

An Isolated Gate Driver Enabled by Simultaneous Wireless Information and Power Transfer

Lukas Hüssen, *Graduate Student Member, IEEE*, Muh-Dey Wei, Renato Negra

Abstract—Wide bandgap semiconductor devices, such as gallium nitride (GaN) or silicon carbide (SiC), are key enablers for innovative power electronic circuit topologies. However, novel types of isolated gate drivers with high common-mode transient immunity and high switching speed need to be developed to put these benefits into action. We propose a radio-frequency gate driver (RFGD) which utilises a single radio-frequency coupling channel and simultaneous wireless information and power transfer (SWIPT) at a 2.4 GHz carrier frequency and harmonic backscattering for monitoring signal transmission. A hybrid prototype is implemented and the measurements are demonstrated. The prototype components are validated individually and the full system showcased by driving the gate of an eGaN-FET in a common-drain circuit. The transferred switching signal is pulse-width modulated at 62.5 kHz and modulates a 2.6 kHz sine wave. The prototype draws its entire power from a single 282 mW RF input signal at this operating point. The common-mode transient immunity (RFGD) of the prototype is discussed and related to the signal-to-noise ratio (SNR) of the receiver path.

Index Terms—Common Mode Transient Immunity (CMTI), Harmonic Backscattering, Isolated Gate Driver, Simultaneous Wireless Information and Power Transfer (SWIPT)

I. INTRODUCTION

High-power semiconductor switches enable efficient conversion of electric energy within power electronic circuits. Semiconductor switches using novel semiconductor materials, such as silicon carbide (SiC) and gallium nitride (GaN), have high power densities and can be operated at high switching frequencies [1]. In many power electronic applications, a high switching frequency is desirable as it decreases the size of the passive components and therefore increases the overall power density while also minimising cost and material [2]. However, elevated switching frequencies put a high demand on the control of the semiconductor switch [3]–[5].

When the source node of the semiconductor switch is not at a fixed voltage, an isolated gate driver is needed to control it while galvanically isolating the control circuitry from the semiconductor gate. Both the power needed to switch the device as well as the switching signal need to be transferred through this isolation barrier. The main measure to assess the quality of the isolation is the common-mode transient immunity (CMTI) [6]. The CMTI gives the rate at which the source node potential can rise with respect to the control

This paper is an expanded version of the paper entitled "Power Efficient Simultaneous Wireless Information Transcepcion and Power Reception Through an Isolation Barrier" from the 2024 IEEE Wireless Power Transfer Technology Conference and Expo, Kyoto, Japan, 8-11 May, 2024.

Lukas Hüssen, Muh-Dey Wei and Renato Negra are with the Chair of High Frequency Electronics, RWTH Aachen University, 52074 Aachen, Germany

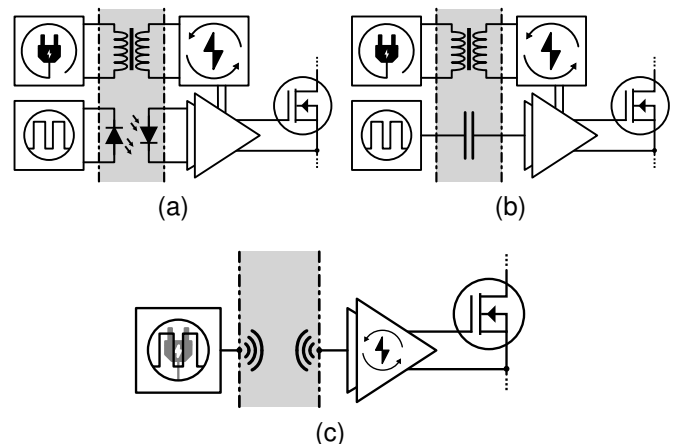


Fig. 1. Isolated gate driver concepts: (a) magnetic-optical (b) magnetic-capacitive and (c) wireless.

circuit ground without affecting the proper operation of the gate driver.

The industry standard of an isolated gate driver is shown in Fig. 1a. The power is transferred via a magnetically coupled DC/DC converter while the switching signal is transferred optically. It has become the standard, as optical signal transfer offers high voltage isolation [7]. However, the magnetic core is undesirable because it is bulky when implemented as discrete component and lossy when fully integrated, wherefore capacitively coupled DC/DC converters are in development [8]. Furthermore, optical coupling requires photonic circuitry and can thereby not be implemented in a purely electronic process. Fully optical coupling of power and signal has been demonstrated, but the receivable power in available technologies is low due to photodiode limitations [9].

To circumvent the need of the photonic circuit, in recent years the concept shown in Fig. 1b has gained popularity. Here, the power is transferred magnetically, still, while the switching signal is capacitively coupled. This is done via differential signal coupling [10] or via modulating a megahertz carrier signal [11]. This approach advances the previous one as it can be implemented in a single integrated circuit (IC). However, the power and signal paths are still separated.

In the wireless gate driver concept, as in Fig. 1c, both signal and power are transferred wirelessly via a high frequency carrier signal. In [12] a 10 kHz PWM control signal is sent via a 13.56 MHz carrier frequency with on-off-keying (OOK) modulation while the power is transferred at 10 MHz. In a drive-by-microwave gate driver, two OOK-paths, one active

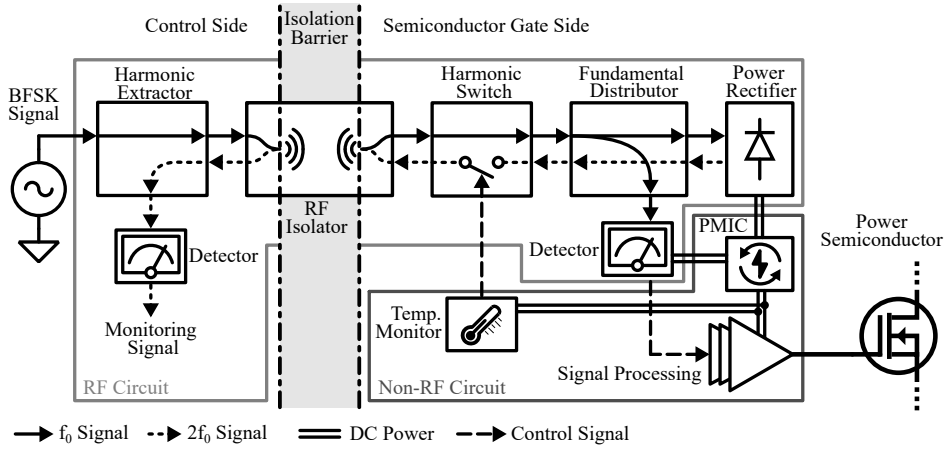


Fig. 2. Block diagram of the proposed radio-frequency gate driver.

