A SEGMENTED OUTPUT STAGE H-BRIDGE IC WITH TUNABLE GATE
DRIVER

by

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Graduate Department of Electrical and Computer Engineering
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Abstract

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In this thesis, an integrated H-bridge is presented for continuously optimizing the power conversion efficiency over a wide range of output current. The proposed IC incorporates both segmented output transistors and segmented gate drivers. This fully segmentation approach allows to flexibly change the on-resistance of the output stage power transistors and the output resistance of the gate drivers. Dynamic adjustment of these parameters allows to investigate the best combination of output power transistor size and gate driver size for any given load current. The IC chip is fabricated using TSMC's 0.18 μm BCD Gen-2 process technology. The presented design when operating as a buck converter with a load current between 0.02 A and 4 A achieves efficiency improvements of 32% and 8% at light and heavy load current, respectively. Furthermore, the dynamically adjustable gate driver output resistance allows for suppression of the peak CEMI by 5.5 dBm.

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Glossary

CEMI Conducted electromagnetic interference

EDMOS Extended drain MOSFET

EMI Electromagnetic interference

LCC Ceramic leadless chip carrier

PFM Pulse frequency modulation

PWM Pulse width modulation

SMPS Switched-mode power supplies

SOA Safe operation area

Chapter 1

Introduction

A DC-DC converter, as shown in Fig. 1.1, is used to convert the one DC voltage to another DC voltage. Linear converter and switched-mode converter are two general methods to achieve the voltage conversion. The power conversion efficiency, defined as the ratio of the output power and the input power, is a good measure to evaluate the converter's performance [1]. Compared to the linear power supply, the switched-mode power supply (SMPS) is more efficient by using the passive components to store the energy.

Figure 2.1 shows a synchronous buck converter. By operating the power switches M1 and M2 alternately during a switching period, a voltage lower than the input voltage is obtained at the output node. The inductor L is used to continuously provide the load current, and the capacitor C_{out} is used to filter out the ripples in the output voltage.

Figure 1.1: Block diagram for DC-DC power conversion [1].

Figure 1.2: Schematic of a synchronous buck converter [1].

1.1 H-bridge Output Stage

The H-bridge topology is a highly versatile switching output stage for numerous modern applications such as Class-D audio amplifiers, motor drivers and DC-DC power conversions.

An H-bridge includes two identical output stages with four power switches, i.e. MOSFETs, IGBTs, etc.. Figure 1.3 shows a simple diagram of an H-bridge, with four n-type MOSFETs (M1, M2, M3, M4). If M1 and M4 are *on*, the other two power switches M2 and M3 are *off*, the load current flows from A to B in the direction. On the other hand, if M2 and M3 are *on*, M1 and M4 are *off*, the load is energized in the reverse direction. Therefore, H-bridge's operation is bidirectional. Table 1.1 summaries the four possible states that an H-bridge can operate. In State 3 and State 4, since of the power switches on the same side are *on* simultaneously, the power supply is shorted to ground and shoot-through happens. This condition is not desirable as since it can ruin the bridge and even the power supply [2].

A half-bridge is more simple, which only includes one side of the output stages with two power switches. Unlike an H-bridge discussed above, the direction of the current flowing through the load can only be in one direction.

As with any switched mode power supply, the overall power efficiency of an H-bridge is dependent on the design and the operation of the output stage and its gate driver. In

Table 1.1: H-bridge Operating States [2].

State	M1	M2	M3	M4	Description
1	<i>on</i>	<i>off</i>	<i>off</i>	<i>on</i>	I_L direction: positive
2	<i>off</i>	<i>on</i>	<i>on</i>	<i>off</i>	I_L direction: reverse
3	<i>on</i>	<i>on</i>	<i>off</i>	<i>off</i>	Shoot-through on the left side
4	<i>off</i>	<i>off</i>	<i>on</i>	<i>on</i>	Shoot-through on the right side

Figure 1.3: An H-bridge consists of four n -type MOSFETs [2].

addition, the nature of high frequency switching in SMPS could lead to the occurrence of electromagnetic interference (EMI). This may cause catastrophic issues in the operation of the SMPS and neighboring circuits. EMI mitigation with an efficient and economical method remains a challenge for IC designers.

1.2 Power Losses in SMPS

It is well known that conduction loss, switching loss and gate driver loss are three of the primary sources of power loss in SMPSs [1]. At light load conditions, the gate driver loss is dominant [3]. By decreasing the size of the output stage transistor, the charging and discharging of the gate capacitance for the power transistor can be reduced. This is

subsequently leads to lower gate driver loss [4]. In addition, a small gate capacitance of the power transistor requires less driving effort from the gate driver, which reduces the unnecessary charging and discharging of the internal capacitive nodes. However, a small size power transistor implies a large on-resistance (R_{on}). When the SMPS is operating at high output current, the increased conduction loss degrades the power conversion efficiency [3]. In this case, a large gate driver is preferred since fast turn on and turn off times (t_{on} and t_{off} , respectively) for the power transistor minimizes the switching loss and the shoot-through current. The observations above indicate that the overall power conversion efficiency of a traditional H-bridge can only be optimized for a particular operating condition.

1.3 EMI in SMPS

Periodically high frequency toggling of voltages and currents in SMPSs can cause EMI. This problem becomes a more serious concern in modern electronics when higher switching frequency (e.g. in 10's MHz range) is used to reduce component sizes and improve the transient performance. Therefore, minimizing the EMI to meet certain electromagnetic compatibility (EMC) standards is required before a SMPS product can be launched into the market. Widely accepted EMC regulations include Federal Communications Commission (FCC), British Standards (BS), Comite International Sepcial des Perturbations Radioelectrique (CISPR) and Verband Deutscher Elektrotechnik (VDE) [5].

The generation, propagation and common mitigation approaches of EMI are discussed in [6] and [5]. Two forms of EMI, including the conducted EMI (CEMI) and the radiated EMI, are generated by DC-DC converters. The CEMI noise only transmits along the current conducting power line, while the radiated EMI causes interference with the nearby electronic devices and systems. Two types of common CEMI mitigation approaches are categoried as follows, post-generation reduction and pre-generation elimination. The first method usually places a snubber circuit at the ringing node to filter the high-frequency noise. This approach is simple but not efficient due to the extra space, power dissipation and the cost for the passive components. As a result, the pre-generation elimination approach is preferred since it eliminates the excessive CEMI before it is generated [6].

1.4 Thesis Objectives and Organization

The objective of this thesis is to develop an integrated power output stage that can continuously optimize the power conversion efficiency of a typical DC-DC converter over

a wide range of output current. The work also investigates the approach to suppress ringing oscillation at the switching node and CEMI without significant power efficiency degradation.

These two goals are demonstrated using an integrated H-bridge with segmented techniques utilized for both the power output transistors and the gate drivers. For any given load current, the best combination of the power output transistor size and the gate driver size can be adjusted on-the-fly. Furthermore, dynamic driving strategy similar to that of [7] is employed in order to reduce the CEMI for the H-bridge IC.

This thesis is organized as follows: Chapter 2 provides a background review on basic SMPS operations, EMI types and segmentation techniques; Chapter 3 presents the proposed H-bridge IC design and the dynamic operation strategy; Chapter 4 describes the IC implementation, the experimental setup, measurement approaches and experimental results for power efficiency optimization and CEMI reduction; Conclusions and future work are discussed in Chapter 5.

Chapter 2

Background Research

This chapter reviews the power losses, EMI suppression and power conversion efficiency enhancement techniques.

