# Optically Switched-Drive-Based Unified Independent *dv/dt* and *di/dt* Control for Turn-Off Transition of Power MOSFETs

Hossein Riazmontazer, Student Member, IEEE, and Sudip K. Mazumder, Senior Member, IEEE

Abstract—To control the switching dynamics of an optically triggered hybrid device, a photonic-control mechanism is outlined. The optically triggered hybrid device comprises a power MOSFET, as the main power semiconductor device (PSD), and a pair of GaAs-based optically triggered power transistors (OTPTs), serving as the driver for the power MOSFET. The switching-transition controller modulates the turn-off transition of the power MOSFET by modulating the optical intensity of the OTPTs. The independent and unified dv/dt and di/dt control of the PSD is achieved using a single control circuit which also predicts the onset of transition between the di/dt and the dv/dt regions of control. Experimental control results validating the OTPT-based dynamical modulation of the turn-off characteristic of the power MOSFET are provided. In this study, the power MOSFET is chosen to be a SiC MOSFET. However, the proposed photonic-control mechanism can be extended to Si power MOSFETs as well.

Index Terms—Active gate drive, di/dt, dv/dt, electromagnetic interference (EMI), gallium arsenide (GaAs), IGBT, optical gate drive, optically triggered power transistor (OTPT), power MOSFET, silicon carbide (SiC), switching loss, switching transients.

## I. INTRODUCTION

S the industry moves toward to high-frequency and highpower-density applications, controlling the switchingtransition dynamics of power semiconductor devices (PSDs) attains importance. Those applications include: switched mode power supplies, motor drives, solar inverters, battery chargers etc., [1]–[6]. As the switching frequency increases, the duration of the switching transition needs to be decreased to reduce the switching losses. However, this leads to higher di/dt and dv/dt, which in turn, causes higher electromagnetic interference (EMI) and device stress. Unification of the dv/dt and di/dt controls enables a high-frequency power-electronics system to achieve controllability and performance optimization over a wide operating range. Even though dv/dt and di/dt are intertwined,

Manuscript received September 29, 2013; revised March 28, 2014; accepted May 15, 2014. Date of publication May 30, 2014; date of current version November 3, 2014. This work was supported in part by the U.S. National Science Foundation Awards (Nos. 1002369 and 0823983), received by Prof. S. K. Mazumder at the University of Illinois, Chicago. However, any opinions, findings, conclusions, or recommendations expressed herein are those of the authors and do not necessarily reflect the views of the NSF. Recommended for publication by Associate Editor Y. C. Liang.

The authors are with the Laboratory for Energy and Switching Electronics Systems, Department of Electrical and Computer Engineering, University of Illinois at Chicago, Chicago, IL 60607 USA (e-mail: hossein.riazmontazer@gmail.com; mazumder@uic.edu).

Color versions of one or more of the figures in this paper are available online at http://ieeexplore.ieee.org.

Digital Object Identifier 10.1109/TPEL.2014.2327014

ability to control them independently provides a pathway toward total performance optimization with regard to switching loss, device stress, and EMI for any given operating condition.

In recent years, there has been growing interest to control the turn-off transition behavior of electrically triggered (ET) PSDs in the literature [7]–[17]. Modulation of the gate resistance has been proposed in [7]-[9]. In this scheme, different gate resistors are inserted in the discharging path of the gate circuit of the PSD during switching transition. Because only discrete values for the gate resistances are attainable, the optimized switching performance is not achievable over a wide operating range. Other approaches like controlling the gate current of the PSD or introducing an intermediate voltage to the gate of the PSD at selected intervals during turn-off transition has been proposed in [11]–[15]. However, work on control of both dv/dt and di/dt has been limited [16], [17]. Independent control of turn-off dv/dt and di/dt of a PSD, by controlling the gate current, is outlined in [16]. However, unification of the control approach has not been achieved because two individual circuits are used to control di/dt and dv/dt. Unified closed-loop di/dt-and-dv/dt control mechanism has been proposed in [17]. The di/dt and dv/dt are individually controlled by a reference voltage and circuit-dependent feedback gains. However, using the same reference voltage for both di/dt and dv/dt along with the constant feedback gains in each switching-transition period makes it difficult to adjust dv/dt without bounding adjustability of di/dt and vice-versa.

Although controlling the switching transitions of ET PSDs has been explored, limited work has been conducted on the switching-transition control of optically triggered (OT) PSDs [18]–[20]. Recent work on monolithic and hybrid OT PSDs [21]–[27] have demonstrated the feasibility of using a single optical link for both pulse-width modulation (PWM) and switching-transition control of a PSD using a controller, which is spatially separated from the PSD power stage. However, the optical-to-electrical conversion delay is appreciable as compared to the total duration of the switching transition which adversely affects the performance of the transition controller with feedback [21], [23].

In this paper, turn-off switching-transition control of an OT hybrid PSD, comprising two GaAs-based optically triggered power transistors (OTPTs) and a SiC MOSFET, is outlined. The outlined mechanism for optical control can be extended to Si power MOSFETs as well because of similarities in device behavioral dynamics [28], [29]. The OTPTs are placed in the charging (turn-on) and discharging (turn-off) paths of the gate of the SiC



Fig. 1. Test circuit and control block diagram. V1 and V2, respectively, control  $dv_{\rm DS}/dt$  and  $di_D/dt$  of M1 in the  $dv_{\rm DS}/dt$ - and  $di_D/dt$ -control regions of operation as illustrated in Fig. 2. The threshold condition, for the onset of transition between the  $dv_{\rm DS}/dt$ - and  $di_D/dt$ -control regions, is provided in Section II-A.

MOSFET. Unified turn-off dv/dt and di/dt control are achieved using a single circuit by modulating the intensity of the optical beam that triggers the OTPT, which controls the turn-off of the SiC MOSFET. A laser driver is designed to dynamically adjust the optical intensities for dv/dt and di/dt control. The independent control of turn-off dv/dt and di/dt is achieved by means of a control circuit which compensates for the total delay in the control loop. It also predicts the moment of transition between dv/dt and di/dt regions of control.

The rest of the paper is organized as follows: in Section II, initially the principle of optical control of turn-off dv/dt and turn-off di/dt of an OT SiC MOSFET is illustrated. Subsequently, the threshold condition and control concept for predicting the onset of transition from the dv/dt to the di/dt control region, which ensures the independent controllability of the two regions of control, is outlined. Then, mathematical analyses for availability of the independent control of turn-off dv/dt and di/dt are provided. Finally, the experimental results validating feasibility of the proposed control under different operating conditions are demonstrated in Section III. It is noted that, in the remaining paper, the terms dv/dt and di/dt are used to represent the rate of change of voltage across and the rate of change of current through a general PSD. While,  $dv_{\rm DS}/dt$  and  $di_D/dt$  are used to represent the rate of change of voltage across and the rate of change of current through the SiC MOSFET (M1) shown in Fig. 1.

## II. TURN-OFF TRANSITION BEHAVIOR AND CONTROL

The standard clamped-inductive test circuit and control block diagram for optical transition control are shown in Fig. 1. The test circuit comprises a bridge leg with the hybrid device package (comprising M1 and the two OTPTs) placed in the low side and a self-gated SiC MOSFET (M2) in the high side. MOSFET M2 has characteristics similar to the characteristics of the SiC MOSFET in the hybrid package. OTPT1 and OTPT2 work complementarily and turn the SiC MOSFET (M1) on and off, respectively.



Fig. 2. Turn-off behavior of the MOSFET and control circuit key waveforms. The output currents  $L_1$  and  $L_2$  of the laser driver are proportional to the voltage commands  $V_1$  and  $V_2$ , which dictate the  $dv_{\rm DS}/dt$  and  $di_D/dt$  dynamics of M1 in the  $dv_{\rm DS}/dt$  and  $di_D/dt$  control regions.

