A Versatile 200-V Capacitor-Coupled Level Shifter for Fully Floating Multi-MHz Gate Drivers

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Abstract— This paper presents a novel level-shifter circuit for high-frequency high-voltage (HV) gate-drives. The proposed level shifter (LS) is designed based on a capacitive-coupler/current mirror/ latch structure which helps to extend operation voltage of a floating supply into the negative range, achieves sub-ns and constant delay, and consumes very low power from the floating supply. Additionally, common-mode noise cancellers based on a cross-current mirror and transmission gates are also presented to enhance the dV/dt immunity of the LS against slewing of the floating ground. Implemented in 0.18 μm HV BCD-on-SOI (bipolar-CMOS-DMOS on silicon-on-insulator) process, the post-layout simulation of the proposed design shows a delay of 680 ps, 200 V/ns of dV_{SSF}/dt slew rate immunity. It dissipates no static power and only 8.1 pJ/transition from the floating supply, improving FoM¹ and FoM² of the proposed LS by 3 times and 11.7 times compared to respective state-of-the-art works.

Keywords—Capacitive-coupled level shifter, BCD-on-SOI, floating level shifter, gate driver, GaN, level shifter, sub-ns delay.

I. INTRODUCTION

The level shifter (LS) circuit of the gate driver (GD) is an essential block that shifts the ground referenced pulse-width modulation (PWM) signal to different voltage domains. Fig. 1(a) shows a common design of GD used for half-bridge (HB) based DC-DC converters with the LS is highlighted [1]. The LS transfers a PWM signal from the low side to the high-side buffer for driving the high-side power switch. The four main challenges for the LS design in this context are:

(Ch1) *Low propagation delay*: the delay of the GD is mainly determined by the delay of the LS. For multi-MHz and/or high conversion ratio DC-DC converters, a LS with sub-ns delay can fully exploit the high switching speed of GaN FETs [2-3].

(Ch2) *High dV_{SSF}/dt immunity*: modern high-switching DC-DC converters use N-type MOS/GaN FETs as they offer low $R_{\text{DS,ON}} \times Q_g$. When driving the HS N-type FETs as shown in Fig. 1(a), the LS would suffer from high dV_{SW}/dt (dV_{SSF}/dt) that can erroneously flip the its output. For high-voltage (HV) applications using GaN FETs, dV_{SSF}/dt can be up to 100 V/ns; thus, a LS with $dV_{\text{SSF}}/dt > 100$ V/ns immunity is needed [3].

(Ch3) *Low energy consumption*: the LS consumes power from the floating supply (modeled as C_B). The LS with high power consumption results in a high loss and thermal issues of the GD. Modern gate drivers prefer fully on-chip floating supplies whose power capability is limited. Therefore, the LS should consume as little power as possible [4].

(Ch4) *Negative V_{SSF} tolerance*: there are applications where V_{SSF} is negative. For example, during the dead-time $V_{\text{SW}} = V_{\text{SSF}} = -V_D$ (forward voltage of body diode). In

general, $V_{\text{D,MOSFET}} < 1.5$ V while $V_{\text{D,GaN}}$ can be up to 4.5 V [5]; thus, the remaining headroom $V_{\text{DDF}} - V_{\text{SSL}}$ is small such that it causes longer delay or even failure of the LS. The scenario becomes more critical if V_{SSF} is more negative with the effect of current sensing resistor R_s and/or parasitic L_{par} . With new emerging DC-DC converters such as hybrid ones [6-8], V_{SW} can be in positive or negative domain as shown in Fig. 1(b). This extends the challenge (Ch4) such that the LS should work even with a large negative potential of the floating ground V_{SSF} .

Fig. 1. High-voltage gate driver for: (a) HB-based DC-DC converters, (b) DC-DC converters with different floating sides.