2.1 SMPS and Sychoronus DC-DC Buck Converter

Compared to linear power supplies, SMPS can provide much higher power conversion efficiency [8]. Common types of SMPS include, buck, boost, buck-boost, Ćuk, and transformer-isolated such as fly-back converter [1]. In this thesis, the focus is to examine the switches' behavior during the transition periods and investigate the influence on the overall conversion performance in practical applications. The buck converter is selected for its simple operating mechanism.

A typical synchronous DC-DC buck converter is as shown in Fig. 2.1. The high-side (M1) and the low-side (M2) power transistors turn on and off alternately. Figure 2.2 illustrates the voltage and current waveforms for a basic operation condition. The conversion ratio $M(D)$ for the buck converter when operating in the continuous conduction mode (*CCM*), can be expressed by

$$M(D) = \frac{V_{out}}{V_{DD}} = D \quad (2.1)$$

The ripples on the inductor current and the output voltage are given by (2.2) and (2.3), respectively. The expressions indicate that for a specific ripple tolerance, high switching frequency can reduce the passive components size, which implies small circuit board space for a converter.

Figure 2.1: A synchronous buck converter consists of two n -type MOSFETs [1].

$$\Delta i_L = \frac{(V_{DD} - V_{out}) \cdot D}{2Lf_s} \quad (2.2)$$

$$\Delta v = \frac{\Delta i}{8Cf_s} \quad (2.3)$$

2.2 Different Power Losses in a Buck Converter

The efficiency of a synchronous buck converter can be expressed as

$$\eta = \frac{P_{out}}{P_{in}} \quad (2.4)$$

For an ideal buck converter (as shown in Fig. 2.1), the theoretical power conversion efficiency can be 100%. This means that the input power is totally transmitted to the load without any loss. However, in a real circuit, the efficiency is always degraded by power losses P_{loss} .

There are three main sources of P_{loss} in a synchronous buck converter: the conduction loss P_{cond} , the switching loss P_{sw} and the gate driver loss P_{drv} [9]. Other fixed power losses P_{fix} , such as the power consumed by the controller, are required to bias the circuit and independent of the switching frequency. P_{fix} is often neglected at heavy load or high

Figure 2.2: Voltage and current waveforms in buck-mode converter operation [1].

switching frequency situation due to its small value when compared to the other three.

2.2.1 Conduction Loss

According to Section 2.1, the high-side and low-side transistors carry the conducting inductor current alternately in a switching period. In a buck converter, when a transistor is *on*, it can be viewed as a resistor with effective on-resistance of R_{on} . The conduction loss due to this on-resistance can be calculated as

$$P_{cond} = I_L^2 \cdot R_{on_{hs}} \cdot D + I_L^2 \cdot R_{on_{ls}} \cdot (1 - D) \quad (2.5)$$

Assuming that the high-side and low-side transistors are identical, (2.5) can be simplified as follows

$$P_{cond} = I_L^2 \cdot R_{on} \quad (2.6)$$

$$R_{on} = R_{sh} \cdot \frac{L}{W} \quad (2.7)$$

where R_{sh} is the sheet resistance of the channel. L is the gate length and W is the width of the transistor. Since R_{sh} is determined by the fabrication technology and the minimum gate length is usually used during IC design, these two parameters can be considered as fixed features [9]. Therefore, the transistor width W is the main variable for the on-resistance R_{on} . Equations (2.6) and (2.7) show that the conduction loss P_{cond} and the on-resistance R_{on} have an inversely proportional relationship to the transistor width W . Therefore, a large W for the power transistor implies a smaller conduction power loss.

2.2.2 Switching Loss

Switching loss is caused by the loss in the power transistor during the transition period. The transistor cannot be switched between *on* and *off* states instantaneously. It requires a certain time to complete the switching action. As shown in Fig. 2.3, during the turn-on transition, both current and voltage values are non-zero. The overlap results in a transistor switching loss P_{sw} [1]. During the turn-off transition, the power loss is similar to the waveforms as shown in Fig. 2.3. Therefore, the total switching loss is given by

$$P_{sw} = f_{sw} \cdot \left(\int_{turn-on} (v_{ds}(t) \cdot i_{ds}(t)) + \int_{turn-off} (v_{ds}(t) \cdot i_{ds}(t)) \right) \quad (2.8)$$

Assuming that the voltage and current waveforms are linear, the switching loss can

Figure 2.3: Waveforms during the turn-on transition of a power transistor [1].

be calculated by [1]

$$P_{sw} = f_{sw} \cdot V_{ds} \cdot I_L \cdot \left(\frac{t_{on} + t_{off}}{2} \right) \quad (2.9)$$

2.2.3 Gate Driver Loss

Gate driving loss P_{drv} due to the charging and discharging of the capacitance internal to the gate driver is another important power loss.

$$P_{drv} = f_{sw} \cdot V_G^2 \cdot (C_{gs.hs} + C_{gs.ls}) \quad (2.10)$$

$$C_{gs} = C_{gs.sp} \cdot W \cdot L \quad (2.11)$$

where $C_{gs.hs}$ and $C_{gs.ls}$ refer to the input gate capacitance of high-side and low-side transistors, respectively. Since $C_{gs.sp}$ is another technology specific parameter which cannot be altered by the IC designers [9], C_{gs} is proportional to the transistor width W . Therefore, a power transistor of minimum usable size is preferred so that the gate driver loss can be reduced.

Recapping the relationship between R_{on} and W , a trade-off exists between R_{on} and C_{gs} . For a specific fabrication, a figure of merit (FOM) defined as the product of on-resistance R_{on} and gate charge Q_g can be used to find an optimized transistor width, so that the total power loss can be minimized.

2.3 Impact of Packaging on IC Performance

The bare unpackaged silicon die is very vulnerable to the environment. High temperature and humidity from the outside environment can lead to oxidization and malfunction of the chip. Therefore, packaging is an essential part of the IC design. The IC package not only acts like a protector, but also as a carrier for the chip. A good packaging should have characteristics of small size, low cost, mechanical reliability, and high electrical performance [10].

2.3.1 IC Package Type and Trends

The IC package type can be classified in different ways, e.g. by material, by mounting method, by lead type, etc.. Table 2.1 is a summary of different IC package types [10], [11].

Driven by the increasing demand of high number of I/O pins, high power density, high reliability and performance requirements, the IC packaging technology has enjoyed

Table 2.1: Summary of IC Packages

	Description	Features	Examples
Materials			
Plastic	plastic package	cheap	PLCC
Ceramic	ceramic package	good heat dissipation	PGA LCC
Mounting Method			
Through Hole	drill hole needed	Cheap;Easy to assembly	DIP ZIP
Surface Mount	one side attched only	Space saving;	SOJ QFN BGA
Lead Type			
Leaded	Gull-wing; J-Lead	Easy to solder	QFP SOJ
Leadless	Lead-free	Higher I/O counts	LCC QFN
Pin Arrangement			
Small Outline	Dual line pin	widely used	SOIC TSOP
Quad Flat Pack	Pins on four sides	Higher I/0 counts	QFN QFP LCC
Grid Array	Flip-chip Assembly	Compact size	BGA

significant advancement. The history of the IC packaging can be traced back to Dual In-Line Package (DIP) in 1970's. Through-hole mounting is a classical approach, which is inexpensive and easy to implement. With higher level of integration density, this kind of packaging is gradually replaced by the surface mount technology (SMT). SMT has the benefits of lower parasitics, higher reliability and space saving for complex PCB design. Figure 2.4 shows that miniaturization and higher pin-count is the trend for IC packaging development [10]. Table 2.2 provides the definition for package type abbreviations mentioned in this section.