As indicated in Fig. 2, when the turn-off command is initiated by the PWM signal at  $t_0$ , the laser driver provides the current level  $L_1$  (proportional to the external voltage control command  $V_1$  shown in Fig. 1) for the laser with its wavelength centered at 808 nm. The laser delivers an optical power corresponding to the current level  $L_1$  to the base region of the OTPT2 via an optical link. OTPT2 then turns-on, after some delay, at  $t_d$ , allowing the gate charge of M1 to be discharged through it. The turn-on delay of OTPT as a function of optical power has been measured using the resistive circuit of Fig. 3 employing the point to point method. The results are then plotted in Fig. 3. The higher optical intensity results in smaller turn-on delay due to the higher rate of photo-generated carrier density inside the OTPT as shown in Fig. 3. More information about the characteristics and behavior of OTPT is provided in [30] and [31]. After the OTPT is turned on, the gate-to-source voltage  $(v_{\rm GS})$  of M1 starts to fall until it reaches the Miller plateau voltage ( $V_{\text{Miller}}$ ); subsequently, the drain-to-source voltage  $(v_{DS})$  of M1 begins to rise. The slope of  $v_{\rm DS}$  is approximated using the following relation:

$$dv_{\rm DS}/dt \approx \frac{v_{\rm GS,TH} + i_D/g_{fs}}{R_G C_{\rm GD}} = \frac{V_{\rm Miller}}{R_G C_{\rm GD}}.$$
 (1)

In (1),  $g_{fs}$  is the forward transconductance of M1,  $i_D$  is the drain current,  $C_{GD}$  is the gate-to-drain capacitance of M1 also known as Miller capacitance,  $R_G$  is the gate resistance, and  $v_{GS,TH}$  is the threshold voltage of M1. However, the gateto-drain capacitance of MOSFETs is a nonlinear function of  $v_{DS}$ .  $C_{GD}$  of M1 is approximated as a two-step function of



Fig. 3. (a) Resistive load test circuit and (b) turn-on delay of OTPT versus the optical power, using the resistive-load circuit.  $V_{\rm Bias} = 10 V$ ,  $R_{\rm Load} = 200 \Omega$ , frequency = 50 kHz, and duty cycle = 50%.



Fig. 4. Gate-to-drain capacitance of M1 (also known as Miller capacitance) as a function of drain-to-source voltage.

drain-to-source voltage of M1 as shown in Fig. 4:

$$C_{\rm GD} = \begin{cases} C_{\rm GD,avg1}, & v_{\rm DS} < v_{\rm DS1} \\ C_{\rm GD,avg2}, & v_{\rm DS} \ge v_{\rm DS1}. \end{cases}$$
(2)

Equation (2) is mostly true for other types of power MOS-FETs and even IGBTs, as well [32]. Typically,  $C_{\text{GD,avg1}}$  is dramatically higher than  $C_{\text{GD,avg2}}$ . Therefore, slope of the drain-to-source voltage of M1 before  $v_{\text{DS}}$  reaches the turning point  $v_{\text{DS1}}$  (corresponding to the time  $t_1$  in Fig. 2), is significantly lower as compared to the duration in which  $v_{\text{DS}} \ge v_{\text{DS1}}$ . The interval in which  $v_{\text{DS}}$  is lower than  $v_{\text{DS1}}$  is referred as ohmic region, as shown in Fig. 2. Furthermore, the interval in which  $v_{\text{DS}}$  is higher than  $v_{\text{DS1}}$  and lower than the bus voltage



Fig. 5. Resistance of OTPT versus the optical power, using the resistive-load circuit of Fig. 3.

 $(V_{\text{Bus}})$  is referred as  $dv_{\text{DS}}/dt$  control region (corresponding to the interval between  $t_1$  and  $t_3$  in Fig. 2).

According to (1), the  $dv_{\rm DS}/dt$  can be controlled by varying the resistance in the discharging path of the gate. The latter in turn is adjusted by changing the optical intensity of OTPT2. Current level  $L_1$  of the laser driver sets the optical intensity in the  $dv_{\rm DS}/dt$  control region. Resistance of OTPT as a function of optical power is measured using the resistive circuit of Fig. 3 and data are plotted in Fig. 5.

When  $v_{\rm DS}$  matches the bus voltage ( $V_{\rm Bus}$ ) at  $t_3$ , current  $i_D$  falls which causes an overvoltage across M1 due to the parasitic inductances in the commutation path. This region is referred to as the  $di_D/dt$  control region. The overvoltage ( $\Delta v_{ov}$ ) across the drain-to-source terminals of M1 is given by the following expression:

$$\Delta v_{ov} = L_c \cdot \frac{di_D}{dt} \tag{3}$$

where  $L_c$  is the sum of the parasitic inductances in the commutation path, which includes the parasitic inductances of M1 and M2, bus-parasitic inductance, and trace inductances. In the  $di_D/dt$  control region,  $di_D/dt$  is given by the following expression:

$$\frac{di_D}{dt} \approx -\frac{V_{\rm GS,TH} + i_D/2g_{fs}}{R_G C_{iss}/g_{fs} + L_{\sigma}}.$$
(4)

In (4),  $C_{iss}$  is the input capacitance of M1 and  $L_{\sigma}$  is the sum of parasitic inductances seen from the source of M1. Following (4), the  $di_D/dt$  is controlled by varying the resistance of the discharging path of the gate of M1. This resistance is adjusted by changing the optical intensity of OTPT2, as shown in Fig. 3. Current level ( $L_2$ ) of the laser driver sets the optical intensity in the  $di_D/dt$  control region and  $L_2$  is proportional to the external voltage control command  $V_2$ , as shown in Fig. 1.

Of course, while the control of  $dv_{\rm DS}/dt$  and  $di_D/dt$  in their respective regions of operation is important, a seamless transition between the  $dv_{\rm DS}/dt$  and  $di_D/dt$  control regions is equally



Fig. 6. Effect of late transition between the  $dv_{\rm DS}/dt$  and  $di_D/dt$  control regions on device stress and switching loss.

important. In the following section, we derive this threshold condition for transition between the two control regions and outline its implementation.

## A. Threshold Condition for Transition Between the $dv_{\rm DS}/dt$ and the $di_D/dt$ Control Regions

The transition between  $L_1$  and  $L_2$ , thereby transitioning from the  $dv_{\rm DS}/dt$  to the  $di_D/dt$  control region, is initiated by the control circuit illustrated in Fig. 1. This transition guarantees the independent control of  $di_D/dt$  and  $dv_{\rm DS}/dt$ . Following, [7], [14], [16], an easy way to detect the onset of the  $di_D/dt$  control region is to detect the change in the  $di_D/dt$  from near zero to a significantly larger value. If this approach is adopted, the onset of transition is initiated later than the desired instant due to control-loop and OTPT-related delays. So, one may lose the control over  $di_D/dt$  in all or a part of this  $di_D/dt$  control region, which may lead to excessive device stress or switching loss as illustrated in Fig. 6. Another approach [8] for predicting the onset of the  $di_D/dt$  control region is based on detecting the saturation region of the voltage corresponding to  $dv_{\rm DS}/dt$ control region in Fig. 2 and initiating the transition between the  $dv_{\rm DS}/dt$  and the  $di_D/dt$  control regions after a fixed delay. However, the error in the prediction of onset of transition is significant in applications where the  $dv_{\rm DS}/dt$  varies over a wide range.

Therefore, in the proposed scheme, to ensure the independent control of  $dv_{\rm DS}/dt$  and  $di_D/dt$ , a simple control circuit is designed which predicts the onset of transition between the  $dv_{\rm DS}/dt$  and  $di_D/dt$  control regions based on the actual  $dv_{\rm DS}/dt$  and the scaled bus-voltage reference  $(v_{\rm ref})$ . The  $di_D/dt$  control region onsets at  $t_3$ , when  $v_{\rm DS}$  reaches the bus voltage  $(V_{\rm Bus})$  as shown in Figs. 2 and 7. However, considering a combined constant delay of  $\Delta t$  seconds due to the feedback and due to the delay in the actuation of OTPT2, the control regions at time  $t_2(=t_3 - \Delta t)$ . An accurate onset of transition ensures the independent controllability of the  $dv_{\rm DS}/dt$  and  $di_D/dt$  control regions.



Fig. 7. Signals  $v_{\rm DS}$  and  $dv_{\rm DS}/dt$ , at the desired time  $(t_2)$  of transition for two different  $dv_{\rm DS}/dt$ , considering the constant delay of  $\Delta t$  seconds in the feedback loop.