at the on-state and the other active at the off-state, were implemented to achieve constant power delivery using an OOK modulated signal at 5.8 GHz [13]. However, in both concepts two paths are needed, either separated by frequency or by independent coupling, which limits the possibilities on how to implement the isolation barrier. In [14], [15] the switching signal is modulated onto a low megahertz frequency carrier using amplitude-shift-keying (ASK) modulation. And in [16] OOK was used on a 1 GHz carrier enabling full integration of the coupler in an IC. The power and signal paths are combined in these implementations. However, the use of signals with varying envelope restricts the available power at the semiconductor gate, as it is depending on the switching signal transferred.

We propose a radio-frequency gate driver (RFGD) using simultaneous wireless information and power transfer (SWIPT) [17] via a single radio-frequency coupling channel and harmonic backscattering. The switching signal is binary frequency-shift-keying (BFSK) modulated onto a 2.4 GHz RF-signal. Both the switching signal and the power are independently recovered at the semiconductor gate and used to switch the semiconductor. Excess recovered power is further used to monitor the semiconductor and a monitoring signal is transferred to the control side by harmonic backscattering. Constant, independent power supply is enabled by the BFSK signal. The RF-signal at a gigahertz frequency enables small component sizes and the possibility for future integration in a single IC. Transferring this signal via a single coupling channel enables the possibility for both near field resonant coupling, as well as far-field radiation implementations.

This work is an expansion of the work presented in [18], where the RF-circuit used for SWIPT and harmonic backscattering is introduced. The rectifier from [19] and the capacitive dualband coupler from [20] are used in this RF-circuit. We expand the previous work by putting the concept to practical use within the RFGD. For this, components of the RF-circuit have been revised or added to enhance the performance within the RFGD application, as described in detail in section III. The non-RF circuit parts needed within the RFGD for power management and signal processing are build up, and the RFGD

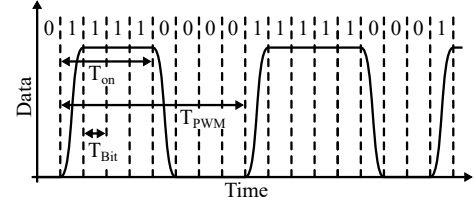


Fig. 3. Approximation of a PWM signal by a fixed data rate binary signal.

is applied to an eGaN-FET in common-drain configuration. The performance of the RFGD is assessed by means of CMTI in this configuration.

II. RADIO-FREQUENCY GATE DRIVER CONCEPT AND SIGNAL THEORY

The block diagram of the proposed radio-frequency gate driver is shown in Fig. 2. The switching signal is modulated onto the RF input signal using BFSK. The on-state of the signal refers to a fixed frequency, f_{on} , and the off-state to a slightly shifted frequency, f_{off} . The frequency swing, Δf , is the difference between the two frequencies, $\Delta f = |f_{off} - f_{on}|$. The signal differs from BFSK signals in communication systems, as it does not have a fixed data rate. Instead, it resembles the switching signal which is pulse-width modulated (PWM) in most power electronic circuits.

This PWM signal can be approximated by a constant data-rate signal, as shown in Fig. 3. For a BFSK signal with a constant duration per bit, T_{Bit} , the modulation index, η_{mod} , is defined as $\eta_{mod} = T_{Bit} \Delta f$. The PWM duty cycle, D , can be calculated from the PWM interval, T_{PWM} , and the on-time, T_{on} , as $D = \frac{T_{on}}{T_{PWM}}$. The minimum duration of T_{Bit} equals the minimum time it takes to switch between f_{on} and f_{off} , which in turn limits the minimum T_{on} regarding the RF-signal path to T_{Bit} . The maximum resolution of the Duty cycle, D , due to the approximation as constant data rate signal is therefore $\frac{T_{PWM}}{T_{Bit}}$ steps.

The BFSK signal is referred to as f_0 signal in Fig. 2. First, it passes through the harmonic extractor and then through the RF isolator. The RF isolator ensures the galvanic isolation between

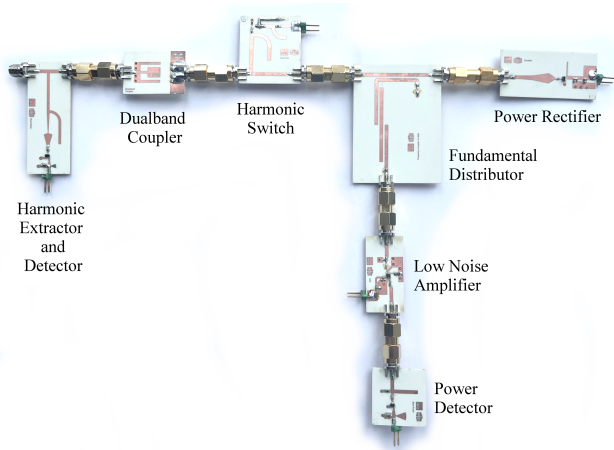


Fig. 4. The RF-structure enabling the RFGD, revised from [18].

the control side and the semiconductor gate side while letting as much of the f_0 and $2f_0$ signals pass as possible. Next, the f_0 signal proceeds through the harmonic switch to reach the fundamental distributor. The fundamental distributor feeds the majority of the f_0 signal to the power rectifier, while coupling a portion of the signal to a detector. At the power rectifier the RF-power is converted to DC-power, which is conditioned by a power management integrated circuit (PMIC). The PMIC feeds the power to all circuit components on the gate side. The switching signal from the detector is processed and amplified to set the proper gate drive voltage of the power semiconductor.

