A Power-electronics-based Modular Nanosecond Pulse Generation System for Plasma-assisted Ignition

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Plasma-assisted ignition technology has been proposed to boost the combustion efficiency of scramjets during high-speed flight. One technique utilizes high-voltage nanosecond-duration pulses, which can generate free radicals, thereby initiating ignition earlier in the combustion chamber and improving fuel efficiency. A high-voltage nanosecond pulse generator is an integral part of the system. In this study, a modular nanosecond pulse generation system, utilizing multiple high-speed high-voltage MOSFETs, is developed and tested. The modular system can generate width-adjustable pulses (from 20 ns to 50 ns) with a fast rise time (< 6 ns), fast fall time (< 6 ns), and variable amplitude using multiple switch cells. The system is also scalable in voltage and current. By employing the inductive voltage adder, the system is configured in two different ways: two switch cells coupled in parallel and two switch cells coupled in series. These configurations demonstrate scalability and the fundamental sizing constraints.

Keywords: nanosecond pulse generation, inductive voltage adder

1. Introduction

The supersonic combustion ramjet (scramjet) requires the fuel and air to mix rapidly with subsequent ignition. The time required to ignite the fuel/air mix has an important impact on the overall efficiency of the scramjet⁽¹⁾, because the ignition delay time is of the same order of magnitude as the residence time in the combustor. Reducing the ignition delay time improves combustion efficiency, as more of the heat release chemistry, which produces thrust, occurs within the combustion chamber. Millisecond improvements in the time to ignite the fuel/air mix enhances the output power considerably because of the short-time the gas remains in the chamber before the hypersonic flow sweeps the gas out of the chamber⁽²⁾. Several methodologies have been proposed to reduce the ignition delay time; one is plasma-assisted hypersonic ignition, which generates reactive species (atomic oxygen, atomic hydrogen or hydroxyl radical) to the fuel during the combustion process⁽³⁾⁽⁴⁾. These species have the effect of accelerating the reaction rate and thus reduce the ignition delay time. The plasma-assisted hypersonic ignition technology has the benefit of shortening the ignition delay time especially in high speed combustion.

It is also important to ensure that these radicals are produced without significantly heating the flow. The thrust produced by the pressure rise due to combustion is strongly related to the temperature difference induced in the flow by the combustion reactions. Thus, a method for generating radicals in a non-thermal plasma is highly desirable.

One such method is the repetitively pulsed nanosecond-duration discharge. Other successful methods include aerodynamic stabilization⁽⁵⁾, electron-beam-controlled discharges^{(6)–(8)} and repetitively pulsed nanosecond-duration discharge⁽⁹⁾⁽¹⁰⁾. From a practical point of view, external ionization by nanosecond-duration pulses offers key advantages compared with other methods because short duration pulse generators are easy to operate and maintain⁽³⁾.

The plasma is generated by high-voltage nanosecond-duration pulses. The high voltage will form electrons with energies of 10–15 eV, elevating the probability of breaking down more molecules in either air or fuel⁽¹¹⁾. In other words, more reactive chemicals are generated. The plasma-assisted hypersonic ignition technology therefore has the potential of reducing ignition delay, increasing peak pressure and raising the heat release ratio in comparison with conventional ignition technologies⁽¹²⁾.

There are several popular methods of generating a high-voltage pulse, Marx generator, inductive voltage adder, Blumlein line pulse forming network and magnetic pulse compression⁽¹³⁾. The Inductive Voltage Adder (IVA) is the most suitable high voltage pulse generation topology in this application because of its minor pulse delay, variable pulse duration and peak amplitude. It is reported that an IVA based pulsed power generator could deliver a pulse of 4MV/60 ns⁽¹⁴⁾. Another IVA based pulse generator employs a transformer and a variety of semiconductor devices to expand peak voltage of output pulse⁽¹⁵⁾. However, the property of scalability through modularization for both current and voltage rating has not been investigated. The proposed pulse generation system investigates this property and this differentiates it from other IVA-based pulse generation system.

This paper investigates a power-electronics-based

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technology for improvement to hypersonic ignition technology. The overall objective is to develop a high-voltage, modular nanosecond pulse generation system. It should be capable of delivering a nearly square-shaped voltage pulse with the following parameters, fast rise time (< 6 ns), fast fall time (< 6 ns), variable pulse duration ranging from 20 ns to 50 ns and modular architecture such that the system is scalable in voltage and current. In addition, this paper also focuses on examining the effect of series-and parallel-coupling of the scalable system.