2.3.2 Package Parasitics

For integrated chip testing on printed circuit board (PCB), there are two levels of connection between the silicon die and board: chip to package and package to PCB. Parasitic capacitance, inductance and resistance are introduced to the system from wire-bonding and PCB traces. The influence from these parasitics is becoming a serious issue for integrated output stage with power transistors of small R_{on} [13]. In addition, these parasitic elements can affect the overall electrical performance of the chip, by unwanted oscillations at various circuit nodes [14]. Voltage and current ringings during the switching transition may result in power device operating out of safe operation area (SOA), and

Figure 2.4: IC Package Development [10]

Table 2.2: Abbreviations for Different IC Packages [12].

Abbreviation	Package Type Description
BGA	Ball Grid Array
LCC	Ceramic Leadless Chip Carrier
CSP	Chip Scale Package
DIP	Dual-in-Line Package
PGA	Pin Grid Array
PLCC	Plastic Leaded Chip Carrier
QFJ	Quad Flat J-Leaded Package
QFN	Quad Flat No-Leaded
QFP	Quad Flat Pack
SOIC	Small Outline IC Package
SOJ	Small Outline J-Leaded Package
SON	Small Outline Non-Leaded Package
SOP	Small Outline Package
SSOP	Shrink Plastic Dual-in-Line Package
TSOP	Thin Small Outline Package
ZIP	Zigzag-in-Line Package

Figure 2.5: A simulation setup for parasitic inductances investigation.

subsequently reduce the lifetime of the device. In addition, the oscillations in the current waveform produce conducted EMI, and degrades the EMC of the products. The impact of ringing oscillation will be discussed in Section 2.4

2.3.3 Impact of Parasitics on Device Reliability

In order to acquire a better understanding on the impact of the parasitic components, the parasitic inductances from the DIP package are estimated and added to the model of a DC-DC buck converter in simulation. Figure 2.5 shows the simulation setup: the parasitic inductances L_{vdd} , L_{gnd} and L_{sw} are estimated based on an output stage IC mounted in a typical DIP package. Assuming that the output node of the gate driver is directly connected to the gate terminal of the power transistor, parasitic inductances from this path is neglected. The output stage consists of two identical n -MOS transistors, M1 and M2. The drain-source voltage V_{ds} , the drain current I_{ds} , and the switching node (SN) voltage V_{sw} are recorded during the switching period of each device in the simulation. The impacts of three variables were individually investigated by increasing the value of L_{vdd} , L_{gnd} and L_{sw} .

Figure 2.6 plots the transient waveforms for the high-side transistor when it turns on. The operating area, which focuses on the locus of the current and voltage transition

Figure 2.6: Transient waveforms at the rising edge of V_{sw} .

behavior of the power transistor is plotted on the right in the same graph. Ringings are observed at the beginning of the transition. In order to investigate the influence of the parasitic inductance, the value of each inductance is varied and re-simulated. The operating area for the high-side transistor is summarized in Fig. 2.7. With a doubling in the size of L_{vdd} , more V_{ds} ringing is produced, but I_{ds} oscillation is reduced at the same time. With a doubling in the size of L_{gnd} , more V_{ds} ringing is produced at the beginning of the transition period, but becomes less at the end of the transition. The value of L_{sw} does not cause a big difference since it can be ignored when compared to the large output inductor in the buck converter.

Similar plots for the low-side transistor at the rising edge of V_{sw} are as shown in Fig. 2.8. Figures 2.9 and 2.10 show the transition behavior for the high-side transistor and the low-side transistor at the falling edge of V_{sw} , respectively. Simulation results show that L_{vdd} leads to ringing oscillation at the switching node; L_{gnd} leads to ringing oscillation during pull up; L_{sw} , however, has little effects on the transistor operation. Excessive ringing oscillation at the switching node degrades the performance of the circuit and possibly damage the transistor if the I - V curve goes outside the safe operation area (SOA) for some time during switching. Current oscillation, on the other hand, can results in EMI issues. Therefore, an appropriate package is a vital factor that guarantees the performance of the chip.

Figure 2.7: Influence of the parasitic inductances on the high-side transistor at the rising edge of V_{sw} .

2.4 EMI in SMPS

SMPS integrated circuits generate significant amount of electromagnetic noise due to its high frequency switching. Negative impacts including performance degradation, or even system malfunction become a big concern for the IC designers [15]. Therefore, understanding the EMI generating mechanism in SMPS and the corresponding mitigation techniques can help IC designers to reduce the effect of the disturbance.

2.4.1 EMI Types and Effects on SMPS

Generally, electromagnetic interference (EMI) in SMPS can be classified into two types: conducted EMI and radiated EMI [16].

The conducted EMI is produced and propagated along the current conducting path. In a SMPS system, voltage at the switching node and input current exhibit the fast switching characteristics, which bring about unwanted harmonics and high frequency noise. Conducted EMI noise generated from high frequency switching is then coupled to the other circuits connected on the same supply line. Radiated EMI is transmitted into all directions in the form of electromagnetic waves. It disturbs the nearby devices which

Figure 2.8: Influence of the parasitic inductances on the low-side transistor at the rising edge of V_{sw} .

are exposed to this electromagnetic field. However, an effective antenna is not easy to be defined or recognized. Therefore, a bond wire inside the IC package or a PCB trace can be a source of radiated EMI [17].

2.4.2 Sources of EMI in Buck Converter

EMI in DC-DC buck converters mainly results from the fast switching of voltages and currents. As shown in Fig. 2.11, the pulse-width-modulation (PWM) signal at the gate node of the power transistor, the voltage ringing at the switching node and the input current oscillation on the input power line are square waveforms with high switching frequency. These three are the main sources of EMI in the buck converter circuit.

PWM is the most common approach to drive the output stage in the buck converter. However, due to the input capacitance C_{gs} of the power device, the parasitic inductance from the PCB layout and the output resistance R_{out} of the gate driver, the RLC circuit generates the voltage ringing at the gate node of the power transistor and subsequently at the switching node in the buck converter. As shown in Fig. 2.11, if the input current is periodically flowing into the buck converter with a duty ratio of D and output current of I , the Fourier transformation of the input current shows EMI at frequencies well the

Figure 2.9: Influence of the parasitic inductances on the high-side transistor at the falling edge of V_{sw} .

PWM frequency and its harmonics [18].

$$f(t) = DI + \sum_{i=1}^{\infty} \left(\frac{2I}{k\pi} \sin(Dk\pi) \cos(k\omega t) \right) \quad (2.12)$$

2.5 Output Stage in SMPS

An output stage normally contains two power switches [19]. Two common output stage configurations and the output stage IC design are described in this section.

2.5.1 Output Stage Configurations

Fig. 2.12 illustrates the two most widely used output stage configurations: totem-pole output stage and push-pull complementary output stage. The totem-pole output stage consists of two stacked n -type MOSFET (n -MOS), while the push-pull complementary output stage consists of one p -type MOSFET (p -MOS) and one n -MO [19].

For both configurations, the low-side transistor is n -MOS with source and body connected to the ground, drain connected to the switching V_X node. The choice between

Figure 2.10: Influence of the parasitic inductances on the low-side transistor at the falling edge of V_{sw} .

Figure 2.11: EMI sources in a typical DC-DC buck converter.

