



ONR Short Pulse Research, Evaluation, and non-SWaP Demonstration for C-sUAV Study (OSPRES)

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The technical objective of the ONR Short Pulse Research, Evaluation, and non-SWaP Demonstration for C-sUAS Study (OSPRES) Grant program is to execute high-risk, high-payoff efforts that address elementary gaps to achieving a modular, scalable & electronically steerable high power microwave (HPM) based defense system for the counter UAS mission, with focus on the short-pulse high-average-power space, as well as the responsivity of the target to RF stimuli, while educating the next generation of pulsed power and defense-minded scientists, engineers, and innovators.					
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1 OVERVIEW

OSPRES Grant Goal: address and transition technologies and capabilities that enable the OSPRES Grant Objective using the OSPRES Grant Approach, while educating the next generation of pulsed power and defense minded, stewards and innovators.

OSPRES Grant Objective: to execute high-risk, high-payoff efforts that mitigate, fill or rectify one or more grand-challenge or elementary gaps or deficiencies needed to achieve a modular, scalable and electronically steerable high-power microwave (HPM) based defense system for the counter unmanned aerial system (cUAS) mission. The OSPRES HPM system and sub-system development/evaluation efforts are focused on the short-pulse high-average-power space and includes the kill chain considerations including the target and its responsivity to radiofrequency stimuli.

OSPRES Metrics of Success: developed technologies and capabilities that are published in the peer review, protected as intellectual property, and/or, transitioned to the OSPRES Contract effort and beyond for integration with other DoD needs or dual-commercial-use, as enabling/integrable capability(ies).

OSPRES Grant Approach: A fail-fast philosophy is maintained, where if a technology or capability is deemed infeasible to be demonstrated or validated during the project period-of-performance, it shall be culled, results to date and basis of infeasibility documented, and the next major gap/deficiency area addressed by that part of the project sub-team. Students whose thesis depends on following through with a full answer during the life of the award will be given special dispensation to continue their work.

OSPRES Grant Security: All efforts should be fundamental and publicly releasable in nature when taken individually. This report has been reviewed for operational security, compilation, proprietary and pre-decisional information concerns.

Contents of this Report: This report compiles monthly status reports (MSRs) from October 2019 through September 2020.

ONR HPM 2019 Program - Monthly Status Report (MSR) – OCTOBER2019
Anthony Caruso – Univ. Missouri – Kansas City
N00014-17-1-3016 [OSPRES Grant]
ONR Short Pulse Research, Evaluation and non-SWaP Demonstration for C-sUAV
Study

11/15/2019

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OSPRES Project Description: Develop tunable High Power Microwave (HPM) sources and electronically steerable radiators whose waveform(s) enable control over all levels-of-lethality on cUAV and similar targets, with the least power density.

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Executive Summary:

During the reporting period the GaN PCSS subgroup performed pulsed testing of 500 μ m gapped GaN:C devices. They determined that the maximum amount of incident 355nm laser light with a pulse duration of 1ns was 20 μ J to achieve less than 1 Ω of on-state resistance. Above 20 μ J incident energies, no notable improvements in device efficiency is observed. Above 20 μ J, the 500 μ m gap device compared well with the theoretical limit where V_{charge} is equivalent to pulse voltage out. If GaN:C devices can be optimized for lower laser energies (in linear mode) then the laser requirements of GaN:C will fall below that of the Si PCSS devices currently being tested. This will open the restraints imposed on laser requirements and may drive decrease in laser system costs.

The RF coupling subgroup at UMKC determined that at large distances from a radiating source, the incidence angle between source and UAV can be assumed to be between 0 and 45 degrees, with 0 degrees being nearly equivalent to a 90 degree electric field orientation. Using this information they studied the time domain of UAV model response to 3 different RF pulse shapes which were chosen as the pulses matched the 1st excitation modes of the UAV models in frequency and bandwidth. Using a pulse close to 200MHz in center frequency between 0 and 45 degrees incident angle a drop of 0.05mA of induced current was seen. Further work will help to inform whether low order modes or higher order modes are preferred to couple to UAV frames at large distances from an RF radiating source.

Electrically small antenna work continued to investigate the directivity (far-field) to quality (near field) (D/Q) factor of the antennas to understand the losses and steerability of the microstrip patch antenna designs in an array format. Using CMA and FEKO the group calculated gain as a function of steerable angle of the microstrip antenna in a 10x10 array with center frequencies of 478, 601, and 724MHz. It is shown that an antenna design around 478MHz showed minimal loss at angles up to 40 degrees.

The device modeling subgroup has added electron beam irradiation defects into models of Si PCSS devices to calculate device charge storage and carrier lifetime as a function of defect concentration. Consistency was obtained with models of devices having 7.3 defects/cm³ with experimental charge storage time data from Si PCSS devices with identical defect concentrations. Further work will refine the simulation models will improve theoretical models consistency with experimental data.

Current Month Progress:**I. Source Development (Myers)****A. PCSS Power Conditioning - IGBT (Khan)****PCSS Power Conditioning Circuit**

During the month of September and October in 2019, we fabricated an in-house power conditioning circuit based on an HV Si MOSFET from BEHLKE. The power conditioning circuit was designed in EAGLE CAD. We have incorporated two current measuring techniques in the circuit design to assess their compatibility in the HV pulsed power circuit. The passive network design for damping unwanted reflections and oscillations in the pulse forming network has been optimized. EMI reduction measures in our experimental setup to counteract the EMI induced from the rapid switching of the PCSS have been implemented. We tested the circuit to a DC supply voltage of 2 kV at a repetition frequency of 2 kHz. Utilizing a new laser system, we have been able to run the experiments at a repetition frequency of 50 kHz and at DC supply voltages up to 2 kV.

A power conditioning circuit was designed in EAGLE CAD using commercial off-the-shelf (COTS) components and an HV Si MOSFET (HTS 121) from BEHLKE. FR4 was selected as the PCB material. Wide traces with a copper thickness of 4 oz have been used in the design coupled with board thickness of 0.12 of an inch. Provision of several jumpers has been included in the board design for modifying the circuit without the need of fabricating a new board altogether. An LEM current sensor and a 150 W rated 0.1 Ω shunt resistor are connected in series with the HV switch to measure current through the circuit. We used non-inductive power resistors in the circuit to reduce parasitic inductance of the power conditioning circuit. The fabricated board is illustrated in Figure A1.



Figure A1. In-house prototype of HV power conditioning circuit for pulse forming network.

Differential probes have been used to measure the drain to source voltage (V_{DS}) of the HV MOSFET switch using test points in the PCB. Additional test points have been

included in the design to measure the voltage across a bulk capacitor before the switch, output of the LEM current sensor and output across the shunt resistor.

Initially, we tested the power conditioning circuit at repetition frequencies of 1 kHz and 2 kHz up to a DC supply voltage of around 1200 V. For a supply voltage of 200 V, the experimental values of the output voltages across a matched load reach close to the theoretical value of 100 V for both of the repetition frequencies. The experimental results are shown in Figure A2 and Figure A3 respectively.

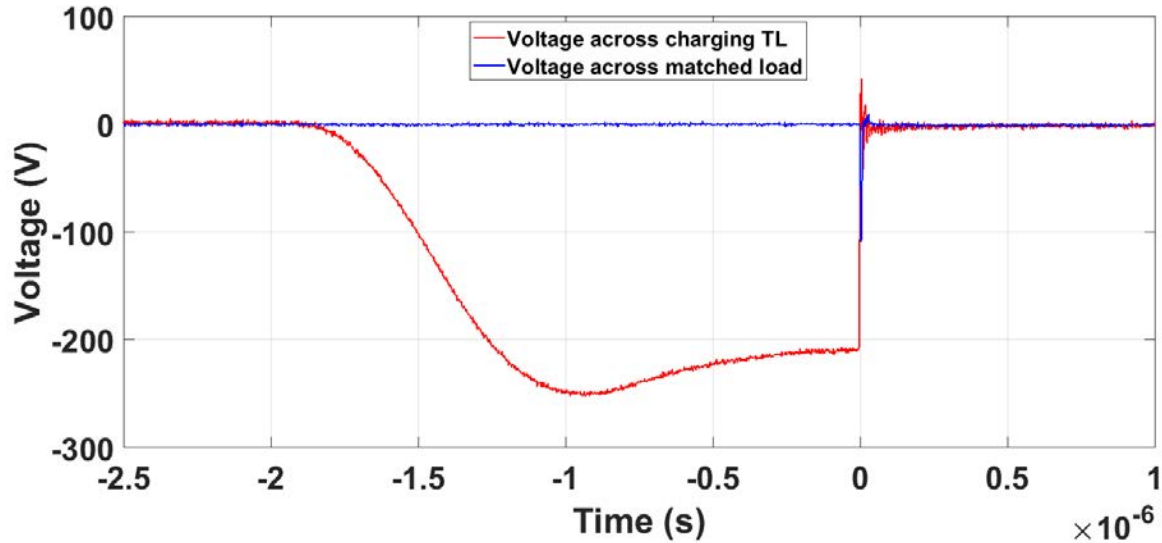


Figure A2. Experimental results using HV switch for a supply voltage of 200 V at a repetition frequency of 1 kHz.

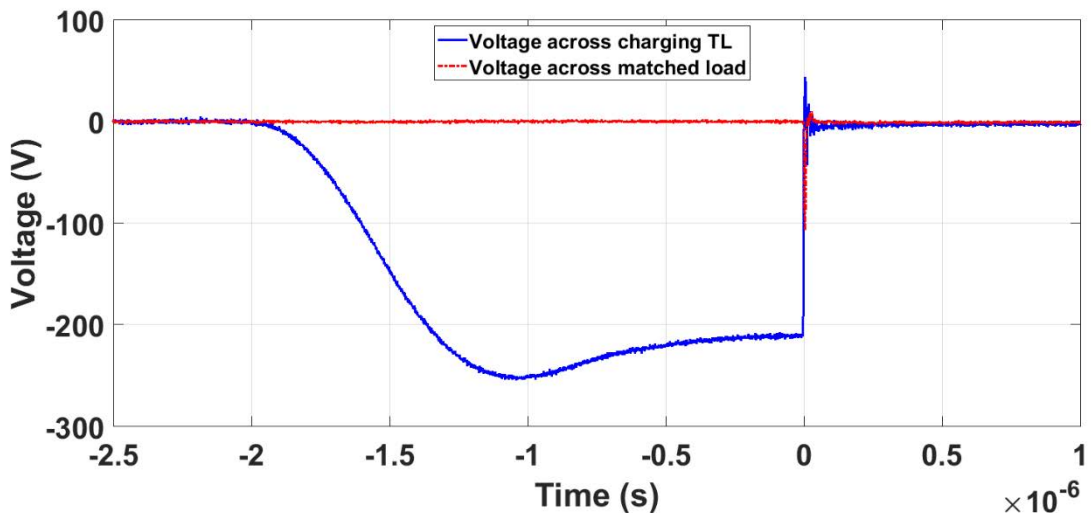


Figure A3. Experimental results using HV switch for a supply voltage of 200 V at a repetition frequency of 2 kHz.

We have increased the bulk capacitance and reduced the damping resistance of the power conditioning circuit. The increased bulk capacitance (20 nF) helps in maintaining steady voltage level across the bulk capacitor throughout the experiment. During the event of the PCSS triggering, the dip in bulk capacitor voltage is minimized by a lower RC time constant and higher bulk capacitance.

It was observed that at a higher supply voltage of around 1300 V and at a repetition frequency of 2 kHz, the DC power supply for the control side of the HV MOSFET switches to overcurrent protection mode. EMI induced from the rapid switching of the PCSS at a high voltage level has been principally culpable for the malfunctioning of the power supply. The DC power supply was in close proximity to the PCSS when the malfunction was observed. We also observed significant EMI nearby the PCSS which was a primary reason for the malfunction. A cardboard box with conductive copper tape was used to create a Faraday cage around the PCB. The copper shield has been shorted to the earth ground. Both the PCB and the DC power supply of the control side of the switch have been moved further back from the PCSS where the EMI strength is significantly weaker. The increased distance and Faraday cage have effectively mitigated the EMI issues to an acceptable level.