Now, using Fig. 7, one can show that for any given  $dv_{\rm DS}/dt$  the following equality holds:

$$v_{\mathrm{DS},t2} + (dv_{\mathrm{DS},t2}/dt) \cdot \Delta t = V_{\mathrm{Bus}}.$$
 (5)

In (5),  $v_{\text{DS},t2}$  and  $dv_{\text{DS},t2}/dt$  represent, respectively, the values of  $v_{\text{DS}}$  and  $dv_{\text{DS}}/dt$  at time  $t_2$ . Assuming the scaling factors of  $\alpha_1$ ,  $\alpha_2$ , and  $\alpha_3$  associated with sensing  $V_{\text{Bus}}$ ,  $v_{\text{DS}}$ , and  $dv_{\text{DS}}/dt$ , respectively, the following equations hold:

$$v_{\rm ref} = \alpha_1 . V_{\rm Bus} \tag{6}$$

$$\psi_{\mathrm{DS},t2} = \alpha_2 . v_{\mathrm{DS},t2} \tag{7}$$

$$d\dot{v}_{\mathrm{DS},t2}/dt = \alpha_3.(dv_{\mathrm{DS},t2}/dt).$$
 (8)

In (7),  $\psi_{\text{DS},t2}$  represents the value of sensed  $v_{\text{DS}}$  at time  $t_2$  with a scaling factor of  $\alpha_2$ . In (8),  $d\psi_{\text{DS},t2}/dt$  represents the value of sensed  $dv_{\text{DS}}/dt$  at  $t_2$  with a scaling factor of  $\alpha_3$ . Substituting (6)–(8) into (5) yields the following relation:

$$\frac{\psi_{\mathrm{DS},t2}}{\alpha_2} + \frac{d\psi_{\mathrm{DS},t2}/dt}{\alpha_3} \cdot \Delta t = \frac{v_{\mathrm{ref}}}{\alpha_1}.$$
(9)



Fig. 8. Schematics of the control circuit and the laser driver.

 TABLE I

 TRUTH TABLE OF THE CONTROL CIRCUIT OF FIG. 8

Comparator	$\overline{\mathrm{PWM}}$	Q	$\bar{\mathrm{Q}}$	LEV1	Lev2
1	0	1	0	1	0
0	1	0	1	0	0
1	1	1	1	0	0
0	0	Х	Х	Х	Х

Because  $\Delta t$  is considered to be a constant, (8) can be rewritten as follows:

$$\beta_1 . \dot{v}_{\mathrm{DS},t2} + \beta_2 . (d\dot{v}_{\mathrm{DS},t2}/dt) = v_{\mathrm{ref}}.$$
 (10)

In (10),  $\beta_1$  equals to  $(\alpha_1/\alpha_2)$  and  $\beta_2$  equals to  $(\alpha_1.\Delta t/\alpha_3)$ . Using (10), the threshold condition for transition from the  $dv_{\rm DS}/dt$  to di/dt control region is found to be

$$(\beta_1.\dot{v}_{\rm DS} + \beta_2.(d\dot{v}_{\rm DS}/dt)) - v_{\rm ref} \ge \varepsilon.$$
(11)

In (11),  $\psi_{\rm DS}$  and  $d\psi_{\rm DS}/dt$  represent, respectively, the sensed values of  $v_{\rm DS}$  and  $dv_{\rm DS}/dt$  at any time with scaling factors of  $\alpha_2$  and  $\alpha_3$  whereas  $\varepsilon$  represents a very small positive value. Essentially, (11) indicates that there is a time t (ideally  $t = t_2 = t_3 - \Delta t$ ) at which the difference between  $\beta_1 . \psi_{\rm DS} + \beta_2 . (d\psi_{\rm DS}/dt)$  and  $v_{\rm ref}$  is either zero or very close to zero. This concept is used to design a controller which compensates for the delay in the feedback loop and ensures a seamless transition between  $dv_{\rm DS}/dt$  and  $di_D/dt$  control regions.

The control circuit and laser driver schematics are shown in Fig. 8. The coefficients  $\beta_1$  and  $\beta_2$  are considered to be equal to make the design of the control circuit easier. Therefore, the coefficients  $\alpha_1$  and  $\alpha_2$  have the following relation:

$$\beta_1 = \beta_2 = \beta \Rightarrow \frac{\alpha_1}{\alpha_2} = \frac{\alpha_1}{\alpha_3} \Delta t \Rightarrow \alpha_2 = \frac{\alpha_3}{\Delta t}.$$
 (12)

The sensed  $v_{\rm DS}$  and  $dv_{\rm DS}/dt$  are scaled with the proper coefficient  $\beta$  and added using the OP1, as shown in Fig. 8, where  $\beta = R_1/3.(1 + R_3/R_2)$ . The output of OP1 is then compared with  $v_{\rm ref}$  using a comparator to monitor if the threshold condition in (11) is met. If (11) is satisfied, the control circuit initiates the transition from  $dv_{\rm DS}/dt$  to  $di_D/dt$  control region by setting Lev1 to logic state 0 and Lev2 to logic state 1 using the D-FF and AND operators of Fig. 8.

Because of the negative  $dv_{\rm DS}/dt$  in the  $di_D/dt$  control region, the threshold condition might not be satisfied in the  $di_D/dt$  control region. Therefore, a D flip-flop is used to prevent undesirable fluctuations of the logic states of signals Lev1 and Lev2 in the  $di_D/dt$  control region. The truth table for the control circuit can be found in Table I. In Table I, X means no change in the state of the signal. Any negligible error in initiating the onset of transition (i.e.,  $\varepsilon \neq 0$  and instead  $\varepsilon \approx 0$ ) is due to the nonidealities in the circuit elements, nonlinearities, and error in the estimation of the total delay of the feedback loop [18].

Subsequent to the change in the logic states of Lev1 and Lev2, the laser driver changes its output current  $(I_{\text{Laser}})$  from  $L_1$  to  $L_2$  (which is proportional to  $V_1$  and  $V_2$ ) and is given by the following relation:

$$I_{\text{Laser}} = \begin{cases} \gamma.V_1 = L_1, & \text{when Lev1 signal is high} \\ \gamma.V_2 = L_2, & \text{when Lev2 signal is high} \\ 0, & \text{when PWM signal is low.} \end{cases}$$
(13)

In (13),  $\gamma$  is a circuit-dependent constant and it is equal to  $\gamma = 0.1/R_6$ . The outputs of the flip-flop Q and  $\overline{Q}$ , in Fig. 8, work complementarily. Furthermore, the Lev1 and Lev2 signals are derived using the AND operation of the PWM signal with  $\overline{Q}$  and Q, respectively. Therefore, Lev1 and Lev2 signals in (12) are complement to each other in the duration when the PWM signal is high and they are both low when PWM signal is low, as illustrated in Fig. 2 and Table I. Following (1), (4), and (14), one can adjust  $dv_{\text{DS}}/dt$  and  $di_D/dt$  by respectively controlling  $V_1$  and  $V_2$ , which in turn control the output current of the laser driver to magnitudes of  $L_1$  and  $L_2$ . Modulating the optical intensity by the proposed laser driver, along with the implementation of the threshold condition (11), enables one to attain the unified  $dv_{\text{DS}}/dt$  and  $di_D/dt$  control.

## B. Availability of Independent $dv_{\rm DS}/dt$ and $di_D/dt$ Control

In Section II-A, the threshold condition for independent control of turn-off  $dv_{\rm DS}/dt$  and  $di_D/dt$  was derived. Subsequently, the control circuit was designed based on the aforementioned threshold condition. However, the threshold condition was derived considering the following assumptions:

- 1) The  $\Delta t$  (total delay in the feedback loop and OTPT) is fixed.
- 2) The  $dv_{\rm DS}/dt$  is fixed from the time at which the transition command initiates  $(t_2)$  up to the desired moment of transition at  $t_3$ . Essentially the  $dv_{\rm DS}/dt$  is fixed during the delay time of OTPT.
- 3)  $\Delta t$  is less than the duration of  $dv_{\rm DS}/dt$  control region  $(\Delta t < (t_3 t_1))$ .

Based on the mathematical analysis and threshold condition in Section II-A, the control circuit can independently control the turn-off  $dv_{\rm DS}/dt$  and  $di_D/dt$  as long as the aforementioned



Fig. 9. Measured transition delay of resistance of the OTPT when the optical intensity changes from  $P_1$  to  $P_2$ .

assumptions are valid. Therefore, the domain of validity of the aforementioned assumptions shall be analyzed to specify the boundaries for the availability of independent  $dv_{\rm DS}/dt$  and  $di_D/dt$  control.