2. System Design

2.1 Literature Review The energy can be stored chemically (battery), mechanically (flywheel), or electrically (capacitor). The stored energy is shaped by a pulse forming system and released rapidly to the load by activating switches. The switch is a critical component of any high-voltage and compact pulsed power system.

With the evolution of semiconductor technology, solid state devices have attracted more and more attention for pulse power applications due to their excellent properties such as fast turn-on and turn-off time, low maintenance and cost. This leads to the retirement of traditional switches (spark gaps) in some applications. The most widely used semiconductor switches include MOSFET, IGBT, Thyristors and Silicon carbide (SiC).

The MOSFET can be easily switched on by charging the gate and switched off by discharging the gate with the Metal-Oxide-Semiconductor gate structure equivalent to a capacitance⁽¹⁶⁾. The advantage of fast operation speed qualifies the MOSFET as a candidate for MHz repetitive pulsed power system. Since turn-on resistance increases with the blocking voltage due to the increase in the path length of current through the semiconductor, the MOSFET has a practical voltage limitation of typically 1 kV and tends to be employed with a lower supply voltages. The switching characteristics of the MOSFET are affected by the intrinsic capacitance, resistance, gate charge and the reverse recovery characteristics of the body diode.

The IGBT has a similar structure to a MOSFET but with a P+ layer. Therefore, it operates in a similar way at turn-on. The injection of holes into the drift region significantly increases the conductivity to current flow, which is the advantage of IGBTs over MOSFETs⁽¹⁷⁾. The maximum voltage and current ratings are around 8 kV and 1000 A, respectively but not simultaneously. The internal structure of the IGBT includes a bipolar-junction-transistor, which provides a nearly constant voltage drop. As a result, for high-voltage ratings, the on-state losses are relatively small compared to the MOSFET at higher blocking voltages. From this point of view, the IGBT is suitable for high supply voltages. The only problem, however, is that it exhibits a tail current. In comparison to MOSFETs, IGBTs have slower turn-on and turn-off speeds but higher power capability.

Silicon carbide (SiC) is the most promising devices to be commercialized in the near future. Compared to Si devices, SiC devices have superior performances resulting from advanced electrical and material properties of SiC. At the present time, SiC JFETs are the most mature device. They have outstanding performance in their high voltage blocking

capability, low voltage drop in their on-state, less switching time and thermal resistance. This is very attractive for applications that demand high power density or high operating temperature. However, SiC JFETs are normally on and only a negative voltage could switch the device off; therefore, conventional gate drivers are not suitable.

The thyristor, employing a highly-distributed gate structure, can be turned on via the gate terminal but can only be switched off when the device conducting current falls to zero; thus, the switching speed is inferior to that of MOSFET and IGBT. It offers the advantage of high voltage rating (around tens of kV). The current rating is related to the size of the thyristor wafer. The gate turn-off thyristor (GTO) is an enhanced version of a thyristor, which can be forced off by diverting 20% typically of the collector current out of the gate. The advantage of high voltage and current ratings makes it suitable for low speed pulsed power⁽¹⁸⁾.

The superior operation speed of the MOSFET makes it the most suitable candidate for achieving the design criteria of this work, nanosecond switching speed. Although the power capability is inferior compared with other switches, a modular approach using MOSFET-based cells, that can be stacked, could provide a competitive technique.

2.2 Structure of the System The overall functional block diagram of the modular nanosecond pulse generation system employing inductive voltage adder is shown in Fig. 1. It utilizes several switch cells (in this case, 2) which are supplied by a common DC voltage supply, V_{dc} . Each switch cell includes a high-speed gate driver, a set of DC capacitors, a HV MOSFET switch and ancillary components. The gate driver receives the variable pulse-width logic signal from the controller unit so as to control rapidly the state of the MOSFET. Once the MOSFET turns on, a high-voltage output will be induced at the secondary windings, because of the rapid change of magnetic flux in the primary winding. Figure 2(a) shows the transformer arrangement used to series-couple output voltages from the two cells using two toroidal cores with a common secondary conductor. The cells can also be parallel coupled where each cell drives a primary wound around one toroidal core with one secondary turn, as shown in Fig. 2(b). In each case, a single primary and secondary turn is used to minimize leakage inductance (which rises to N^2). Leakage inductance is a primary limitation on achieving nanosecond rise times as it increases the L/R time constant of the circuit. The transformer reset circuit is vital for demagnetizing (re-setting) the transformer cores prior to the next pulse.

