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2017

Cai, A. Q., & Siek, L. (2017). A 2-kW, 95% Efficiency Inductive Power Transfer System Using Gallium Nitride Gate Injection Transistors. IEEE Journal of Emerging and Selected Topics in Power Electronics, 5(1), 458-468.

<https://hdl.handle.net/10356/83228>

<https://doi.org/10.1109/JESTPE.2016.2632743>

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A 2kW, 95% Efficiency Inductive Power Transfer System Using Gallium Nitride Gate Injection Transistors

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Abstract—This paper presents an Inductive Power Transfer (IPT) system targeting at Electric Vehicles (EV) and Hybrid Electric Vehicles (HEV). IPT systems provide significant benefits over conventional plug-in chargers. However, in order for IPT to be adopted for EV charging, efficiency is a key Figure of Merit (FOM) which needs to be achieved. This paper presents a high frequency inverter using Gallium Nitride (GaN) power transistors which have the benefit of low on-resistance and gate charge to reduce the switching and conduction loss. The switching characteristics of the GaN GIT are studied and the inverter is designed to ensure low switching losses, while keeping overshoot and slew rates under control. An efficiency centric mode of operation is proposed to improve the efficiency of the system, while ensuring sufficient power transfer. The system efficiency peaks at 95% at 100 kHz operation and 92% at 250 kHz operation for a coil gap of 80mm at 2kW output power. At a coil gap of 150mm, the system obtains above 90% efficiency at 1.3 kW. The inductive power transfer system is compared to a similar system using SiC power transistors and outperforms it by 1% at 2kW.

Keywords—Enhancement mode GaN, Gate Injection Transistor, Electric vehicles, Wireless power transfer, Inductive power transfer.

I. INTRODUCTION

Increased environmental awareness on the effects of global warming has spurred the emergence for sustainable technologies leading to innovations in Electric Vehicles (EVs) and Hybrid Electric Vehicles (HEVs). Compared to traditional Internal Combustion Engine (ICE) cars, EVs and HEVs have the advantage of higher energy conversion efficiency, less pollution and smoother engine performance [1], [2]. However, EVs and HEVs are limited by battery related issues such as limited driving range and long charge time [2]. In addition, car batteries are big and bulky [3]. This has prompted researchers to explore alternative to battery charging such as Inductive Power Transfer (IPT) and battery swapping.

Inductive Power Transfer (IPT), a method of wirelessly transferring power using magnetic resonance coupling, has been emerging over the last few years as a convenient and safe charging alternative for EV and HEV users [4]. The pair of transmitting and receiving coils is tuned to resonate at the same frequency using a compensation network. However, IPT encounters efficiency challenges compared to conventional plug-in charging due to losses in the coil [4]. Therefore, in

order to facilitate easy IPT charging for EVs and HEVs, focus needs to be placed on improving the efficiency of the IPT system.

In the recent years, wide-bandgap semiconductor devices such as Gallium Nitride (GaN) Gate Injection Transistor (GIT) [5] had been gaining attention due to their lower gate charge and on-state resistance which translates to lower switching and conduction loss compared to conventional Silicon (Si) power transistors [6]. Commercial normally off GaN samples are available as enhancement mode GaN by manufacturers such as Panasonic, GaN systems and Efficient Power Conversion (EPC), or Cascode mode GaN developed by Transphorm and International Rectifiers [7].

The advantage of adopting GaN GIT comes in terms of higher efficiency, higher reliability and power density. With lower switching and conducting losses, efficiency of the power electronics can be increased. The lower losses in the semiconductor device allow utilization of smaller and cheaper heat dissipation components [8]. The reliability of the system will improve because the stress on the components due to temperature is reduced [8]. Finally, the lower gate charge of GaN GIT allows higher frequency operation, resulting in smaller passive filter components and increased power density.

Recent work has showcased GaN power transistor application in various power electronics such as high frequency resonant converters [9], low power integrated DC-DC converters [10] and inverter application for wireless power transfer [11], photovoltaic applications [8] and motor driver applications [12], [13].

This paper is presented in 7 sections. Section II will present a description of the system. Section III will explain the GaN GIT driver design. Section IV will study the characteristics of the compensation. Section V will discuss on the efficiency centric mode of operation. Section VI will illustrate the experimental results and finally the conclusion of this work.

II. SYSTEM DESCRIPTION

A. Overall system

The EV/HEV wireless charging system is illustrated in Fig.1. Power drawn from the AC grid is rectified into a DC voltage. Since the application draws greater than 75W, a

Fig. 1. Wireless charging system

Power Factor Correction (PFC) stage is necessary to improve the power factor. A high frequency inverter converts the DC voltage into AC voltage so that it can be transmitted using the magnetic resonant wireless power transfer coil. On the secondary side, the output of the resonant tank is connected to a SiC diode rectifier which rectifies the AC signal into a DC voltage.

The focus of this work is on the IPT system and its ability to efficiently transmit power. This paper proposes a GaN GIT based high frequency power full bridge inverter which converts the DC power into an AC square wave using Phase Shift Modulation (PSM). PSM operates in a way that the two power transistors in the same leg are operated at 50% duty cycle at 180° phase shift from each other and the two legs of the inverter operate at certain phase differences which vary the duty cycle of the output voltage of the inverter [14]. By varying the phase difference, the inverter can vary the output power.