Figure 2.12: Common output stage configurations: (a) totem pole output stage, (b) push-pull output stage.

these two output stage configurations is mainly dependent on the target output power. For a given R_{on} for the power transistor, the p -MOS power transistor occupies a larger area than an n -MOS due to its lower hole carrier mobility. Considering the increasing requirement of high power level and small device area, the totem-pole configuration with two n -MOS power transistors is more popular in integrated SMPS designs. However, the totem-pole output configuration has an intrinsic drawback: the floating source of the high-side transistor. As shown in Fig. 2.12, since the source of the high-side n -MOS is tied to the output node V_X , the source voltage is floating and toggling between the ground and V_{DD} at the switching frequency. In order to turn on the high-side n -MOS, the high-side gate driver needs to provide a driving voltage higher than the supply voltage. Therefore, a bootstrap circuit (as shown in Fig. 2.13) is required to bias the high-side gate driver circuit.

2.5.2 Integrated Output Stage Design

There are many parameters need to be considered for an output stage IC design. The key factors include the output stage configuration, the on-resistance R_{on} , the input capacitance C_{gs} , the switching speed f_s , the device area, etc..

Normally, for low output power applications, the push-pull complementary output stage is more appropriate due to its simplicity. The low hole carrier mobility issue does not matter much when duty ratio D is low since the p -MOS on the high-side is

Figure 2.13: A general bootstrap circuit.

conducting current for a very short time. On the contrary, for high voltage and heavy load applications, the totem-pole configuration is preferred due to its smaller area for a given R_{on} requirement [19].

On-resistance R_{on} , input capacitance C_{gs} , device area and other related parameters mentioned above must be taken into account when choosing the size of the power transistors. Since the sheet resistance R_{sh} is determined by the technology process, according to (2.7), the on-resistance of an n -MOS is inversely proportional to the width W of the transistor provided minimum gate length L is used. However, (2.14) indicates a large transistor width would lead to large C_{gs} , which means a long charging or discharging time for a specific driving strength. So, a small on-resistance R_{on} and a low input capacitance C_{gs} cannot be achieved at the same time. Power loss calculations based on the analytical model as discussed in Section 2.2, and circuit simulation can help to find an optimized transistor width to minimize the power loss for a specific load current.

$$R_{ds,on} = \left(\frac{\rho}{t}\right)\left(\frac{L}{W}\right) = R_s \frac{L}{W} \quad (2.13)$$

$$C_{gs} = WL\left(\frac{\epsilon_{ox}}{t_{ox}}\right) = WLC_{ox} \quad (2.14)$$

The target switching speed f_s is another important design specification. It is dependent on the input capacitance C_{gs} and the gate driver driving ability. This part will be further discussed in Section 2.6.

2.6 Gate Driver for the Power Transistor

Gate driver design is a critical part for integrated power transistors. In order to turn the power transistors on and off, the gate driver must supply enough current to charge and discharge the gate capacitance C_{gs} . As shown in Fig. 2.14, the gate driver topology can be categorized into three groups: voltage driven gate driver, current driven gate driver, and resonance driven gate driver [20].

Among these three types of gate driver topologies, the current driven gate driver is seldom used due to a high efficiency current source requirement [20]. The resonant driven topology is popular for high frequency applications. It has the advantage to recover the power loss when charging and discharging the gate capacitance [21]. However, the extra passive components is not desirable for space and cost saving purpose in the integrated output stage design. Therefore, the voltage driven gate driver is the most widely used topology. In the simplified equivalent circuit as shown in Fig. 2.14 (a), the input capacitance of the power transistor C_{gs} is charged by a DC voltage source through a effective resistance R_{out} .

Figure 2.15 shows the schematic of voltage driven gate driver topology, which contains an inverter chain. The gate driver loss due to the charging and discharging of capacitances internal to the gate driver is given by [22]

$$P_{gd} = f_s \cdot V_G \cdot \left(\frac{f^N - 1}{f - 1} \right) \cdot (C_{in} + C_{out}) + P_{static} \quad (2.15)$$

Equation (2.15) indicates that a long inverter chain consumes more power when compared to a short one. The chain length, namely the stage count N , is determined by the input capacitance C_{gs} of the power transistor seen at the output node of the gate driver.

2.7 Segmented Topologies

According to Section 2.5.2, the size of the output stage is determined by the desired output power level [23]. The power loss, dominated by the conduction loss and the gate driving loss can be minimized for a specific operating condition. However, in order to continuously optimize the power conversion efficiency of a DC-DC buck converter over a wide range of load current, an output stage with transistors of fixed size is not desired. Segmentation is one approach to achieve the size changing capability.

The idea of segmented topology was first presented by [4]. This topology was employed for the output stage which allows to digitally adjust the width of the output transistor

Figure 2.14: Simplified circuit for (a) voltage driven (b) current driven (c) resonant driven gate driver topologies [20].

Figure 2.15: General schematic for a voltage driven gate driver [22].

Figure 2.16: A circuit configuration showing the segmented output stage [4].

on-the-fly (as shown in Fig. 2.16). Since increasing the W/L ratio can reduce the R_{on} of the MOSFET and subsequently reduce the conduction loss P_{cond} , enabling maximum output stage segments is beneficial at heavy load. On the other hand, observations have shown that the gate driver loss dominates the power loss at low current. An output stage with small transistor size exhibits an advantage in reducing P_{sw} due to a low C_{gs} .

Research on segmented topology has been applied on the gate driver IC design to drive a practical IGBT by [22], and further developed on EMI mitigation by [7]. Figure 2.17 shows the general block diagram for the segmented gate driver circuits. Results show that the segmented gate driver has the ability to change the output resistance, which subsequently affects the switching speed, the switching node ringing and the input current oscillation. In addition, a gate driver with dynamic output resistance R_{out} allows the suppression of the EMI noise without degrading the overall power conversion efficiency [7].

Figure 2.17: A simplified diagram for the segmented gate driver [22].

2.8 Summary

This chapter outlines the SMPS fundamentals, EMI in SMPSs, existing segmentation techniques employed in the integrated circuit. In Chapter 3, detailed design process for an H-bridge IC, which incorporates segmented output stages and segmented gate drivers, is presented.

Chapter 3

Segmentation Technique for the H-Bridge Design

This chapter describes the segmentation technique used for the integrated H-bridge design. This feature allows the on-the-fly optimization of the power conversion efficiency over a wide range of load current. The influence of the power transistor's R_{on} and the gate driver's R_{out} on the power conversion efficiency is investigated via simulation. This is followed by a discussion on the optimization strategy.

3.1 Influence of R_{on} of the Output Stage

Based on the discussion in Section 2.2, the conduction loss in a DC-DC buck converter is dependent on the on-resistance R_{on} of the power transistor. Equation (2.6) shows that for a given output current, the conduction loss increases with R_{on} , which is pre-determined by the width of the power transistor according to (2.7). Under heavy load conditions, where the conduction loss dominates the total power loss, an output stage with large size power transistors is preferred. However, a large power transistor in the output stage implies a large input capacitance C_{gs} , which requires more driving effort from the gate driver. It also consumes more power on charging and discharging the power transistors. An increase switching loss from the gate driver, as indicated by (2.10), degrades the overall power conversion efficiency at light load. The trade-off between small input gate capacitance C_{gs} and small on-resistance R_{on} of the power transistors limits the performance of the traditional output stage over a wide range of load current.

In this section, a segmentation technique that allows the on-the-fly changes of the effective transistor size based on the load current condition is introduced. The high-side and low-side power devices are both n -type extended drain MOSFETs (EDMOS). Each

Figure 3.1: A segmented power transistor which contains four identical n -EDMOS devices.

