voltage of the main transistors reaches zero, the inherent diode of the main transistors \((D_{Sm1}, D_{Sm2})\) conducts. In this way, the conditions for turning on the main transistors in ZVS mode are provided. Due to the low capacity of the parasitic capacitor, this time period is very short and ends quickly [Figure 5(b)]. At this stage, \(i_{Lk}\) is expressed by Equation (9).

\[
i_{Lk} = i_{Lk}(t_4) \cos \left[\omega_1 (t - t_4)\right] + \frac{V_{OFF} - V_{CO1}}{Z_1} \sin \left[\omega_1 (t - t_4)\right]
\]  

(9)

where \(\omega_1 = 1/\sqrt{C_{O1}(L_{R})}\), \(Z_1 = \sqrt{L_{R}/C_{O1}}\).

State VI \((t_5-t_6)\): At time \(t_5\), the \((D_{Sm1}, D_{Sm2})\) diodes conduct the negative \(i_P\) current and the primary winding voltage of the transformer is equal to \(V_{dc}\). Also, both the primary winding current and the magnetization current \(i_{NP}\) and \(i_{Lm}\) are negative, leading to zero. In the secondary part, leakage inductance current \(i_{Lk}\) of the power transformer is also a negative value that goes to zero. Thus, \(i_{Lk}\) is expressed by Equation (10). In addition, the capacitor voltage \(V_{CO1}\) is still increasing and \(V_{CO2}\) is decreasing. This stage ends when the leakage inductance current \(i_{Lk}\) of the power transformer reaches zero. This naturally shuts OFF the \(D_{O1}\) diode. Furthermore, due to the fact that at this moment, the output voltage of the power transformer and the \(C_{O2}\) capacitor are equal, the \(D_{O2}\) diode is turned ON in normal and soft conditions [Figure 5(c)].

\[
i_{Lk} = i_{Lk}(t_5) + \frac{V_{OFF} - V_{CO1}}{L_{k}} (t - t_5)
\]  

(10)

Then, the converter returns to the first stage and the cycle is repeated. Figure 6 shows the shape of the waves corresponding to the six stages. Also, Figure 7 summarizes the six-state operation.
Switched on as ZVS.

\[ F_r < F_s \]  \hspace{1cm} (11)

\[ F_r = \frac{1}{2\pi \sqrt{C_r \times L_r}} \]  \hspace{1cm} (12)

\[ C_r = \frac{C_{O1} + C_{O2}}{N^2} \]  \hspace{1cm} (13)

\[ L_r = \frac{L_K}{N^2} \]  \hspace{1cm} (14)

\[ \frac{1}{2} \times \frac{L_K}{N^2} \times i_p^2 \geq \frac{1}{2} \times C_{\text{sm}} \times V_{\text{DSm}}^2 \]  \hspace{1cm} (15)

### 3.3 Clamp capacitor

The resonant frequency of the clamp transistor path is determined by the leakage inductance of the \( L_K \) power transformer and the clamp capacitor. The resonance frequency must be small enough so that no resonance occurs when the transistor is off [14]. The clamp capacitor is present when the resonant frequency period is half [18, 19]. Therefore, the resonance oscillation must be larger than the maximum shutdown time of the main transistors. Thus, the capacitance of the resonator capacitor is calculated by Equation (16).

In Equation (16), \( D_2 \) is the duty cycle of the clamp transistor \( S_C \).

\[ C_C = \frac{(1 - D_2)^2}{\pi^2 \times L_r \times F_s^2} \]  \hspace{1cm} (16)