Fig. 1. Diagram of the proposed prototype design showing two switch cells series coupled and core reset circuit

Fig. 2. The arrangement of two switch cells coupled in (a) series with two cores where each core has one primary turn driven by its own cell and a common secondary turn and (b) parallel with one core where each of the two cells drive their own single primary turn and the secondary has a single turn

Fig. 3. The schematic diagram of the controller unit showing the pulse generation unit and pulse adjustment circuit

2.3 The Controller Unit The controller unit is designed to generate a debounced and adjustable-width signal. Figure 3 is a schematic diagram of the controller unit, which uses a dual retriggerable monostable multivibrator and a Schmitt input buffer. The retriggerable monostable multivibrator (U_1) produces a debounced signal once a test is initiated via switch S_1 .

The Schmitt input buffer (U_2) and its associated circuitry are used to reduce the duration of the pulse to 20 ns. The Schmitt input has adequate hysteresis to generate a jitter-free output. The variable resistor R_3 is used to change the transition rate of the input signal, which means the output pulse duration can be altered by varying R_3 , from 10 ns to 50 ns.

2.4 The Switch Cell Figure 4 shows the schematic diagram of the switch cell, the transformer and the R_{load} ($40\ \Omega$). The switch cell utilizes a gate driver to control a HV power MOSFET with a short duration pulse. The MOSFET

Fig. 4. The schematic diagram of the gate drive and HV MOSFET circuit board

is powered by a +700 V V_{dc} supply in this case. The source is connected to the circuit ground and the drain is connected to the primary winding of the transformer (T_1).

An IXYS nMOS DE475-102N21A (U_3) rated at 1000 V, 21 A and 30 MHz is selected because of its nanosecond switching speed (typically 5 ns), low turn on resistance R_g ($0.41\ \Omega$), and low stray inductance at the drain and the source terminals, L_G (1 nH) and L_S (0.5 nH).

An ultra-fast recovery diode D_1 (RHRG30120) is in parallel with the primary winding so as to freewheel the magnetizing and leakage current when the switch turns off. The anode of the diode should also be placed as close as possible to the drain of the MOSFET, so as to reduce the overshoot voltage due to commutation current flowing through the stray inductance.

The performance of the MOSFET switch in the system is significantly affected by the gate drive circuitry, especially when it comes to achieving the nanosecond rise and fall times. A CMOS high speed high current gate driver IXYS IXRF630 (U_2) is used because it can source and sink sufficient peak current, 30 A, to fully charge the MOSFET C_{gs} within 4 ns. Also, the minimum pulse width (8 ns), generated by the gate driver, is shorter than the design requirement (20 ns). The IXRF630 driver is a surface-mount device and incorporates patented RF layout techniques and a similar low inductance design to the MOSFET package.

The gate driver is powered by a +15 V supply, V_{cc} . It is bypassed via a group of surface mount capacitors (C_4 through C_{19}), which are symmetrically located on the two sides of U_2 . These surface mount capacitors can provide the 30 A gate drive current in 2–6 ns and are low inductance, low equivalent series resistance and high pulse current service capacitors. The capacitors avoid sags or ringing on the gate signal. Three different capacitance values are selected, $0.01\ \mu\text{F}$, $0.47\ \mu\text{F}$ and $4.7\ \mu\text{F}$.

2.5 Transformer Design The transformer uses nanoscale crystalline toroidal core (FINEMET F1AH0898) with o.d. 104 mm i.d. 76 mm and height 25 mm. This material is selected because of its superior magnetic properties in the application of nano-second pulse generation, including high saturation magnetic flux density (B_s) of 1.23 T, high permeability and low core loss. One layer of PVC insulation tape is wrapped around the primary and secondary foils fabricated from copper strip of thickness 0.4 mm and width 22 mm. The following relations are used to estimate the magnetizing inductance (L_m), the leakage inductance ($L_{leakage}$), the interwinding capacitance for the single layer winding (C_{ds}) and

Table 1. Specification of Parameter Values Used in Simulation

Parameter	Name	Value
V_{dc}	High voltage supply	700 V
R_{load}	Load resistance	40 Ω
$L_{leakage}$	Leakage inductance	30 nH
L_{trace}	Trace inductance	100 nH
$R_{winding}$	Winding Resistance	0.8 Ω
L_m	Magnetizing inductance	0.8 μ H
C_s	Stray capacitance	20 pF
C_d	Inter winding capacitance	25 pF

the stray capacitance (C_s) between the foil winding and the core. The calculated values of C_d , C_s , $L_{leakage}$ and L_m are given in Table 1.