This high frequency IPT inverter is able to achieve soft-switching because of the LC resonant tank on the power train. Zero voltage Turn-on (ZVT) can be achieved in the inverter by operating the inverter's switching frequency above the resonant frequency of the LC resonant tank, resulting in the current lagging the voltage due to an inductive load is present.

In this work, ZVT was not used when designing the system because coil efficiency deteriorates as we move away from the resonant frequency. The inverter is operate at resonant frequency, which results a resistive reflected load (hard-switching). The resonant tank results in the GaN GIT in the inverter to be switched on and off when the drain to source current is small which reduces the turn-on and turn-off loss. So while we do not have a ZVT, we managed to cut down the switching losses. The rectifier works at Zero Current Switching (ZCS), due to the sine wave current from the coils.

In order to ensure high efficiency, GaN GITs are introduced into the high frequency inverter to reduce switching and conduction loss. Optimization of the GaN GIT gate drive design will be explained in detail in section III. The IPT coil's characteristics vary with the coupling and the load. Hence, a novel mode of operation is proposed to optimize the efficiency for the required amount of power which will be explained in section V.

B. Coil Parameters

Coil design considerations are to ensure good efficiency, high power transfer [15], while being light weight. The power

TABLE I. IPT COIL AND CAPACITOR SPECIFICATIONS

Specifications	100kHz	250kHz
Internal radius	75mm	75mm
External radius	150 mm	150 mm
Type of wire	210/33AWG	210/33AWG
Number of turns	28	21
Compensation type	Series-Series	Series-Series
Primary and secondary inductance	234μH	120μH
Primary and secondary capacitance	11nF	3.29nF

Fig. 2. IPT coil setup

is transferred across the coils using magnetic resonance. The coil adopts a series-series configuration where the capacitor and inductor on the primary and secondary sides are connected in series. Two sets of resonant coil and capacitors are designed for 2 different operating frequencies of 100 kHz and 250 kHz operation, and their specifications are shown in Table I. Figure 2 shows the experimental setup. The compensation characteristics will be further explained in chapter IV.

III. HIGH SLEW RATE GATE DRIVER OPTIMISATION

A. Power device selection

A common Figure of Merit (FOM) adopted by power semiconductor devices is $R_{on}Q_g$ that accounts for the switching and conduction loss, and is representative of the material and

TABLE II. FOM COMPARISON BETWEEN SEMICONDUCTOR DEVICES

Material	GaN	Si	SiC	SiC
Technology	Gate Injection Transistor	Super Junction MOSFET	MOSFET	MOSFET
Breakdown voltage (V)	600	700	650	1200
Rated current (A)	15	18	29	40
R_{on} (mΩ)	71	125	120	80
Q_g (nC)	8	35	61	106
$R_{on}Q_g$ (nΩC)	0.568	4.38	7.32	8.48
Q_{rr}/Q_{Ds} (nC)	53	7000	53	60

technology [16]. A comparison of FOM among state-of-the-art transistors using GaN, SiC and Si are compared and shown in Table II. GaN GIT has the lowest FOM due material and HEMT characteristics.

Reverse recovery charge is another specification important for bridge related topologies as it results in reverse recovery loss and increases dead time requirements. Due to the absence of reverse recovery diode, GaN devices do not have a Q_{rr} , but it still needs to remove charge originating from C_{DS} . From Table II, the GaN GIT has a Q_{DS} approximately 70 times lower than Si SJ MOSFET. The GaN GIT has comparable values with SiC MOSFET even though the GaN device is tested at more than 2 times I_{DS} slew rate ($di_{DS}/dt = 400A/\mu s$ (GaN) vs. $di_{DS}/dt = 150A/\mu s$ (SiC)) as Q_{rr}/Q_{DS} directly proportionate to the I_{DS} slew rate [17].

B. GaN GIT gate driver design

The high frequency inverter is driven by 4 GaN GITs in a full bridge configuration. The half bridge section of the GaN GIT driver circuit used in this project is shown in Fig. 3 using 600V enhancement mode GaN GIT in TO-220 package. The GaN GIT is a non-insulated gate device which requires a forward current gate during conduction to maintain low on-resistance. Another concern about enhancement mode devices is noise margin due to low threshold voltage. Therefore, the divided voltage capacitor type gate drive circuit is used to control the gate drive current and negative voltage turn off which creates a larger noise margin to prevent false turn on [18].

Selection considerations for isolated gate driver ICs are Common Mode Transient Immunity (CMTI), voltage isolation and driving current capability. The GaN GIT is driven by ADum3223 isolated half bridge driver which has a source and sink capability of 4A, 3kV isolation and CMTI capability of 50V/ns. For this paper, we choose to set the slew rate to 50V/ns to suit the requirements of the isolated gate driver, which was the state of the art during the time of this project.