After mitigating the EMI issues, we tested the power conditioning circuit at 2 kHz repetition frequency and a DC supply voltage of 2 kV. The output voltages across the matched load have been close to the theoretical values (half of the supply voltage) for input voltages up to 2 kV. The power conditioning circuit was tested with a new laser setup that can operate at a repetition frequency of 50 kHz and higher. The setup has been tested at a DC supply voltage level of 2 kV and a repetition frequency of 50 kHz. At higher repetition frequency, switching loss becomes the dominant loss mechanism of the HV switch. At 50 kHz, the switch operates in burst mode with a 2 ms interval between consecutive bursts. The output pulse-count in a burst decreases with increasing supply voltage. Upwards of 40 pulses in a single burst for an input supply voltage of 500 V have been observed. At 2 kV, the pulse count in a single burst has reduced to around 28 pulses. However, the interval duration between two successive bursts has remained almost the same (2 ms). The experimental results at a repetition frequency of 50 kHz and a supply voltage of 700 V are shown in Figure A4 and Figure A5.

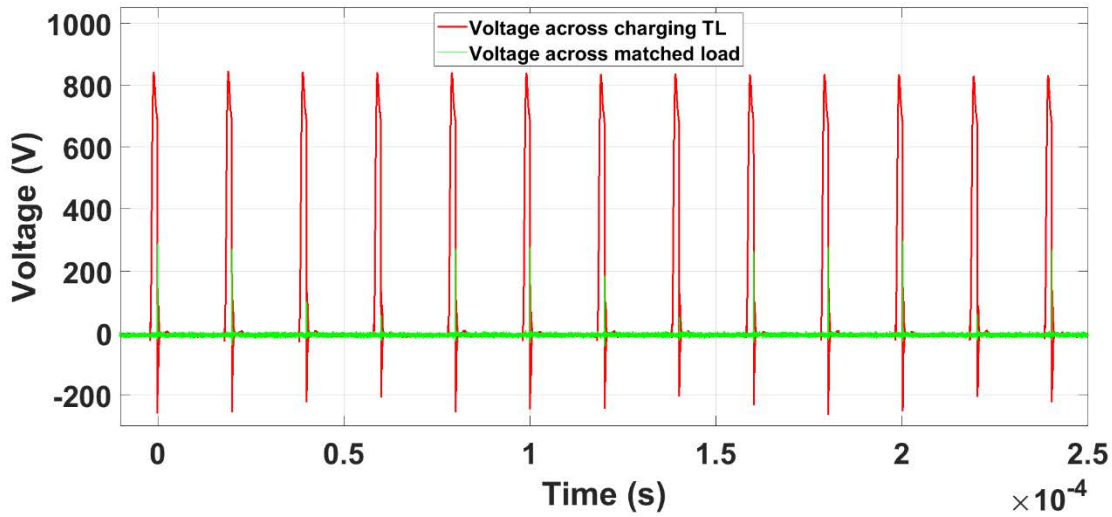


Figure A4. Experimental results using HV switch for a supply voltage of 700 V at a repetition frequency of 50 kHz.

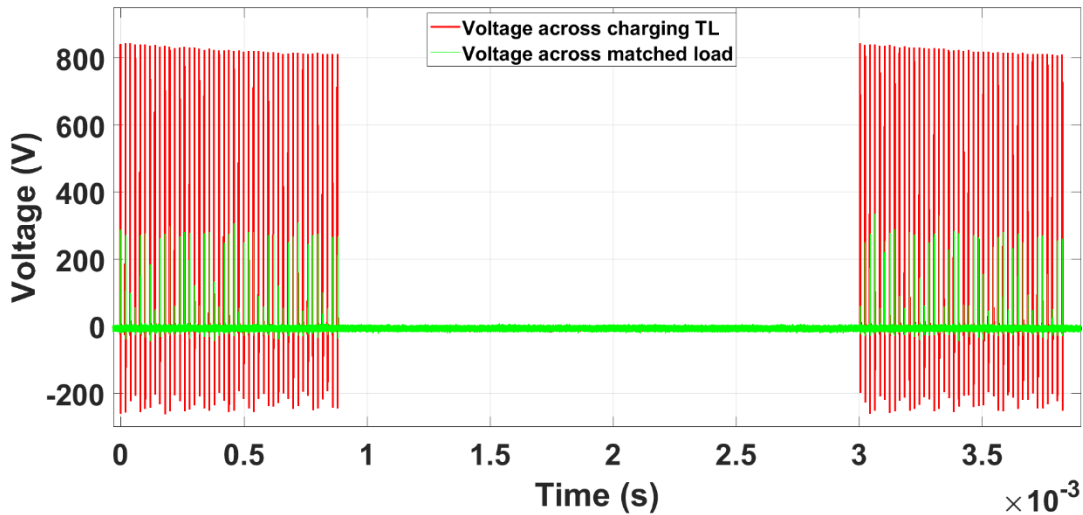


Figure A5. Experimental results using HV switch for a supply voltage of 700 V at a repetition frequency of 50 kHz (zoomed out).

The rated continuous switching frequency of the HV switch is 20 kHz. With the addition of forced-air cooling or direct liquid cooling the average power handling capacity of the switch can be increased resulting in an increased maximum continuous operating frequency. At an operating voltage of 1.5 kV at 50 kHz repetition frequency, we have observed the temperature of a resistor, connected in series with a flyback diode, has surged to 80 °C. We are planning to replace this resistor with one in a TO-247 package and attach it to a heat sink for improved heat dissipation.

We are planning to introduce forced-air cooling aided thermal management mechanism for the power conditioning board to increase the average power handling capability and

output pulse count. We are also exploring some HV SiC MOSFET switches that can operate at a continuous switching frequency up to 100 kHz with sufficient cooling options. The low ON-resistance of SiC BEHLKE switches can minimize the conduction loss of the switch, and increase the output pulse count at a higher switching frequency (100 kHz).

Immediate Tasks for the PCSS Subgroup

- i. Incorporating forced-air cooling mechanism to improve the average power handling capability of the Si BEHLKE switch.
- ii. Operating the power conditioning circuit at a higher voltage (> 3 kV) and at a repetition frequency of 2 kHz.
- iii. Operating the power conditioning circuit at a lower voltage (< 3 kV) and at a higher repetition frequency (> 50 kHz).
- iv. Designing a high-voltage MOSFET and corresponding gate driver.

Expected Products of PCSS Subgroup

- i. Designing HV SiC MOSFET based power conditioning circuit.
- ii. Incorporating efficient thermal management system with the power conditioning circuit.
- iii. Designing a single gate driver-based HV series connected switch.
- iv. Exploring different power conditioning circuit topologies to find out the most suitable one for the intended application.
- v. Understanding the reliability of the HV switch used in the power conditioner circuit.

B. Switches - GaN:C (Thompson)

Switches - GaN:C (Thompson)

The dependence of incident laser energy on on-state resistance was further investigated using a 500 μm gap length GaN:C device. The device was vacuum encapsulated with Sylgard 184 Silicone Elastomer, to prevent arcing, and mounted on the PCB shown in the previous report. The 1-ns Wedge HB laser with the THG setup was used, and the beam was shaped using a cylindrical lens to ensure it covered the entire active area of the device under test. The energy density was measured and the total energy incident on the device's active area was calculated. The device was tested at 200 Hz starting at 100 V and increased until the device failed at 900V.

The relationship between output (peak) voltage and charge voltage was approximately linear, as shown in figure B1. The average on-state resistance reduced as the incident energy increased (Figure B2), and did not change as the voltage was increased. This approximately linear relationship allows for extrapolation of the device's behavior outside of the measured range. On-state resistances of less than $1\ \Omega$ were achieved at 20 μJ and greater energies. Above this energy, there is no additional measurable gain in output voltage with the 500 μm device.

Failure at 900 V was due to the power limit of the resistors on the PCB during testing. To allow for higher voltages, more resistors will be used in parallel to allow for power

distribution amongst more resistors while maintaining the same total effective resistance in the circuit.

Future pulse testing will be done using devices with various gap lengths to see how the gap length affects the amount of incident energy necessary to reach maximum gain. Pulse testing will also be done at higher voltages to show if the linear relationship between charge voltage and peak output voltage continues at higher charge voltages. A vacuum oven has been obtained and vacuum curing of the Sylgard 184 encapsulation material will be more efficient as the devices no longer need to be sent out for encapsulation. New devices from NRL with different carbon concentrations will be used in pulse testing as well to determine how the concentration affects device performance.

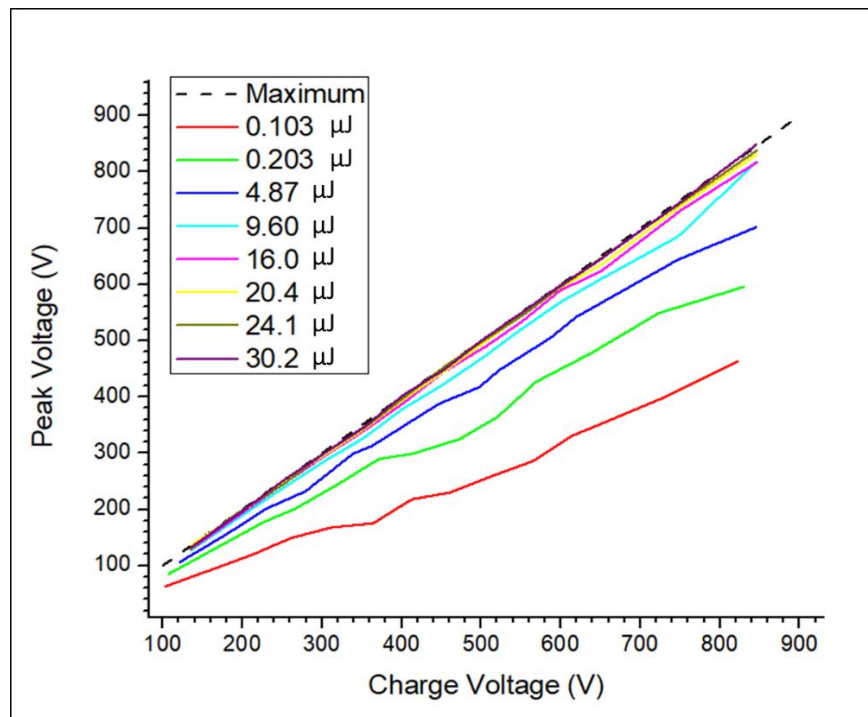


Figure B2: Peak voltage as a function of charge voltage at different incident energies for a 500 μ m gap length device.

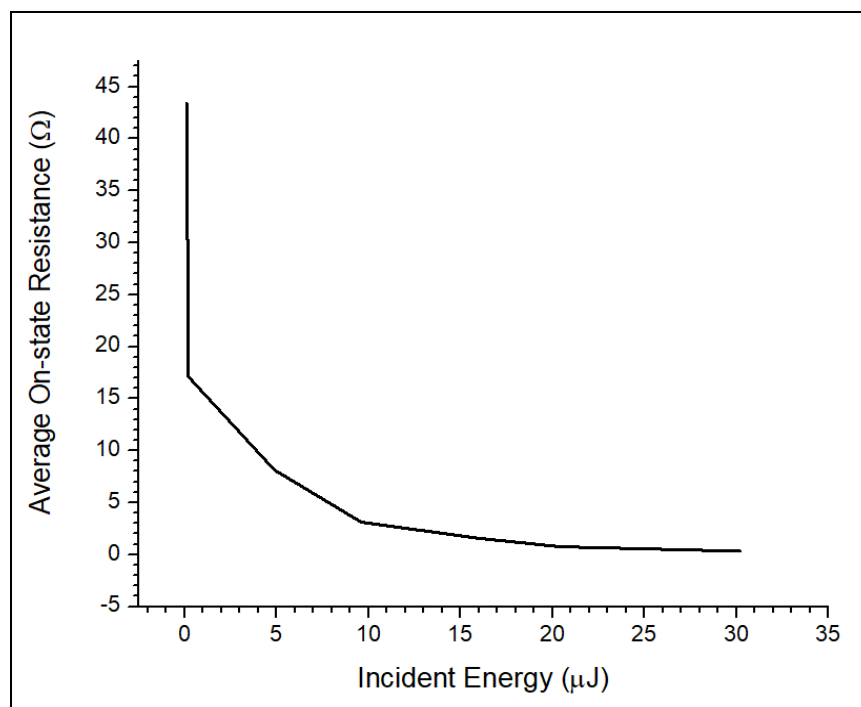


Figure B3: Average on-state resistance as a function of incident energy for a 500 μm gap length device.