The magnetizing inductance is estimated using ⁽¹⁹⁾

$$L_m = \frac{N^2 t \mu_r \mu_0}{2\pi} \ln \frac{b}{a} \dots \dots \dots (1)$$

Where μ_r is the relative permeability of the magnetic material, μ_0 is the permeability of free space, b is the outer diameter of the toroidal core (m), a is the inner diameter of the toroidal core (m), t is the thickness of the toroidal core (m) and N is the number of the primary turns.

The leakage inductance is estimated using ⁽²⁰⁾

$$L_{leakage} = \frac{2\pi MLTN^2 \left(c + \frac{d_1 + d_2}{3} \right)}{L_c} \times 10^{-8} H \dots \dots (2)$$

Where a is the height of the winding in meter, d_1 is the thickness of the secondary winding, d_2 is the thickness of primary winding, c is the distance between the primary and secondary windings, L_c is the thickness of the core and MLT is the mean length of a turn.

The interwinding capacitance is ⁽²¹⁾

$$C_d = \frac{\epsilon \epsilon_0 S}{2h_1 + h_2} F \dots \dots \dots (3)$$

Where h_1 , and h_2 are the insulation thickness of PVC and air respectively, ϵ and ϵ_0 are the relative and free space permeability respectively, S is the surface area of the edges of the secondary foil (m^2).

The stray capacitance is ⁽²¹⁾

$$C_s = \frac{\epsilon \epsilon_0 S}{h_1 + h_2} F \dots \dots \dots (4)$$

Where S is the surface area of the foil adjacent to the core.

The modular nanosecond pulse generation system with two switch cells is represented using the LTSpice software, as shown in Fig. 5. Because the single layer secondary is wound near a one-layer primary and the distance between them is small, the transformer is modeled by a lumped circuit ⁽²²⁾. It includes the winding resistance ($R_{winding}$), the primary inductances L_p , the secondary inductance L_s , the magnetizing inductance L_m , the leakage inductance ($L_{leakage}$) and the interwinding capacitance C_d . C_d and C_s models the interwinding capacitance of P and S as well as stray capacitance. The winding resistance $R_{winding}$ is measured at DC in the laboratory. L_{load} accounts for the parasitic inductance of R_{load} . Table 1 specifies these circuit elements. The lumped circuit model simplifies the analysis of the circuit.

(a) Series coupled two switch cells.

(b) Parallel coupled two switch cells.

Fig. 5. The transformer model of the system with two switch cells

In the series-coupled configuration, L_{m1} and L_{m2} are connected in series in the equivalent circuit so that $2V_{dc}$ is magnetically coupled to the secondary. However, L_{m1} and L_{m2} are parallel connected in parallel-coupled configuration, which means only one V_{dc} is coupled to the secondary. This configuration helps reduce leakage inductance and increase drive current capability, compared with the method of connecting switches in parallel and increasing the number of primary turns.

3. Experimental Results

The following experiments demonstrate that the nanosecond pulse generation system is scalable in voltage and current. Coupling switch cells in parallel can increase the peak current capability of the system for a given MOSFET current rating. Coupling switch cells in series amplifies the input supply by the number of switch cells.

Each configuration is examined under two load conditions: the transformer is unloaded (open circuit) and loaded with a 40 Ω resistor. The same experimental parameters are used for each condition: $V_{dc} = 700$ V and a 20 ns gate signal is fed into the gate driver.

The prototype of the single switch cell is shown in Fig. 6. The DC voltage supply (700 V) is employed to power the switch cell; the transformer is loaded with a 40 Ω resistor. It should be noted that the transformer core is reset by the DC reset circuit before each test is performed.