### 3.4 Converter in the steady-state

In the early part of the circuit, the voltage and current stress of the main transistors are expressed as Equations (17) and (18) [20]. In Equation (17), \( D_1 \) is the duty cycle of the main transistors. In Equation (18), \( I_{\text{sm}} \) is the maximum current of the main transistors, \( V_{O-C} \) is the output voltage of the converter and \( R_{I_{\text{min}}} \) is the minimum load (the minimum load is in the oscillating area which is about 700 \( \Omega \)). The voltage of the clamp capacitor is also expressed by Equation (19). Besides, the voltage of the secondary winding of the power transformer is calculated by Equation (20). In Equation (20), \( V_{S/T} \) is the voltage of the secondary coil of the power transformer. The voltage gain of the DC/DC converter of the MDC is calculated by Equation (21). Also, the voltage stress of diodes \( D_{O1} \) and \( D_{O2} \) is expressed by Equation (22). In Equations (21) and (22), \( M_{V_{\text{dc}}} \) is the converter voltage gain and \( V_{P/T} \) maximum reverse voltage of the output diodes.

\[ V_{dc} = \frac{V_{dc}}{2(1 - D_1)} \]  \hspace{1cm} (17)
\[ I_{sw} = I_{Isw} + \frac{1}{2} \Delta I_{w} \]
\[ = \frac{1}{N (1 - D_1)} \times \frac{V_{O \_max} + N (1 - D_1)}{2L_{sw} F_3} V_{O \_C} \quad (18) \]

\[ V_{CI} = \frac{V_{dc}}{1 - D_1} \quad (19) \]
\[ V_{OF} = \frac{DV_{dc}}{N (1 - D_1)} \quad (20) \]

\[ M_{V_{dc}} = \frac{V_{O \_C}}{V_{dc}} = \frac{D_1}{N (1 - D_1)} \quad (21) \]
\[ V_{gU/D} = \frac{D_1 V_{dc}}{N (1 - D_1)} \quad (22) \]

3.5 | Filament power supply

To diffuse the microwave, the magnetron cathode must have a suitable temperature. The filament feed is designed with specifications (3/15 V, 10 A). In this study, a reducing fly-back converter was used for the filament power supply. The input voltage of the converter is 350 V and the output voltage is 3.15 V. If the power is transferred to the filament synchronous with the magnetron positioning in the oscillating area, it reduces additional losses in the filament. The magnetron beam feed is not isolated from the filament feed and the feed is controlled concerning the cathode.

3.6 | Transformer design

In MDC design, the transformer is very important. The transformer design is based on maximum magnetron power (1 kW). The primary voltage of the transformer is 350 V and the secondary voltage is 2 kV. Assume that the transformer losses are negligible. In addition to the transformer design, the maximum value of \( i_s \) and \( v_o \) is taken into account. The maximum output voltage value is calculated by Equation (23). The conversion ratio of the primary transformer coil to its secondary coil is obtained by Equations (24)–(26) [21]:

\[ V_{O \_max} = 2 \times V_{Sec \_max} \quad (23) \]

\[ \frac{D_1}{N} = \frac{n_s}{n_p} D_1 = \frac{V_{O \_max}}{2 \times V_{fhi \_max}} \quad (24) \]

\[ V_{fhi \_max} = V_{dc} \quad (25) \]

\[ N = \frac{2D_1 \times V_{dc}}{V_{O \_max}} \quad (26) \]

Calculations of transformer currents are calculated by Equations (27)–(29).

\[ \frac{i_p}{i_s} = \frac{n_s}{n_p} = \frac{1}{N} \quad (27) \]

\[ i_p = \frac{i_s}{N} \quad (28) \]

\[ i_p = \frac{V_{O \_max} \times i_s}{2D_1 \times V_{dc}} \quad (29) \]

4 | PROPOSED CONTROL BLOCK DIAGRAM

The main idea of the paper is presented in the control circuit section: “phase shifting the clamp transistor and increasing the duty cycle of the main transistors”, the voltage gain increases and the current stress reduces.

4.1 | Results of phase shifting

The use of active clamp topology balances the magnetization current. In this study, while maintaining the balance of the transformer magnetization current, the phase shift clamp transistor is fitted and it is possible to increase the duty cycle of the main transistors. Thus, the following results are obtained.