Immediate Tasks of Fundamental GaN:C PCSS Subgroup

- i. Continued pulse testing to find on-state resistance dependence on incident energy and gap length.
- ii. Pulse testing using the new devices from NRL – Carbon concentration of $8\text{E}16$ vs. $5\text{E}17$ with $50\text{ }\mu\text{m}$ and $1000\text{ }\mu\text{m}$ gap length devices.

Expected Products of Fundamental GaN:C PCSS Subgroup

- i. Results showing continued linear relationship between charge voltage and peak output voltage at different incident energies and gap lengths.
- ii. Results showing relationship of carbon concentration and on-state resistance.

C. Switches - Figure of Merit (Bhamidipati)

High Power FoM:

In the earlier reports, the power losses incurred within the device during the on-state for Si, SiC and GaN have been reported. It was seen that SiC incurred higher losses when calculated using equations derived from Nunnally's carrier generation model varying incident laser intensity, device geometries, fixed recombination time(t_r), and constant carrier mobility.

In continued efforts to derive the material FoM, the power loss equations have been derived for the turn-on and turn-off stages in the device operation by taking temporal laser profile into consideration, omitting size assumptions and approximations as shown in equations (a) and (b). The power dissipated in the device during these stages of operation adds heat to the lattice and drives device damage at high peak powers and repetition rates. Equations (a) and (b) will be numerically analyzed with time varying Gaussian laser profiles, geometric variations, recombination/transit times, temperature, dopant dependent mobility.

$$P_{\text{turn-on}} = \frac{E_c^2 L^3}{R_{\text{Load}} I_0} \int_{t_1}^{t_2} \frac{\left(1 - \exp\left(-\frac{t}{T_{\text{net}}}\right)\right)}{\exp\left(-2.77\left(\frac{t}{t_p}\right)^2\right) \left[2 + \frac{R_{\text{Load}} q \mu_T I_0 T_{\text{net}}}{h\nu L}\right] - 2R_{\text{Load}} q \mu_T I_0 T_{\text{net}} \left[\exp\left(-5.54\left(\frac{t}{t_p}\right)^2 - \left(\frac{t}{T_{\text{net}}}\right)\right)\right] / (h\nu L)} dt \quad (a)$$

and

$$P_{\text{turn-off}} = I_0 T_{\text{net}} q \mu_T E_c^2 L^3 \int_{t_3}^{t_4} \frac{\left[\exp\left(-2.77\left(\frac{t}{t_p}\right)^2 - \frac{t}{T_{\text{net}}}\right)\right]}{h\nu L^3 + 2R_{\text{Load}} q \mu_T I_0 T_{\text{net}} \exp\left(-2.77\left(\frac{t}{t_p}\right)^2 - \frac{t}{T_{\text{net}}}\right)} dt \quad (b)$$

Where

E_c – critical field

L – thickness of the device (switch gap)

d, w – length and width of the device

$h\nu$ – energy carried in each photon

R_{Load} – load resistance

q – charge of the electron

μ_T – mobility of holes and electrons

t_r – inherent device recombination time

t_T – transit time of carriers

t_p – FWHM duration of the laser pulse

$T_{\text{net}} = t_r * t_T / (t_r + t_T)$

I_0 – peak optical intensity

For on-state loss calculations, the FoM is inconclusive incorporating losses only as SiC has a higher thermal conductivity that could keep the lattice temperature low in spite of the incurred losses. Therefore, the power losses should be combined with the thermal conductivities to determine an optimal material. Along with the material figure of merit, our approach is also focused on developing an application specific FoM to be included in device classification. The flow chart in figure C1 below shows the flow of this classification process. The approach in defining an application specific FoM is still in the process. We are currently looking to weight factors such as power throughput, repetition rates, cost, and cooling requirements (size). The best material for various applications may not necessarily need to withstand high temperatures, repetition rates and stress levels. Also, factors such as peak power, switching mode (mono, burst mode, quasi burst, CW), PRF requirements, cooling requirements (size), cost factors that influence SWaP-C² of the system will be investigated and prioritized based on final application of the device. The factors will be weighed in the order of their influence on the system performance, and implementing cost vs. effectiveness analysis would provide a cost effective solution for PCSS material for a given application.

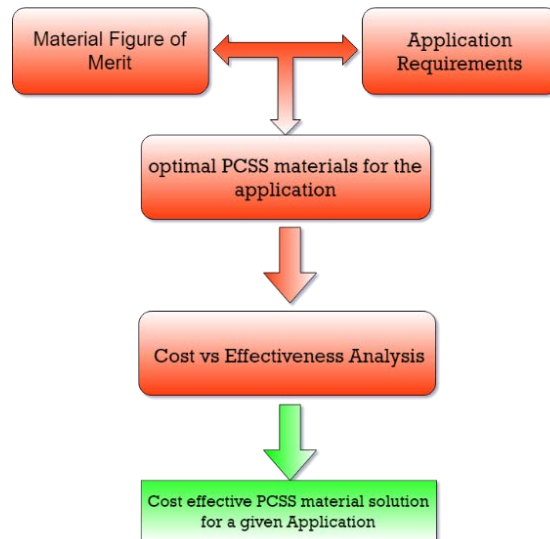


Figure C1: Flow chart showing the approach for application based solution for material selection.

Immediate Tasks for Figure of Merit subgroup:

- i. Numerical analysis for turn-on and turn-off power loss equations to quantify the switching losses and combine with thermal conductivity for material FoM.
- ii. Effectively weight the influencing factors for materials' performance to determine application specific FoM that would drive SWaP-C² implementation of the system.
- iii. Cost vs. effectiveness analysis for materials based on application requirements.

II. Effects, Coupling and UAS Developments (Currie)

D. RF Coupling (Hassan)

Time-domain analysis for UAVs

During this month we studied the effect of the excitation pulse and the drone detection tower distance effect on the prediction and control of the coupling to the drone.

The comparison between the perfect conducting and dielectric quadcopter drones and the effect of each material on the coupling to subsystems of the drone has been illustrated in our previous MSRs. The first-order approximation of the drone has been proposed and studied extensively to show the effect of frame orientation on the coupling to the drone's subsystems.

The first-order approximation of the quadcopter UAV's subsystems, typically wires and PCB, is shown in Figure D1. We also performed the CMA for the structure. The modal significance of the structure is shown in Figure D2. To illustrate our study point, Figure D3 presents the possible scenarios of detecting and taking action with incoming drones to a system.

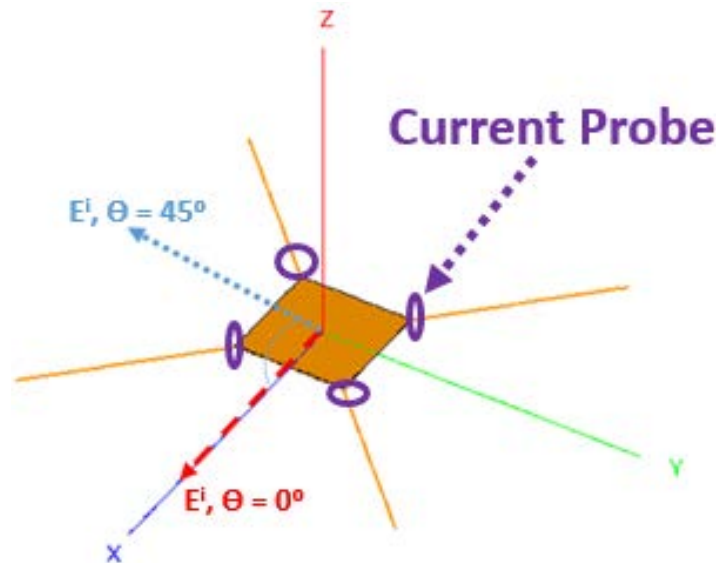


Figure D1. Sketch of the proposed UAV's subsystem model showing the 4 locations of current probes and the incident electric field direction.

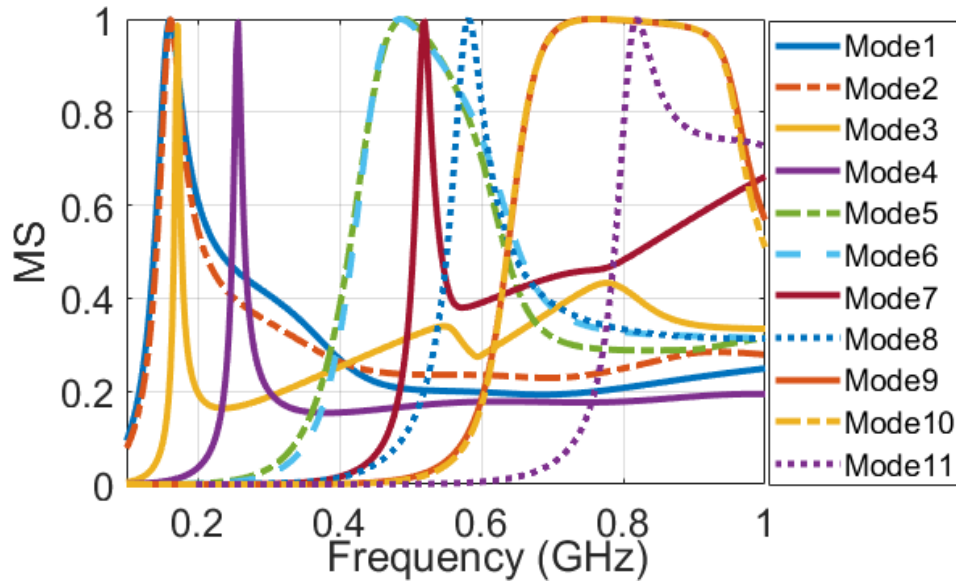


Figure D2. Modal significance of the proposed UAV's subsystem model.

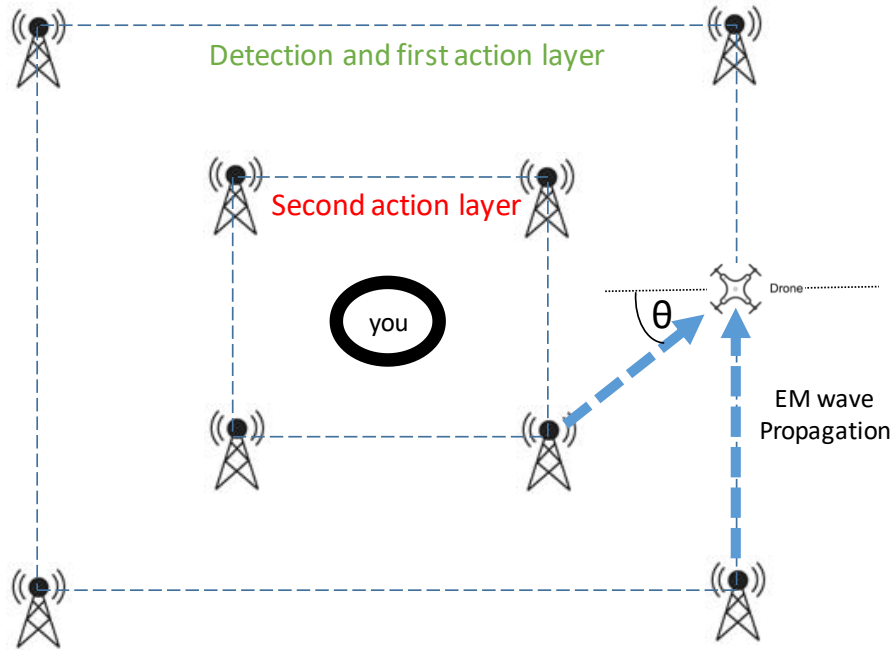


Figure D3. Sketch of detection and action tower layers of the drone's detection system.

From Figure D3 if the drone is far from a system, the elevation angle “ θ ” of the propagation field can be hypothetically 0° which have a corresponding 90° electric field orientation (this value has too many limitations because of the range limitation of the detection system). As the drone is approaching a system, the practical range of the elevation angle will be from 0° to 45° . Based on this, the investigation can be limited to those angles.

The other factor that controls the coupling to the drone is the time response of the excitation pulse. Three different pulses, shown in Figure D4, have been used to excite the model presented in Figure D1.