Table 3.1: Characteristics of One n -EDMOS Segment

Parameter	Value	Units
On-resistance	26.6	$m\Omega$
C_{gs}	230	pF
C_{db}	7.7	nF
Threshold Voltage	1.14	V
Size ($W \times L$)	$168,000 \times 1.3$	μm^2

power transistor is divided into 4 identical segments with independent gate drivers. As shown in Fig. 3.1, the four segments are connected in parallel: source and drain are shared, while four gates are controlled by its individual gate signals.

The simulated R_{on} of each EDMOS, is $26.6 m\Omega$. Table 3.1 lists the key characteristics of one n -EDMOS. This segmentation arrangement allows the entire power transistor to exhibit four possible effective on-resistance: $26.6 m\Omega$, $12.8 m\Omega$, $8.4 m\Omega$ and $6.3 m\Omega$ by enabling one to four segments, respectively.

In order to investigate the influence of R_{on} on the power conversion performance, a DC-DC buck converter as discussed in Section 2.1 with two proposed n -type segmented EDMOS transistors is simulated using *HSpice*. The circuit parameters are as listed in Table 3.2. The voltage and current waveforms at the input node and output node are monitored in order to evaluate the power conversion efficiency.

Table 3.2: Buck Converter Testing Parameters for Different R_{on}

Parameter	Value	Units
Input Voltage V_{DD}	20	V
Output Voltage V_{out}	4	V
Load Current I_{out}	0.2 - 10	A
Switching Frequency f_s	500	kHz
Output Inductor L	47	μ H
Output Capacitance C_{out}	2	μ C
Gate Voltage V_G	5	V
Parasitic Inductances L_{vdd}, L_{gnd}, L_{vx}	500	pF
Enabled Output Stage Segment(s)	1 to 4	-

Figure 3.2 shows the simulated power conversion efficiency curves for four different on-resistance of the power transistor when current ranges from 0.1 A - 10 A. The gate driver loss and other fixed loss dominate the power loss at light load conditions. When the load current increases, the conversion efficiency first increases until it reaches a peak value. After that, the conduction loss dominates the power loss and degrades the conversion efficiency. From another perspective, at a particular load current, a specific output stage with appropriate effective R_{on} can maximize efficiency. At a low output current condition of 100 mA, the output stage with only one segment enabled (*Configuration A*) achieves 4.2 % efficiency improvement when compared to the all segments enabled situation (*Configuration B*). At a high output current of 10 A, *Configuration B* shows 7.6 % improvement in the efficiency when compared to *Configuration A*.

3.2 Influence of R_{out} of the Gate Driver

As discussed in Section 2.2, the switching loss is another main source of power loss in the DC-DC buck converter. Equation (2.9) shows that decreasing transition time t_{on} , t_{off} can help reduce the switching loss. According to (2.10), the gate driver loss is proportional to the input gate capacitance C_{gs} of the power transistor. For a voltage driven strategy, as discussed in Section 2.6, the transition time is determined by two parameters: the output resistance R_{out} of the gate driver and the input gate capacitance C_{gs} of the power transistor. If a gate driver with small R_{out} is used, the power transistor can be turned on and off fast so that switching loss is reduced. The observations above indicate that keeping R_{out} as small as possible is an ideal approach for achieving high

Figure 3.2: Efficiency versus load current for different enabled output stage segments.

conversion efficiency when delivering large output current.

However, as mentioned in Section 2.4, the R_{out} of the gate driver, the parasitic inductance existing on the path from gate driver to the gate node of the power transistor, and the C_{gs} of the power transistor together forms a RLC circuit. Depending on the values of R_{out} , L and C_{gs} , ringing oscillation could appear at the gate node of the power transistor. Figure 3.3 shows a simple equivalent RLC circuit [24]. Equations (3.1), (3.2) and (3.3) are the mathematical expressions of this second-order system. The damping factor ζ is proportional to the output resistance. A small R_{out} for the gate driver exhibits fast rise and fall times, but leads to larger ringing at the gate node, which is subsequently amplified at the switching node of the DC-DC buck converter. The discussion above explains the phenomenon that CEMI occurs during the switching transition.

$$g(s) = \frac{1}{s^2LC + sRC + 1} \quad (3.1)$$

$$\zeta = \frac{R}{2} \sqrt{\frac{L}{C}} \quad (3.2)$$

$$\omega_n = \frac{1}{\sqrt{LC}} \quad (3.3)$$

Figure 3.3: Equivalent RLC circuit for the gate driver.

Figure 3.4: Segmented gate driver design.

In [7], the authors proposed a driving strategy to dynamically adjust the driving strength of the gate driver during the switching of the power MOSFET. Experimental results proved that by using low R_{out} during the transition and then switching to large R_{out} can effectively balance the trade-off between fast switching speed and low EMI. Low output resistance allows the gate node of the power transistor to be pulled up and down in a short time. In the second stage, high R_{out} is selected to over-damp the ringing and overshoot, which help to eliminate CEMI before it is generated.

In this thesis, an integrated segmented gate driver is designed to drive an on-chip n -type EDMOS. The gate driver proposed here has a push-pull complementary design, which has a p -type MOSFET on the high-side and an n -type MOSFET on the low-side. Each gate driver is divided into 4 segments with size ratios of 10:24:64:160. As shown in Fig. 3.4, the four gate driver segments are connected in parallel similar to the segmented output stage design. Each segment has independent enable signal at the input node. The sizing factors are $\times 10$, $\times 24$, $\times 64$, and $\times 160$, for each gate driver segment.

Table 3.3: Buck Converter Testing Parameters for Different R_{out}

Parameter	Value	Units
Input Voltage V_{DD}	20	V
Output Voltage V_{out}	4	V
Load Current I_{out}	0.2 - 10	A
Switching Frequency f_s	500	kHz
Output Inductor L	47	μ H
Output Capacitance C_{out}	2	μ C
Gate Voltage V_G	5	V
Parasitic Inductances L_{vdd}, L_{gnd}, L_{vx}	500	pF
Enabled Output Stage Segment(s)	1	-
Gate Driver Sizing Factor	$\times 24, \times 64, \times 160, \times 258$	-

In order to investigate the impact of the driving strength (inversely proportional to the R_{out}) of the gate driver on the power conversion efficiency, the same buck converter circuit described in Section 3.1 is studied using *HSpice*. The operating condition is as listed in Table 3.3. The power conversion efficiencies for different driving strength over a range of load current are as shown in Fig. 3.5. For a given power transistor size and gate driving strength, the transition period (t_{on} and t_{off}) is fixed. According to (2.9), the MOSFET switching loss increases when load current becomes larger. A gate driver with low C_{gs} is preferred in this case. Nevertheless, a gate driver with high driving strength implies a large output stage, which requires a long inverter chain. The inverter chain consumes power for charging and discharging of the internal capacitance, which is independent of the load current according to (2.15). Under light load conditions, this gate driver power loss exceeds the other losses and degrades the overall efficiency. Figure 3.5 shows that the buck converter circuit with low driving strength exhibits 4.4 % higher power conversion efficiency when compared to the one with strong driving strength.

3.3 Optimized Operating Strategy of H-bridge

Traditional H-bridge designs are optimized to achieve high power conversion efficiency for a specific set of operating conditions, such as switching frequency, conversion ratio and load current. Such design greatly limits the practical applications of the H-bridge IC, since a constant, unchanging operating condition is not common. Simulation results

Figure 3.5: Efficiency versus load current for different gate driving strengths.