In Section II-A, it is assumed that the  $\Delta t$  is fixed. However,  $\Delta t$  varies proportionally to the difference between the optical intensities in the  $dv_{\rm DS}/dt$  and the  $di_D/dt$  control regions. In this control scheme, the turn-on and turn-off delays of OTPT are not important. However, the important delay is defined as the total time that it takes for the resistance of OTPT to change from the value R1 (corresponding to the optical intensity  $P_1$ and the laser current of L1) and reaches the final resistance value of R2 (corresponding to the optical intensity  $P_2$  and the laser current of  $L_2$ ). In order to derive the transition delay of the OTPT when it is subject to a step change in its receiving optical intensity, OTPT is tested using the resistive-load circuit of Fig. 3. In this setup, OTPT receives the optical intensity of  $P_1$ through the laser and optical link which causes the voltage drop of  $V_{d1}$  across the OTPT. Subsequently, the optical intensity is varied from  $P_1$  to  $P_2$  which makes the voltage drop  $V_{d2}$  across the OTPT. The delay is measured between the time at which the optical intensity is changed to  $P_2$  and the time at which the voltage drop across OTPT reaches 90% of its final value ( $V_{d2}$ ). The measured delay time for different values of  $P_1$  and  $P_2$  is derived using the point to point method, and plotted in Fig. 9.

If optical intensities  $P_1$  and  $P_2$  are equal  $(P_1 = P_2)$ , the transition delay is essentially zero. However, the value of transition delay in this case is selected such that the plot of Fig. 9 is smooth.

Now consider the case in which the optical intensity  $P_1$  is applied to the OTPT in the  $dv_{\rm DS}/dt$  control region, to control the turn-off  $dv_{\rm DS}/dt$ . Similarly, the optical intensity  $P_2$  is applied in the  $di_D/dt$  control region to control the turn-off  $di_D/dt$ . Therefore, the transition delay for the resistance of OTPT2 to change from R1 (corresponding to the optical intensity  $P_1$ ) and reaches the value of R2 (corresponding to the optical intensity  $P_2$ ) is  $\Delta t_1$ , based on Fig. 9. Considering the implemented transition delay in the control circuit is equal to  $\Delta t$ , the transition error because of the variable transition delay of OTPT2,  $(E_{r1})$ , is defined by the following expression:

$$E_{r1} = \Delta t_1 + \Delta t_c - \Delta t. \tag{14}$$

In (14),  $\Delta t_c$  is the delay of the control circuit. If  $E_{r1} > 0$ , it affects the controllability over the  $di_D/dt$  in the  $di_D/dt$  control region, and if  $E_{r1} < 0$ , it affects the controllability over the  $dv_{\rm DS}/dt$  in the  $dv_{\rm DS}/dt$  control region. Therefore, the proportional error due to the variable transition delay of OTPT2,  $(E_{r1}\%)$ , is defined as

$$E_{r1}\% = \frac{E_{r1}}{t_{\text{rise}}} = \frac{E_{r1}}{i_D} \cdot dt_D / dt \times 100\%, \text{ if } E_{r1} > 0$$
$$E_{r1}\% = \frac{E_{r1}}{t_{\text{fall}}} = \frac{E_{r1} \cdot dv_{\text{DS}} / dt}{V_{\text{Bus}} - v_{\text{DS}1}} \times 100\%, \text{ if } E_{r1} < 0.$$
(15)

In (15),  $t_{\rm rise}$  is the rise time of the drain-to-source voltage of M1, and  $t_{\rm fall}$  is the fall time of the drain current of M1. In this study, the independent transition control of turn-off  $dv_{\rm DS}/dt$  and turn-off  $di_D/dt$  is valid, if  $E_{r1}\% \leq 10\%$ .

The control circuit in Section II-A is designed using the threshold condition and considering a fixed  $dv_{\rm DS}/dt$  in the  $dv_{\rm DS}/dt$  control region. However, because of nonlinear behavior of  $C_{\rm GD}$  the actual  $dv_{\rm DS}/dt$  in the  $dv_{\rm DS}/dt$  control region is not fixed. Therefore, the actual  $dv_{\rm DS}/dt$  has some deviation from the average  $dv_{\rm DS}/dt$  which is defined by  $C_{\rm GD,avg2}$ . This deviation might results in a transition error, especially in the case of a high  $dv_{\rm DS}/dt$  in which the value of  $v_{\rm DS}$  at  $t_2$  ( $v_{\rm DS,t2}$ ) is close to  $v_{\rm DS1}$ .  $t_2$  is the time at which the transition from the  $dv_{\rm DS}/dt$  to the  $di_D/dt$  control region is initiated by the control circuit, as shown in Fig. 2.

Following (2) and Fig. 4, the average  $C_{\rm GD}$  in the  $dv_{\rm DS}/dt$  control region is equal to  $C_{\rm GD,avg2}$ . Therefore, the average  $dv_{\rm DS}/dt$  which is defined by  $C_{\rm GD,avg2}$ , is equal to  $dv_{\rm DS,avg}/dt$ . Hence, the drain-to-source voltage of M1 at which the threshold condition is satisfied and transition is initiated is derived by the following expression, using (5):

$$v_{\rm DS2} = V_{\rm Bus} - \left( \frac{dv_{\rm DS,avg}}{dt} \right) \cdot \Delta t.$$
(16)

However, if  $v_{\text{DS2}}$  is close to  $v_{\text{DS1}}$ , the value of  $C_{\text{GD}}$  at  $v_{\text{DS2}}(C_{\text{GD2}})$  is higher than the value of  $C_{\text{GD,avg2}}$ , as shown in Fig. 4. Because  $C_{\text{GD2}}$  is higher than  $C_{\text{GD,avg2}}$ , the slope of the drain-to-source voltage of M1 at  $v_{\text{DS2}}$  is lower than the  $dv_{\text{DS,avg}}/dt$ , according to (2). Therefore, the value of  $dv_{\text{DS}}/dt$  at  $v_{\text{DS2}}$  is not high enough to satisfy the threshold condition. As a result, the threshold condition is satisfied and transition is initiated, after some delay, at  $v_{\text{DS3}}(v_{\text{DS3}} > v_{\text{DS2}})$ .  $v_{\text{DS3}}$  is derived using the following expression:

$$v_{\rm DS3} = V_{\rm Bus} - \left( \frac{dv_{\rm DS3}}{dt} \right) \cdot \Delta t. \tag{17}$$

In (17),  $dv_{\rm DS3}/dt$  is the  $dv_{\rm DS}/dt$  at voltage  $v_{\rm DS3}$  corresponding to  $C_{\rm GD3}$  in Fig. 4. The transition is initiated  $\Delta t$  seconds after  $v_{\rm DS}$  reaches  $v_{\rm DS3}$ . However, the actual time that it takes for  $v_{\rm DS}$  to reach  $V_{\rm Bus}$ , ( $\Delta t_2$ ), (beginning of the  $di_D/dt$  control region) is derived using the following equation:

$$\Delta t_2 = \frac{dv_{\rm DS3}/dt}{dv_{\rm DS,avg}/dt} \,.\,\Delta t. \tag{18}$$

In (18), it is assumed that the average  $dv_{\rm DS}/dt$  from  $v_{\rm DS3}$  to  $V_{\rm Bus}$  is equal to  $dv_{\rm DS,avg}/dt$ . Therefore, the transition error due to variable  $dv_{\rm DS}/dt$  in the  $dv_{\rm DS}/dt$  control region,  $(Er_2)$ , is equal to:

$$Er_{2} = \Delta t - \Delta t_{2} = \left(1 - \frac{dv_{\text{DS3}}/dt}{dv_{\text{DS,avg}}/dt}\right) \cdot \Delta t$$
  

$$Er_{2} = \left(1 - \frac{C_{\text{GD,avg2}}}{C_{\text{GD3}}}\right) \cdot \Delta t.$$
(19)

Because  $C_{\text{GD3}}$  is always greater than  $C_{\text{GD,avg2}}$  in high  $dv_{\text{DS,avg}}/dts$ ,  $Er_2$  is positive. Positive  $Er_2$  means that, the transition error does not affect the controllability over the  $dv_{\text{DS}}/dt$  in the  $dv_{\text{DS}}/dt$  control region, but it affects the controllability over the  $di_D/dt$  in the  $di_D/dt$  control region. If the amount of the transition error  $(Er_2)$  is considerable comparing to  $t_{\text{fall}}$  of M1, the independent controllability of  $dv_{\text{DS}}/dt$  and  $di_D/dt$  is not granted. Therefore, the proportional error due to variable  $dv_{\text{DS}}/dt$  in the  $dv_{\text{DS}}/dt$  control region,  $(Er_2\%)$ , is defined as follows:

$$Er_2\% = \frac{Er_2}{t_{\text{fall}}} \times 100\% = Er_2 \times \frac{di_D/dt}{i_D} \times 100\%.$$
 (20)

In this study,  $Er_2\% \leq 10\%$  is considered as an acceptable error for independent controllability of  $dv_{\rm DS}/dt$  and  $di_D/dt$ .