As the power rectifier contains a nonlinear element to enable rectification, it also creates power at integer multiples of the rectified frequencies, the harmonic frequencies. This power is used to enable the transmission of a monitoring signal from the semiconductor gate side to the control side, called $2f_0$ signal in Fig. 2. The $2f_0$ signal contains the second harmonic of both the on- and off-state frequencies, $2f_{on}$ and $2f_{off}$. After generation at the power rectifier it passes through the fundamental distributor to reach the harmonic switch. The harmonic switch either lets the $2f_0$ signal pass or stops it depending on the state of the monitoring signal. The monitoring signal in this case refers to the temperature, but could be any quantity of interest. Thereby, an on-off-keying (OOK) monitoring signal is created which is detected at the control side after passing through the RF-isolator and being routed to the detector by the harmonic extractor. The system components can be separated in two major categories; Components that operate within the RF-signal path and the ones that are not. Components within the RF-signal path need to be taken care of by RF circuit design.

III. RF CIRCUIT DESIGN

The prototype RFGDs components within the RF signal path are shown in Fig. 4. This RF-structure was proposed in [18], demonstrating above 60 % maximum RF-to-DC conversion efficiency, including coupler and filtering components. Four changes were adopted compared to the original design.

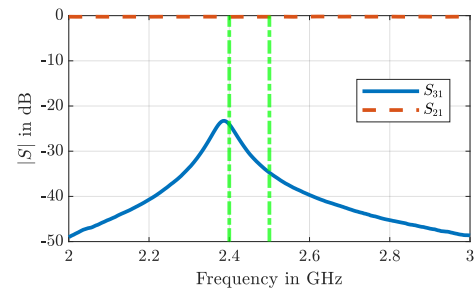


Fig. 5. Measured S-parameters of the fundamental filter.

Firstly, the harmonic switch is based on a PIN-diode and not an RF-switch. Secondly, The harmonic extractor and detector have been united to a smaller design on a single PCB and the filter was revised to avoid the use of a directional coupler. Both revisions will be introduced in section III-D. Furthermore, a smaller size power detector was used and a low-noise amplifier (LNA) added to the fundamental detector path to improve signal quality. This will be further discussed in section IV-C. The system components were designed to operate at 2.4 GHz to 2.5 GHz fundamental f_0 frequency and subsequently 4.8 GHz to 5 GHz harmonic $2f_0$ frequency. All RF-system components were implemented on a Rogers RO4350B substrate with 0.762 mm thickness. The layout and simulation was performed using Keysight Advanced Design Systems. The circuit boards were etched and assembled at the laboratory of the chair of high frequency electronics. S-parameter measurements were performed using a Rohde & Schwarz ZVA50 vector network analyser. The operation of the single RF-components will be introduced in this section.

A. Power Rectifier

The power rectifier used was demonstrated in [19]. It is a diode based rectifier in voltage-doubler configuration matched by a non-uniform transmission line (NUTL). The NUTL allows for uncompromised travel of the generated second harmonic, as necessary for harmonic backscattering, while maintaining a high RF-to-DC conversion efficiency.

B. Dualband Capacitive Coupler

The RF-isolator is realised by a dualband resonant coupler as presented in [20]. For demonstration purposes it is convenient to use this coupler, as it operates based on capacitive resonant coupling and therefore has high efficiency at both the fundamental and harmonic frequencies. However, the RFGD concept is not limited to near field capacitive coupling.

C. f_0 Signal Transmission

The fundamental distributor is comprised of a 20 dB directional coupler followed by a coupled line bandpass filter. Its measured S-parameters are given in Fig. 5. The frequency band of interest from 2.4 GHz to 2.5 GHz is marked. S_{21} gives the ratio of power delivered from the input to the power rectifier and is better than 0.27 dB over the desired bandwidth, ensuring that the majority of RF-power can be converted to DC power

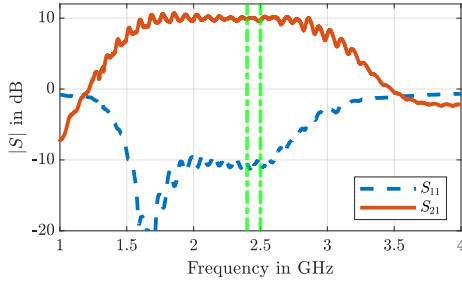


Fig. 6. Measured S-parameters of the low-noise amplifier.

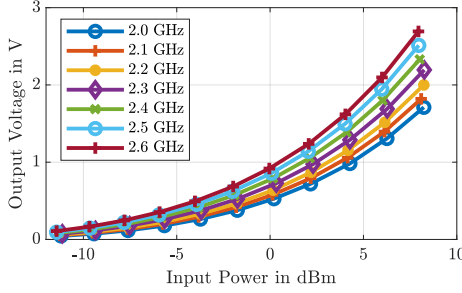


Fig. 7. Measured voltage response of the fundamental detector.

via the power rectifier. S_{31} gives the ratio of power delivered to the low noise amplifier. In the frequency band of interest it is strongly dependent on the frequency. It plunges from 24.0 dB at 2.4 GHz to 34.7 dB at 2.5 GHz. This frequency dependence enables the demodulation of the BFSK signal.

The low-noise amplifier is a MAX2644EXT chip by Analog Devices on a custom PCB. The measured S-parameters are given in Fig. 6. It has a gain, S_{21} , of about 10 dB in the band of interest and is matched, S_{11} , to about 10 dB.

The power detector uses BAT15 diodes from Infineon in voltage doubler configuration. The matching is done as a combination of resistive and transmission line matching for a broadband behaviour, the measurements show better than 16 dB matching over the desired band. The measured voltage response of the fundamental detector is given in Fig. 7. The power detectors output voltage is directly proportional to the input power, but only slightly depending on the input frequency within the desired band.

Combining the behaviour of the power detector, the fundamental distributor and the low-noise amplifier, the output voltage is highly dependent on the frequency, whereby the BFSK signal can be effectively demodulated.