The functionality of the RC gate drive circuit will be explained with respect to the high side GaN GIT driver section of Fig. 3. $R2a$ serves to control the forward gate current of the GaN GIT during turn on conduction and is decided using (1). The supply voltage for the gate driver is denoted by VDD . The forward gate current of the GaN GIT is set to 20mA for this setup. Resistor $R1a$ is designed to set the slew rate of the GaN GIT and is designed based on (2). The resistor sizing for $R1a$ is smaller than $R2a$ by at least an order of 1. This is to supply a large current to turn on the device fast. The capacitor $C1a$ serves to separate the high current route and the low current route and to enable negative voltage turn off. $C1a$ is designed using (3). Passive components $C1b$, $R1b$ and $R2b$ perform the same function but for the low side GaN GIT.

$$R2a = \frac{VDD - V_{gs,forward}}{I_{gs,forward}} \quad (1)$$

$$R1a = \frac{(VDD - V_{gs,plateau}) \times V_{powersupply}}{\frac{dV}{dt} \times Q_{gd}} \quad (2)$$

$$\frac{Q_{g,forward}}{VDD - V_{gs,forward}} < C1a < \frac{Q_{g,max}}{VDD - V_{gs,forward}} \quad (3)$$

Fig. 3. Half bridge GaN GIT driver schematic

Fig. 4. GaN GIT RC gate driver functionality

Fig.4 shows the time chart for the RC gate drive circuit. When the GaN device is turned on at $T1$, current flows through $R1a$ and $R2a$, but a larger amount of current flows through $R1a$ due to the lower resistance. When the device is fully charged up at $T2$, the capacitor $C1a$, which is connected in series with $R1a$, blocks the current flow. This results in current flowing through $R2a$ that keeps the device in conduction.

When the device is turned off at $T3$, the positive end of the capacitor is connected to GND and the negative end is connected to the gate. This results in negative voltage turn off, which allows for fast turn off and creates a voltage margin in case of false turn on. We designed the negative turn off voltage at -4 volts.

C. Half bridge circuit loss modelling

The losses in a half bridge circuit constitute conduction loss, switching loss, ringing loss and dead time loss as shown in equation (4), using a simplified model which studies the losses like buck type converter. The total conduction loss and switching loss can be shown (5) and (6), respectively. Reduction in conduction loss is limited by device parameters R_{on} and application requirement I_o . Reduction in switching loss can be achieved by increasing the slew rate to reduce rise and fall time of the GaN GIT.

The ringing loss is obtained from [19] and modified for GaN GIT as shown in (7), which does not have Q_{rr} but still has

to discharge C_{DS} , which is small compared to Q_{oss} . The turn on ringing loss are affected by the Q_{oss} and turn off losses are affected by slew rates and parasitic source drain inductances that increase the peak voltage. The GaN device will have a voltage drop during dead time that results in dead time loss shown in (8). The absence of reverse recovery diode implies that dead time can be reduced, hence reducing dead time loss. Switching, ringing and dead time losses are proportional to frequency.

$$P_{half\ bridge\ loss} = P_{top\ \&\ bot,cond\ loss} + P_{top\ \&\ bot,sw\ loss} + P_{ring,turn\ on\ \&\ turn\ off} + P_{dead\ time} \quad (4)$$

$$P_{top\ \&\ bot,cond\ loss} = I_{DS}^2 R_{on,top} D + I_{DS}^2 R_{on,bot} (1 - D) \quad (5)$$

$$P_{sw,turn\ on\ \&\ turn\ off} = \frac{1}{2} f V_{DS} I_{DS} (t_{rise,top} + t_{fall,top} + t_{rise,bot} + t_{fall,bot}) \quad (6)$$

$$P_{ring,turn\ on\ \&\ turn\ off} = V_{in} f \left(Q_{DS,top} + \frac{1}{2} Q_{oss,bot} \right) + \frac{1}{2} f (Q_{Vpeak} (V_{peak} - 2V_{in}) + Q_{Vin} V_{in}) \quad (7)$$

$$P_{dead\ time} = V_{SD} \times I_{SD} \times t_{SD} \times f = (V_{GD,th} - V_{GS} + I_{DS} \times R_{on}) \times I_{SD} \times t_{SD} \times f \quad (8)$$

D. Experimental results and GaN GIT gate driver optimization

Double pulse switching characteristic test, using an

Fig. 5. Experimental setup for inductive load circuit for double pulse test

Fig. 6. Switching experimental results of 2-pulse test results. (a) Turn-off transition for TO-220 GaN GIT (b) Turn-on transition for TO-220 GaN GIT

inductive load circuit shown in Fig.5, is conducted to evaluate the performance of the GaN GIT under EV wireless charging conditions. The GaN GIT device will be evaluated at drain-source voltage at 400V due to application requirements. The GaN driver voltage is fixed to 12V due to auxiliary power supply. This experiment will measure the value drain-source voltage/current overshoot, drain-source voltage rise/fall time and switching losses energy while varying load current from

a) Turn-off switching energy

b) Turn-on switching energy

c) Total switching energy

Fig.7. Evaluation results for switching energy per cycle. (a) Turn-off switching energy (b) Turn-on switching energy (c) Total switching energy

a) Turn off Drain-Source voltage peak

a) Turn on Drain-Source current peak

Fig. 8. Evaluation results for overshoot. (a) Drain source voltage peak vs. absolute rating (b) Drain source current peak vs. absolute rating

2.5A-15A and resistor R1 from 5.1Ω to 36Ω. The switching characteristics waveforms are shown in Fig. 6 using R1=10Ω gate resistor at target application voltage and current of 400V and 10A.