3.1 Two Switch Cells Coupled in Parallel Figure 7 shows the prototype of the nanosecond pulse generation system with two switch cells, magnetically coupled in parallel with only one high-voltage DC supply, V_{dc} (700 V) employed to power both switch cells. The secondary winding, S , is located between two primary windings P_1 and P_2 in order to improve the magnetic coupling and reduce leakage inductance. This method is superior to the method of connecting switches in parallel to increase the number of primary transformer turns. This is because this configuration is that the peak current capability of the system is increased because each primary winding shares half the load current. Also, the equivalent leakage inductance seen by the load (40 Ω) is reduced by half, but the distributed capacitance accounting for

Fig. 6. The prototype of the nanosecond pulse generation system with a single switch cell

Fig. 7. The prototype of two switch cells coupled in parallel

Fig. 8. The measured waveforms under the loaded condition ($40\ \Omega$ resistor) showing V_{ds1} (M1, 200 V/div) with $t_{rise} = 6.6\text{ ns}$, $t_{fall} = 4.4\text{ ns}$, and V_{ds2} (C2, 200 V/div) with $t_{rise} = 5.6\text{ ns}$ and $t_{fall} = 3.9\text{ ns}$

the inter coil capacitance of two primary windings increases.

The loaded condition with a $40\ \Omega$ resistor is examined first. Figure 8 shows both MOSFETs in the switch cells are triggered simultaneously. At turn on, the two voltage trajectories V_{ds1} and V_{ds2} are synchronized and well matched. When the switches are turned off, the turn-off voltage transients are no longer synchronized and are mis-matched. As a result, the pulse width of V_{ds1} is stretched to 24.6 ns, although duration of the gate signal is 20 ns. The second primary winding affects the distributed capacitance between the windings, leading to the mismatch between V_{ds1} and V_{ds2} .

Figures 9(a) and 9(b) show the waveforms of $I_{primary1}$ and $I_{primary2}$, as well as V_{ds1} and V_{ds2} . The two primary windings share equal current with maximum amplitude of 28 A. They are composed of I_m and a half of I_{load} , indicating I_m is $\sim 20\text{ A}$.

The I_{load} and V_{out} are illustrated in Fig. 9(c). V_{out} experiences a slight overshoot and reaches 650 V with a width of 23.4 ns and a small overshoot. Both the fall time (5.6 ns) and

(a) V_{ds1} (M4, 200 V/div) and $I_{primary1}$ (M3, 5 A/div)

(b) V_{ds1} (M4, 200 V/div) and $I_{primary1}$ (M2, 5 A/div)

(c) V_{out} (M1, 200 V/div) and I_{load} (C4, 5 A/div)

Fig. 9. Experimental results of the two switch cells coupled in parallel under the loaded condition ($40\ \Omega$ resistor)

(a) V_{ds1} (M1, 200 V/div) with $t_{rise} = 8.9\text{ ns}$ and $t_{fall} = 4.4\text{ ns}$ and $I_{primary1}$ (M3, 5 A/div)

(b) V_{ds2} (M2, 200 V/div) and $I_{primary2}$ (M4, 5 A/div)

(c) V_{out} (C2, 200 V/div)

Fig. 10. Experimental results of the two switch cells coupled in parallel under no load condition

rise time (3.5 ns) are smaller than 6 ns. The shape of I_{load} suffers from distortion; that is, longer rise time and fall time in comparison with V_{out} . Part of the reason for this is the limited bandwidth (100 MHz) of the current probe. The peak load current is 17 A.

Figure 10 shows the experimental results under the no-load condition. The pulse widths of V_{ds1} and V_{ds2} are also affected by the distributed capacitance and extended to 24.1 ns and 24.9 ns, respectively. The rise time of V_{ds1} and V_{ds2} are 2 ns longer than that in the loaded condition because of the

reduced primary current (≈ 18 A). A reduced primary current contributes to slower MOSFET turn-off as C_{ds} is charged slowly by the primary current.

Figure 10(c) shows V_{out} with a width of 26.1 ns has a fall time of 5.6 ns and a rise time of 2.9 ns, which is shortest transient time in all the experiments, as expected. The peak in V_{out} is almost 700 V.

The experimental results prove that the two switch cells coupled in parallel can generate a 23.4 ns voltage pulse, which satisfies the requirement of rise time and fall time. The parallel coupling relieves the current burden on the MOSFET. It increases the peak current capability of the system for a given MOSFET current rating. The increased equivalent distributed capacitance in the parallel configuration is considered as the dominant factor causing the distortion of the output pulse response, including mismatching of V_{ds1} and V_{ds2} at turn off and extended pulse width of V_{out} .