4.1.1 | Increment of the voltage gain

Increase the voltage gain of the converter according to Equation (21). As a result, the transformer round ratio decreases. Therefore, a smaller magnetic core is used in the converter.

4.1.2 | Current stress reduction

Reduce the transformer primary winding current according to Equation (18). This reduces the diameter of the primary winding of the transformer, the dimensions of the magnetic core and the current stress of the main switches.

Of course, increasing the duty cycle of the main transistors increases the stress voltage of the switches according to Equation (22). The use of two main switches in the forward converter structure reduces the effects of the problem. Figure 8 shows the waveforms caused by the phase shift.

4.2 | MDC control functions

The control circuit has three main functions: (1) controlling the magnetron average power, (2) controlling the magnetron beam current, and (3) controlling the filament power supply.
4.2.1 Control of magnetron average power

If the magnetron is constantly in the oscillating area, the magnetron will be damaged. Therefore, while operating, some time should be set aside for cooling it [10]. Thus, there are three working modes for MDC.

Non-oscillating area
The magnetron beam voltage is less than 3800 V and the magnetron is in the non-oscillating area [Disable mode or Cooling time (see Figure 9)].

Oscillating area
Magnetron beam has a voltage between 3800–4000 V. In this case, the magnetron is in the oscillating area [Enable mode (see Figure 9)].

Discharge area
Magnetron beam voltage exceeds 4000 V, the converter enters the discharge area and is ready for re-operation.

In the Disable mode, the converter transistors are off and no power is transmitted. In the Enable mode, the converter transistors are activated and the power is transferred to the load. At the end of the Enabled mode, the output capacitors are discharged. Therefore, the MDC’s performance is a combination of the gate signals [Figure 9(a)] and the Enable mode signal [Figure 9(b)]. Finally, the modulated wave signal is generated to activate the converter by combining the two signals [Figure 9(c)].

The average power of a magnetron is determined by adjusting the area (Enable mode) throughout $T_m$ by the average power controller [Figure 9(b)]. The Enable mode of the converter ($t_{En}$) determines the magnetron output average power. The value of $t_{En}$ is obtained by Equation (31). In Equation (30), $F_m$ is the frequency of the pulse repetition and $T_m$ its periodicity. Also, $T_m$, according to Equation (31), consists of two parts, $t_{En}$ when the MDC converter is active and $t_{Dis}$ when the MDC converter is disabled.

$$F_m = \frac{1}{T_m} \quad (30)$$

$$T_m = t_{En} + t_{Dis} \quad (31)$$

The average power of a magnetron is determined by adjusting the area (Enable mode) throughout $T_m$ by the average power controller [Figure 9(b)]. The Enable mode of the converter ($t_{En}$) determines the magnetron output average power. The value of $t_{En}$ is obtained by Equation (31). In Equation (32), $D_E$ is the converter Enable mode, $P_{O-max}$ is the maximum magnetron power in a period ($T_m$) and $P_{O-av}$ is the average power of the magnetron tube.

$$t_{En} = D_E \cdot T_m = \frac{P_{O-av}}{P_{O-max}} \quad (32)$$

Using the time control of "$S_{En} = 1\)", the average power of the magnetron is controlled (see Figure 10). The $S_{En}$ signal becomes binary value 1 at time $t_{En}$ and binary value 0 at time $t_{Dis}$. The $S_{En}$ signal thus has two modes Equation (33):

$$S_{En} = \begin{cases} 
1 & \text{Converter is ON} \\
0 & \text{Converter is OFF} 
\end{cases} \quad (33)$$
4.2.2 Controlling the magnetron beam current

The main transistors ($S_{m1}, S_{m2}$) gate signal is obtained by Equation (34) and the $S_C$ transistor gate signal is obtained by Equation (35).

$$S_{m1} = S_{m2} = S_g \cdot S_{En} \quad (34)$$

$$S_C = S_{Ph} \cdot S_{En} \quad (35)$$

In Equations (34) and (35), $S_g$ is the gate signal controlling the PI output current, and $S_{Ph}$ is the $S_g$ signal, which is phase-shifted $t_d$. $t_d$ is the phase shift of the clamp transistor. By increasing $t_d$, the voltage stress of the clamp transistor as well as the converter voltage gain increase. Therefore, a compromise must be made between increasing the voltage stress of the clamp transistor and increasing the gain of the converter voltage. The magnetron beam current is also controlled by the PWM signal duty cycle by the PI controller ($I_w = 285$ mA) (see Figure 10).