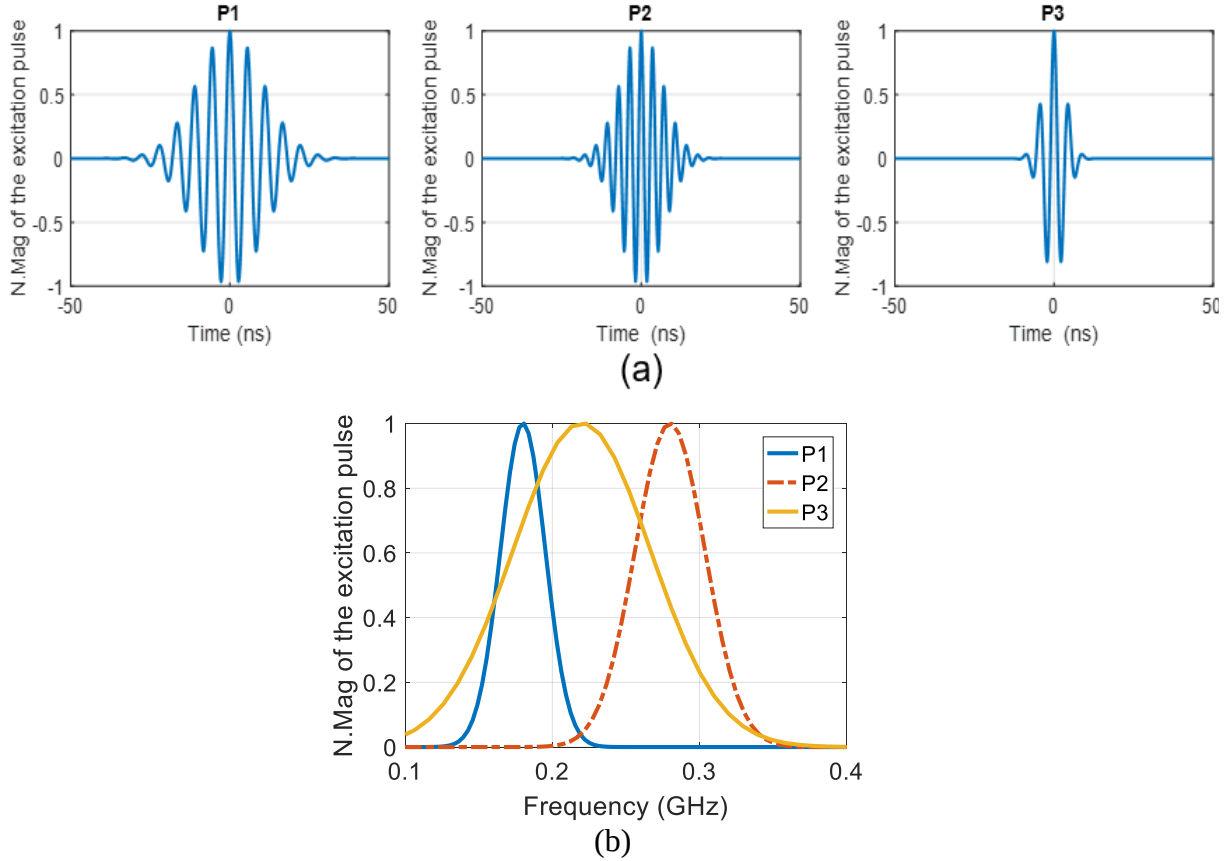


Figure D4. (a) Time-domain pulses used to excite the model in Fig.E1. (b) The corresponding frequency domain pulses.

The pulses were chosen to match the frequency range and the bandwidth of the first modes of the frame. The first pulse (P1) is intended to excite Mode 1, the second pulse is supposed to excite Mode 4 and the third pulse should excite both Mode 1 and Mode 4. As we illustrated in the previous MSR, the Modal fields will be the most important factor to control the coupling to the frame. The modal fields of the modes that will be studied herein, Mode 1 and Mode 4, are shown in Figure D5.

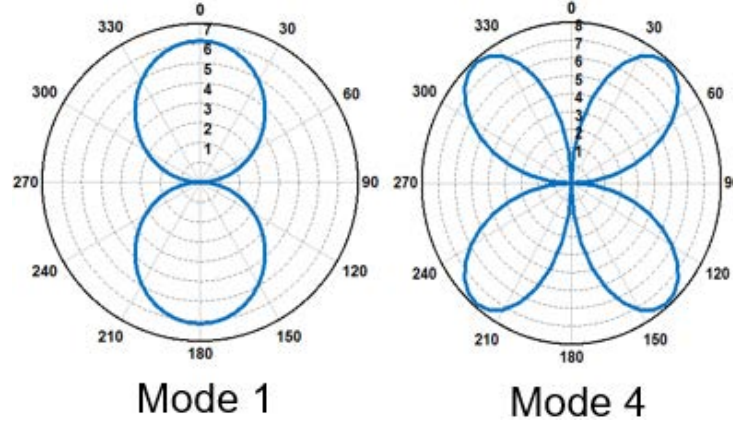


Figure D5. Modal fields of Mode 1 and Mode 4.

We utilized the knowledge of CMA predictions in determining the excitation pulse required to cause a specified effect. If the structure is excited by the first pulse (P1) the same pulse should be generated at the clamp point on the model in Figure D1. However, as Figure D5 illustrates, the maximum induced current at the model will occur when the elevation angle of the incident field $\theta = 0^\circ$ and will decrease gradually as θ increases. This is proven in Figure D6 that shows the induced current on the model for different incidence angles.

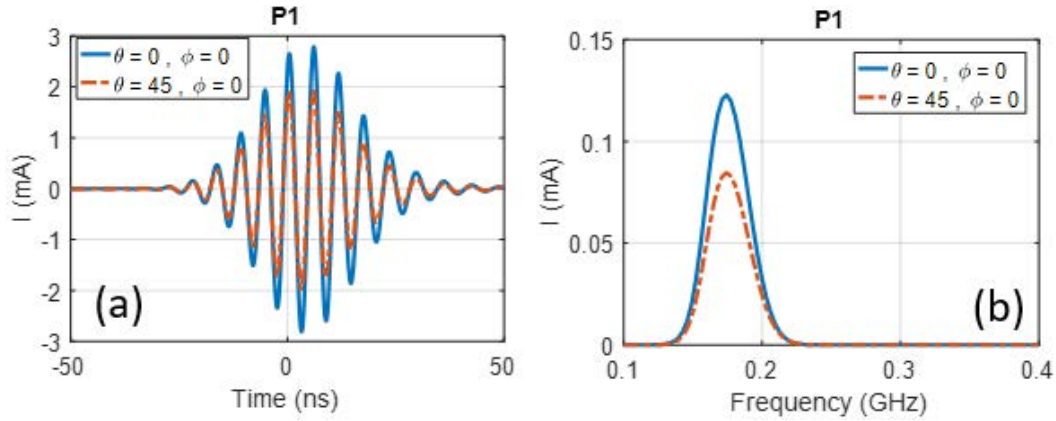


Figure D6. The induced current on the model when excited by the first pulse shown in Fig.E4(a).

On the other hand, as Figure D5 illustrates, the maximum induced current on the model at the resonance frequency of Mode 4 will occur when the elevation angle of the incident field $\theta = 45^\circ$ and will change according to the change in θ as in Figure D5. This is shown in Figure D7 which shows the induced current when the model is excited by the second pulse for different incidence angles. The induced current when $\theta = 0$ is very low because the modal field at this angle is almost zero and theoretically the induced current on the model should be zero. However, it has a finite value because the majority of the power of the incident signal is at this frequency.

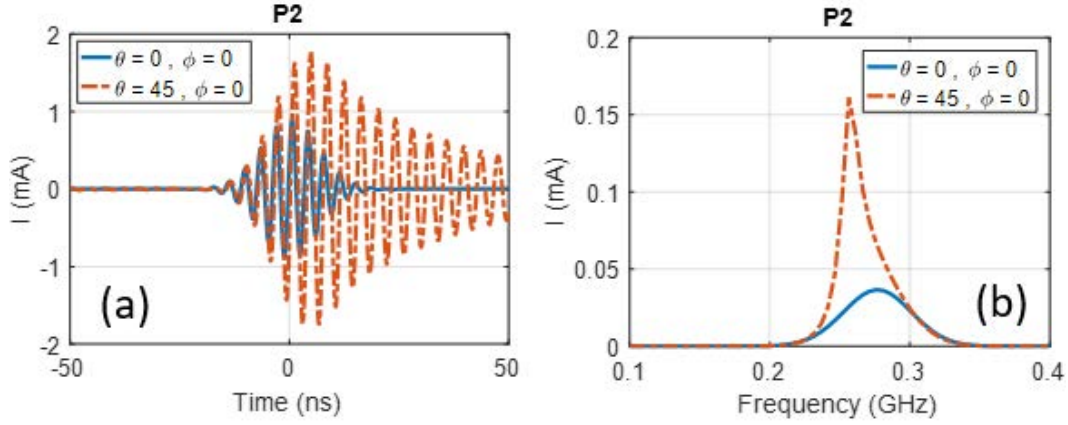


Figure D7. The induced current on the model when excited by the second pulse shown in Fig.E4(a).

If the third pulse, the wider pulse that can cover both Mode 1 and Mode 4, is used to excite the model then the incidence angle will not only affect the magnitude of the induced current but it will also determine which resonance will dominate the induced current as shown in Figure D8.

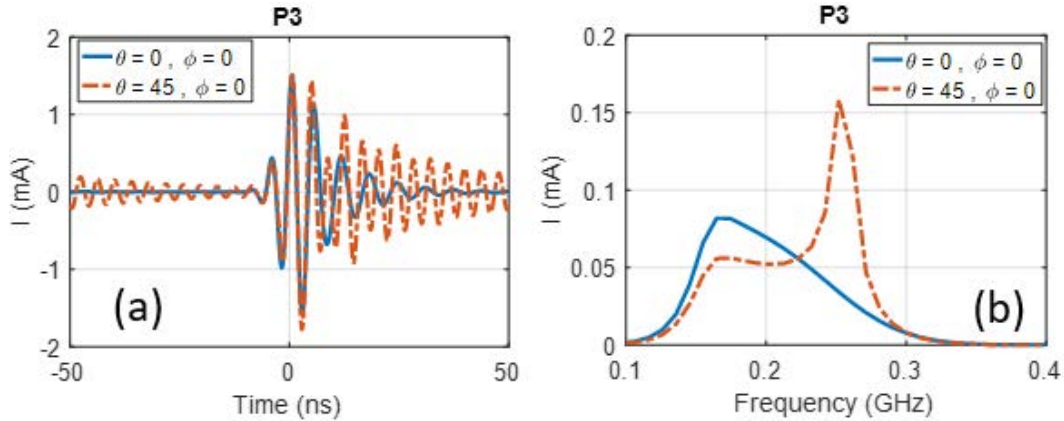


Figure D8. The induced current on the model when excited by the third pulse shown in Fig.E4(a).

When $\theta = 0^\circ$, according to the modal field in Figure D5, the current generated on the model is composed of only Mode 1 and no peaks at the resonance frequency of Mode 4 because the power of the incident power was not confined at this resonance frequency and the modal field is almost zero for this particular incidence angle. Once the field orientation is changed to have $\theta = 45^\circ$, Mode 4 is perfectly excited and it dominates the total response of the induced current. The induced current at the resonance frequency of the first mode reduced because of the effect of the modal fields shown in Figure D5.

In summary, this month we are proposing different scenarios to optimize the detection and action systems. If the low order modes are the most important mode to be affected then the first action layer is enough (if it can wait for the drone to be on the top of it) with narrow source bandwidth requirements.

If the high order modes are the targeted modes, then the first action layer can deal with it at far distances from it.

If both modes are targeted, either use a wideband source which will be costly and will not be able to introduce the coupling efficiency to all the modes simultaneously, not recommended, or design multiple narrow-band sources action layers each one targeting different frequency efficiently. For example in this exercise, it would be better if layer 1 can generate a pulse-like P1 and layer 2 generate P2 and both targeting the drone simultaneously.

Immediate Tasks for RF Coupling Subgroup:

- i. Study different scanned UAV frames with different practical dielectric properties including carbon fibers and Plexiglas.
- ii. Perform an exhaustive statistical analysis of coupling to randomly shaped loops above the UAV frames with dielectric properties.