Gate drive loop inductance should be small to improve the slew rate of the GaN device. One should consider reduction of the source and drain inductances along the power loop to reduce the V_{DS} ringing. A freewheeling diode with a low reverse recovery charge improves the slew rate of the device. For realistic results, adopt a freewheeling diode which have a reverse recovery charge equivalent to the GaN GIT's reverse conduction Q_{DS} .

Fig.7a, 7b and 7c shows the turn-off, turn on and total switching loss energy per cycle respectively. Switching loss energy is reduced as the resistor R1 is reduced because larger gate current reduces rising and fall time. Switching loss energy increases with current. These two observations are in agreement with (6).

Another factor for consideration should be the overshoot. Fig 8a and 8b shows the turn-off voltage peak and turn-on current peak respectively of the GaN GIT for varying current levels and varying resistor, R1. Overshoot increases as drain-source current increases and as R1 decreases as higher slew

a) VDS slew rate during turn-off transition

b) VDS slew rate during turn-on transition

Fig.9. Evaluation results for slew rate. (a) VDS slew rate during turn-off transition (b) VDS slew rate during turn-on transition

rates cause larger overshoot. The overshoot test show that the peak does not exceed the absolute value and hence the device is safe.

Finally, the slew rates of the GaN GIT are plotted in Fig. 9. Fig.9a shows the slew rate for the turn off transition. As R1 is reduced, slew rates increase and the device achieves a maximum slew rate of 67V/ns. This work utilizes the GaN GIT with TO-220 package which results in low slew rates compared to PQFN packages.

The turn on transition slew rate is illustrated in Fig. 9b. Due to the parasitic drain and source inductance of the TO-220 package, a voltage drop is observed in the drain source voltage when drain source current flows through it, affecting the slew rate measurement of V_{DS} at low load (2.5A and 5A condition). Therefore we will adopt higher load conditions (10A and 15A). It is shown that low resistance results in higher slew rates and slew rates drop as current increases.

Based on the evaluation data, the choice of R1 should ensure low total switching energy, peak drain-source voltage and current below the limit and having a slew rate less than 50v/ns. Based on these requirements, 10Ω was chosen.

IV. COIL AND COMPENSATION CHARACTERISTICS

The two critical parameters in IPT design that is related to the feasibility and reliability of the system are power transfer capability and efficiency of the compensation [15]. Depending

on the application, the IPT system needs to be able to deliver the required amount of power, while having ability to be efficient.

For EV charging applications, there are two factors which can affect the key metrics. Firstly, the load at the secondary side of the IPT system will vary. Secondly, the relative position of the coils may vary in terms of vertical and lateral distance due to human error in parking and clearance of vehicle from the ground. This will influence the coil coupling and change the reflected load of the secondary to the primary side.

A. Compensation network modelling

There has been much work conducted on the modelling of compensation [20], [21]. The IPT compensation circuit can be modelled like a transformer without a core as shown in Fig. 10a, which can be simplified to into Fig.10b for easy manipulation. By taking Kirchhoff's Voltage Law around the primary and secondary loops respectively, one is able to obtain (9) and (10), while (11) and (12) can be obtained by taking the primary (not including source impedance) and secondary impedance. With some manipulation, one will be able to get (13). These equations are crucial for the calculation of the power transfer and efficiency.

$$VIN = \left(Z_s + \frac{1}{j\omega C_1} + R_1 + j\omega L_1 \right) I_1 - j\omega M I_2 \quad (9)$$

$$0 = -j\omega M I_1 + (Z_{Load} + R_2 + \frac{1}{j\omega C_2} + j\omega L_2) I_2 \quad (10)$$

$$Z_{11} = \frac{1}{j\omega C_1} + R_1 + j\omega L_1 \quad (11)$$

$$Z_{22} = Z_{Load} + R_2 + \frac{1}{j\omega C_2} + j\omega L_2 \quad (12)$$

$$Z_{ab} = R_1 + j \left(\omega L_1 - \frac{1}{\omega C_1} \right) + \frac{\omega^2 M^2}{Z_{Load} + R_2 + j \left(\omega L_2 - \frac{1}{\omega C_2} \right)} = Z_{11} + \frac{\omega^2 M^2}{Z_{22}} \quad (13)$$

B. Power transfer and efficiency equation

Using the equations found earlier, the power transfer capability of the compensation network is shown in (14). With reference to Fig. 10a, the output power can be calculated by multiplying the square of the secondary side current by the output impedance. Many works [15] [22] have provided the calculation for the secondary current with respect to the

primary current for series-series compensation. The efficiency of the compensation can be found by dividing the output power from (14) by the input power. The input power can be found by multiplying the square of the primary side current by the sum of the source impedance and reflected load, making reference to Fig.10b, resulting in (15).

$$P_{out} = I_2^2 Z_{Load} = \left(\frac{\omega M I_1}{Z_{22}} \right)^2 Z_{Load} \quad (14)$$

$$Efficiency = \frac{P_{out}}{P_{in}} \times 100 = \frac{\left(\frac{\omega M I_1}{Z_{22}} \right)^2 Z_{Load}}{I_1^2 (Z_s + Z_{ab})} \times 100 = \frac{\left(\frac{\omega M}{Z_{22}} \right)^2 Z_{Load}}{(Z_s + Z_{ab})} \times 100 \quad (15)$$

The values of the mutual inductance, M, is obtained through experimental data as shown in Fig.11, based on the 250 kHz compensation network in Table I.