Several level shifters have been proposed [9-17], which can be divided into two main categories: active-coupled LS (AC-LS) [9-15] and capacitive-coupled LS (CC-LS) [16-17]. AC-LSs presented in [9-11] achieve a delay of 3 ns, 2.03 ns and 1.3 ns, respectively. However, these works cannot be used in systems where V_{SSF} slews fast and/or V_{SSF} is negative. AC-LSs with dV_{SSF}/dt slewing immunity are also reported [12-14]. The work in [13] presents a dV_{SSF}/dt immunity up to 120 V/ns but has a high delay of 20 ns. The work in [14] has only 0.37 ns delay but its dV_{SSF}/dt immunity is limited at 30 V/ns. In these works, negative V_{SSF} tolerance is not reported. The recent work in [15] presents an excellent LS for SMPS applications with a 200 V/ns dV_{SSF}/dt immunity, 0.53 ns delay, but minimum V_{SSF} is only -1.5 V. Though [12-15] can be applied to conventional HB-based DC-DC converters, even few negative volts of V_{SSF} would affect their reliable operation. These AC-LSs also cannot be used for GDs for emerging converters, as V_{SSF} is in negative domain [6-8]. Digital isolators are widely used as LSs to transmit the PWM signal, whether V_{SSF} is negative or positive. However, this solution adds more complexity, cost, and size to the GD designs [18]. Furthermore, high-power consumption leads to a large C_B which is difficult to integrate C_B on-chip as the power consumption is high and scaled-up with V_{DDF} in these AC-LSs. Regarding CC-LS designs, reported works in [4, 16] also achieve sub-ns delays but are used for static V_{SSF} . A CC-LS with slew-rate immunity is reported in [17]

with 1.45 ns delay and consumes only 4.1 pJ/transition, but dV_{SSF}/dt is limited to 6 V/ns. In short, the reported works [9-17] are not able to satisfy all four design challenges; thus, limiting their use to specific applications.

To tackle these issues, this paper presents a versatile LS for HV gate drivers that simultaneously achieves four design criteria (Ch1-Ch4). The proposed LS achieves 680 ps delay, 200 V/ns dV/dt immunity, 8.1 pJ/transition and can work under negative floating supply. Therefore, it can be applied to both conventional DC-DC converters and emerging ones. The remainder of the paper is organized as follows: Section II presents the proposed LS. The performance of the proposed LS is verified with intensive post-layout simulation results in Section III. Section IV concludes the paper.

II. PROPOSED CAPACITIVE-COUPLE LEVEL SHIFTERS

A. Basic capacitive-coupled level shifter

Fig. 2 shows the proposed basic CC-LS (BCC-LS) which is composed of three parts: (1) a pulse generator at low side, (2) two symmetrical current mirrors, a back-to-back inverter-based dynamic latch, and two buffer inverters at the floating side, (3) two coupling capacitors C_1 , C_2 that connects the low side and the floating side. All devices are standard low-voltage CMOS devices, except the HV coupling capacitors.

Fig. 2. The proposed basic capacitive-coupled level shifter with rising-edge operation. (Red dashed boxes are deep N-wells for isolations).

The operation of the proposed BCC-LS is also shown in Fig. 2 (for the rising-edge operation) and its key timing waveform is shown in Fig. 3. On the rising transition of the input signal, a short pulse is generated at the output V_{PR} of the pulse generator while the output V_{PF} is maintained at V_{DDL} . As V_{PF} does not change, the current mirrors PM3 and PM4 are inactive. Assume that C_1 , C_2 is much higher than C_{par} ($C_{par} = C_{ds,PM1} + C_{gs,PM1,2}$), a falling edge of the short pulse V_{PR} pulls the voltage V_A down from V_{DDF} to $V_{DDL} - V_{DDL}$; thus, generating a current I_{PM1} on the C_{C1} -PM1. This current is copied through the 1:N current mirror PM1-PM2 to charge the node C. As the triggering current I_{PM2} is enough to pull up node C to exceed the threshold voltage of the latch, the positive feedback of the latch accelerates node C to V_{DDF} and node D to V_{SSF} . Even when the triggering current reduces to zero, nodes C and D of the latch are held, ensuring the output is kept until the next change at the input IN. The inverter flips the voltage at node D so that the input

and output signal are equal. In the same fashion, at the falling edge of the input signal, the right half is activated and a current I_{PM4} is generated to change the state of the output to V_{SSF} . It should be noted that at the rising edge of the short pulse at V_{PR} and V_{PF} of the pulse generator, capacitors C_1 and C_2 are discharged to the floating supply capacitor C_B through the body diode of PM1 and PM3, thus clamping V_A and V_B at a maximum level of $V_{DDF} + V_D$ where V_D is forward voltage of the body diode D_{B1} , D_{B2} .