3.2 Two Switch Cells Coupled in Series The experimental set up of two switch cells coupled in series is shown in Fig. 11. The same V_{dc} (700 V) and load resistor (40 Ω) are employed. Each FINEMET core associated with a switch cell is stacked to inductively add the cell output voltages. Therefore, V_{out} can be made a multiple of V_{dc} (in this case, 2). However, the leakage inductance is increased due to the increased length of foil in the practical set-up and therefore the equivalent leakage inductance seen by the load resistor is doubled due to series coupling.

Figure 12 shows the drain to source voltage when the system is tested under the loaded condition. V_{ds1} and V_{ds2} are synchronized and well matched during both the turn-on and turn-off, unlike the parallel-coupled two switch cell configuration. This means the influence of distributed capacitance is

Fig. 11. The prototype of two switch cells coupled in series

Fig. 12. The measured waveforms under loaded condition showing V_{ds1} (C2, 200 V/div) with $t_{rise} = 3.9$ ns, $t_{fall} = 3.6$ ns, width = 18.2 ns, and V_{ds2} (M2, 200 V/div) with $t_{rise} = 4.2$ ns, $t_{fall} = 3.7$ ns, width = 18.2 ns

not as significant as in the parallel configuration. The pulse width is around 18 ns.

Unfortunately, V_{out} experiences distortion, as shown in Fig. 13(a). It has a rise time of 7.7 ns and fall time of 6.3 ns, which are longer than that in parallel configuration. The increase in rise time is caused by the doubled equivalent leakage inductance seen by load resistor, which leads to a doubled load current time constant.

The doubling effect of the series coupled configuration is proved by the waveforms of output load voltage (V_{out}), primary current, ($I_{primary1}$ and $I_{primary2}$), as well as load current (I_{load}), shown in Figs. 13(a), (b) (c) and (d). V_{out} reaches 1044 V. I_{load} is approximately 22 A. Also, $I_{primary1}$ (36.2 A) and $I_{primary2}$ (37.6 A) are close to expectation (40 A).

Figure 14 shows the measured waveforms when the load is disconnected. V_{out} reaches 1200 V in 3.3 ns at turn-on and 0 V in 5.4 ns at turn-off. The MOSFET is turned on and off in 5.3 ns and 5.2 ns, respectively. The magnetizing current, I_{m1} , which equals to $I_{primary}$, is still 19 A. It is demonstrated that, when the transformer is disconnected from the load, the series coupling configuration can deliver a narrow pulse (18.2 ns) with fast rise time (< 6 ns) and fall time (< 6 ns), which meets the design criteria.

When the load is connected, the series coupling can still generate a voltage pulse as requested but only when the R_{load} is increased high enough to maintain an appropriate time constant with the increased leakage inductance. This is shown in Fig. 15. A 1100 V/18 ns pulse with rise time of 4.3 ns fall time of 4.1 ns is produced when $R_{load} = 120 \Omega$. This confirms that the increased leakage inductance is a major factor affecting

(a) V_{out} (C2, 200 V/div)

(b) I_{load} (C4, 10 A/div)

(c) V_{ds1} (M1, 200 V/div) and $I_{primary1}$ (C4, 10 A/div)

(d) V_{ds2} (M4, 200 V/div) and $I_{primary2}$ (M3, 10 A/div)

Fig. 13. Experimental results of the two switch cells coupled in series in loaded condition (40 Ω resistor)

Fig. 14. The measured waveforms in no load condition

Fig. 15. The demonstration of generating V_{out} , which has improved rise time and fall time with $R = 120 \Omega$. V_{out} (M3, 200 V/div), V_{ds} (M1 and M2, 200 V/div) and I_{load} (M4, 5 A/div) are displayedTable 2. Comparison of V_{ds} and V_{out} with parallel coupling

Parameter	Waveforms (without load)			Waveforms (with load)		
	V_{ds1} (ns)	V_{ds2} (ns)	V_{out} (ns)	V_{ds1} (ns)	V_{ds2} (ns)	V_{out} (ns)
Rise time	6.7	8.9	2.9	6.6	5.6	3.5
Fall time	4.2	4.4	5.6	4.4	3.9	5.6
Duration	24.9	24.1	26.3	24.6	22.6	23.4