The output power and efficiency equations (13) and (14), with M values from Fig. 11, are simulated in Matlab against frequency by varying distance and load as shown in Figure 12 and 13 respectively. In the first simulation as shown in Fig. 12, the load is fixed and the distance is varied. Fig.12a and Fig.12b illustrates the output power and efficiency respectively. As the coils are brought closer, coil coupling increases from under-coupled to critically-coupled and finally over-coupled.

At under-coupled condition, maximum power transfer occurs at resonant frequency. As the coils become over-coupled in Fig 12a, maximum power transfer occurs at 2 peaks that move away from the resonant frequency, which is known as frequency bifurcation. The boundary condition is called critically-coupled.

Fig. 13a and Fig.13b shows the output power and shows the efficiency respectively of the simulation conducted by varying loads and fixing the coupling. By increasing the load, the coupling of the coil is increased according to Fig. 13a.

For an under-coupled or critically coupled system, maximum power transfer and efficiency occurs at resonant frequency. For an over-coupled system, the frequency bifurcation phenomenon occurs and the maximum output power occurs at 2 frequencies away from the resonant frequency. Maximum efficiency occurs at the resonant frequency. As coils become more over-coupled, 3 efficiency

Fig. 10. (a) IPT circuit (b) IPT simplified model

Fig. 11. Experimental results of mutual inductance and coupling against distance for 250 kHz coil

Fig. 12. Simulation results of (a) Output power against frequency for varying distance (b) Efficiency against frequency with varying distance

Fig. 13. Simulation results of (a) Output power against frequency for varying load (b) Efficiency against frequency with varying load

peaks will be observed.

C. Evaluation on coils

Experimental evaluation is conducted on the coils as shown in Fig. 14 to verify the simulation results. The coils are tested at the coil gap of 180mm. The setup of evaluation is similar to the high load, medium distance condition in Fig. 13. In the experiment, the bifurcation phenomenon is observed in the output power with variation with frequency implying over-coupled conditions. The graphs of the simulation and evaluation are very similar with small variations due to deviation in measurement equipment and simulation parameters.

V. EFFICIENCY CENTRIC MODE OF OPERATION

A. Control objective

For HEV and EV charging, the competing technology against IPT is the current norm which is wired plug-in charging, which has the clear advantage of high efficiency [23]. For wireless power transfer to become a feasible alternative, the control objective is to achieve high power transfer efficiency, while being able to deliver sufficient power.

As explained in the previous section, maximum efficiency and maximum power transfer is influenced by the coupling of the coils. In the case of critical or under-coupling, maximum

Fig. 14. Evaluation result of 250 kHz compensation network performance

efficiency and maximum power transfer occurs at the resonant frequency, For over-coupled coils, as in the case of this project, maximum efficiency occurs at the resonant frequency and the maximum power transfer occurs at two peaks which are away from the resonant frequency. Hence, it is crucial for the system to identify the coupling characteristics before starting.

B. Frequency sweep

The system begins an initialization by sweeping the switching frequency across a range of frequencies to understand the coupling characteristics. The diagram of the frequency sweep is shown in Fig. 15. With reference to Fig.16 for the explanation of the frequency sweep, the system is first initialized at the starting frequency, $freq_start$, the state machine is set to 1 and the output power of the inverter is measured. Frequency is swept downwards towards the ending frequency, $freq_end$ and the output power at the inverter is measured at each increment. The measured power is compared to the power measured at the previous increment. At a point of inflection, the state is increase and the frequency of the previous increment is recorded.

When the sweep is complete, the system checks the state machine. If the state machine end at state 4, the system is over-coupled, this means that the bifurcation process occurs as shown in Fig.15a. There would be 3 recorded points of inflection. The first is Peak Power Frequency 1 (PPF1), the second would be the Resonant Frequency (RF) and the third will be Peak Power Frequency 2 (PPF2).

If it ends at state 2, it means that the coil is either critically coupled or under-coupled as shown in Fig.15b. There would only be one point of inflection which is the Resonant Frequency, RF. If the state machine ends at anything other than the previous 2 conditions, there is an error and the sweep is conducted again.

C. Sufficient power transfer

The process flow chart is shown in Fig. 16 illustrates the mode of operation for sufficient power transfer. Initial starting frequency is at the resonant frequency. The duty cycle of the inverter is increased if more power is needed to reach the target power and reduced if the output power is larger than the target power. There are two levels of tuning, coarse and fine.

Fig. 15. Frequency sweep diagram for (a) over-coupled condition (b) Under-coupled or critically coupled condition

Fig. 16. Process flow chart of frequency sweep

At a large difference, the systems utilize a coarse increment to improve the response. As error power gets closer, it utilizes the fine tuning to increase the accuracy.