Fig. 3. Simulated waveform of the BCC-LS at $V_{SSF} = -20$ V.

Fig. 4. Simulated delays, energy consumption of the BCC-LS at different floating ground potential.

Fig. 4 shows the pre-layout simulation results of the proposed BCC-LS in terms of rising/falling propagation delays (T_R/T_F) and energy per transition (E_T) versus floating ground potential using 180 nm SOI CMOS process. In this test scheme (shown by the inset), floating supply is set to the fixed values of 5 V ($V_{DDF} - V_{SSF} = 5$ V) then a DC offset is swept from -20 V to 20 V. It can be seen that the propagation delays T_R/T_F are 575 ps/390 ps, respectively, while the energy per transition E_T is only 5.4 pJ. Especially, the level shifter works whether the potential of V_{SSF} is negative or positive, which is not the case with AC-LS reported in [9-15]. The insensitive delays versus V_{SSF} help to achieve delay matching between different floating sides. Furthermore, the energy consumption of the proposed BCC-LS is also a constant and independent of V_{SSF} . The low delays result from fast level shifter structure using high-bandwidth capacitive couplers and current mirrors, low power consumption and negative V_{SSF} tolerance is a consequence of using capacitive-coupling technique for the proposed level shifter.

In summary, 3/4 key specifications defined in Section I for a versatile LS can be achieved with BCC-LS: sub-ns delay

(Ch1), low power consumption (Ch3), and wide negative V_{SSF} tolerance (Ch4). The low power consumption leads to relaxed requirements for the floating-side power supply, thus making a fully integrated floating supply feasible.

B. High dV_{SSF}/dt capacitive-coupled level shifter

The proposed BCC-LS is well suited for gate drivers where V_{SSF} is static such as those driving P-type FETs. However, for deployment in gate drivers driving n-type FETs, the requirement (Ch2) should be satisfied. That is a robust operation under fast dV_{SSF}/dt slewing. With the BCC-LS shown in Fig. 2, when V_{SSF} slews up/down with a fast rate, slew-induced common-mode current I_{cm} appears on both capacitive-coupling branches as the C_1 and C_2 need to be charged/discharged. When V_{SSF} slews up, and assume that node C is at V_{SSF} and node D is at V_{DDF} , slew-induced current I_{cm} is only copied by the current mirror PM1-PM2 because $V_{ds,PM1} = V_{DDF}$ while $V_{ds,PM4} = 0$. If the induced common-mode current $I_{PM2} = N \times I_{cm}$ is high enough, node C can be charged up and the output of the LS flips erroneously. When V_{SSF} slews down, induced I_{cm} currents are in the reverse direction and discharge to the floating supply V_{DDF} through the body diodes of PM1 and PM3. This does not cause any effect on the output of the LS. By simulation, the maximum V_{SSF}/dt of the basic BCC-LS is only 12 V/ns. To cope with the high dV_{SSF}/dt requirement, a modified version of the BCC-LS, so-called MCCL-LS is proposed as shown in Fig. 5. The dV_{SSF}/dt immunity of the modified LS is enhanced by using two cross-current mirrors (CCMs) [15] and two fast transmission gates (TGs). Brief operation of the MCC-LS is explained below:

+ *Differential mode (shifting mode)* – Fig. 5(a): At the rising edge of the input IN, a short pulse is generated at the output of the pulse generator V_{PR} . This short pulse pulls down the voltage at node A1 and generates a current I_{PM1} through the diode-connected transistor PM1. While V_{PF} of the pulse generator is maintained at V_{DDL} , the right half (C_2 , PM3, PM4, PM6, NM3-NM4) is inactive. The current PM1 is mirrored by PM2-PM5, NM1-NM2, goes through the TGs to charge node D2 and discharge node D1 (in differential mode, TGs are always ON as either A1 or A2 is at V_{DDF} , and either B1 or B2 is at V_{SSF}). As node D2 is charged, source/sink triggering currents I_{NM6}/I_{PM10} are generated. As a consequence, node F1 is charged via PM10 while node F2 is discharged via NM6; thus, triggering the latch and flip the output OUT to V_{DDF} . It should be noted that since the latch is charged/discharged by PM10/NM6 simultaneously, the latch is triggered faster as compared to BCC-LS shown in Fig. 2. The same working fashion is applied at the falling edge of the input IN to pull OUT down to V_{SSF} .