Following, the procedure of calculating the quantitative boundaries of  $dv_{\rm DS}/dt$  and  $di_D/dt$  for having the independent controllability of  $dv_{\rm DS}/dt$  and  $di_D/dt$  is described. Initially, the maximum applicable  $di_D/dt$ ,  $(di_{D,\max}/dt)$ , is derived for a given load current and bus voltage, using (3) and considering the maximum allowable overvoltage. Then, the minimum fall time of the drain current of M1,  $(t_{\rm fall,min})$ , is calculated using the following expression:

$$t_{\rm fall,min} = \frac{i_D}{di_{D,\max}/dt}.$$
 (21)

Subsequently,  $P_2$  is calculated using (4) and Fig. 5. Knowing  $P_2$ ,  $\Delta t$  is selected using Fig. 9, such that the conditions in (14) is satisfied. Afterward,  $Er_2$  is calculated knowing the maximum admissible  $Er_2\%$  for  $t_{\rm fall,min}$ , using (20). Then, the value of  $C_{\rm GD3}$  is derived, using (19), and the corresponding voltage for  $C_{\rm GD3}$  ( $v_{\rm DS3}$ ) is derived, using Fig. 4. Therefore, the maximum allowable  $dv_{\rm DS}/dt$  is equal to

$$dv_{\rm DS,max}/dt = \frac{V_{\rm Bus} - v_{\rm DS3}}{(1 - Er_2\%/100) \cdot \Delta t}.$$
 (22)

If  $dv_{\rm DS,max}$  is selected properly, one will not face the condition in which  $\Delta t$  is less than the duration of  $dv_{\rm DS}/dt$  control region ( $\Delta t < (t_3 - t_1)$ ). However, if this condition happens the independent controllability of turn-off  $dv_{\rm DS}/dt$  and  $di_D/dt$  is not granted. The flowchart of the procedure of calculating the quantitative boundaries of  $dv_{\rm DS}/dt$  and  $di_D/dt$  is shown in Fig. 10.

#### **III. EXPERIMENTAL RESULTS**

The standard clamped-inductive test circuit of Fig. 1 along with the proposed control circuit are designed and fabricated as shown in Fig. 11. The fabricated set up includes the power



Fig. 10. Flowchart of the procedure of calculating the quantitative boundaries of  $dv_{\rm DS}/dt$  and  $di_D/dt$  for independent control of turn-off  $dv_{\rm DS}/dt$  and  $di_D/dt$ .

circuit, the hybrid device package, power-supply circuits and the laser on the top side of the board and the control circuit, laser driver and sensing circuits on the bottom side of the board. The implemented SiC power MOSFET is CMF10120D with break-down voltage (BV) of 1200 V, current rating of 24 A,  $C_{GD} = 7 \text{ pF}$ , Ciss = 928 pF and Coss = 68 pF. A 2 W, 808 nm fiber-coupled laser is used to trigger the OTPT2, as shown in Fig. 11. The threshold current of the laser is 0.4 A, and its output optical power as a function of input current is shown in Fig. 12. An additional inductor is used in the commutation path to simulate the effect of the leakage inductance in the isolated dc/dc converters such as cuk, flyback, and forward. Voltage sensing circuit has a band-width (BW) of 200 MHz and delay of 5 ns. The experimental waveforms are measured using Tektronix DPO7104, which has the BW of 1 GHz. A 25 MHz differential voltage-probe along with a 50 MHz current sensors are used to provide the signals for the oscilloscope. Subsequently, the measured data are plotted using MATLAB software.

Experimental results for the independent optical control of the  $di_D/dt$  of M1 with fixed  $dv_{\rm DS}/dt$  are shown in Fig. 13. The output current of the laser-driver  $(I_{\text{Laser}})$  remains the same in the  $dv_{\rm DS}/dt$  control region. This leads to the same resistance for OTPT2, which is placed in the discharging path of the gate of M1, for all the cases. As the result, the current through OTPT2 and the turn-off  $dv_{\rm DS}/dt$  in the  $dv_{\rm DS}/dt$  control region is kept similar for all of the cases following (1). A step change in the output current of the laser driver initiates the onset of transition from the  $dv_{\rm DS}/dt$  to the  $di_D/dt$  control region. The proper time of transition ensures independent controllability in the  $dv_{\rm DS}/dt$  and  $di_D/dt$  control regions. Therefore, one is able to control the slope of the drain current in the  $di_D/dt$  control region without affecting the controllability over slope of the drain-to-source voltage in the  $dv_{\rm DS}/dt$  control region. This is illustrated in Fig. 13. As current level  $(L_2)$  of the laser driver in the  $di_D/dt$  control region decreases the turn-off  $di_D/dt$ , the



(a)



(b)

Fig. 11. Fabricated test set up: (a) top side of the board which includes the power circuit, laser and power-supply circuits; and (b) bottom side of the board which includes the control circuit, sensing circuits, and laser driver.

voltage overshoot and oscillation reduce as well. Fig. 13 validates that, the  $di_D/dt$  of M1 is dynamically controlled by controlling the current flowing through OTPT2 (i.e.,  $I_{OTPT2}$ ). The latter is dependent on the optical intensity of OTPT2, which in turn, is dependent on the output-current level of the laser driver.

The gate-to-drain and input capacitances of the SiC power MOSFETs are dramatically lower than their Si counterparts with the same rating (i.e., compare the  $C_{\rm GD} = 77 \,\mathrm{pF}$  and  $C_{iss} = 21 \,nF$  of IXFL32N120P, to  $C_{\rm GD}$  and  $C_{iss}$  of M1). Therefore, a higher gate resistance (which can be 10 to 20 times higher) is needed for a SiC MOSFET to ensure that it has the same  $dv_{\rm DS}/dt$  as compared to its Si counterpart, following (3). A higher gate resistance in the case of having the same  $dv_{\rm DS}/dt$ , results in a dramatically lower gate current. Furthermore, during the discharging time of the  $C_{iss}$  of M1, OTPT acts like a constant current-source and prevents the current spike of the gate current which is a common phenomenon in the conventional gate drive circuits, as shown in Fig. 13. This behavior also results in a



Fig. 12. Output optical power of the laser at the end of the fiber-optic cable as a function of laser current  $(I_{Laser})$ .



Fig. 13. Measured turn-off waveform of  $v_{\rm DS}$ ,  $i_D$ ,  $I_{\rm laser}$ , and  $I_{\rm OTPT2}$  with varied  $di_D/dt$  (of 30, 20, 10, and 5 A/ $\mu$ s) and the fixed  $dv_{\rm DS}/dt$  (of 600 V/ $\mu$ s).

longer turn-off delay for M1. If the traditional gate drive with the fixed gate resistance is used, the gate voltage exponentially decreases by the following equation until it reaches the Miller voltage ( $V_{\text{Miller}}$ ):

$$v_{\rm GS} = V_{CC} \cdot e^{-t} R_G C_{iss} \,. \tag{23}$$

In (23),  $V_{CC}$  is the gate bias voltage. Therefore, the delay time for this case is derived by the following equation:

$$t_{d-R} = R_G C_{iss} . \ln \left( V_{CC} / V_{\text{Miller}} \right).$$
<sup>(24)</sup>

In (24),  $t_{d-R}$  is the turn-off delay time of M1 with the conventional fixed-gate-resistance method. On the other hand, if the OTPT is used in the gate circuit of M1, the gate current  $(i_G)$  is derived using following relation:

$$i_G = \frac{V_{CC} - V_{\text{Miller}}}{t_{d-O}} \cdot C_{iss} = \frac{V_{\text{Miller}}}{R_G}.$$
 (25)

In (25),  $t_{d-O}$  is the turn-off delay time of M1 in the case of using the proposed optical approach. Therefore,  $t_{d-O}$  is derived using the following equation:

$$t_{d-O} = R_G C_{iss} \cdot \frac{V_{CC} - V_{\text{Miller}}}{V_{\text{Miller}}}.$$
 (26)



Fig. 14. Measured turn-off waveforms of  $v_{\rm DS}$ ,  $i_D$ , and  $I_{\rm laser}$  with varied  $dv_{\rm DS}/dt$  (of 530, 330, and 225 V/ $\mu$ s) and a fixed  $di_D/dt$  (of 17 A/ $\mu$ s).