Table 3.4: Proposed Operating Strategy

Load Current		Output Stage Size	
I_{out}		<i>Small</i>	<i>Big</i>
Gate Driver	<i>High</i>	Light	-
R_{out}	<i>Low</i>	-	Heavy

and analysis from Sections 3.1 and 3.2 show that if the load current changes, the output stage power transistor and the gate driver should be changed accordingly in order to continuously maintain the optimal power conversion efficiency.

The proposed segmentation technique for both the output stage and the gate drivers is one way to achieve this functionality. For operation at high load current, all segments of the output stage are enabled and the gate drivers also operate with all segments enabled, minimizing the gate driver output resistance. For operation with low load current, a single segment of the output stage is enabled and the gate drivers operate with only one single segment is to maximize the gate driver output resistance. The proposed operating strategy is summarized in Table 3.4.

The high load current strategy outlined above utilizes a very low R_{out} for gate drivers. Unfortunately, a low gate driver R_{out} can lead to excessive ringing at the gate node of the transistor, which subsequently produces ringing at the switching nodes of the H-bridge.

Figure 3.6: Timing diagram for the driving strategy.

Therefore, this operating strategy can potentially generate excessive CEMI. In order to minimize ringing at the switching node, an operation similar to the one proposed in [7] is employed. The gate driver output resistance R_{out} is set high for a short period while the power transistor turns on and off. The dynamic driving strategy is shown in Fig. 3.6. As the high-side gate receives a switching signal, R_{out} is first set to be low to decrease the rise-time and then set to be high after the high-side gate is turned on. In this process, two critical timing T_{pre} and T_{post} require careful tuning based on different gate strength combinations.

3.4 System Topologies and Schematic Design

This section described the system topologies and the detailed schematic design of two important functional blocks.

3.4.1 System Topologies

In order to verify the proposed optimization operating strategy as outlined in the previous section, a segmented H-bridge IC is designed as shown in Fig. 3.7.

As mentioned in Section 3.1, each power device of the proposed segmented output stage is equally partitioned into four identical segments. Figure 3.8 illustrates the arrangements for the power transistors and the gate drivers. For each independent tunable gate driver, four parallel output stages are designed with size ratio of 10:24:64:160.

In order to enable every possible combinations of segmented output stage and tunable gate driver, a digital circuit which stores the digital control code is also integrated, as

Figure 3.7: Block Diagram of the Integrated Power H-bridge.

Figure 3.8: Block diagram of one n -EDMOS with segmented gate driver arrangement.

Figure 3.9: Block diagram for driving the low-side n -EDMOS transistor.

shown in Fig. 3.9. The digital circuits utilizes 1.8 V low voltage device. A level shifter circuit is required to convert the 1.8 V digital signal to a 5 V signal.

3.4.2 Schematic Design

Digital Data Shifting

In order to digitally control the behavior of every gate driver segments and output stage segment, the control signal is shifted through the entire chip serially to decrease the pin-count requirement.

Normally, a shifted register circuit containing a chain of D flip-flops with an reset option, as shown in Fig. 3.10, can be used. A potential problem exist, in which the serial data and clock signal propagate in the same direction, as shown in Fig. 3.11. At time t_0 , $DATA[0]$ is shifted to node $D1$, and the clock signal propagates to node $C1$. t_{DC} and t_{DD} refer to the delay times for the D flip-flop and the buffer, respectively. If t_{DC} is longer than t_{DD} , $DATA[0]$ is shifted to node D2 at time t_2 . This situation is unwanted, since at time t_2 , D2 is supposed to hold the data from last clock cycle. This shifting error may occur in every clock cycle.

This problem can be solved by shifting the digital signal and the triggering clock signal from two ends of the D flip-flop chain. Figure 3.12 shows the unit cell schematic of control signal shifting circuit. The state select signal SEL is used to enable the output code $EN[0:3]$ to adjust R_{out} on-the-fly. Two states can be dynamically switched using this topology.

Segmented Gate Driver

Figure 3.13 shows the general schematic for one segment the of gate driver circuit. It contains an non-overlap circuit and a pull-push complementary output stage. The non-overlap circuit is designed to minimize the overshoot current of the output stage.

Figure 3.10: Data shift in the same direction as the clock signal.

Figure 3.11: Hold time problem.

Figure 3.12: Unit cell of the data shifting circuit.

Input signals D , $EN[0:3]$ and $INIT$ are used to flexibly enable or disable the output stage of one gate driver segment. D is an external gating signal which determines the behavior of the gate driver for one power transistor segment. The $EN[0:3]$ and $INIT$ signals are generated by the digital circuit and then converted to the required 5 V level. If a segment of the gate driver is enabled, the resistance R_{out} seen at the output node is determined by the size of the gate driver output stage size. If a segment is disabled, the resistance seen at the output node exhibits high impedances and the R_{out} of the entire gate driver is dependent on how many segments are enabled.

There is a special condition when all the gate driver segments received a *zero* EN signal, which makes the final R_{out} shows high impedance. This is an unwanted situation during operation, since the floating output voltage of the gate driver leads to unpredictable power MOSFET states. Therefore, a initialize signal $INIT$ is used to control the gate driver behavior under such condition. When $EN[0:3]$ is [0000], $INIT$ is set to 0, in which case the low-side transistor in the output stage of the gate driver is forced to be *on*. This means the voltage at the output node of the gate driver is low, so the connected power device can be guaranteed to be *off*. $INIT$ is set to high except for the situation when $EN[0:3]$ is [0000]. Table 3.5 illustrates the outcome for different input signal conditions.

3.5 Summary

In this chapter, the influence of the power transistor R_{on} and the gate driver R_{out} on the power conversion efficiency is examined by simulation. An operating strategy is to

Figure 3.13: General schematic of each gate driver segment.

Table 3.5: Gate Driver Segment Behaviors.

Segment Behavior		D	
		0	1
Enable Signal	0	High Z	High Z
	1	off	on
INIT = 0		off	

allow the continuous optimization of the power conversion efficiency over a wide range of load current. In Chapter 4, the implementation and experimental measurements are presented.

Chapter 4

Implementation and Experimental Results

This chapter outlines the IC implementation of the proposed H-bridge IC design, and the experimental measurement in practical circuit applications. Layout design for the implementation of the segmentation technique is provided. The experimental setup is then described. Finally, measurement results for the proposed operating strategy from two aspects, namely power conversion efficiency optimization and EMI reduction, are provided.

4.1 IC Layout Design and Package Specifications

The proposed segmented H-bridge IC is fabricated using TSMC's 0.18 μm BCD Gen-2 process technology. 1 layer of polysilicon and 5 layers of metal with the ultra-thick top metal back-end technology are utilized in fabrication. The micrograph in Fig. 4.1 shows the full layout of the H-bridge IC. The chip occupies an area of $5\text{ mm} \times 5\text{ mm}$. Two identical output stages with independent input voltage nodes are placed symmetrical to each other on two sides of the chip. The low voltage digital circuit and the segmented gate drivers are located in the middle. Table 4.1 shows the area requirement for the primary functional blocks.

A simple diagram of the segmented power transistor layout pattern is shown in Fig. 4.2. Each segment of the power transistor, represented by different number, is placed around a common centroid pattern to avoid mismatch and improve linearity. The segmented gate drivers are designed with a similar layout pattern.

The final IC is mounted into a LCC84 package, as shown in Fig. 4.3 and 4.4. Table 4.2 shows the specifications for this package.

Figure 4.1: A micrograph of the full layout for the H-bridge IC.