D. $2f_0$ Harmonic Backscattering

The $2f_0$ harmonic signal at 4.8 GHz to 5 GHz generated by the power rectifier is modulated by the harmonic switch, as shown in Fig. 8. The principle is the same as in [18], where it is explained in detail. The difference lays within the generation of the open or short at the intersection of the transmission lines 2 and 3. Instead of an RF-switch a PIN-diode is used. The modulation signal forward biases the diode via the current limiting resistor, R_i , when the signal is high, creating an RF-short. When it is low, the diode behaves like

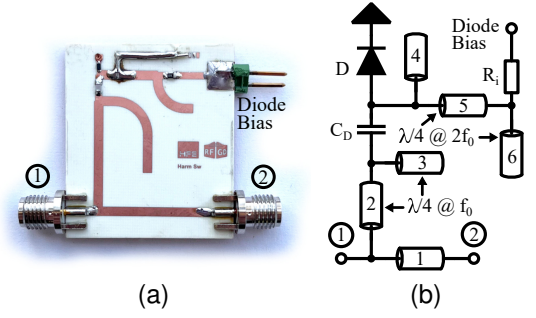


Fig. 8. Harmonic switch (a) photograph and (b) schematic.

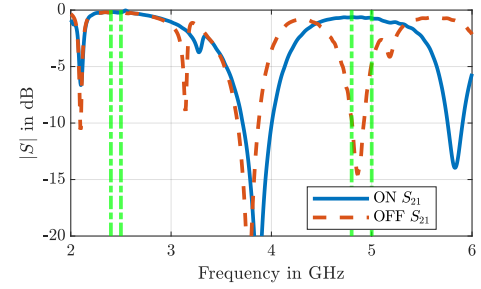


Fig. 9. Measured S-Parameters of the harmonic switch.

an open. The modulation signal input is RF-isolated at $2f_0$ by the transmission lines 5 and 6. The capacitor, C_D , blocks the DC-component of the modulation signal and transmission line 4 ensures proper matching of the desired impedances. The use of the diode instead of an RF-switch has several benefits. Firstly, the control voltage suffices to drive the switch and no additional supply voltage is needed. Secondly, the diode can be switched faster than the RF-switch. Also, the measurements show improved switching behaviour, as given in Fig. 9. The measured transfer behaviour between the two RF-connectors, $|S_{21}|$, is given over frequency for on- and off-state. Within the fundamental band, the behaviour does not change between on- and off-state to not disturb the transmission of the f_0 signal. At the harmonic band, the frequency band over which more than 10 dB difference between on- and off-state is achieved is broadened compared to the implementation using an RF-switch, leading to a stronger $2f_0$ modulation.

On the control side, the OOK modulated harmonic signal is detected by the harmonic extractor shown in Fig. 10. Lines 3 and 4 ensure that none of the f_0 signal and only $2f_0$ frequency components reach the diode based detector circuit. The diode based detector is optimised for a strong voltage response at the harmonic frequency by utilising an NUTL and uses BAT15 diodes from Infineon in voltage doubler configuration.

The measured output voltage over input power and frequency of the harmonic detector while one RF-connector is terminated to the 50Ω system impedance is given in Fig. 11a. At the f_0 band the maximum output voltage measured is 1 mV which indicates that the rejection of the fundamental frequency by lines 3 and 4 is working. At the $2f_0$ band the output voltage follows the input power as desired. It is also dependent on the harmonic frequency, which is not desirable as we do not want

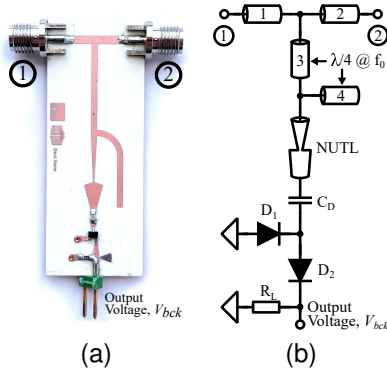


Fig. 10. Harmonic extractor (a) photograph and (b) schematic.

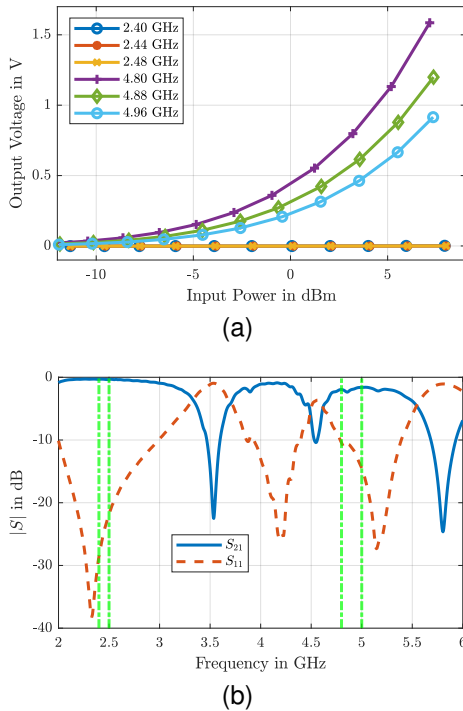


Fig. 11. Measured harmonic extractor (a) output voltage and (b) S-parameters.

to detect the fundamental BFSK signal but the harmonic OOK signal. However, a change of 10 dB in the harmonic signal amplitude causes a larger voltage swing than the frequency shift over the entire desired band, which indicates that proper operation of the detector within the RFGD system can be expected. The measured S_{21} between the two RF-connectors, given in Fig. 11b, confirms that the feedthrough of the f_0 signal is given as it only causes a maximum of 0.36 dB of losses. S_{11} confirms that the circuit is matched better than 20 dB at the fundamental. At $2f_0$, the circuit is matched to about 10 dB and causes about 2 dB of losses, which is fine as the $2f_0$ signal chain anyway ends at the harmonic extractor.

IV. GATE DRIVER DEMONSTRATION

The non-RF circuit components are implemented on an FR-4 PCB as shown in Fig. 12. The LTC3130 Buck/Boost converter by Analog Devices is used as PMIC connected to

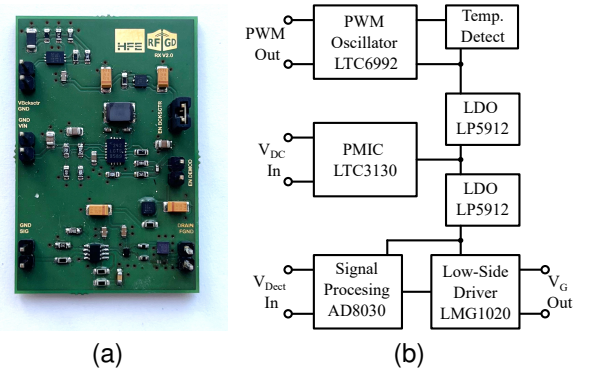


Fig. 12. (a) Photograph and (b) block diagram of the non-RF circuit.