E. UAS Developments (Fields)

See Contract MSR for the month of October for updates on UAS developments

III. Antenna Development (Crow)

F. Antennas - Electrically Small Antennas (Chatterjee/Barman)

Overview: In this report, two separate problems have been addressed. First is the element design and its optimization. For antenna element optimization, D/Q method has been investigated, where D and Q refer to the Directivity and Quality Factor of the antenna respectively. The quality factor (Q) of the designed microstrip patch antenna is calculated using various approximate formulas available in literature. Secondly, one of the most important limiting factors of array design, i.e., scan blindness has been investigated and a method to predict blind scan angles is studied.

Element Design and Optimization: Owing to their simple designs and tractable performance, microstrip patch antennas are widely used as antenna elements in a planar phased array. Electrically Small Antennas (ESAs) can be easily designed using rectangular patches by selecting a substrate of higher permittivity. Furthermore, a simple

rectangular patch can be used to obtain $> 30\%$ 2:1 VSWR bandwidth by strategically placing the probe. This concept has been utilized to design a patch at 600 MHz.

We have considered a rectangular patch with the probe located along the diagonal at $2/3^{\text{rd}}$ distance from the center. A 3.8 cm thick TMM 10i substrate ($\epsilon_r = 9.8$, $\tan \delta = 0.0020$) is used for the design. Other antenna parameters are enlisted in Table F1. Around 41% 2:1 VSWR bandwidth is achieved with 2.86 dB of maximum gain (on an infinite ground plane). Figure F1 shows the antenna geometry and its characteristics.

Table F1. Antenna Parameters

L (mm)	W (mm)	r_p (mm)	(x_p, y_p)	f_L (MHz)	f_C (MHz)	f_U (MHz)	% BW	Max. Gain (dB)	ka
57.6652	86.4978	3	(19.5, 28.8326)	478	601	724	40.93	2.86	0.6547

Antenna Optimization: The D/Q method is a multi-parameter optimization technique first hypothesized by Gustafsson [1], [2] and later investigated by Shahpari *et al.* [3]. Here D and Q stand for Directivity and Quality Factor of the antenna respectively. This method can be utilized to optimize both the far-field (Directivity) as well as the near-field (Quality Factor) parameters of an antenna simultaneously. Since the designed antenna has $ka < 1$, it can classify as a nearly electrically small antenna. Hence, to calculate the quality factor of such an antenna we have used different approximate formulas provided in the literature [4] (Table F2).

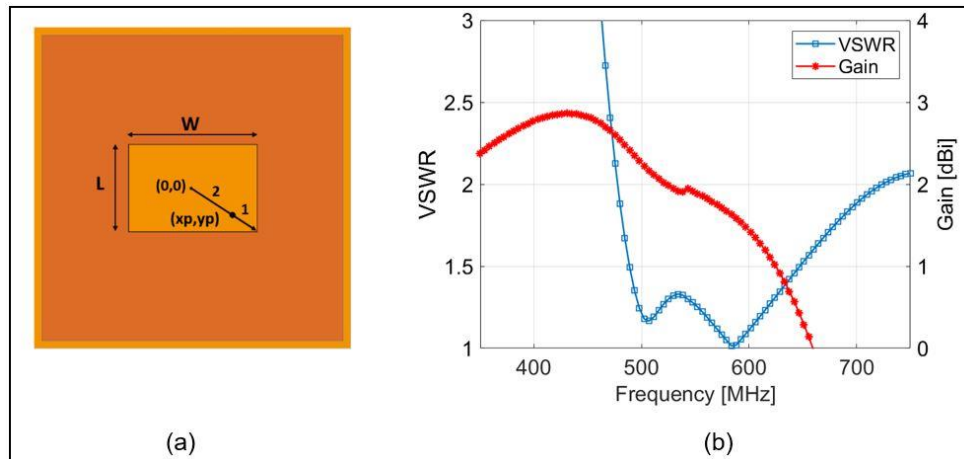


Figure F1. (a) Geometry of the microstrip patch antenna designed at 600 MHz (b) VSWR and gain of the designed antenna.

Table F2. Approximate Expressions for Quality Factor of an Electrically Small Antenna

Reference [4]	Minimum Quality Factor
Harrington	$\frac{1}{2} \left[\frac{1}{(ka)} + \frac{1}{(ka)^3} \right]$
McLean	$\frac{1}{2} \left[\frac{2}{(ka)} + \frac{1}{(ka)^3} \right]$
Thal	$\frac{1}{(ka)^3}$

The above formulas take into account only the electrical length of the antenna (ka). Quality Factor of the antenna can also be calculated using its input impedance as given by Yaghjian *et al.* [5]:

$$Q(\omega_0) \approx \frac{2\sqrt{\beta}}{\text{FBW}_v(\omega_0)} \quad (1)$$

where, $\text{FBW}_v(\omega_0) = \frac{\omega_- - \omega_+}{\omega_0}$ is the matched VSWR Fractional Bandwidth and β is given by,

$$4\beta R_0(\omega_0) = \frac{X_0^2(\omega_{\pm}) + [R_0(\omega_{\pm}) - R_0(\omega_0)]^2}{R_0(\omega_{\pm})} \quad (2)$$

The input impedance of the designed antenna is extracted from FEKO and is used to calculate the quality factor using eq. (1). The quality factor obtained using the input impedance is now compared with the other approximate relations given in Table F2. Figure F2 shows the comparison between different Q's of the microstrip patch antenna designed at 600 MHz.

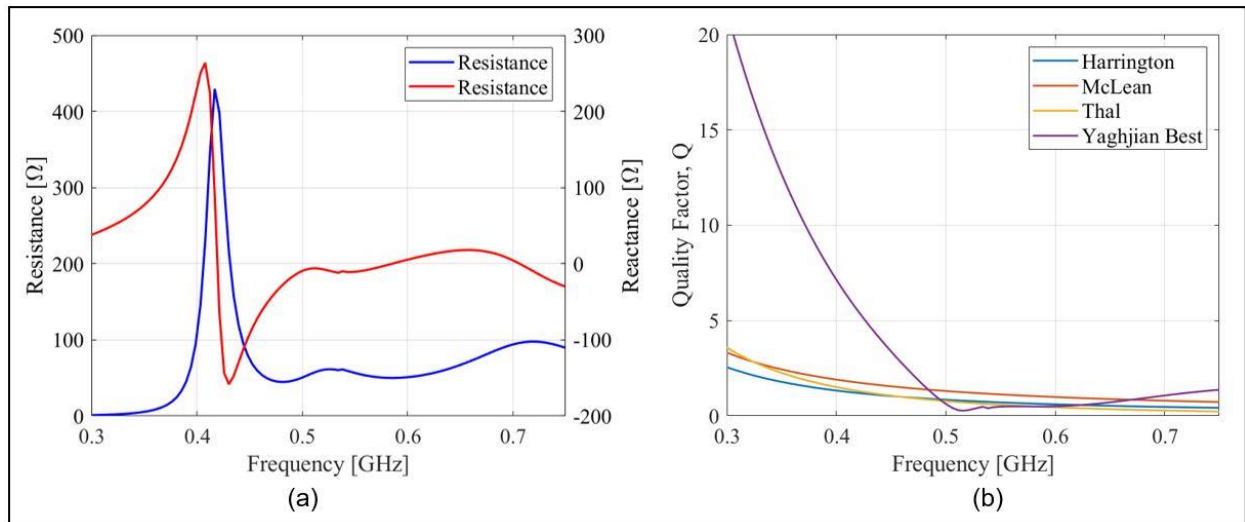


Figure F2. (a) Impedance and (b) Quality Factor of the designed antenna.

The directivity of the antenna will be obtained through FEKO and the D/Q ratio will be calculated. This value will be used to optimize the microstrip antennas and hence the overall array in the future.

Array Trade Space Analysis: Scan Blindness: One of the major objectives of the trade space analysis is to obtain the desired power density at a fixed distance from the Phased Array Antenna for different scan angles. The power density $S(\theta, \phi)$ of the array at a certain distance R can be given by [6]:

$$S(\theta, \phi) = \frac{1}{4\pi} \frac{P_{in}}{R^2} G^R(\theta, \phi) \quad (2)$$

where, P_{in} refers to the input power to the array and $G^R(\theta, \phi)$ is the array realized gain, which involves reflection as well as dissipative losses. The realized gain of the array depends on directivity $D(\theta, \phi)$ and reflection coefficient Γ as [6]:

$$G^R(\theta, \phi) = \epsilon_L (1 - |\Gamma|^2) D(\theta, \phi) \quad (3)$$

Eq. (3) points out that even if the array has a considerable directivity, the realized gain can be significantly reduced due to high reflection at a particular scan angle [7]. This phenomenon is referred to as scan blindness. Scan blindness limits the steerability of the array. For a transmitting array, the reflected power can incur severe damage to the transmitter if the beam is steered to the blind scan angle. Therefore, predicting the blind scan angles is of extreme importance.

A method to predict the scan blindness in an infinite planar array of dipoles has been presented in [8]. The dipoles, each of length L and width W , are placed in the X-Y plane at inter-element spacing along X-axis and b along Y-axis. The geometry of the infinite array is shown in Figure F3.

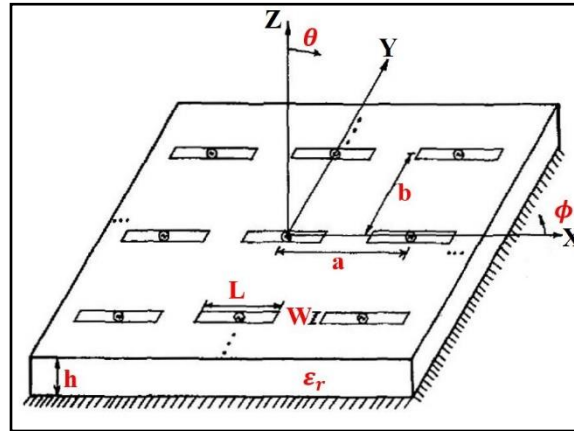


Figure F3. Geometry of the infinite array used in [8].

Scan blindness in the planar array can occur when the wavenumber β coincides with the surface wave propagation constant β_{sw} . This can be mathematically expressed as [8]:

$$\left(\frac{\beta_{sw}}{k_0}\right)^2 = \left(\frac{k_x}{k_0}\right)^2 + \left(\frac{k_y}{k_0}\right)^2 = \left(\frac{m}{a/\lambda} + u\right)^2 + \left(\frac{n}{b/\lambda} + v\right)^2 \quad (4)$$

where, $k_x = \left(\frac{2\pi m}{a} + k_0 u\right)$ and $k_y = \left(\frac{2\pi n}{b} + k_0 v\right)$ (5)

$$u = \sin \theta \cos \phi \quad (6)$$

$$v = \sin \theta \sin \phi \quad (7)$$

The R.H.S of eq. (4) refers to the Floquet mode propagation constant γ_z , which should be real and positive for the scan blindness to occur. It is given by:

$$\gamma_z^2(\theta, \phi) = k_0^2 - \Gamma_{mn}^2(\theta, \phi) \quad (8)$$

where,

$$\Gamma_{mn}^2(\theta, \phi) = \left(\frac{m}{a/\lambda} + u\right)^2 + \left(\frac{n}{b/\lambda} + v\right)^2 \quad (9)$$

This method is implemented for a 10×10 planar array of the microstrip patch antenna designed at 600 MHz. The surface wave propagation constant β_{sw} is calculated as follows [10]:

$$\beta_{sw} = \beta_r - j\beta_i \quad (10)$$

where,

$$\frac{\beta_r}{k_0} = 1 + \frac{\alpha_0 \alpha_1 - \epsilon_r^2 + \epsilon_r \sqrt{\epsilon_r^2 - 2\alpha_0 \alpha_1 + \alpha_0^2}}{\epsilon_r^2 - \alpha_1^2} \quad (11)$$

and,

$$\beta_i = (\epsilon_r - 1) \tan \delta (k_0 h s)^2 \quad (12)$$

$$s = \sqrt{\epsilon_r - 1} \quad (13)$$

$$\alpha_0 = s \tan(k_0 h s) \quad (14)$$

$$\alpha_1 = -\frac{1}{s} \left[\tan(k_0 h s) + \frac{k_0 h s}{\cos^2(k_0 h s)} \right] \quad (15)$$

For a lossless medium ($\tan \delta = 0$), the value of $\frac{\beta_{sw}}{k_0}$ is calculated to be 1.1650 at 478 MHz. A solution for eq. (4) is possible in the E-plane ($\phi = 0^\circ$) when $m = -1, n = 0, u = \sin(25.19^\circ) = 0.4256$. However, the Floquet Modes are not excited as γ_z remains imaginary for all scan angles. Hence, the scan blindness is not observed at the low frequency (478 MHz). Figure F4 shows the gain vs. scan angle plot for the antenna at three different frequencies.