Table 4.1: Area Requirement for Primary Functional Blocks.

Parameter	Area (mm ²)
Output Stages	17.78
Gate Drivers	0.87
Level Shifters	0.77
Digital Circuit	0.19

Figure 4.2: Common centroid layout pattern for one power transistor.

Figure 4.3: H-bridge IC mounted in the LCC84 package.

Figure 4.4: Micrograph showing the wire bonding inside the LCC84 package.

Table 4.2: LCC84 Package Specifications.

Item	Description/Specification
Package Type	Leadless Ceramic Carrier (LCC)
Bare Die Size	5 mm $\times$ 5 mm
Pin Number	84
Bonding Wire Material	Au
Bonding Wire Diameter	1.3 mil
Wire Max Current	1 A
Pad Pitch	0.05 mil
HV Pad Number	16
LV I/O Number	32

Figure 4.5: Voltage potentials on Metal 4 layer as simulated using *R3D*.

4.2 On-resistance R_{on} Measurements

In order to verify the segmented output stage design, the power transistor's on-resistance R_{on} is simulated or measured. The ideal *Spectre* model only incorporates the channel resistance. It grossly under-estimate the on-resistance of the power transistor. The post-layout resistance extracted from *R3D* takes metal interconnect resistance into account, but not the bond wire resistance. The actual R_{on} was measured on the PCB after chip fabrication.

R3D is a novel software which can extract the post-layout R_{on} of power semiconductor devices [25]. In addition, *R3D* provides a graphical user interface to display the analytical results such as the current density and the voltage potentials. It allows the users to customize the parameters, i.e. technology process, bond wire contacts location and related parasitic resistance, which help build a realistic model and makes the extracted resistance convincing [25].

Figures 4.5 and 4.6 demonstrates the graphical *R3D* results for only one segment of the power transistor enabled situation. The potential distribution and current density profile are simulated and monitored in these two figures, respectively. Figure 4.7 shows a report of the post-layout resistance extraction.

The simplified representation of the experimental setup to measure the R_{on} is as shown in Fig. 4.8. Since the two power transistors in the output stage are identical,

Figure 4.6: Current density on the diffusion region as simulated using *R3D*.

Table of resistive components contributions to Rdson (in Ohms)				
Component	Type	Res_[Ohm]	Source_[Ohm]	Drain_[Ohm]
Diffusion	MET	0.000216588	0.000199143	1.74448e-05
metal1	MET	0.00245255	0.00234998	0.000102571
metal2	MET	0.00138943	0.00121243	0.000176998
metal3	MET	0.0106841	0.010483	0.000201136
metal4	MET	0.0126057	0.0124383	0.000167403
metal5	MET	0.00749115	0.00095775	0.0065334
ndiff_cont	VIA	1.53717e-08	9.13743e-09	6.23431e-09
VIA1	VIA	0.000219298	0.000192388	2.69106e-05
VIA2	VIA	0.000427785	0.000396766	3.10187e-05
VIA3	VIA	9.9413e-05	6.32407e-05	3.61723e-05
VIA4	VIA	0.000148391	0.000100972	4.74191e-05
s1	TC	0		
s2	TC	0		
d1	TC	0		
d2	TC	0		
Interconnects	ALL	0.0357345	0.028394	0.00734048
channel	DEV	0.0271877		
Total Rdson		0.0629221		
Extracted channel/gate width [um]: 672296				
Average channel resistivity [Ohm*um]: 18278.1				

Figure 4.7: Post-layout resistance extraction report generated from *R3D* simulation.

Table 4.3: Summary of R_{on} Measurements.

R_{on} (m Ω)	Enabled Segment(s)			
	1	2	3	4
<i>Spectre</i>	26.6	12.8	8.4	6.3
<i>R3D</i>	62.9	45.9	38.1	34.9
Measured	134	118	111	108

Figure 4.8: Simplified circuit diagram for R_{on} measurement.

the R_{on} measurement here only focus on the low-side transistor. The high-side power transistor is disabled and the low-side one is turned on. A $1\ \Omega$ power resistor is connected to the switching node of the original output stage. It is well known that MOSFETs have a positive temperature coefficient. If the low-side n -EDMOS is conducting a high current for a long time, the R_{on} goes higher. Therefore, a 1 V voltage pulse is applied at the input node to avoid such heat effect. V_{ds} is measured for different number of enabled segments. The on-resistance R_{on} is calculated by dividing V_{ds} over the current. Table 4.3 lists the results obtained by different models and testing situations. The comparison indicates that for each segment of the power transistor, five layers of metal bring about 25 to 36 $m\Omega$ to the ideal channel resistance, while bonding wires from a practical package and PCB traces bring another 70 $m\Omega$ parasitic resistance. Metal routing in the layout design further increases to the total R_{on} . The parasitic resistance is fixed for a given package and PCB. Its influence on the total resistance increases when all segments of the power transistor are enabled.

Figure 4.9: The buck converter circuit test bench.

4.3 Experimental Setup

In order to verify the performance of the segmented H-bridge, an experimental test bench is setup as shown in Fig. 4.9. For the work presented in this thesis, half of the H-bridge is examined at a time, with the other half disabled. Therefore, the experiments were performed with the H-bridge operating as a half bridge, with the circuit parameters as listed in Table 4.4. In order to measure the power conversion efficiency over a wide range of output current, an electronic load with programmable resistance is used. The input voltage, input current and output voltage are probed using an [Agilent MSO-X 3024A] oscilloscope. The rise-time and overshoot of the switching node voltage is monitored in order to verify the effectiveness of the segmented gate driver. A *DE2-115* FPGA board with an *Altera Cyclone IV* is used to generate the 1.8 V digital control signals. In order to investigate the effect of different gate driver output resistances on the CEMI produced by the system, the spectrum of the input current data are measured between 150 kHz and 50 MHz. The PCB implementation details and whole testing setup can be found in Fig. 4.10 and 4.11, respectively. Two integrated H-bridge chips were used to build the output stage on the PCB since the high-side power transistors were not fully functioning after fabrication.

Figure 4.10: PCB implementation details.

Figure 4.11: Overall experimental system setup.

Table 4.4: Buck Converter Experimental Parameters.

Parameter	Value	Units
Input Voltage V_{DD}	10	V
Output Voltage V_{out}	2	V
Load Current I_{out}	0.02 - 4	A
Switching Frequency f_s	100 ^a	kHz
Output Inductor L	47	μ H
Output Capacitance C_{out}	2	μ C
Gate Voltage V_G	5	V
Enabled Output Stage Segment(s)	1 to 4	-
Driving Strategy $EN[0:3]$	[1111] to [0001]	-

^aA low switching frequency is chosen to accommodate the slow rise and fall times of the gate signal when a single gate drive segment is used.

4.4 Switching Speed and Ringing

Based on test bench as shown in Fig. 4.9, the behavior of the switching node voltage waveform is recorded during the turn-on of the high-side transistor for different gate driver output resistances. The switching node waveforms, as shown in Fig. 4.12, is monitored with a load current of 1 A. It is evident that by decreasing the output resistance of the gate driver, the rise-time of the switching node drops from 472 ns to 33 ns. However, a lower gate driver output resistance also produces a greater amount of ringing at the switching node and increases the voltage overshoot from 0% to 33% ¹. Therefore, this inverse relationship leads to a trade-off between the turn-on time and the overshoot, as shown in Fig. 4.13.