Consequently, the ratio of the turn-off delay times for the two cases is:

$$\frac{t_{d-O}}{t_{d-R}} = \frac{(V_{CC}/V_{\text{Miller}} - 1)}{\ln(V_{CC}/V_{\text{Miller}})}.$$
(27)

Following (27) and depending on the load current of M1, one can conclude that using the proposed approach the turnoff delay of M1 is 1.3–2.5 times longer as compared to the conventional fixed-resistance method, in the case of having the same  $dv_{\rm DS}/dt$ .

The experimental results for the independent optical control of the  $dv_{\rm DS}/dt$  of M1 with the fixed  $i_D/dt$ , are shown in Fig. 14. The turn-off  $dv_{\rm DS}/dt$  increases by increasing the current level of the laser driver in this region. The onset of transition from  $dv_{\rm DS}/dt$  to di/dt control region is set properly by control circuit for different values of  $dv_{\rm DS}/dt$ . This ensures the independent  $dv_{\rm DS}/dt$  and  $di_D/dt$  controls. Therefore, one is able to control the drain-to-source voltage slope in the  $dv_{\rm DS}/dt$  control region without affecting the controllability of slope of the drain current in the  $di_D/dt$  control region, as shown in Fig. 14. The outputcurrent levels of the laser driver remain the same in the  $di_D/dt$ control region which lead to the same resistance for OTPT2. Therefore,  $di_D/dt$  and overvoltage for all the cases are same according to (3) and (4).  $di_D/dt$  can be adjusted regardless of the value of  $dv_{\rm DS}/dt$  in the  $dv_{\rm DS}/dt$  control region by varying the current level of the laser driver.

The test circuit and control block diagram for the high-sidedrive case is shown in Fig. 15. Experimental results for the independent optical control of the  $di_D/dt$  of M1 with the fixed  $dv_{DS}/dt$ , as well as, the independent optical control of the  $dv_{DS}/dt$  of M1 with the fixed  $di_D/dt$  for the case of highside drive are shown respectively in Fig. 16 and Fig. 17. The results are similar to the results of the case of low-side drive, which validates the feasibility of the proposed control for highside drive cases, as long as an isolated voltage for the gate bias of M1 is provided.



Fig. 15. Test circuit and control block diagram for the high-side drive case.



Fig. 16. Measured turn-off waveform of  $v_{\rm DS}$ ,  $i_D$ ,  $I_{\rm laser}$ , and  $I_{\rm OTPT2}$  with varied  $di_D/dt$  (of 16, 11.5, and 9.5 A/ $\mu$ s) and the fixed  $dv_{\rm DS}/dt$  (of 770 V/ $\mu$ s) for the high-side drive case.

To verify the control circuit adaptability in different operating conditions, the  $v_{\text{DS}}$ ,  $i_D$ , and  $I_{\text{Laser}}$  are measured for variation of the bus voltage and load current as depicted in Figs. 18 and 19, respectively. The control circuit naturally initiates the transition from the  $dv_{\text{DS}}/dt$  to the  $di_D/dt$  control region by changing the current levels of the laser driver at the desired points, taking into account the loop delay as discussed earlier. According to (1) and (4), and considering the same load and thermal conditions, the output-current level of the laser driver remains approximately constant in each of the regions of control for different bus voltages to attain the same  $dv_{\text{DS}}/dt$  and  $di_D/dt$ , as shown in Fig. 18. Similarly, to attain the same  $dv_{\text{DS}}/dt$  and  $di_D/dt$  for different load currents,  $I_{\text{Laser}}$  decreases in each region of control as the load current increases, which is depicted in Fig. 19.

The measured turn-off switching losses and energy obtained using a conventional approach to drive the gate and that obtained using the proposed optically switched transition controller are shown in Figs. 20 and 21, respectively. The intensity of the optical beam for OTPT2 is kept constant in the two regions of control for the conventional approach, similar to the proposed approach in [22]. This emulates the discharge of the gate of



Fig. 17. Measured turn-off waveforms of  $v_{\rm DS}$ ,  $i_D$ , and  $I_{\rm laser}$  with varied  $dv_{\rm DS}/dt$  (of 430, 350, and 300 V/µs) and a fixed  $di_D/dt$  (of 20 A/µs) for the high-side drive case.



Fig. 18. Measured turn-off waveform of  $v_{\rm DS}$ ,  $i_D$ , and  $I_{\rm laser}$  for different bus voltages with the fixed  $dv_{\rm DS}/dt$  of 525 V/µs and the fixed  $di_D/dt$  of 20 A/µs.



Fig. 19. Measured turn-off waveforms of  $v_{\rm DS}$ ,  $i_D$ , and  $I_{\rm Laser}$  for different load currents with the fixed  $dv_{\rm DS}/dt$  of 340 V/ $\mu$ s and the fixed  $di_D/dt$  of 20 A/ $\mu$ s.

M1 under condition of fixed gate resistance for  $dv_{\rm DS}/dt$  and  $di_D/dt$  regions of control. In contrast, for the proposed controller, the  $dv_{\rm DS}/dt$  and  $di_D/dt$  are so adjusted such that not only the peak-power loss and switching energy are reduced but the peak overvoltage stress is decreased as well, compared to



Fig. 20. Measured switching loss and energy for conventional approach [24] to drive the gate. In this approach, the intensity of the optical beam for OTPT2 is kept constant in the  $dv_{\rm DS}/dt$  and  $di_D/dt$  regions of control. E = switching energy (200  $\mu$ j/div), P = switching power loss (258 W/div),  $v_{\rm DS}$  is 90 V/div,  $i_D$  is 3 A/div and time division is 800 ns/div.



Fig. 21. Measured switching loss and energy for the proposed opticaltransition controller. The peak-power reduction of 110 W along with the switching-energy reduction of 100  $\mu$ J and 17% reduction of overvoltage stress are achieved using the optical-transition controller as compared to the conventional gate drive. E = switching energy (200  $\mu$ j/div), P = switching power loss (258 W/div),  $v_{\rm DS}$  is 100 V/div,  $i_D$  is 3 A/div and time division is 800 ns/div.

the results obtained using the conventional approach. Using the new controller, the peak-power reduction of 110 W along with the switching-energy reduction of 100  $\mu$ J and 15 V reduction of overvoltage stress are achieved. This is achieved in part by increasing the  $dv_{\rm DS}/dt$ . Furthermore, the  $di_D/dt$  is slightly decreased to reduce the peak voltage stress and minimize the adverse effect of the reduction of  $di_D/dt$  on the switching loss.

In the case of maximum  $di_D/dt$  of 43A/ $\mu$ s in 6 A load current and using the flowchart of Fig. 10, the  $\Delta t$  is selected to be 150 ns and implemented in the control circuit. Therefore, the  $dv_{\text{DSmax}}/dt$  is calculated to be equal to 1030 V/ $\mu$ s in 200 V, based on the explained procedure in Section II-A and using (21). The minimum  $dv_{\text{DSmax}}/dt$  for the case of maximum  $di_D/dt$  is around 200 V/ $\mu$ s, and the minimum  $di_D/dt$  for the case of maximum  $dv_{\text{DSmax}}/dt$  is around 4.7 A/ $\mu$ s, based on Figs. 9, 5, and using (1), (4), and (15).