For a critically-coupled or under-coupled system, operation is optimal at resonant frequency RF, where the maximum efficiency and power transfer occurs.

For an over-coupled system, maximum efficiency occurs at RF and Maximum Power Transfer occurs at PPF1 and PPF2. In order to deliver sufficient power at high efficiency, operating frequency will occur between the region of the resonant frequency and Peak Power Frequency 1 as shown in fig.15a. If the inverter is functioning at maximum duty and the power is not sufficient (output power is less than required power), frequency is increased towards Peak Power Frequency 1(PPF1) to increase output power. If the duty cycle is low, the system will check if it is operating at resonant frequency, and tune itself towards the resonant frequency in order to improve efficiency. If the inverter is already functioning at PPF1, with maximum duty and still unable to deliver enough power, it maintains at that condition as it is at maximum power transfer condition.

VI. EXPERIMENTAL RESULTS

The experimental setup is as shown in Fig. 17. It consists of an inverter, which is connected to the magnetically coupled IPT system. On the secondary side of the coil, the output is rectified and the load is a resistor bank.

The system is evaluated across a distance from 80mm to 150mm. Due to the variation of mutual inductance as the distance between the coil varies, the resistor load is varied from 47Ω at 80mm to 11.5Ω at 150mm so as it ensure the inverter output current is below the current rating of the GaN GIT with consideration of a safety margin. The input to the inverter is supplied at 370V DC.

The system level efficiency result is shown in Fig. 18. At 80mm, the system is tested up to 2.1 kW and it achieves a peak efficiency of 95.6% efficiency at 1.5 kW. As the distance is increased, the peak efficiency of the system falls to 90.4% at 150mm with the main contributor of the efficiency variation is due to efficiency loss in the coil and secondary rectifier. The rationale for testing at 150mm at 1.5kW is to operate the inverter below the current limit of the GaN GIT device.

Fig. 17. Experiment setup

Fig. 18. Experimental results: Efficiency vs. distance

Fig. 19 and 20 shows the efficiency breakdowns of each power conversion stages at 150mm and 80mm respectively operating at 100 kHz. The efficiency of the high frequency power inverter remains relatively constant with varying distance, with the inverter operating at about 97-98% efficiency.

Comparing Fig. 19 and 20, when the load value is changed from 47Ω at 80mm to 11.5Ω at 150mm as mentioned earlier, the secondary side voltage is decreased and current is increased. It is observed that the coil efficiency drops drastically. The reasons of the efficiency drop is due to weaker coupling as the distance increases and higher secondary current resulting in higher copper loss in the coil.

The rectifier block remains highly efficient at 80mm as the output voltage is very high ($> 300V$) due to the larger load resistance, making the losses due to the forward voltage of the diode negligible. However, at further distances (150mm), the lower secondary voltage will make the diode forward voltage drop more significant.

Fig. 21 shows the waveforms of the inverter output current (CH1) and voltage (CH2) as well as the IPT output voltage (CH3) and current (CH4) at 80mm distance. The top half shows the zoom out version of the signals at $20\mu s/div$, while the bottom half displays a close up on the signals at $2\mu s/div$. This setup adopts the 100 kHz compensation network. Using the mode of operation method discussed previously, the operating frequency is tuned towards the resonant frequency of 100 kHz to ensure maximum efficiency.

An efficiency comparison is conducted on the converter with varying frequencies of 100 kHz and 250 kHz at 80mm coil gap. The system is made to transmit the same amount of

Fig. 19. Experimental results: Efficiency breakdown at 150mm

Fig. 20. Experimental results: Efficiency breakdown at 80mm

target power wirelessly. The results are shown in Fig. 22. The efficiency of the converter falls by at 2-3% as frequency increases from 100 kHz to 250 kHz, with efficiency of 95.13% falling to 91.7 at 2kW.

The efficiency drop is caused by the switching loss due to higher frequency operation. For the inverter, switching losses dominate during low power condition and gradually conduction losses start to dominate as shown by the low efficiency at 250 kHz. The coil will experience higher AC resistance as the frequency is increased 2.5 times. The rectifier will suffer from higher reverse recovery losses with higher frequencies.

Fig. 21. Experimental results of inverter and coil. Channel 1: Inverter output current, Channel 2: Inverter output voltage, Channel 3: IPT output voltage, Channel 4: IPT output current

Fig. 23 illustrates the efficiency comparison between the GaN based and SiC based system. The SiC based system was made using the similar components as the GaN based system and switching characteristics is optimized to ensure fairness. The GaN based system performs at 1% higher efficiency compared to the SiC based system attributing from the GaN inverter's 1% improvement over the SiC inverter. The lower on-resistance and gate charge translate to less switching and conduction loss.