4.3 Controlling the filament power supply

Suitable power supply for the filament is one of the important tasks of the control block. The filament is fed at $t_{En}$ time. In this way, the loss of the filament is reduced. The filament power supply is equal to $(3/15, 10$ A). The block diagram of the MDC control is presented in Figure 10.

The MDC design is presented in the flowchart of Figure 11. First, the switching frequency is selected. The resonance frequency is then calculated using Equations (11)–(15). Due to the fact that the converter works in the CCM area, the switching frequency is greater than the resonant frequency to provide ZVS conditions. The value of the clamp capacitor is also calculated by Equation (16). The transformer is designed according to MDC ($V_{dc} = 350$ V, $V_o = 4$ kV, $P_O = 1$ kW, $F_S = 50$ kHz) considerations. If the value of the designed circuit elements does not provide the ZVS conditions, the design of the elements will be modified. Finally, the output current PI controller controls the amount of the current. Table 1 presents the proposed MDC components.

5 SIMULATION AND EXPERIMENTAL RESULTS

MDC performance is verified using PSCAD simulation software and empirical tests. For this purpose, the magnetron characteristic curve is given in Figure 2, the circuit is shown in Figure 3 and the MDC parameter values are given in Table 1.

| Parameter | Symbol | Value |
|-----------|--------|-------|
| Peak power | $P_{o-Peak}$ | 1000 ± 40 W |
| Mean power | $P_{o-mean}$ | 125–250 W |
| Input voltage | $V_s$ | 220 V,50 Hz |
| Output voltage(PFC) | $V_d$ | 350 V |
| Oscillation threshold | $V_{Ot}$ | 380 V |
| Resistance of oscillation zone | $R_{o-o}$ | 700 Ω |
| Resistance of non-oscillation zone | $R_{o-n}$ | 1 MΩ |
| Switching frequency | $F_s$ | 50 kHz |
| Resonant inductance | $L_r$ | 43.3 μH |
| Magnetic inductance of transformer | $L_m$ | 350 μH |
| DC link capacitor | $C_{dc}$ | 100 μF |
| Clamp capacitor | $C_C$ | 5.6 μF |
| Double rectifier capacitor | $C_{o1}/C_{o2}$ | 8.2 nF |
| Output capacitor | $C_o$ | 47 nF |
| Power transformer turn ratio | $n_{P3}/n_{P1}$ | 15/57 |
| Resonate frequency | $F_r$ | 49.64 kHz |
| Filament voltage | $V_h$ | 3.15 V |
| Filament current | $i_h$ | 10 A |
| Filament transformer turn ratio | $n_{Ph}/n_{Sh}$ | 71/3 |

5.1 Simulation result

Figure 12 shows the waveforms of MDC test points. In the simulation results, Figure 12(a) shows the transistors’ gate
signal. The magnetron beam voltage in the active area is 4 kV at 285 mA [see Figure 12(b) and (c)]. Figure 13(a) shows the voltage waveform of the main transistors at an Enable mode. Figure 13(b) shows the current waveform of the main transistors at an Enable mode to reduce the switching losses at ZVS conditions which are created for both transistors. Also, the maximum current of the main transistors is 17.5 A.

The maximum voltage of the main transistors is about 495 V. Figure 13(c) shows the waveform of the power transformer magnetization current. Figure 14(a) and (b) shows the voltage and current waveforms of the clamp transistor during a period of magnetron activity. Also, the maximum current and voltage of the clamp transistors are -12.5 A, 950 V. Finally, Figure 14(c) shows the waveform of the power transformer primary voltage.