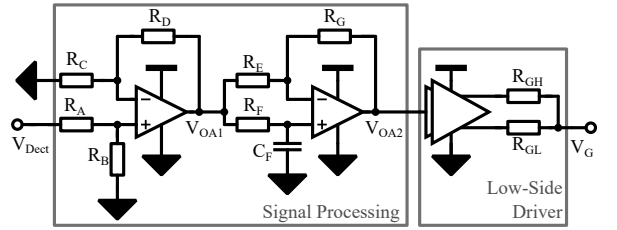


Fig. 13. Signal processing circuit schematic.

TABLE I
SIGNAL PROCESSING CIRCUIT COMPONENT VALUES.

R_A	R_B	R_C	R_D	R_E	R_F	R_G	R_{GL}	R_{GH}	C_F
30 k Ω	10 k Ω	10 k Ω	60 k Ω	0.1 Ω	20 k Ω	1 M Ω	1 Ω	5 Ω	1 μ F

the output of the power rectifier. It is chosen because of its wide input voltage range, high power conversion efficiency over a wide power range and the ability of maximum power point (MPP) control. The MPP control is enabled by setting the minimum input voltage at the PMIC. This is beneficial, as it sets the output voltage of the power rectifier and therefore enables efficient operation of the power rectifier. The output of the PMIC is filtered by two LP5912 LDOs from Texas Instruments to generate the 5 V rails for the other circuit components. The gate drive signal, V_G , is generated from the detector voltage using op-amp signal processing with Analog Devices' AD8030 and a low-side driver LMG1020 by Texas Instruments, the signal processing will be presented in section IV-A. The monitoring signal is generated by an LTC6992 PWM oscillator from Analog Devices which sets the duty cycle of the PWM by an external reference voltage. This voltage is generated using a temperature dependent resistor in a resistive voltage divider. Thereby, the capability to backscatter a real monitoring signal is demonstrated.

A. Signal Processing

In Fig. 13 the signal processing circuit schematic is given. The component values are listed in table I and examples of measured voltages at marked nodes in the circuit are plotted in Fig. 14. The first op-amp of the AD8030 chip is connected as a non-inverting amplifier, its output, V_{OA1} , is a slightly

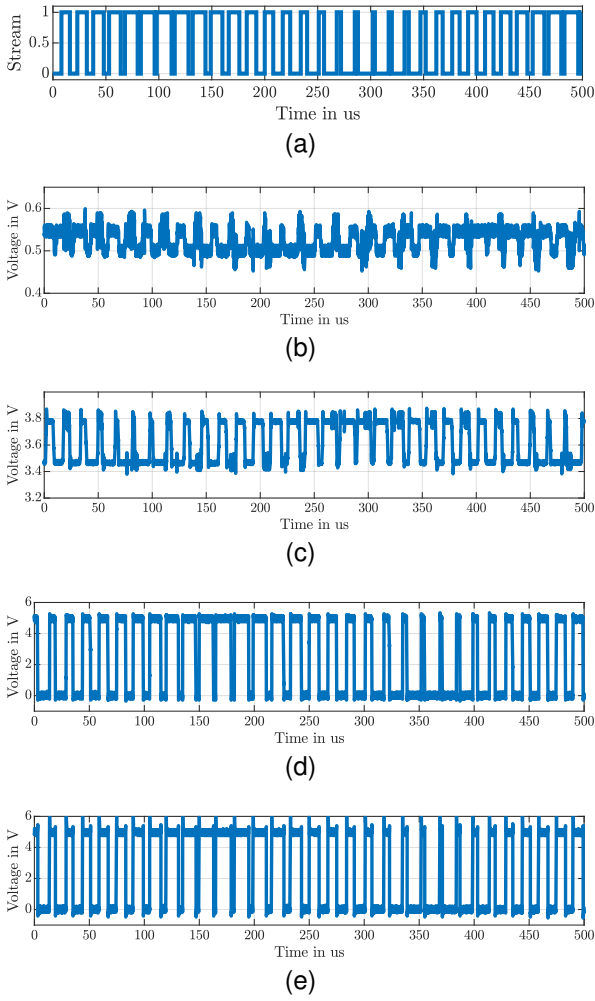


Fig. 14. (a) Input binary signal and measured voltages along the signal processing chain: (b) V_{Dect} , (c) V_{OA1} , (d) V_{OA2} , and (e) V_G .

amplified version of the input signal V_{Dect} . The second op-amp is used as comparator with hysteresis. It compares V_{OA1} to a low-pass filtered version of V_{OA1} created by the RC-Filter composed of R_F and C_F . Therefore the two control voltage states, 0 V and 5 V, are defined at this point. The output, V_{OA2} , is the input for the low-side driver LMG1020. Its output is fed through the damping resistors, R_{GH} and R_{GL} , to the power semiconductor gate as V_G .

B. System Measurements

In all time domain measurements a 62.5 kHz PWM switching signal is used that modulates a 2.6 kHz sine wave, as is shown in Fig. 14a. The measured gate voltage in Fig. 14e gives a maximum duty cycle deviation of 0.2 μ s, which occurs at the minimum duty cycle. The BFSK signal containing this PWM switching signal has a power of 24.5 dBm (282 mW) and its normalised spectrum, measured by a Rohde & Schwarz FSW, is given in Fig. 15. Its parameters according to section II are given in table II. It is the only source of power for the RFGD, including RF and non-RF-circuit, and therefore equals the entire power consumption. The backscatter modulator is set to 10 kHz PWM frequency.