#### **IV. CONCLUSION**

A novel unified independent dv/dt and di/dt control of an optically triggered (OT) hybrid PSD, which contains a SiC MOSFET as the main PSD and two GaAs-based OTPTs as the gate driver, has been outlined. It has been shown that the unified control of turn-off  $dv_{\rm DS}/dt$  and  $di_D/dt$  is achieved by modulating the optical intensity of the OTPT2 using a single

circuit. Independent control of  $dv_{\rm DS}/dt$  and  $di_D/dt$  is achieved by predicting the onset of transition between the corresponding control regions using the unified control circuit. The control circuit performance has been verified through experimental results over different operating conditions including variation of  $dv_{\rm DS}/dt$ ,  $di_D/dt$ , load current, and bus voltage. It has been shown that, the proposed optical controller is able to attain the desired switching-transition behavior using independent control of  $dv_{\rm DS}/dt$  and  $di_D/dt$ , which is not possible in the conventional gate drivers. Although the new optical controller is designed for an OT PSD, the control concept is also applicable for ET PSDs where the delay of the event feedback-loop is significant compared to the total duration of the switching transition.

#### REFERENCES

- [1] S. Bhattacharya, L. Resta, D. M. Divan, D. W. Novotny, and T. A. Lipo, "Experimental comparison of motor bearing currents with PWM hard and soft switched voltage source inverters," in Proc. IEEE 27th Annu. Power Electron. Spec. Conf. Rec., Jun. 23–27, 1996, vol. 2, pp. 1528–1534.
- [2] D. Leuenberger and J. Biela, "Comparison of a soft switched TCM T-Type inverter to hard switched inverters for a 3 phase PV grid interface," in Proc. Power Electron. Motion Control Conf., Sep. 4–6, 2012, pp. LS1d.1-1–LS1d.1-8.
- [3] A. Rahnamaee, J. Milimonfared, and K. Malekian, "Reliability consideration for a high power single switch flyback power supply," in Proc. IEEE 14th Mediterranean Electrotechnical Conf., May 5–7, 2008, pp. 527–533.
- [4] S. Mehrnami and S. K. Mazumder, "Modulation scheme of the differentialmode Ćuk inverter for loss mitigation," in Proc. IEEE Energy Convers. Congr. Expo., Sep. 15–19, 2013, pp. 3419–3425.
- [5] S. K. Mazumder and S. Mehrnami, "A low-device-count single-stage direct-power-conversion solar microinverter for microgrid," in Proc. IEEE 3rd Int. Symp. Power Electron. Distrib. Generation Syst., Jun. 25–28, 2012, pp. 725–730.
- [6] A. Rahnamaee, J. Milimonfared, and K. Malekian, "Reliability and performance improvement by implementing zero-voltage-switching in high power flyback power supply," *Int. Rev. Electr. Eng.*, vol. 4, no. 5, pp. 699–705, 2009.
- [7] Z. Wang, X. Shi, L. M. Tolbrert, and B. J. Blalock, "Switching performance improvement of IGBT modules using an active gate driver," in Proc. IEEE 28th Annu. Appl. Power Electron. Conf. Expo., Long Beach, CA, USA, Mar. 2013, pp. 1266–1273.
- [8] J. E. Makaran, "Gate charge control for MOSFET turn-off in PWM motor drives through empirical means," *IEEE Trans. Power Electron.*, vol. 25, no. 5, pp. 1339–1350, May 2010.
- [9] C. Licitra, S. Musumeci, A. Raciti, A. U. Galluzzo, R. Letor, and M. Melito, "A new driving circuit for IGBT devices," *IEEE Trans. Power Electron.*, vol. 10, no. 3, pp. 373–378, May 1995.
- [10] D. Aggeler, F. Canales, J. Biela, and J. W. Kolar, "Dv /Dt-control methods for the SiC JFET/Si MOSFET cascode," *IEEE Trans. Power Electron.*, vol. 28, no. 8, pp. 4074–4082, Aug. 2013.
- [11] V. John, B. S. Suh, and T. A. Lipo, "High performance active gate drive for high power IGBTs," *IEEE Trans. Ind. Appl.*, vol. 35, no. 5, pp. 1108–1117, Sep./Oct. 1999.
- [12] N. Idir, R. Bausiere, and J. J. Franchaud, "Active gate voltage control of turn-on di/dt and turn-off dv/dt in insulated gate transistors," *IEEE Trans. Power Electron.*, vol. 21, no. 4, pp. 849–855, Jul. 2006.
- [13] B. Wittig and F. Fuchs, "Analysis and comparison of turn-off active gate control methods for low-voltage power MOSFETs with high current ratings," *IEEE Tran. Power Electron.*, vol. 27, no. 3, pp. 1632–1640, Mar. 2012.
- [14] M. Rose, J. Krupar, and H. Hauswald, "Adaptive dv/dt and di/dt control for isolated gate power devices," in Proc. IEEE Energy Convers. Congr. Expo., Sep. 2010, pp. 927–934.
- [15] L. Michel, X. Boucher, A. Chériti, P. Sicard, and F. Sirois, "FPGA implementation of an optimal IGBT gate driver based on posicast control," *IEEE Trans. Power Electron.*, vol. 28, no. 5, pp. 2569–2575, May 2013.
- [16] S. Park and T. M. Jahns, "Flexible dv/dt and di/dt control method for insulated gate power switches," *IEEE Trans. Ind. Appl.*, vol. 39, no. 3, pp. 657–664, May/Jun. 2003.

- [17] Y. Lobsiger and J. W. Kolar, "Closed-loop IGBT gate drive featuring highly dynamic di/dt and dv/dt control," in Proc. IEEE Energy Convers. Congr. Expo., Sep. 2012, pp. 4754–4761.
- [18] H. Riazmontazer and S. K. Mazumder, "Dynamic optical turn-off control of a high-voltage SiC MOSFET," in Proc. IEEE 28th Annu. Appl. Power Electron. Conf. Expo., Long Beach, CA, USA, Mar. 2013, pp. 1266–1273.
- [19] H. Riazmontazer and S. K. Mazumder, "Self-contained control for turnon transition of an optically driven IGBT," in Proc. IEEE 29th Annu. Appl. Power Electron. Conf. Expo., Fort worth, TX, USA, Mar. 2014, pp. 1465–1470.
- [20] S. K. Mazumder, A. Mojab, and H. Riazmontazer, "Optically-switched wide-bandgap power semiconductor devices and device-transition control," in *Physics of Semiconductor Devices*, New York: Springer International Publishing, 2014, pp. 57–65.
- [21] R. Vafaei, N. Rouger, D. N. To, and J. Crebier, "Experimental investigation of an integrated optical interface for power MOSFET drivers," *IEEE Electron Device Lett.*, vol. 33, no. 2, pp. 230–232, Feb. 2012.
- [22] F. Zhao and M. M Islam, "Optically activated SiC power transistors for pulsed-power application," *IEEE Electron Device Lett.*, vol. 31, no. 10, pp. 1146–1148, Oct. 2010.
- [23] T. Sarkar and S. K. Mazumder, "Epitaxial design of a direct optically controlled GaAs/AlGaAs-based heterostructure lateral superjunction power device for fast repetitive switching," *IEEE Trans. Electron Devices*, vol. 54, no. 3, pp. 589–600, Mar. 2007.
- [24] S. K. Mazumder and T. Sarkar, "Optically-activated gate control for power electronics," *IEEE Trans. Power Electron.*, vol. 26, no. 10, pp. 2863–2886, 2011.
- [25] F. Zhao, M. M. Islam, P. Muzykov, A. Bolotnikov, and T. S. Sudarshan, "Optically activated 4H-SiC p-i-n diodes for high-power applications," *IEEE Electron Device Letters*, vol. 30, no. 11, pp. 1182–1184, Nov. 2009.
- [26] A. Meyer, S. K. Mazumder, and H. Riazmontazer, "Optical control of 1200-V and 20-A SiC MOSFET," in Proc. IEEE 27th Annu. Appl. Power Electron. Conf. Expo., Orlando, FL, USA, Feb. 2012, pp. 2530–2533.
- [27] A. Meyer, A. Mojab, and S. K. Mazumder, "Evaluation of first 10-kV optical ETO thyristor operating without any low-voltage control bias," in Proc. IEEE Int. Symp. Power Electron. Distrib. Generation Syst., 2013, pp. 1–5.
- [28] S. Potbhare, "Modeling and characterization of 4H-SiC MOSFETs: High fields, high temperature and transient effects," Ph.D. dissertation, Dept. Elect. and Comput. Eng., Univ. of Maryland, College Park, MD, USA, 2008.
- [29] N. Phankong, T. Yanagi, and T. Hikihara, "Evaluation of inherent elements in a SiC power MOSFET by its equivalent circuit," in Proc. 14th Eur. Conf. Power Electron. Appl., 2011, pp. 1–8.
- [30] T. Sarkar and S. K. Mazumder, "Dynamic power density, wavelength, and switching time modulation of optically-triggered power transistor (OTPT) performance parameters," *Microelectron. J.*, vol. 38, pp. 285–298, 2007.
- [31] T. Sarkar, "Optical intensity modulated gate control of power-electronic system performance parameters," Ph.D. dissertation, Dept. Elect. and Comput. Eng., Univ. of Illinois at Chicago, Chicago, IL, USA, 2009.
- [32] P. J. Grbovic, "An IGBT gate driver for feed-forward control of turnon losses and reverse recovery current," *IEEE Trans. Power Electron.*, vol. 23, no. 2, pp. 643–652, Mar. 2008.