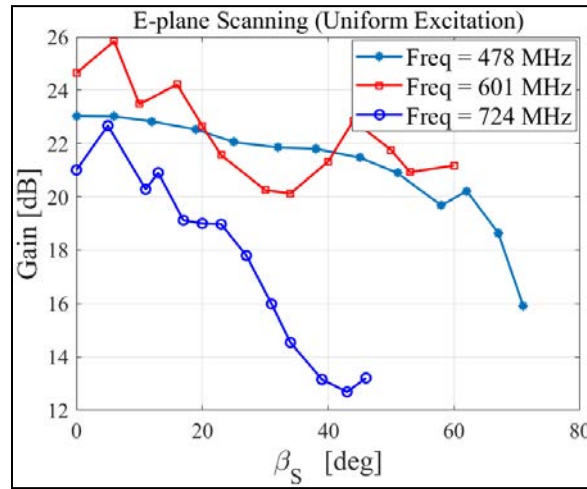


Figure F4. Gain vs. scan angle of the antenna at the center (601 MHz) and two edge (478 MHz and 724 MHz) frequencies.

Summary: A method to optimize the antenna element and hence the entire array has been studied. So far, the quality factor of the antenna has been calculated using its input impedance and other approximate formulas. Further investigations are required to completely obtain the D/Q ratio which can be utilized to optimize the overall antenna.

A method to predict the scan blindness in an array has been investigated. No scan blindness is observed at lower frequencies as the Floquet modes are not excited. Prediction of scan blindness using the reflection coefficient values of the antenna array will be looked into in the future.

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G. Antennas - Log Periodic (Hassan)

We have been exploring different antenna elements that have the potential of satisfying the new design requirements for the program. While frequency-independent log-periodic antennas provide excellent bandwidth at a reasonable size, their time-domain performance for short pulse applications is limited due to dispersion which causes the radiated pulse to be different from that applied to the input of the antenna [1]-[2]. Therefore, we moved on to the helical antenna as a more promising element to achieve the program requirements. A drawback of the helical antenna is the smaller bandwidth and larger size, but advantages are controllable gain, design flexibility, and large power handling capability [3]. In this MSR we summarize the theory behind helical antennas, simulate a few designs, validate the accuracy of our simulations with results previously reported in the literature, and report early antenna prototypes designed to meet the program requirements.

Helical Antenna Design

A helical antenna is shaped by winding a wire with thickness d into a diameter D with a pitch angle α . The pitch angle (α) controls the growth of the wire winding forming a right triangle between the distance between turns (S), helix circumference (C), and the length of one turn (L_0), as shown in Figure G4. The height of the antenna is $L = NS$, whereas the length of wire required is $L_n = NL_0 = N \sqrt{S^2 + C^2}$ where L_0 is the length of one turn. The angle between turns is $\alpha = \tan^{-1} \left(\frac{S}{\pi D} \right)$. Note when $\alpha = 0^\circ$ the helical is flat reducing into a loop antenna, and when $\alpha = 90^\circ$ the helical is extended into a wire radiator that is a helical antenna exists when $0^\circ < \alpha < 90^\circ$.

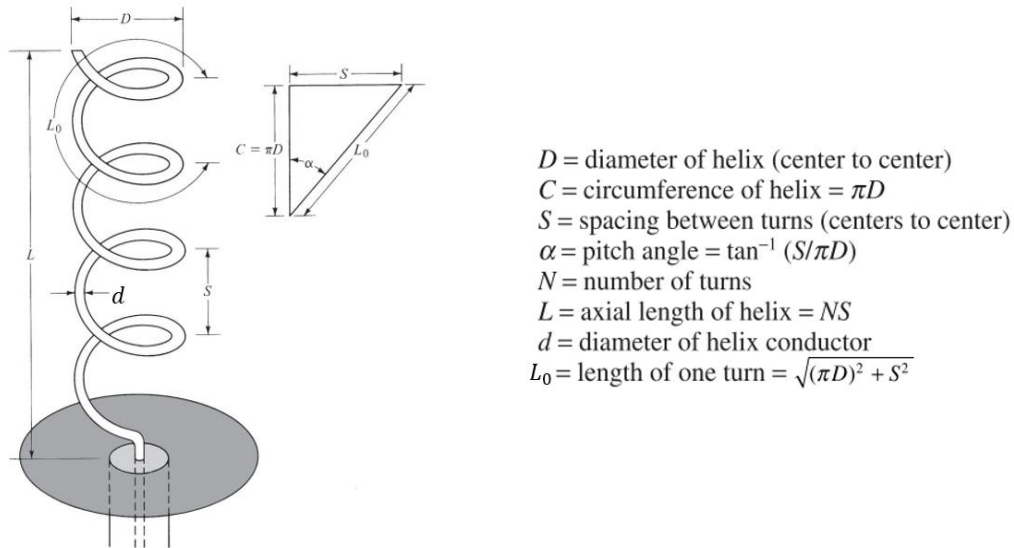


Figure G4. Helical antenna description [4].

The advantages of the helical antenna are: (i) wide bandwidth (45-55%), (ii) controllable beam width (20-60°) and gain (9-19dBi), (iii) high power handling capability, (iv) low complexity and good tolerance to manufacturing variation. Depending on its dimensions, a helical antenna can operate in one of multiple mode with different far-field radiation patterns three of which are displayed in Figure G5, antenna behaves similar to a monopole antenna, the axial mode of operation where the major lobe radiation is in the axial direction (also called end-fire mode), and conical which can be viewed as a combination of the two operating modes with multilobe radiation skewed towards the forward direction.

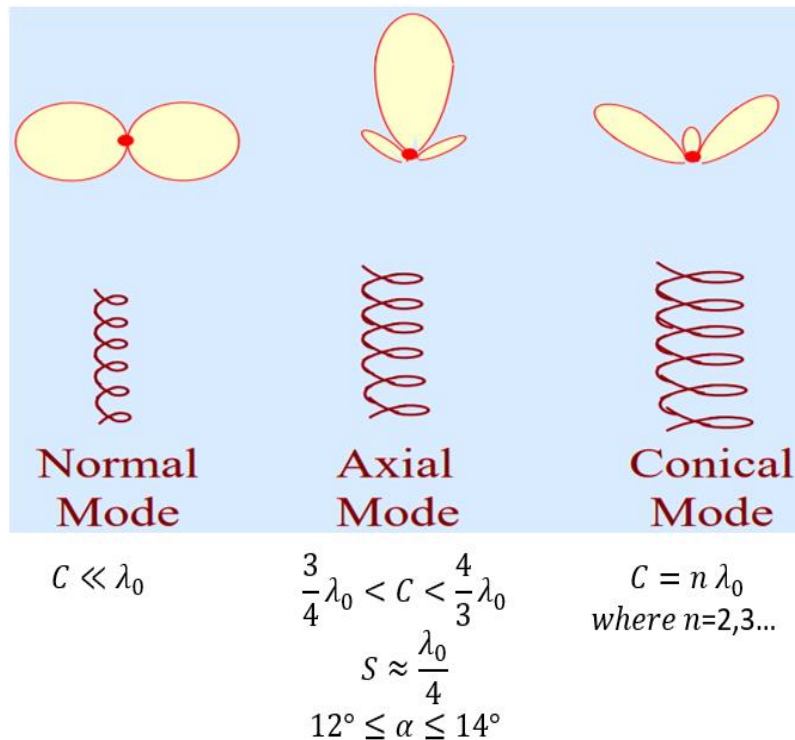


Figure G5. Modes of operation of a helical antenna: normal, axial, and conical [2].

As a validation step, we re-simulated a previously published helical antenna with the dimensions shown in Figure G6, where $N = 15$ turns. The antenna was simulated using FEKO using a coarse mesh with 551 points [8]. The two plots shown in Figure G7 compare the UMKC simulation to the published result they show excellent agreement. The small deviation is believed to be attributed to the definition of feeding port used and meshing differences. The paper does not report a port type, but the UMKC simulations used a wire port. Given the close result attained, we feel comfortable with the validation to move to design specific antennas for this project.

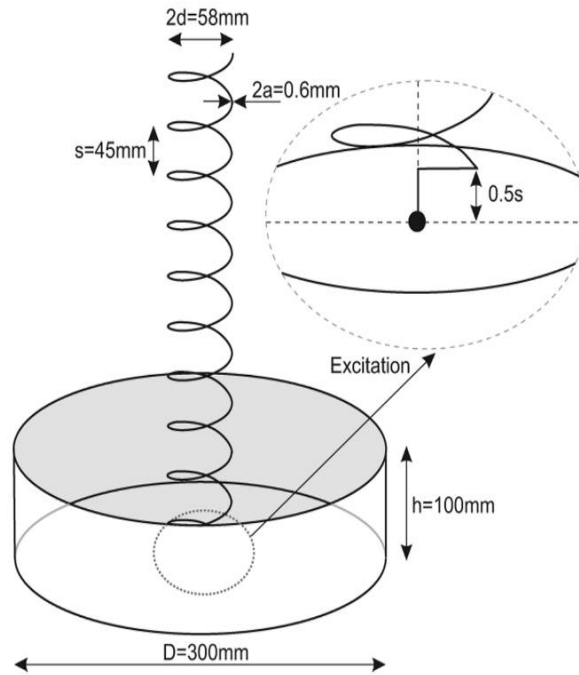


Figure G6. Helical antenna with a cylindrical cup ground plane (reflector) [4]

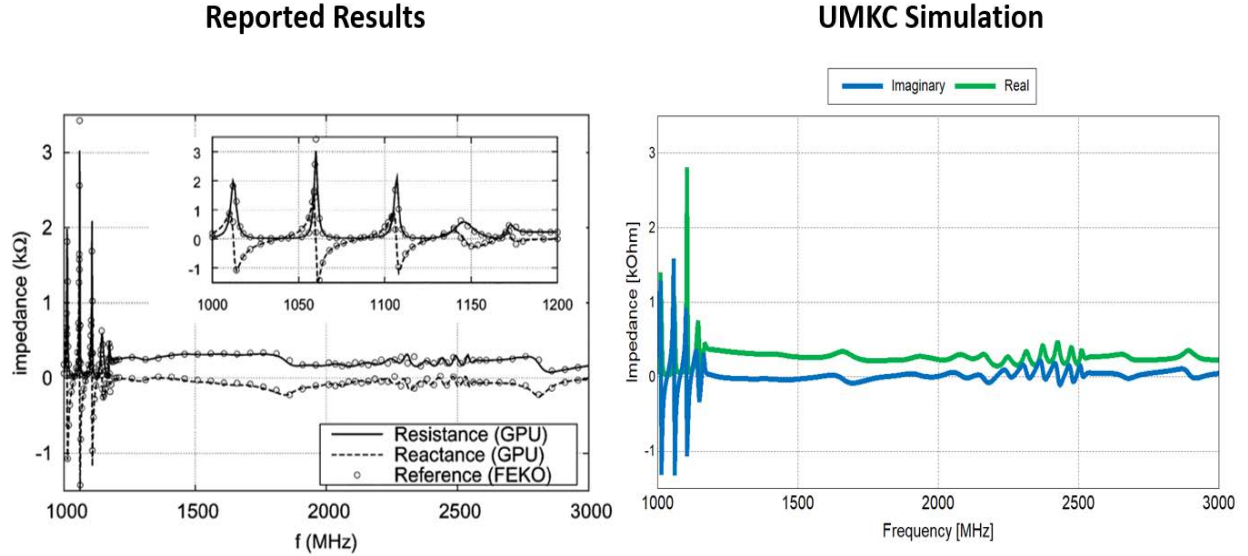


Figure G4. Validation the accuracy of our simulations. (a) Reported input impedance of a helical antenna, and (b) Recreated input impedance of the same antenna [8].