TABLE II
BFSK SIGNAL PARAMETERS.

f_{on}	f_{off}	Δf	T_{Bit}	η_{mod}	T_{PWM}	$T_{on,min}$
2.4 GHz	2.41 GHz	10 MHz	1 μ s	10	16 μ s	2 μ s

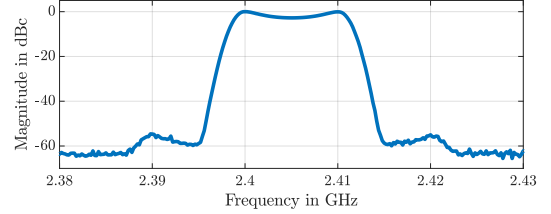


Fig. 15. Measured Spectrum of the BFSK signal.

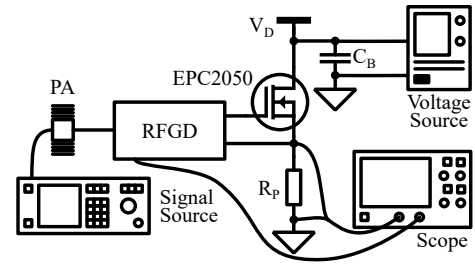


Fig. 16. Measurement setup for gate control of an eGaN-FET using the RFGD.

The circuit used to test the RFGDs gate drive capability is given in Fig. 16. An EPC2050 enhancement mode GaN FET by Efficient Power Conversion is connected in a common drain configuration. The EPC2050, rated at 400 pF input capacitance in the datasheet, is driven by the RFGD, which includes R_{GL} and R_{GH} given in Table I as gate resistors. The drain voltage, V_D , is supplied by a Keithley 2260B-80-27 voltage source and buffered by an electrolytic capacitor, C_B . R_P is a high power resistor of 75 Ω . The RF-signal is supplied to the RFGD by a SMBV100A signal source from Rohde & Schwarz and amplified by a ZVE-8G+ power amplifier from Mini-Circuits. A LeCroy waveRunner 62Xi two-channel oscilloscope is used to measure the voltage across R_P , V_{Rp} , as well as the backscattered monitoring signal, V_{bck} .

The time domain waveforms for $V_D = 20$ V are given in Fig. 17. V_D shows that the f_0 signal path operates as expected. The $2f_0$ backscattered signal is visible in V_{bck} , it is however submerged in the noise of switching transients, creating short peaks in the signal. At this V_D -level the PWM signal will still be detectable by applying proper filtering.

Raising V_D to 84 V yields the waveforms given in Fig. 18. Still, the f_0 switching signal works as expected. The backscatter signal, however, is barely distinguishable in the V_{bck} measurement.

C. CMTI Considerations

The CMTI of a isolated gate driver is assessed by applying a pulse between the ground of the control voltage and the reference voltage on the semiconductor gate side. This pulse

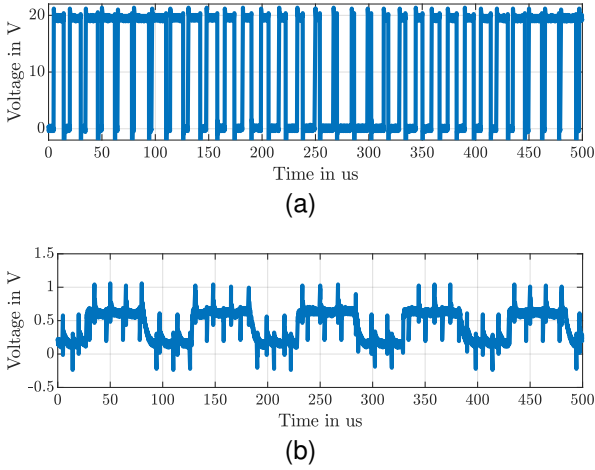


Fig. 17. Measurement results at $V_D = 20$ V: (a) V_{Rp} (b) V_{bck} .

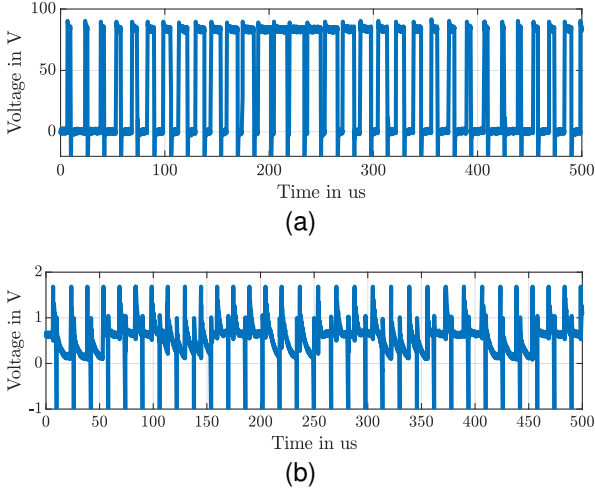


Fig. 18. Measurement results at $V_D = 84$ V: (a) V_{Rp} (b) V_{bck} .

is created by the voltage across the resistance R_p as shown in Fig. 16 for the considerations in this section.

During the measurements presented in section IV-B it becomes clear that the f_0 signal path and the backscattering $2f_0$ signal path have strongly differing CMTI. The switching transients of V_{Rp} at $V_D = 20$ V and $V_D = 84$ V are given in Fig. 19.

From this, the $\frac{dv}{dt}$ transients can be calculated from the 10% to 90% rise and fall times. At $V_D = 20$ V the rise time is 8.2 ns giving $\frac{dv}{dt}_{20, \text{rise}} = \frac{16 \text{ V}}{8.2 \text{ ns}} = 1.9 \frac{\text{kV}}{\mu\text{s}}$ and the fall time 20.8 ns wherefore $\frac{dv}{dt}_{20, \text{fall}} = 0.7 \frac{\text{kV}}{\mu\text{s}}$. At $V_D = 84$ V the transients are expectedly higher with $\frac{dv}{dt}_{84, \text{rise}} = 6.4 \frac{\text{kV}}{\mu\text{s}}$ and $\frac{dv}{dt}_{84, \text{fall}} = 5.6 \frac{\text{kV}}{\mu\text{s}}$. While $\frac{dv}{dt}_{20, \text{rise}} = 1.9 \frac{\text{kV}}{\mu\text{s}}$ is approximately the maximum CMTI at which the backscattered signal could still be detected by proper filtering, the fundamental signals CMTI probably exceeds $\frac{dv}{dt}_{84, \text{rise}} = 6.4 \frac{\text{kV}}{\mu\text{s}}$. Due to the limitation of the maximum V_D available from the voltage source used, no faster transients could be tested, even though the circuit limitations were not reached.