Table 3. Comparison of V_{ds} and V_{out} with series coupling

Parameter	Waveforms (without load)			Waveforms (with load)		
	V_{ds1} (ns)	V_{ds2} (ns)	V_{out} (ns)	V_{ds1} (ns)	V_{ds2} (ns)	V_{out} (ns)
Rise time	5.2	5.2	3.3	3.9	4.2	7.7
Fall time	5.3	5.3	5.4	3.6	3.7	6.3
Duration	21.2	21.2	23.5	18.2	18.2	22.9

the shape of V_{out} and I_{load} .

3.3 Discussion Table 2 and Table 3 compares the waveforms of drain to source voltages (V_{ds1} and V_{ds2}) and output waveform (V_{out}) in terms of rise time, fall time and the duration measured in parallel and series coupled configurations.

It is clear that fall time of V_{ds} , ranging from 3.6 ns to 5.6 ns, has small variation. This means it is the gate current generated by the gate driver that determines the switch on time of the MOSFET, rather than the load condition and the configuration of the system. On the contrary, the rise time of V_{ds} is strongly related to the magnitude of $I_{primary}$, which charges the C_{gd} and C_{ds} capacitance of the MOSFET to turn off the switch. Since it always decreases when load is removed, the

rise time of V_{ds} is always longer than that in the loaded condition in each configuration.

The load time constant, which is dominated by the ratio of equivalent leakage inductance and load resistance (L/R), determines the rise time of V_{out} . Due to the influence of the increased equivalent leakage inductance in series-coupled two switch cell configuration, the rise time (7.7 ns) of V_{out} is more than doubled, compared with the turn-on time of MOSFET (3.7 ns). However, the fall time (6.3 ns) is slightly longer than switch off time (4.2 ns) because the increased peak load current charges the C_{gd} and C_{ds} capacitance of the MOSFET to turn off the MOSFET, which in turn affects the fall time of V_{out} .

V_{out} in the parallel-coupled switch cell configuration has shorter rise time (3.5 ns). It is close to the switch-on time of MOSFET (3.9 ns) as reduced equivalent leakage inductance shortens the rise time; however, the extended duration of V_{out} and V_{ds} is a result of increased distributed capacitance caused by the parallel coupling.

The peak voltage for the single switch cell is ~ 800 V because exist of voltage overshoot and the maximum voltage rating of the DE475-102N21A MOSFET is 1000 V. The scramjet application will require voltage of tens of kV. Therefore, a multiple number of switch cells are required to scale the system up to 10 kV using series coupling. The equivalent load resistance of the system, which is a contributor to the voltage rise time in the scramjet application, is, as yet, unknown. If the load resistance is large, it is possible to deliver output pulse with nanosecond rise time and fall time, even with the expected increase in leakage inductance. If the load resistance is small, the rise time and fall time will be increased and much attention during the design phase will be required to reduce as far as possible the leakage inductance. Nonetheless, the combination of series and parallel coupling may provide a sufficiently fast rise time. More research investigating the equivalent load of the ignition system is required to better predict the load and optimize the system size and design.

The experimental results show that the parallel coupling configuration can be used to reduce the time constant inferred by the rising edge of V_{out} . The series coupling configuration increases the time constant because of increased leakage inductance. Reducing the leakage inductance and distributed capacitance plays a key role in delivering pulses with fast rise and fall times, and demonstrates that when using series coupling it may well be a requirement to also parallel-couple cells to limit the consequent increase in equivalent leakage inductance.

4. Conclusion

A MOSFET-based nanosecond pulse generation system utilizing an inductive voltage adder is proposed, developed and tested. It satisfies the design criteria successfully.

Employing the inductive voltage adder topology, the switch cell can be coupled in series or in parallel to form systems with expanded power capability.

The series-coupled two cell configuration has the feature of amplifying the input supply by the number of switch cells. The increased leakage inductance is responsible for the increased load time constant, which in turn lengthens the rise

time and fall time of the output pulse.

The parallel-coupled two switch cell configuration has the advantage of increasing the peak current capability of the system for a given MOSFET current rating. The increased equivalent distributed capacitance leads to the mismatch of the drain-to-source voltages at turn-off and distortion of the output pulse. However, the load time constant is decreased.

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