The circuit parameters of the converters, the two-switch forward converter (TSFC), the active clamp two-switch forward converter (ACTSFC) and the phase-shifted active clamp two-switch forward converter (PSACTSFC) are compared in Table 2. Here, $L_r$ is resonant inductance, $n_S$ is secondary power transformer coil, $I_{smP}$ is maximum main switch current, $V_{smP}$ is maximum main switch voltage, $C_{O1}, C_{O2}$ are output capacitors.

| Converter | $L_r$ (µH) | $n_S$ | $I_{smP}$ (A) | $V_{smP}$ (V) | $C_{O1}, C_{O2}$ (nF) | $I_{Lm}$ (A) |
|-----------|-------------|------|----------------|---------------|----------------------|--------------|
| TSFC      | 30.2        | 59   | 25.3           | 952.6         | 8.2                  | 35.8         |
| ACTSFC    | 29.6        | 92   | 23.92          | 383.8         | 10                   | 13.85        |
| PSACTSFC  | 29.4        | 69   | 17.5           | 495           | 23.5                 | 14.73        |

$V_o = 4$ kV, $I_o = 285$ mA, $P_{out} = 250$ W
and $I_{LM}$ is the maximum magnetic induction current. By comparing the three converters in Table 2, the advantages of the PSACTFC converter are revealed.

5.2 Experimental test results

The proposed circuit was developed for experimental MDC testing (see Figure 15). The circuit consists of two parts—the control circuit and the power circuit. The main control circuit is a dsPIC microcontroller. The control circuit supplies the signal to the power circuit transistors. The control circuit also protects the power circuit against over-current and over-voltage. There are, of course, measures to protect against over-current and over-voltage in the power circuit.

The power circuit consists of three parts: (1) the first part is the AC/DC converter which rectifies the alternating input voltage, (2) the second one is the PFC converter which corrects the power factor and stabilizes the DC voltage, and (3) the third one is the MDC converter which is the magnetron drive circuit.

The specifications of the dsPIC microcontroller used in the converter are presented in Table 3. The ATP29F100B2 transistor is used for the main switches of the converter, because the voltage and current stress of the converter transistors are 500 V and 18 A, respectively. Also, according to the specifications of the clamp switch, transistor IXTK20N150 is used. For PFC switches, transistor 60R190C6 is a good option.

The required power supply for magnetron filament is 10 A, 3.15 V, so NDF05N50Z transistor is used. The EE30307 magnetic core is also suitable for filament feeding. Finally, the RUR30120 diode is used for output rectifiers. The circuit elements are presented according to the design shown in Table 4.

The control circuit produces a 50 kHz switching frequency to control the output current. The PRF of the magnetron active area is also 1 kHz. By sampling the output voltage and current,
the average output power, magnetron beam voltage, and magnetron beam current can be controlled. The elements used in the sampling circuit are listed in Table 5.

Figure 16 shows the waveforms of the circuit’s major points. Figure 16(a) depicts the gate signal of the transistors in the active area of the magnetron. The repeating period of the magnetron active area is 1 ms, the transistor gate signal is provided for a power of 250 W with a time of 250 µs. The output current PI controller regulates the output current at \( i_0 = 285 \text{ mA} \) and the output voltage at \( V_0 = 4 \text{ kV} \) [see Figure 16(b) and (c)].

The PI controller coefficients are: \( k_{PC} = 0.421 \) and \( k_{IC} = 46 \). Figure 17(a) illustrates the voltage and current curves of the main transistors. Resonant conditions in the main transistors are provided by the output capacitors and leakage inductance of the transformer. The ZVS conditions are shown in the main transistors [see Figure 17(a)].