+ *Common mode (dV_{SSF}/dt noise cancelling mode)* – Fig. 5(b): When V_{SSF} slews fast, induced common mode currents appear on both PM1 and PM3, then copied and cancelled by the first CCM at node C1 and C2 (ideally $I_{cm1} = I_{cm2}$). However, when the induced common mode current is high and due to mismatch (between C_1 , C_2 ; between devices of the first CCM caused by process variations, etc.), the net common mode currents coming out nodes C1 and C2, that is $I_{cm,C1,2} = |I_{cm1} - I_{cm2}| \neq 0$. These currents can be further blocked by the transmission gates TG1 and TG2. When V_{SSF} slews, nodes A1 and A2 are pulled down and nodes B1 and B2 are pulled up. As a consequence, TG1 and TG2 are opened and net common mode currents $I_{cm,C1,2}$ cannot enter nodes D1 and D2 as the TGs' responses are faster than the

current mirrors. As a consequence, the second CCM is not triggered and the latch states are not affected.

Fig. 5. Operation of MCC-LS: (a) Differential mode, (b) Common mode (Red dashed boxes are deep N-wells for isolations).

At high dV_{SSF}/dt rate, I_{cm} is significant and source-gate/source-drain dropped voltages of PM1/PM4 can be higher than maximum rated values of the utilized standard 5 V MOSFETs. This can break down low V_{GS} devices. By using stacked-diode clamping cells (C-cells), C_1 and C_2 are charged via these clamping cells and V_{SG}/V_{SD} of PM1-PM6 are limited within a safe range. The clamping voltages are designed such that none of the clamping cells is conducted in shifting mode. In summary, as TGs and CCMs are included, dV_{SSF}/dt immunity of the proposed MCC-LS is higher than BCC-LS counterpart with the cost of longer delay

III. POST-LAYOUT SIMULATION RESULTS

The proposed MCC-LS was implemented using 0.18 μm HV BCD-on-SOI process supplied by X-FAB. The coupling capacitors C_1 and C_2 are HV devices with a maximum rated voltage of 200 V and the remaining are standard 5 V devices. To reduce the induced common-mode currents during fast dV_{SSF}/dt slewing, coupling capacitors C_1 and C_2 have a small value of 40 fF. Besides, two 300 k Ω resistors R1 and R2 are added between nodes A1 and A2 and V_{DDF} which help to reset these node potentials to V_{DDF} to be ready for the next pulse, especially for high-frequency operation. The layout and test bench are shown in Fig. 6(a) and 6(b), respectively.

Fig. 7(a) shows the post-layout simulation results of the MCC-LS in terms of delays and energy consumption per transition. The rising/falling delays and energy consumption

of the MCC-LS are only 686 ps/660 ps and 8.1 pJ, respectively. This confirms the sub-ns feature of the proposed design. Also, the propagation delays and energy consumption are constant and independent of floating ground V_{SSF} . The dV_{SSF}/dt immunity of MCC-LS is shown in Fig. 7(b). Even after exposing a slewing rate up to 200 V/ns to V_{SSF} node, the output status follows the input perfectly as the common currents entering node D1 and D2 are well-cancelled. This can be confirmed by low common mode voltage noise at node D1. Therefore, the proposed MCC-LS guarantees a malfunction-free operation up to 200 V/ns of dV_{SSF}/dt slew rate. Besides, V_{PR} of the low side is clamped by body diodes of CMOS devices at the output of the NAND gate of the pulse generator while V_{SD} of PM1, PM3 of floating side are clamped by body diodes of PM1, PM3 and C-cells; thus $-V_D < V_{DDL}/V_{DDF} + V_D$. This verifies that low-voltage CMOS devices are always operated within safety regions during V_{SSF} transition.