Hossein Riazmontazer (S'12) was born in Shiraz, Iran, in 1985. He received the B.S. degree in electrical engineering majoring in power from the Iran University of Science and Technology, Tehran, Iran, in 2008 and the M.S. degree in electrical engineering majoring in power electronics from the Amirkabir University of Technology (Tehran Poly-technique), Tehran, Iran, in 2011. He is currently working toward the Ph.D. degree at the University of Illinois at Chicago (UIC), Chicago, IL, USA.

Prior to join the UIC, he was working in industry as a Electrical Design Engineer for about 5 years. Since 2011, he has been a Research Assistant at the Laboratory of Energy and Switching Electronics Systems at the University of Illinois at Chicago. His research interests include but not limited to: gate drive circuits for power semiconductor devices, design, modeling and control of power electronics converters for vehicles, renewable energy systems, motor drives and smart grids.

Mr. Riazmontazer is a Reviewer for the IEEE TRANSACTIONS ON POWERELECTRONICS.



Sudip K. Mazumder (S'97–M'01–SM' 03) received the Ph.D. degree from the Department of Electrical and Computer Engineering of Virginia Tech, VA, USA, in 2001.

He is the Director of the Laboratory for Energy and Switching-Electronics Systems and a Professor in the Department of Electrical and Computer Engineering at the University of Illinois, Chicago (UIC). He has more than 22 years of professional experience and has held R&D and design positions in leading industrial organizations and has served as

a Technical Consultant for several industries. He also serves as the President of NextWatt LLC, a small business organization that he setup in 2008. His current areas of interests include 1) interactive power-electronics/power networks, smart grid, and energy storage; 2) renewable and alternative energy-based power electronics systems for distributed generation and microgrid; and 3) Optically triggered wide-bandgap power-electronics device and control technologies and SiC and GaN device-based high-frequency, high-temperature, and high-voltage power electronics.

Since joining the UIC in 2001, Dr. Mazumder has been awarded about 40 sponsored projects by NSF, DOE, ONR, ARPA-E, CEC, EPA, AFRL, NASA, NAVSEA, and multiple leading industries in aforereferenced areas. He has published more than 170 refereed papers in prestigious journals and conferences and has published one book and eight book chapters. About 50% of his journal papers are published in IEEE transactions with a current impact factor above 4. He has presented 53 invited/plenary/keynote presentations at leading organizations and institutions and currently, he also holds eight issued and two pending patents. He received in 2014 the Inventor of the Year Award at the UIC, in 2013 the University Scholar Award from the University of Illinois, and in 2011, the Teaching Recognition Program Award at the UIC. In 2008 and 2006, he received the prestigious Faculty Research Award from the UIC for outstanding research performance and excellent scholarly activities. He also received the ONR Young Investigator Award and the NSF CAREER Awards in 2005 and 2003, respectively, and IEEE Prize Paper Awards in 2002, 2007, and 2013, respectively, and the IEEE International Future Energy Challenge Award in 2005. He served as the first Editor-in-Chief for International Journal of Power Management Electronics (currently known as Advances in Power Electronics) between 2006 and 2009. He served as the Guest Editor-in-Chief for the IEEE TRANSACTIONS ON POWER ELECTRONICS SPECIAL ISSUE ON HIGH-FREQUENCY-LINK POWER-CONVERSION SYSTEMS (2013-2014). He also serves as a Guest Associate Editor for the IEEE JOURNAL OF EMERGING AND SELECTED TOPICS IN POWER ELECTRONICS SPECIAL ISSUE ON CLEAN ENERGY SYSTEMS INTEGRATION. Currently, he also serves as an Associate Editor for the IEEE TRANSACTIONS ON INDUSTRIAL INFORMATICS (since 2014), the IEEE TRANSACTIONS ON POWER ELECTRONICS (since 2009), the IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS (since 2003), the IEEE TRANSACTIONS ON AEROSPACE AND ELECTRONICS SYSTEMS (SINCE 2008). He is also an Editorial Board Member for Advances in Power

Electronics since 2009. Previously, he has also served as an Associate Editor for the IEEE TRANSACTIONS ON CIRCUITS AND SYSTEMS AND IEEE POWER ELECTRONICS LETTER. He has also served as the Guest Coeditor for the following Transaction Special Issues: IEEE TRANSACTIONS ON POWER ELECTRONICS SPECIAL ISSUE ON POWER ELECTRONICS IN DC DISTRIBUTION SYSTEMS (2011-2013) and Advances in Power Electronics Special Issue on Advances in POWER ELECTRONICS FOR RENEWABLE ENERGY (2010–2011). He is currently serving as a Plenary Chair for the 2015 IEEE Energy Conversion Congress and Exposition. He is also serving as the Technical Awards Committee Chair for IEEE Power Electronics Society (PELS) Technical Committee on Sustainable Energy Systems (SES). He is also serving as a Chair for IEEE PELS Subcommittee on Distributed Generation and Renewable Energy and for IEEE PELS Subcommittee on Power Semiconductors, and Member for IEEE PELS Standards Subcommittee and IEEE IES Network Control Systems and IEEE IES Power Electronics Technical Committees. In 2010, he served as the Chair, Student/Industry Coordination Activities for IEEE Energy Conversion Congress and Exposition, which is the largest conference in power electronics today in North America. He serves/served as a Steering Committee Member for 2014 Power Electronics and Application Conference and Exposition and as an Advisory Committee Member for 2012 IEEE India International Conference on Power Electronics as well as for 2010 IEEE International Symposium on Power Electronics for Distributed Generation Systems. He also served as the Cochair for SES. In addition, he is serving/has served as Technical Program Committee Member for numerous IEEE sponsored and other reputed conferences including IEEE Energy Conversion Congress and Exposition, IEEE Applied Power Electronics Conference and Exposition, IEEE Industrial Electronics Conference, IEEE International Symposium on Power Electronics for Distributed Generation Systems. He was one of the five leading researchers invited by the inaugural 2012 Clean Energy Trust Show Case, an event that connects entrepreneurs, investors, and researchers who can work together to commercialize the latest clean technology, to deliver his vision on Smart Grid. Between 2010 and 2011, he also served as an Advisory Council Member for Vice Chancellor for Research's Urban Resilience and Global Environment at the UIC. In 2009 and 2010, he also served as the Expert Representative on Smart Grid for the UIC at the Midwestern Great Lakes Alliance for Sustainable Energy Research (GLASER) initiative. In 2008, he was invited by DOE to participate along with several leading industries and selected academic professionals regarding High MW Power Converter for next generation power grid. In 2008, he was invited by NSF to participate in a unique workshop (comprising leading industries and research experts) leading to decision on nation's specific R&D focus on energy and energy distribution over the next 10 and 50 years. He has also been invited to serve as the Working Group Committee Member for IEEE P1676, which focuses on Control Architecture for High Power Electronics (>1 MW) used in Electric Power Transmission and Distribution Systems. In 2009, he was also part of the team that wrote the NSF and National Coordination Office for Networking and Information Report on Research Directions for Future Cyber-Physical Energy Systems. He has also served as a panel reviewer and reviewer for NSF, DOE, ARPA-E, CRDF, and AAAS.