In order to find an appropriate operating point, the output resistance R_{out} is dynamically changed during the rising transition of the switching node. Two settings of R_{out} are selected to demonstrate the benefits of the proposed driving strategy. R_{out} is pre-set to a low value (all gate driver segments enabled with $\times 258$ driving strength) before the switching node starts to rise. The high-side power transistor is then turns on with the same fast speed provided by the $\times 258$ driving strength. Once the gate voltage transition is completed, a high R_{out} is selected so that the ringing at the gate node can be damped. The switching node waveform for the proposed driving strategy can be found in Fig. 4.14, in which the overshoot and ringing is suppressed by 66% while the rise-time

¹Overshoot voltage is defined as the ratio of the overshoot voltage value to the stable state step voltage.

Figure 4.12: Switching node waveforms for different constant driving strength.

is maintained at 84 ns.

4.5 Power Conversion Efficiency Measurements

The optimization strategy aims to find the best combination of gate driver configuration and output stage configuration to maximize the efficiency for a given current load. Two variables, namely R_{on} of the output transistor and R_{out} of the integrated gate driver, are investigated in this thesis. Experiments to record the power conversion efficiency versus load current are conducted in order to verify the simulation results in practical circuits.

Figure 4.15 plots the power conversion efficiency curves for three typical driving strengths when a half-bridge is operated with all output stage segments enabled. At light loads, small driving strength, which means high R_{out} , benefits from less power loss on gate driver circuit. At heavy load current, strong driving strength shows its advantage on switching loss reduction because of the short turn-on and turn-off transition time. A similar plots for the comparison of the efficiency with different output stage size is as shown in Fig. 4.16. A large power transistor is preferred for high load current, due to the small R_{on} . At light loads, a small size output stage achieves higher power conversion efficiency since a small C_{gs} reduces the gate driver loss according to (2.10).

Based on the discussions and experimental results mentioned above, the final H-bridge operation optimization is performed for load currents between 0.02 A and 4 A. The power

Figure 4.13: Rise time versus overshoot percentage.

Figure 4.14: Switching node waveform for the proposed strategy.

Figure 4.15: Power conversion efficiency versus load current for different gate driving strengths.

Figure 4.16: Power conversion efficiency versus load current for different output stage sizes.

Figure 4.17: Efficiency versus load current for a half-bridge with the proposed operating strategy.

conversion efficiency for two typical combinations of output stage configuration and gate driver configuration is as plotted in Fig. 4.17. These two configurations are *Configuration A* (when the output stage is configured such that only one segment is enabled and $\times 64$ driving strength is used) and *Configuration B* (all output transistor segments and all gate driver segments are enabled).

For *Configuration A*, improvement in efficiency from 6% to 12% is observed when compared to *Configuration B*, for a load current of 0.02 A. *Configuration A* exhibits a 25% efficiency improvement at light load. However, for higher load currents, the advantages of *Configuration A* is diminished due to higher conduction loss and switching loss. The highest recorded efficiency of 86% occurs when the load current is 2 A and is achieved by *Configuration B*, which shows an 8% efficiency improvement at 4 A when compared to *Configuration A*. The results in Fig. 4.17 demonstrate that for a given load current, a specific H-bridge configuration will maximize efficiency. Furthermore, as the design is digitally controlled, the combination of output stage and gate drive segments can be continuously optimized as the load current changes.

Figure 4.18: CEMI spectrum for two typical driving strength at 1 A load current.

4.6 Switching Node Ringing on EMI Issue

In order to investigate the conducted emissions of the strategy discussed in the previous section, the input current of the H-bridge was monitored using a [TektronixTCPA300] current probe. The CEMI was then determined by calculating the spectrum of the input current between 150 kHz and 50 MHz. When the half-bridge, with all output stage segments enabled, is operated with the parameters as listed in Table 4.4, the CEMI spectrum for the highest driving strength ($\times 258$) and lowest one ($\times 10$) at 1 A load current is plotted in Fig. 4.18. The peak CEMI within a range of 20 MHz to 40 MHz is 5.2 dBm.

The proposed driving strategy for overshoot suppression is presented in Section 4.4. This section focuses on the CEMI mitigation at high switching frequency. Experimented half-bridge is operated at 1 MHz, with target load current of 1 A. As described before, during the rising transition of the switching node, R_{out} of the high-side gate driver is first set to a low value for fast switching and then switched to a high value to damp the ringing. By using the proposed gate driving strategy (dynamically switched from $\times 258$ to $\times 10$), an input current overshoot suppression of 6 A was observed, as shown in Fig. 4.19, when compared to the constant highest gate driving strength condition. The CEMI spectrum comparison can be found in Fig. 4.20. This figure clearly shows that within the frequency range of 20 MHz to 40 MHz, the peak CEMI has a reduction of 5.5

Figure 4.19: Input Current waveforms for different driving strategies.

dBm. In addition, the power conversion efficiency for the dynamic driving strategy is 81%. Compared to the 83% of the constant driving mode, the proposed strategy is not significantly degraded.

4.7 Summary

In this section, the segmented H-bridge IC with tunable gate driver is implemented and tested in a typical buck converter. The improvement in the power conversion efficiency is achieved by employing the proposed operation strategy. In addition, CEMI reduction is observed when dynamic driving strength is used. In Chapter 5, the conclusions of the research and future work are provided.

Figure 4.20: CEMI spectrum for different driving strategies.

Chapter 5

Conclusions

The focus of this thesis is to optimize the output stage performance over a wide range of current, by operating with the best combination of segmented output stage and segmented gate drivers. EMI reduction by dynamically adjusting the output resistance of the gate driver is also investigated in this thesis.

5.1 Conclusions

This thesis investigates a segmented H-bridge design technique whereby the best combination of segmented output stage and gate driver are used to maximize the power conversion efficiency over a range of current loads. The design is capable of improving the power conversion efficiency by 25% under light load and 8% under heavy load conditions. Appropriate driving strength is also examined in this thesis for switching node ringing suppression. A reduction in CEMI by 5.5 dBm without any significant power conversion efficiency degradation is obtained.

A paper titled “*A segmented Output Stage H-Bridge IC with Tunable Gate Driver*” was accepted at “The 26th International Symposium On Power Semiconductor Devices and ICs”.

5.2 Future Work

The segmented H-bridge design proposed in this thesis could be further studied to include the following topics.

- **Broader load current range:** The size for the unit segment of the output stage power transistor and the gate driver could be further optimized to suit for a broader

range of load current. Decrease the unit segment size, but scale the total number of segments may help examine the efficiency at even lower or higher output power (i.e. less than 0.02 A or larger than 10 A).

- **Pulse frequency modulation (PFM):** PFM can be utilized to help further improve the power conversion efficiency at light load conditions. The gate driver loss could be significantly reduced when H-bridge is operating in PFM mode.
- **More states for driving strategy:** The current design allows the switching between two states of the driving strength. Further work could be expanded into four or more states such that more than two settings of R_{out} could be selected during a transition period. In addition, impact of different T_{pre} and T_{post} setting at the rising edge and the falling edge of the switching node can be examined.
- **More practical applications:** Only half-bridge is experimented so far. The testing can be expended to full H-bridge operation applied in more practical situations, e.g. motor drive, Class-D amplifiers, etc.. Benefits of the segmentation techniques can be further demonstrated in this way.
- **Smart close loop system control on chip:** The current design is experimented in a open loop circuit. If a close loop digital control could be implemented on chip, the best operating combination and driving strategy can be self-tuned, on-the-fly. In addition, this feature could greatly eliminates the propagation delay and noise brought by the long path from PCB to IC.

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