Fig. 6. (a) Layout of the proposed MCC-LS level shifter, (b) Setup for post simulation of level shifters.

Fig. 7. Post-sim results of MCC-LS: (a) Propagation delays and energy consumption, (b) dV_{SSF}/dt immunity at 200 V/ns, (@ $f_{IN} = 20$ MHz, OUT is referenced to V_{SSF}).

Table I summarizes the proposed LS's performance (MCC-LS) along with the prior arts mentioned in Section 2. Among these works targeted for high-voltage gate drivers, the proposed LS can transmit signals with a delay below 700 ps, which confirms the sub-ns delay beside the works in [14-15]. The dV_{SSF}/dt immunity of the proposed design is

200 V/ns, which is highest and is comparable to [15]. The energy consumption $E_T < 8.1$ pJ/transition and stays constant regardless of the floating supply V_{DDF} potential. Only [17] has lower E_T , but its delay is higher and dV_{SSF}/dt is much lower. At higher floating supply potential, the benefit of the proposed LS in terms of E_T becomes more significant as compared to AC-LSs in [14-16]. By extrapolation, the LS using active-coupled technique in [16] has E_T of 285 pJ that is higher than E_T of this work by a factor of 35 at 200 V of V_{DDF} . The low energy consumption makes the proposed LS very suitable for the implementation of fully integrated floating supplies for gate drivers. By using capacitive coupling technique, the operation range of the proposed LS is not limited in positive range of V_{DDF} . In particular, the proposed work can shift the signal down to -200 V of the floating rail.

TABLE I: COMPARISON WITH PREVIOUS WORKS

REF	$V_{DDF,max}$ (V)	E_T (pJ)	$V_{SSF,min}$ (V)	T_d (ns)	dV/dt (V/ns)	$FoM^1/cFoM^2$	Process (μm)
[4]	40	160	N/A	2	40	0.1/64	0.5
[14]	20	27.6	N/A	0.37	30	0.1/23	0.18
[15]	50	30.3	-1.5	0.53	200	0.06/55	0.18
[17]	50	4.1	N/A	1.45	6	0.16/20	0.18
This work	200	8.1	-200	0.67	200	0.02/4.7	0.18

^a $T_d = (T_{d,R} + T_{d,F})/2$; ^b FoM^1 : ns/($\mu m \cdot V$); ^c FoM^2 : (pJ·ns)/($\mu m^3 \cdot V$)

To evaluate the overall performance of the proposed LS in terms of delay, energy consumption, process dependence, and transfer voltage, two figure of merits (FoMs) are commonly used as benchmarks [14, 17]. FoM^1 takes into consideration of transfer delay, process node length and transfer voltage while FoM^2 includes those factors and also reflects power dissipation. The smaller the FoM^1 , the faster the LS and the smaller the FoM^2 , the LS is faster and more energy efficient. At 200 V of floating supply potential V_{DDF} , the proposed LS has a FoM^1 of 0.02 ns/($\mu m \cdot V$) and an excellent FoM^2 of 4.7 (pJ·ns)/($\mu m^3 \cdot V$), thanks to the low delay and low E_T . These figures of merits can be considered the lowest as compared to the reported works.

IV. CONCLUSION

Although many high-speed, high dV_{SSF}/dt immunity level shifters have existed, the limited negative V_{SSF} tolerance and high energy consumption are the remained drawbacks. This paper has presented a novel capacitive-coupler/current mirror/ latch structure for floating high-voltage gate drivers. The proposed design achieves sub-ns of less than 700 ps of propagation delays and a dV_{SSF}/dt up to 200 V/ns. Especially, the proposed LS can operate even with negative floating supply and consumes a very low energy consumption of only 8.1 pJ/transition regardless of the value of floating power supply rail. Therefore, the proposed level shifter can be used for gate drivers for both of conventional and emerging DC-DC converters. Measurements will be conducted in the future to confirm the performance of the proposed level shifter.

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