The current stress of the main transistor is 14.1 A and the voltage stress of the main transistor is 495 V. Figure 17(b) shows the voltage and current curves of the clamp transistor. The current stress of the clamp transistor is 4.92 A. Meanwhile, the current stress of the Clamp Transistor Anti-Parallel Diode is 9.97 A and the voltage stress of the clamp transistor is 980 V. Finally, Figure 17(c) shows the transformer primary current waveform. The primary coil current stress of the power transformer is 10.85 A.

The input voltage \( (V_{in}) \) and input current \( (i_{in}) \) waveforms are depicted in Figure 18. Due to the presence of PFC, the input voltage and current were synchronized and the converter power factor became more than 95%.

Figure 19 shows the amount of loss in each converter element. The main converter losses are related to transistors. The creation of ZVS conditions in the converter, using the output capacitors and inductance leakage of transformer, has reduced the losses of transistors and transformer.

Table 6 signifies the efficiency of the converter in two modes of 125 and 250 W average power. Also, in Table 7, the converter provided is compared with other references. Therefore, MDC has fewer transistors, better transformer winding ratios and fewer losses than other references.

### 6 | CONCLUSION

This paper presented a new method for magnetron tube driving. The design results are validated using simulations and laboratory tests. The design process was described including the magnetron characteristics, power circuit performance, design and operation principles of the control circuit, active clamp circuit as well as series resonator circuit. The magnetron driving power supply specifications were considered according to the instantaneous and intermediate power requirements and 4 kV voltage. Power transformer leakage inductance and resonant capacitors ensured soft switching (ZVS) conditions. Using

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**TABLE 5** Sampling circuit elements

| Parameter          | Symbol | Value  |
|--------------------|--------|--------|
| Feedback capacitor 1 | \( C_F \) | 1.8 nF |
| Feedback capacitor 2 | \( C_v \) | 680 nF |
| Feedback resistance 1 | \( R_{F1} \) | 10 MΩ  |
| Voltage feedback resistance | \( R_v \) | 10 K   |
| Current feedback resistance | \( R_i \) | 0.1 Ω  |

**TABLE 6** MDC efficiency

| Output average power | 125 | 250 |
|----------------------|-----|-----|
| Output maximum power | 1000| 1000|
| Enable time (µs)     | 125 | 250 |
| PFC efficiency (%)   | 95.21| 96.43|
| MDC efficiency (%)   | 95.73| 96.52|
| Total efficiency (%) | 91.14| 93.1 |

**TABLE 7** Comparing the MDC with other references

| Comparison factor | Proposed | [9] | [11] | [12] |
|-------------------|----------|-----|------|------|
| TRR*1             | 0.2174   | 0.1613| 0.267| 0.1  |
| ZVS               | Yes      | Yes | Yes | Yes |
| PFC               | Yes      | Yes | No  | –    |
| Efficiency (%)    | 96.52    | 96.44| –   | –    |
| NS*2              | 3        | 4   | 4   | 4    |

*1TRR = Transformer round ratio.
*2NS = Number of switches.

The current stress of the clamp transistor is 4.92 A. Meanwhile, the current stress of the Clamp Transistor Anti-Parallel Diode is 9.97 A and the voltage stress of the clamp transistor is 980 V. Finally, Figure 17(c) shows the transformer primary current waveform. The primary coil current stress of the power transformer is 10.85 A.
the active clamp structure, the balancing in the magnetization current was restored and the main transistors’ current stress was reduced. According to this study, by shifting the phase of the clamp transistor, it is possible to increase the gain of the voltage while maintaining balance in the magnetization current. Thus, by a phase shift of clamp transistor, the turn-on time of the main transistors can be increased to 56%. This allowed the voltage gain to increase from 5.22 to 5.85. This idea reduced the number of secondary winding rounds to 69. Therefore, the size, weight and cost of the power transformer core are reduced. Also, the main transistor current stress was decreased from 23.92 to 14.1 A. The converter was experimentally tested under conditions of 125 and 250 W average output power. The efficiency of the converter at an average power of 250 W improves to 96.52%. Therefore, the converter provided is suitable for low-power sulfur plasma tube.

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