Table III shows a comparison table between this work and recently published works. This work demonstrates higher or comparable system level efficiency compared to other works

Fig. 22. Experimental results: Efficiency vs. input power for varying frequencies

Fig. 23. Experimental results: Efficiency comparison between GaN and SiC

with low switching frequency and smaller coil gap to coil size ratio. A lower switching frequency will result in lower switching loss, but SAE J2954 proposes 85 kHz. This work was developed prior to this knowledge and it would serve as a good example to showcase possibilities of GaN devices to functions beyond this frequency range. A lower coil gap to coil size ratio will result in better coupling and higher efficiency. However, one needs to consider the practicality of a large coil on the car, which can be bulky. Therefore, the authors believe that designing the coil at 300mm diameter is a reasonable decision. Other works have been able to achieve higher output power because of the availability of higher current rating Si and SiC devices in the market. The author believes that as current ratings of GaN devices start to increase, we would start to see more works using GaN HEMTs for EV wireless power transfer. A higher current rating GaN HEMT will have lower $R_{DS(on)}$, which is advantageous for higher power levels where conduction losses dominate.

VII. CONCLUSION

In this work, a practical high efficiency wireless power transfer system for EV charging application capable of transmitting 2 kW of power across 80 mm at 95% efficiency was developed. GaN GIT is adopted to improve the efficiency of the inverter and the gate drive circuit for the GaN GIT is optimized for low switching loss. Through the development process, a control algorithm is developed to make it function at the optimal frequency to provide maximum efficiency for the required output power. Experimental results demonstrate the efficiency advantage of adopting GaN GIT in high frequency applications and potential of replacing competing power transistor materials.

ACKNOWLEDGMENTS

The authors would like to acknowledge the funding support from NTU-A*STAR Silicon Technologies Centre of Excellence (Si-COE) under the program grant No. 11235100003. The authors would like to acknowledge the

TABLE III. COMPARISON TABLE

	This work		EPE/PMEC2012 [1]		APEC 2014 [23]	ITEC2013 [24]	
Power transistor type	GaN GIT		SiC MOSFET		SiC MOSFET	Si SJ MOSFET	
Input voltage, VIN (V)	370V DC		AC 100V 60 Hz		-	150V DC	
Coil size	300 mm diameter (Circle)		170 mm diameter (Circle)		-	700mm diameter (Circle)	
Frequency	100kHz, 250kHz		20-30 kHz		23.5 kHz, 48 kHz, 81 kHz	42.5 kHz	
Coil gap	80 mm	150 mm	100mm	150mm	-	100mm	150mm
Output power	2.1kW @ 100kHz	1.5kW @ 100kHz	670W	500W	3.3kW	1.5kW	
System peak Efficiency	95.6% @ 100kHz 91.7% @ 250kHz	90.4% @ 100kHz	84%	81%	91.77% @ 23 kHz 91.54% @ 49 kHz 88.70% @ 80.75 kHz	86%	82%

financial support and the provision of GaN GIT samples by Panasonic Industrial Devices Semiconductor Development Asia (PIDSCDA). The authors are grateful for the financial support provided by the Economic Development Board (EDB) of Singapore and would like to express appreciation to Energy Research Institute @ NTU (ERI@N) for providing the facilities and technical support for this project. Last but not least, the authors would like to express gratitude towards VIRTUS IC design center.