The higher CMTI within the fundamental signal path suggests that the LNA in the signal chain is of great impor-

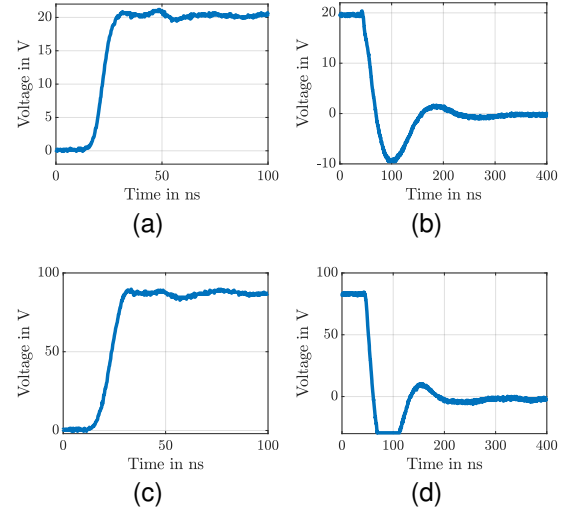


Fig. 19. Measured V_{Rp} voltage transients: (a) rising and (b) falling edge at $V_D = 20$ V, and (c) rising and (d) falling edge at $V_D = 84$ V.

tance to the CMTI. Which was confirmed by measuring the performance of the f_0 signal path without the LNA. In this configuration, the fundamental signal chain only worked reliably up to $V_D = 30$ V. An LNA is the first receiver stage in RF-communication systems to keep the signal to noise-ratio (SNR) as low as possible. In the RFGD it fulfills the same purpose. The dv/dt -transients from switching the power semiconductor translate to noise in the signal path, which is spread across the frequency spectrum. Therefore, the increase in SNR by addition of the LNA directly translates to an increase in CMTI. The receiver SNR will therefore dictate the CMTI in the RFGD.

This applies in both the fundamental f_0 and the $2f_0$ backscattering path. The introduction of the harmonic extractor, as discussed in section III-D, gives a high received voltage level, however it does not improve the SNR. It is expected that the addition of an LNA to this circuit will yield significant improvements in the resilience of the $2f_0$ backscattering receiver path.

The CMTI performance is furthermore expected to be determined by the RF-isolator. The dualband capacitive resonant coupler behaves quite broadband, as can be seen in [20]. As the voltage slopes that determine the CMTI cause disturbances at various frequencies spread across the spectrum, designing a more narrowband coupler with better band-pass filtering properties can be expected to improve the CMTI performance further. This could also be achieved by using far field transmission through dualband antennas.

V. CONCLUSION AND OUTLOOK

In this work a prototype for the RFGD is presented. It is proven that an isolated gate driver enabled by SWIPT and harmonic backscattering works. The CMTI implications of the design are discussed. It can be expected that two main factors will determine the CMTI of the RFGD. Firstly, the SNR of the signal chain and, secondly, the design of the RF-isolator. The maximum CMTI capability of the RFGD

prototype still has to be tested, as the given value is a lower boundary approximation. For this, an LNA should be added to the backscattered signal reception path and a high voltage source is needed for measurement.

To demonstrate the RFGD concept concisely, the full capability of the concept could not yet be exploited. The voltage on the receiver side is 5 V which is suitable for eGaN-FETs, higher voltages to enable other types of devices, e.g. 15 V for SiC MOSFETs, can be realised with small adjustments to the receiver design. The BFSK signal is generated by a signal source in this work. It can be replaced by a voltage controlled oscillator (VCO), which switches between two frequencies, followed by a power efficient switch mode amplifier. It is obvious that the prototype has comparatively large dimensions for an isolated gate driver. However, all components can be realised equivalently in an RFIC in the future leading to a compact design. The single RF coupling channel opens up the possibility of transmitting over far field, which means that the maximum voltage isolation capability could be easily increased together with the CMTI. Furthermore, the signal chain is not limited to BFSK modulation. Any constant envelope modulation scheme is thinkable, and a more advanced RF signal chain can aid the CMTI as well as enabling control of multiple devices.

Furthermore, the SWIPT and harmonic backscattering RF-structure is not limited to be used as isolated gate driver but can be imagined to enable efficient unidirectional power and bidirectional signal transmission through an isolation barrier in many more applications.