Using the design methodology reported in [6], we designed three antennas to operate at the center frequencies of 400 MHz, 750 MHz, and 1000 MHz. Instead of the conical ground plane reported in the papers, we used a flat ground plane as a first step. Different ground plane configurations will be studied in future MSRs. The dimensions of the designed antennas are summarized in Figure G5. Figure G8 shows the Voltage Standing Wave Ratio (VSWR) of the 400 MHz antenna and Figure G7 shows the corresponding gain. The antennas in Fig. 5 do not satisfy all the requirements for this program yet, but they are a good starting point for further improvements. In future MSRs we will use different optimizations techniques, such as the genetic algorithm and Multi-Objective Optimization (MOO), to improve and optimize the helical antenna designs.

Parameter	Description	400 MHz	750 MHz	1000 MHz
N	Number of Turns	6	8	8
D	Wire Diameter	3 mm	3 mm	3 mm
Dg	Ground Plane Diameter	749.5 mm	400 mm	300 mm
D	Helix Diameter	238.5 mm	127.2 mm	95.43 mm
Hf	Feed Height	30 mm	16 mm	12 mm
α	Pitch Angle	13°	13°	13°

Figure G7. Dimensions of designed antennas

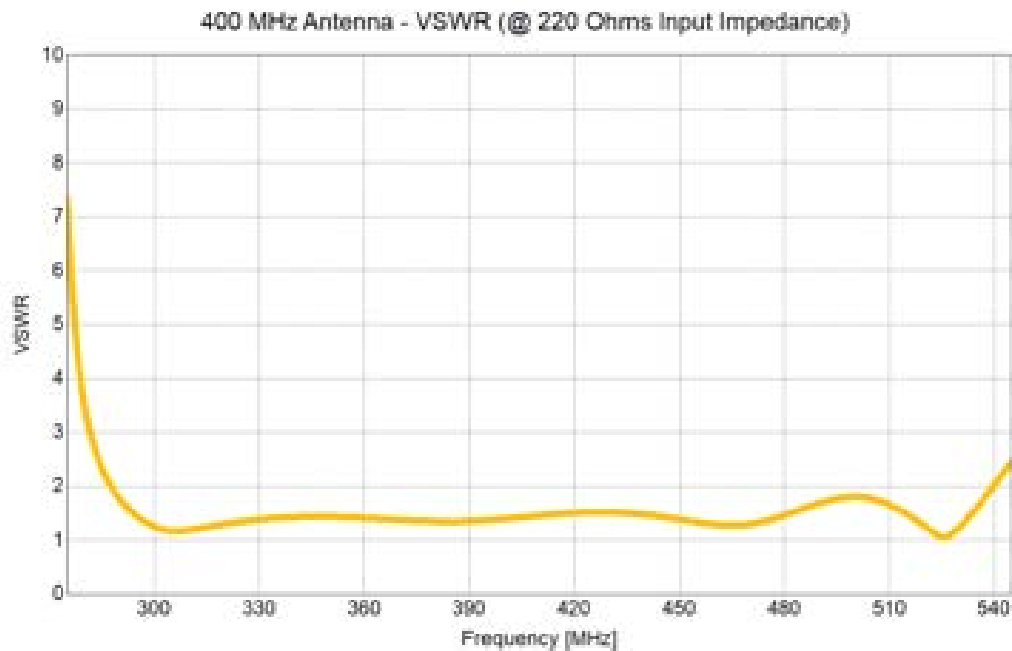


Figure G8. Designed antennas input impedance and non-50-ohm VSWR

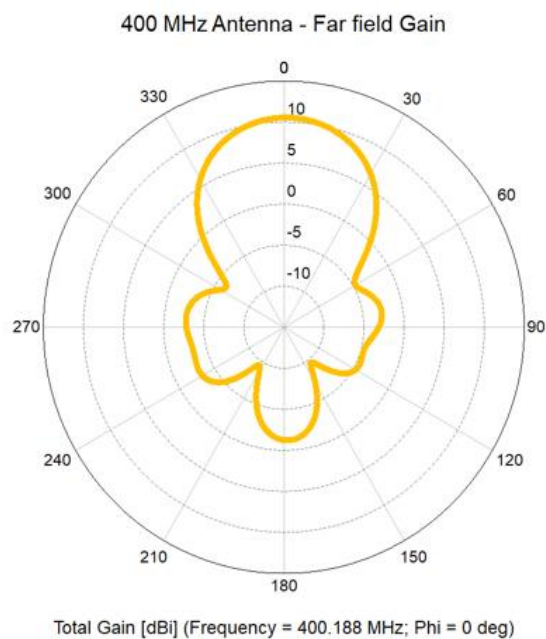


Figure G9. Designed antennas boresight Gain at operating frequency

Immediate Tasks for the Log Periodic Antenna subgroup

- i. Explore design parameters and the different ways to optimize the helical antenna
- ii. Come up with a trade space analysis of antenna design parameters
- iii. Explore a systematic and repeatable method to improve input impedance mismatch

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IV. Research and Development Support (Smith)

H. Device and Materials Modelling (Eifler)

Over the reporting period, the SiPCSS simulation and modeling team has focused on replicating reverse recovery testing currently being conducted on SiPCSS devices to improve the accuracy of transient SiPCSS models in Silvaco TCAD. Until recently, the reverse recovery time of transient models of the SiPCSS were less than those measured empirically by as much as an order of magnitude. An example of the reverse recovery profiles simulated previously are shown in Figure H1, along with a recovery profile measured on a device irradiated at 20 MR. In operation carrier generation occurs primarily via photo-absorption in the device, whereas in reverse recovery time testing, carriers are injected during the forward bias. Despite these differences, the simplicity of the reverse recovery time serves as a comparison tool to improve fidelity of recombination models used in TCAD. However, the default recombination models have been insufficient in replicating the recombination process in real devices.

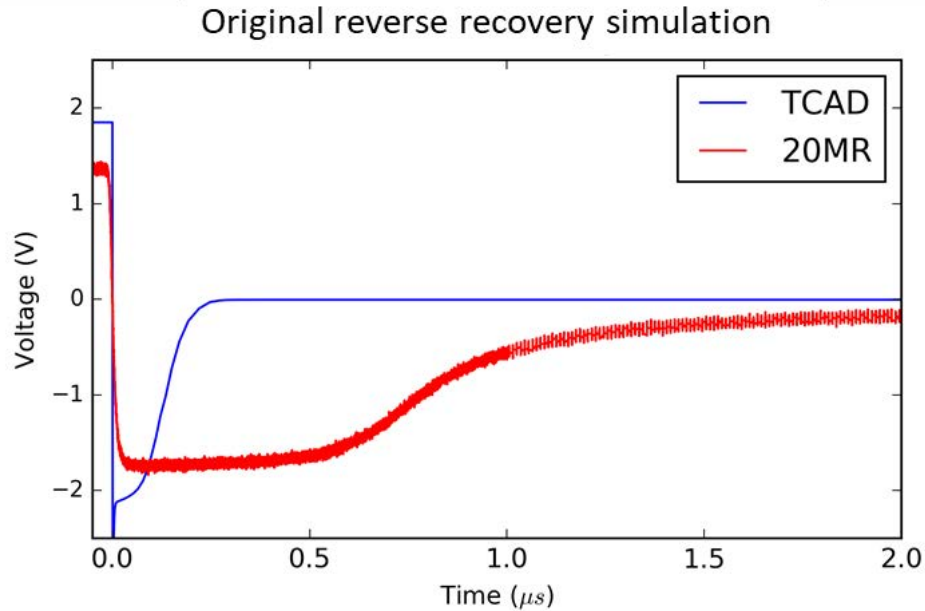


Figure H1. Original simulated reverse recovery profile compared to experimental data collected on a device irradiated with 3 MeV electrons (20 MR dose). Reverse recovery time differed between experiment and simulation by nearly an order of magnitude.

To overcome these differences, several reverse recovery simulations were conducted with the addition of traps and compared to experimental lab tests. Lifetime within simulation was adjusted by changing the defect concentration of traps with specific energy states corresponding to electron-irradiation induced defects. The results of these simulations, including a measurement from experiment, are shown in Figure H2. Four defect concentrations distributed uniformly throughout the device were simulated: (yellow) 3×10^{12} , (orange) $7.8 \times 10^{12} \text{ cm}^{-3}$, (grey) $9 \times 10^{12} \text{ cm}^{-3}$, and (dark blue) $13 \times 10^{12} \text{ cm}^{-3}$. A measured reverse recovery profile of a 10-MR device was also included for comparison (light blue). Values were selected based on trial and error in order to be representative of the experimental results from irradiated device.

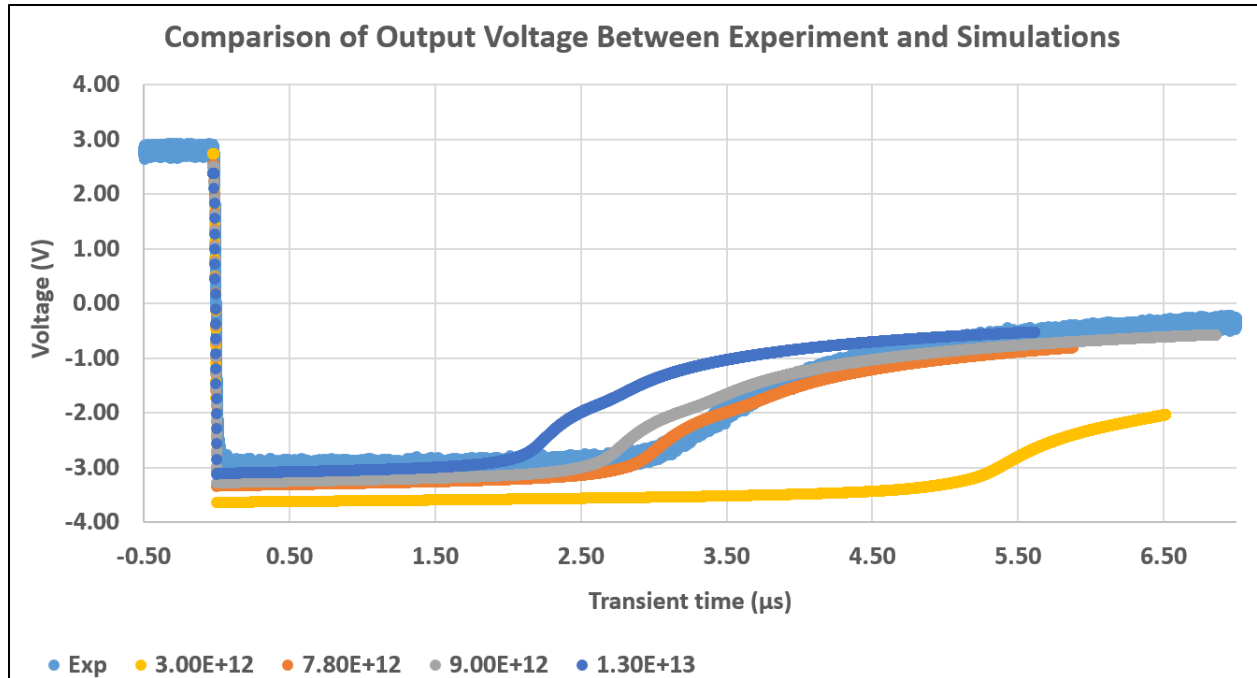


Figure H2. Reverse recovery profiles for experiments and simulations for various defect concentrations. Legend corresponds to defect concentration.

While there are nuanced differences in the measured and simulated reverse recovery profile (e.g., slightly increased slope during the charge storage time, or flat top, and greater reverse current in simulations), all simulated profiles more accurately represent experiment than previous models. Of the defect concentrations simulated, a concentration of 7.8×10^{12} defects/cm³ (orange), the charge storage time most accurately matched experiment.

While further refinement of the defect states and concentrations will be made, a model representative of the measured recovery profile has been generated. Additional simulations are needed to generate estimates of the defect concentrations as a function of dose in order to optimize the dose and electron energy. Moreover, now that a model has been generated which approximates the reverse recovery profile of real devices, additional modeling to optimize critical design parameters can begin.

Critical Actions for SiPCSS Simulation and Modeling Subgroup:

- i. Determine defect concentrations for devices irradiated with 3 MeV electrons at a dose ranging from 1 MR to 20 MR.
- ii. Improve the fidelity of the SiPCSS reverse recovery model and incorporate non-uniform defect states, including lower energy electron irradiation to target the device junction.
- iii. Use SiPCSS model to evaluate the performance parameters (e.g., on-state resistance, recovery time, required optical energy)

I. GaN:C Calculation (Rulis)

This month, students were able to compute XANES/ELNES spectra for the neutral state defect models of GaN:C, which were produced last month. For the charged state defect models, the current methods for XANES/ELNES calculation is not possible without further development. We explored possible routes for computing charge-state XANES/ELNES spectra which would push the state of the art in this area.