REFERENCES

- [1] H. Omori, Y. Iga, T. Morizane, N. Kimura, K. Nakagawa and M. Nakaoka, "A novel wireless EV charger using SiC single-ended quasi-resonant inverter for home use," in *Power Electronics and Motion Control Conference (EPE/PEMC), 2012 15th International*, 4-6 Sept. 2012.
- [2] US department of Energy, "All-Electric Vehicles (EVs)," US department of Energy, 08 April 2014. [Online]. Available: <https://www.fueleconomy.gov/feg/evtech.shtml#end-notes>. [Accessed 09 April 2014].
- [3] S. Li and C. C. Mi, "Wireless Power Transfer for Electric Vehicle Applications," in *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 3, no. 1, pp. 4-17, March 2015.
- [4] H. H. Wu, A. Gilchrist, K. Sealy and D. Bronson, "A 90 percent efficient 5kW inductive charger for EVs," in *2012 IEEE Energy Conversion Congress and Exposition (ECCE)*, Raleigh, NC, 15-20 Sept. 2012, pp. 275-282..
- [5] Y. Uemoto, M. Hikita, H. Ueno, H. Matsuo, H. Ishida, M. Yanagihara, T. Ueda, T. Tanaka and D. Ueda, "Gate Injection Transistor (GIT)—A Normally-Off AlGaIn/GaN Power Transistor Using Conductivity Modulation," in *IEEE Transactions on Electron Devices*, vol. 54, no. 12, pp. 3393-3399, Dec. 2007.
- [6] D. Ueda, M. Hikita, S. Nakazawa, K. Nakazawa, H. Ishida, M. Yanagihara, K. Inoue, T. Ueda, Y. Uemoto, T. Tanaka and T. Egawa, "Present and future prospects of gan-based power electronics," in *Solid-State and Integrated-Circuit Technology, 2008. ICSICT 2008. 9th International Conference on, Beijing, 2008*, 1078-1081,, 20-23 Oct. 2008.
- [7] E. A. Jones, F. F. Wang and D. Costinett, "Review of Commercial GaN Power Devices and GaN-Based Converter Design Challenges," in *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 4, no. 3, pp. 707-719, Sept. 2016.
- [8] L. Garcia-Rodriguez, V. Jones, J. C. Balda, E. Lindstrom, A. Oliva and J. Gonzalez-Llorente, "Design of a GaN-based microinverter for photovoltaic systems," in *2014 IEEE 5th International Symposium on Power Electronics for Distributed Generation Systems (PEDG)*, Galway, 24-27 June 2014, pp. 1-6..
- [9] W. Zhang, Z. Xu, Z. Zhang, F. Wang, L. M. Tolbert and B. J. Blalock, "Evaluation of 600 V cascode GaN HEMT in device characterization and all-GaN-based LLC resonant converter," in *2013 IEEE Energy Conversion Congress and Exposition*, , Denver, CO, , 15-19 Sept. 2013, pp. 3571-3578..
- [10] S. Ujita, Y. Kinoshita, H. Umeda, T. Morita, S. Tamura, M. Ishida and T. Ueda, "A compact GaN-based DC-DC converter IC with high-speed gate drivers enabling high efficiencies," in *2014 IEEE 26th International Symposium on Power Semiconductor Devices & IC's (ISPSD)*, Waikoloa, HI, 15-19 June 2014, pp. 51-54..
- [11] A. Cai, A. Pereira, R. Tanzania, Y. K. Tan and L. Siek, "A high frequency, high efficiency GaN HFET based inductive power transfer system," in *2015 IEEE Applied Power Electronics Conference and Exposition (APEC)*, Charlotte, NC,, 15-19 March 2015, pp.3094-3100.
- [12] K. Shirabe, M. Swamy, J.-K. Kang, M. Hisatsune, Y. Wu, D. Kebort and J. Honea, "Efficiency Comparison Between Si-IGBT-Based Drive and GaN-Based Drive," in *IEEE Transactions on Industry Applications*, vol. 50, no. 1, pp. 566-572, Jan.-Feb. 2014.
- [13] D. Han, A. Ogale, S. Li, Y. Li and B. Sarlioglu, "Efficiency characterization and thermal study of GaN based 1 kW inverter," in *2014 IEEE Applied Power Electronics Conference and Exposition - APEC 2014*, Fort Worth, TX, 16-20 March 2014, pp. 2344-2350.
- [14] H. R. Rahnamaee, U. K. Madawala and D. J. Thrimawithana, "A multi-level converter for high power-high frequency IPT systems," in *2014 IEEE 5th International Symposium on Power Electronics for Distributed Generation Systems (PEDG)*, Galway, 24-27 June 2014, pp. 1-6..
- [15] M. Kiani and M. Ghovanloo, "A Figure-of-Merit for Designing High-Performance Inductive Power Transmission Links," in *IEEE Transactions on Industrial Electronics*, vol. 60, no. 11, pp. 5292-5305, Nov. 2013.
- [16] C. Mößlacher and O. Guillemant, "AN 2012-03: Simple design techniques for optimizing efficiency and overvoltage spike of synchronous rectification in DC to DC converters," Infineon Technologies Austria AG, Austria, March 2012.
- [17] B. Tien and C. Hu, "Determination of carrier lifetime from rectifier ramp recovery waveform," in *Electron Device Letters, IEEE*, vol. 9, no. 10, pp. 553-555, Oct. 1988.
- [18] H. Umegami, Y. Nozaki, M. Yamamoto and O. Machida, "A novel high efficiency gate drive circuit for normally off type GaN-FET," in *IEEE Transactions on Industry Applications*, vol. 50, no. 1, pp. 593-599, Jan.-Feb. 2014.
- [19] Y. Ren, M. Xu, J. Zhou and F. C. Lee, "Analytical loss model of power MOSFET," in *Power Electronics, IEEE Transactions on*, vol. 21, no. 2, pp. 310-319, March 2006.
- [20] J. Sallan, J. L. Villa, A. Llombart and J. F. Sanz, "Optimal Design of ICPT Systems Applied to Electric Vehicle Battery Charge," in *IEEE Transactions on Industrial Electronics*, vol. 56, no. 6, pp. 2140-2149, June 2009.
- [21] C. S. Wang, G. Covic and O. H. Stielau, "Power transfer capability and bifurcation phenomena of loosely coupled inductive power transfer systems," in *IEEE Transactions on Industrial Electronics*, vol. 51, no. 1, pp. 148-157, Feb. 2004.
- [22] R. Chen, C. Zheng, Z. Zahid, E. Faraci, W. Yu, J. S. Lai, M. Senesky, D. Anderson and G. Lisi, "Analysis and parameters optimization of a contactless IPT system for EV charger," in *2014 IEEE Applied Power Electronics Conference and Exposition - APEC 2014*, Fort Worth, TX, 16-20 March 2014, pp. 1654-1661.
- [23] O. Onar, M. Chinthavali, S. Campbell, P. Ning, C. P. White and J. M. Miller, "A SiC MOSFET based inverter for wireless power transfer applications," in *2014 IEEE Applied Power Electronics Conference and Exposition - APEC 2014*, Fort Worth, TX,, 16-20 March 2014, pp. 1690-1696.
- [24] C. Duan, C. Jiang, A. Taylor and K. Bai, "Design of a zero-voltage-switching large-air-gap wireless charger with low electrical stress for Plug-in Hybrid Electric Vehicles," in *Transportation Electrification Conference and Expo (ITEC), 2013 IEEE*, Detroit, MI, 16-19 June 2013, pp. 1-5.

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