REFERENCES

- [1] M. Buffolo, D. Favero, A. Marcuzzi, C. De Santi, G. Meneghesso, E. Zanoni, and M. Meneghini, "Review and outlook on GaN and SiC power devices: Industrial state-of-the-art, applications, and perspectives," *IEEE Trans. on Electron Devices*, vol. 71, no. 3, pp. 1344–1355, 2024.
- [2] D. J. Perreault, J. Hu, J. M. Rivas, Y. Han, O. Leitermann, R. C. Pilawa-Podgurski, A. Sagneri, and C. R. Sullivan, "Opportunities and challenges in very high frequency power conversion," in *24th Annu. IEEE Applied Power Electron. Conf. and Expo.*, 2009, pp. 1–14.
- [3] T. Ivaniš and M. Kovačić, "An overview of advanced gate driver concepts for SiC semiconductors," in *Int. Conf. on Elect. Drives and Power Electron. (EDPE)*, 2023, pp. 1–9.
- [4] X. Mu, Y. Jiang, M.-K. Law, P.-I. Mak, and R. P. Martins, "Design challenges and considerations of non-isolated gate driver for GaN-based converters," in *IEEE 65th Int. Midwest Symp. on Circuits and Syst. (MWSCAS)*, 2022, pp. 1–4.
- [5] X. Zan and A.-T. Avestruz, "Isolated ultrafast gate driver with variable duty cycle for pulse and VHF power electronics," *IEEE Trans. on Power Electron.*, vol. 35, no. 12, pp. 12 678–12 685, 2020.
- [6] F. Vladoianu, M. Strelec, I. Hurez, V. Anghel, and G. Brezeanu, "Changing the paradigm in common mode transient immunity (CMTI) testing of dual channel galvanically isolated gate drivers," in *2021 Int. Semicond. Conf. (CAS)*, 2021, pp. 125–128.
- [7] D. Colin and N. Rouger, "High speed optical gate driver for wide band gap power transistors," in *IEEE Energy Convers. Congr. and Expo. (ECCE)*, 2016, pp. 1–6.
- [8] J. Tang, L. Zhao, and C. Huang, "A 68.3% efficiency reconfigurable 400-/800-mW capacitive isolated DC-DC converter with common-mode transient immunity and fast dynamic response by through-power-link hysteretic control," in *IEEE Int. Solid-State Circuits Conf. (ISSCC)*, vol. 65, 2022, pp. 1–3.
- [9] H. Fujita and M. Ishigaki, "A resonant gate-drive circuit with optically-isolated control signal and power supply for fast-switching and high-voltage power semiconductor devices," in *Int. Power Electron. Conf. (ECCE ASIA)*, 2010, pp. 1895–1901.
- [10] V. Lorentz, R. Schwarz, T. Heckel, M. März, and L. Frey, "Integrated galvanically isolated MOSFET and IGBT gate-driver circuit with switching speed control," in *41st Annu. Conf. of the IEEE Ind. Electron. Soc. (IECON)*, 2015, pp. 000 544–000 549.
- [11] I. Khan, M. Zhang, and N. Llacer, "Signal transmission for an isolated gate driver using capacitive coupling technique," in *19th IEEE Int. New Circuits and Syst. Conf. (NEWCAS)*, 2021, pp. 1–4.
- [12] J.-J. Lin, "Design and implementation of a wirelessly powered and controlled gate driver," *IEEE J. of Radio Freq. Identification*, vol. 5, no. 1, pp. 40–45, 2021.
- [13] S. Nagai, N. Negoro, T. Fukuda, N. Otsuka, H. Sakai, T. Ueda, T. Tanaka, and D. Ueda, "A DC-isolated gate drive IC with drive-by-microwave technology for power switching devices," in *IEEE Int. Solid-State Circuits Conf. IEEE*, 2012.
- [14] K. Sung, Y. Ouchi, S. Amagai, K. Sagara, Y. Kawasaki, and H. Sung, "Gate drive method using wireless multiplex transmission of power and signal," in *Int. Power Electron. Conf.*, 2022, pp. 2082–2087.
- [15] V. T. Nguyen, G. Veera Bharath, and G. Gohil, "Design of isolated gate driver power supply in medium voltage converters using high frequency and compact wireless power transfer," in *IEEE Energy Convers. Congr. and Expo. (ECCE Asia)*, 2019, pp. 135–142.
- [16] J. Lin and K.-C. Lin, "A monolithic isolated gate driver with on-chip transformer," in *IEEE Int. Future Energy Electron. Conf. and ECCE Asia*, 2017, pp. 1256–1259.
- [17] J. Hu, Y. Zhao, and K. Yang, "Modulation and coding design for simultaneous wireless information and power transfer," *IEEE Commun. Mag.*, vol. 57, no. 5, pp. 124–130, 2019.
- [18] L. Hüssen, M.-D. Wei, and R. Negra, "Power efficient simultaneous wireless information transceiption and power reception through an isolation barrier," in *IEEE Wireless Power Technol. Conf. and Expo (WPTCE)*, 2024, pp. 872–875.
- [19] —, "Broadband high-efficiency microwave rectifier with nonuniform transmission-line input matching for harmonic backscattering applications," in *IEEE Int. Microw. Symp.*, 2024, pp. 118–121.
- [20] —, "Design and synthesis of a dualband capacitive resonant coupler for harmonic backscattering through an isolation barrier," in *Eur. Microw. Conf.*, 2024, accepted for presentation.



Lukas Hüssen (Graduate Student Member, IEEE) received the B.Sc. and M.Sc. degrees in electrical power engineering from RWTH Aachen University, Aachen, Germany, in 2017 and 2019, respectively. He is currently pursuing the Ph.D. degree at the Chair of High Frequency Electronics, RWTH Aachen University, Aachen, Germany.

His current research interests include the interaction between high-frequency electronics and neighbouring electrical engineering domains, such as power electronics.



Muh-Dey Wei received the M.Sc. and Ph.D. degrees in electrical engineering from National Chung Cheng University, Chiayi, Taiwan, in 2006 and 2011, respectively.

During his Ph.D., he spent 18 months with Mixed Signal CMOS Circuits (MSCC) and Chair of Integrated Analog Circuits and RF Systems (IAS), RWTH Aachen University, Aachen, Germany, where he is currently a Chief Engineer with the Chair of High Frequency Electronics (HFE). His current research interests are wireless power transfer, rectifiers, 2-D material, advanced transceivers, microwave frequency sources, and power amplifiers.

Dr. Wei was a recipient of the Deutscher Akademischer Austauschdienst (DAAD) Scholarship.



Renato Negra received the M.Sc. degree in telematics from Graz University of Technology, Graz, Austria, in 1999, and the Ph.D. degree in electrical engineering from ETH Zurich, Zurich, Switzerland, in 2006, with a focus on power-efficient linear amplification of wireless communication signals.

During his M.Sc., he spend one year as an Exchange Student at the Norwegian University of Science and Technology (NTNU), Trondheim, Norway. From 1998 to 2000, he was with Alcatel Space Norway AS (now NorSpace AS), Horten, Norway,

where he was involved in the design and characterization of space-qualified RF equipment. In 2000, he joined the Laboratory for Electromagnetic Fields and Microwave Electronics, ETH Zurich. From 2006 to 2008, he was a Post-Doctoral Fellow with the iRadio Laboratory, University of Calgary, Calgary, AB, Canada, working on switching-mode power amplifiers and advanced wireless transmitters. From June 2008 to December 2013, he was an Assistant Professor of mixed-signal CMOS circuits with the Ultra High Speed Mobile Information and Communication (UMIC) Research Centre, RWTH Aachen University, Aachen, Germany, where he has been with the Chair of High Frequency Electronics, since December 2013. His research interests are high-frequency circuits and systems in both silicon, III–V, and 2-D-material technologies. Particular interests include power amplifiers and transmitter architectures.