Some of the spectra that were computed for the bulk crystal components were compared to spectra that had been computed in prior publications [1]. An example is shown in Figure I1.

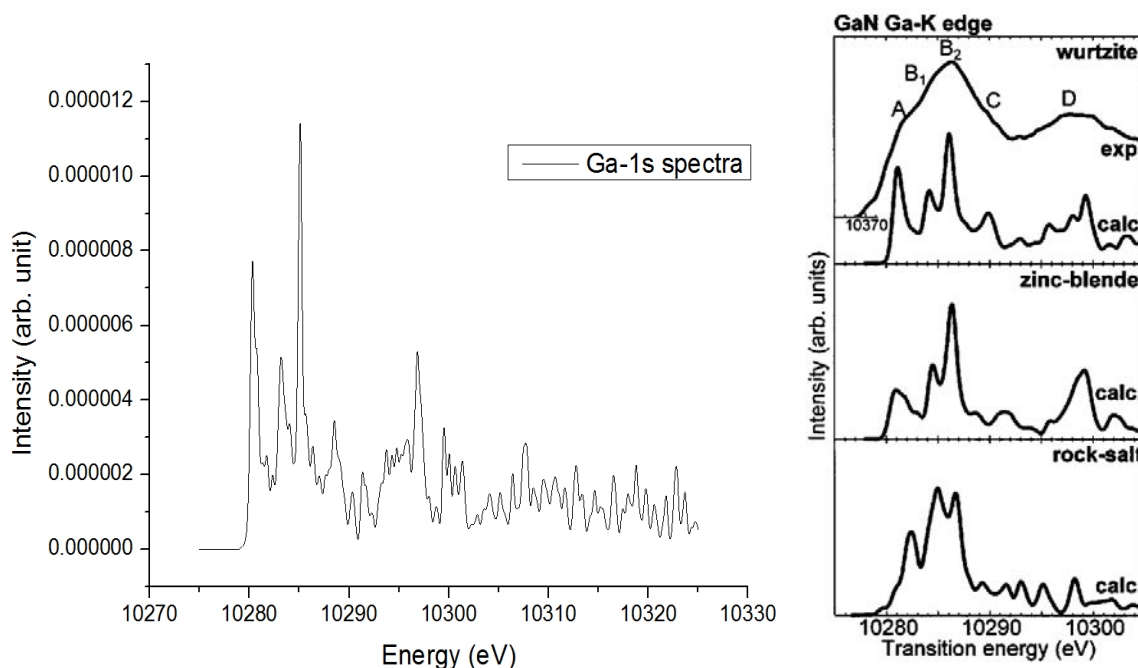


Figure I1. Left: Recently completed computation of Ga-K edge ELNES/XANES spectrum; Right: computed and measured Ga-K edge spectrum from different GaN crystal forms. The first ~40 eV of the spectrum on the left agrees with the top computed.

However, Ga and N spectra are of limited value when trying to understand the nature of C dopants in the GaN matrix because of the overwhelming bulk signal versus any influence on Ga or N due to the small number of C dopants. Hence in the next figure we have discovered literature [2] that includes (but does not focus on) the C-K edge in GaN. Our computed spectra compares favorably, but the resolution and background signal in the experimental data make detailed comparison difficult. In Figure I2, there is a peak appearing at 290 eV which is due to the presence of unintentionally impure carbon that is the result of the fabrication process.

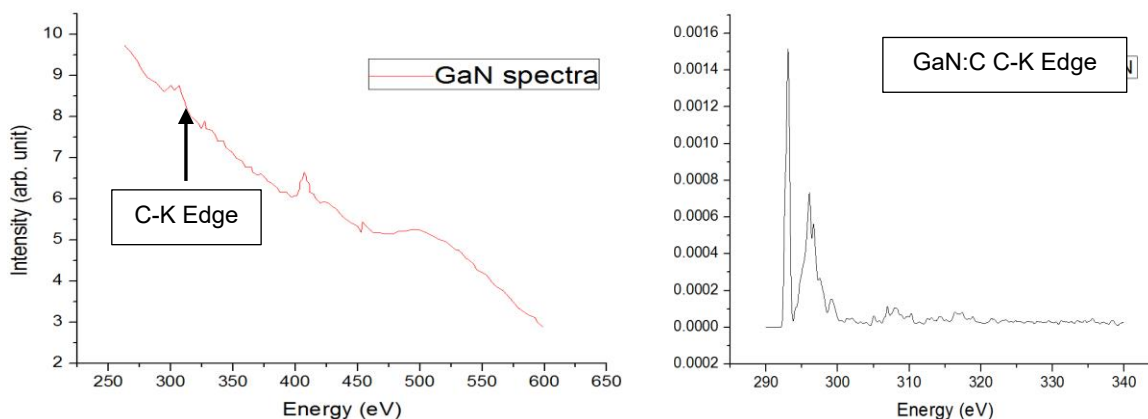


Figure 12. Left: Experimental ELNES spectrum of GaN:C showing C-K and N-K edges. Right: Compute GaN:C C-K edge computed from neutral.

Additionally, we have taken pristine and used samples of the GaN:C PCSS device switches and started coordination with a collaborator that will obtain experimental ELNES C-K edge spectra. Some special sample preparation work will need to be performed to obtain high quality C-K edge spectra because the samples GaN:C is grown on a thin AlN layer that is on a SiC substrate. The influence of the SiC C-K edge must be accounted for. This activity is ongoing.

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Issues:

[1] **Scope** – N/A

[2] **Budget** – N/A

[3] **Schedule** – N/A

Interactions with Others:

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

Major Procurement Actions: N/A**Travel:****Travel Plans**

Please note that all papers or presentations on any aspect of the ONR HPM program must be reviewed by the ONR Program Officer (or in some cases ONR Public Release Process) regardless of classification level or audience.

Planned Travel Details	Purpose	Attendees	Estimated Costs
SARA		Matt Wisher	

Pending or Upcoming Public Release Requests:

N/A

ONR HPM 2019 Program - Monthly Status Report (MSR) – NOVEMBER2019
Anthony Caruso – Univ. Missouri – Kansas City
N00014-17-1-3016 [OSPRES Grant]
ONR Short Pulse Research, Evaluation and non-SWaP Demonstration for C-sUAV
Study

12/15/2019

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OSPRES Project Description: Measure, analyze and summarize the optimal high power microwave waveform (pulse width, pulse repetition frequency, center frequency, polarization, bandwidth) that simultaneously yields the lowest power density AND lowest possible size, weight, monetary cost and cooling high power microwave system (i.e., driver and radiator) that can produce such waveforms. Modularity and scalability are also critical requirements of the overall design/downselect criteria, that are considered as part of the tradespace study analysis (i.e., within the context of the optimal waveform).

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Executive Summary:

- Analysis shows that varying the materials used for UAV frames can significantly affect electronic speed controller (ESC) loads. Conducting materials result in upwards of 300% increase in coupled current versus dielectric frames. Increased coupling current is an increase in coupling efficiency reducing the amount of incident RF power from a source necessary to induce coupling. A database correlating frame materials to effects of/from the frame on coupled current and interference with UAV wires and components is being populated from which a validating test plan can be constructed.
- Commercially available pulse charger options based on Si switches fail to perform at 2-kV with 50-kHz switching rate. Two new designs utilizing SiC have been purchased and will be integrated and tested when received.
- Electrically small antenna simulations show necessary power input into an array to achieve various power densities up to 5-W/cm² at a distance of 1-km based on the gain of an individual antenna. At a gain of 30dB all power densities between 2 and 5-W/cm² require more than 200 MW of input power. At 35 dB and above all power densities require less than 200 MW of input power. A suitable gain value to strive for in designs and optimization may be 35 dB to reduce input power required of the sources.

Current Month Progress:

I. Source Development (Myers)

A. PCSS Power Conditioning - IGBT (Khan)

- Experimentally verified that COTs HV Si switches fail to perform at 2-kV with 50-kHz switching rate due to built-in over temperature protection. New SiC-based HV switches have been purchased and will be integrated and tested when received.
- We have conducted a literature study to come up with a new HV switch topologies as an alternative to COTs switches.
- We have adopted two tentative designs for HV switches and began initial simulations.

B. Switches - GaN:C (Thompson)

- Reduced PCB capacitance to validate reduced recovery time effects pulse shaping (reduces recovery time).
- Continued writing manuscript draft for generating long-term cost and feasibility assessment of GaN-based switching technologies: *Can Semi-insulating GaN be Cost Competitive?* Now 60% complete.
- Researched manufacturing methods and cost trends of SiC for manuscript.

C. Switches - Figure of Merit (Bhamidipati)

- Initial power dissipation and temperature rise during turn-on and turn-off losses have been calculated for different materials to contribute towards material FoM as a function of electric field, geometry, pulse width and optical intensity.
- Application specific requirements were proposed and are being weighed in the order of their influence on the device performance factors for the application specific FoM.
- FoM manuscript progress, now at 70% completion.

II. Effects, Coupling and UAS Developments (Currie)

D. RF Coupling (Hassan)

- A simple wire configuration that emulates the connections between the Electronic Speed Controller (ESC) and the UAV microcontroller above frames composed of different practical materials was studied. UAV frames composed of carbon fibers, Plexiglas, Nylon66, Polyethylene, and perfect electric conducting frames were investigated.
- Analysis shows that the ESC loads, due to their impedance and relative location, are significantly affected by the modes of the UAV frame. The current coupled to the ESC at the resonance frequency of the conducting frame can reach 300% of the current coupled to the ESC in air or above a dielectric frame.

- As a result of these tests, a database of frame materials quantifying the effect of the frame on coupling and interference with UAV wires and components is being populated.

E. UAS Developments (Fields)

- Expanding capability of dogfighting simulation system to include larger multi rotor vehicles.
- Developed a test plan to provide initial indications of UAV degradation from GTEM test environment.
- Finished construction, software loading, and flight checks for 30 UAV assets that were shipped to NM for testing in January.

III. Antenna Development (Crow)

F. Antennas - Electrically Small Antennas (Chatterjee/Barman)

- *D/Q Element Optimization Technique* [1]: A method to optimize the antenna performances using the multi-parameter optimization technique (also known as the D/Q method) has been investigated. Using this technique it is possible to optimize the far-field (Directivity) and near-field (Quality Factor) parameters of the antenna simultaneously.
- *Calculation of Quality Factor*: Two types of Quality Factors are used: the exact Quality Factor of the antennas (calculated using the antenna input impedance [2]) and the approximate Quality Factor (applicable for ESAs [3]). Figure F1 compares some of the models for quality factors as a function of frequency.

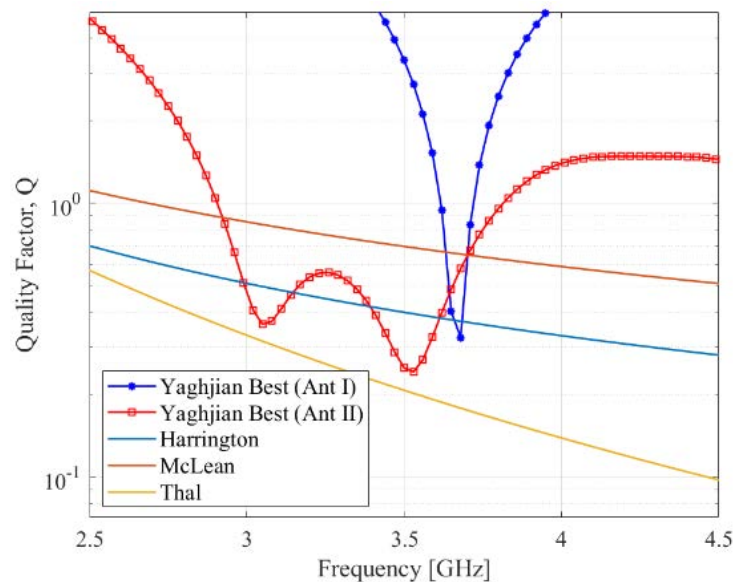


Figure F1. Comparison of Exact and Approximate Quality Factors of the antenna

- **Array Trade Space Analysis:** An investigation into the required input peak power and the array gain to achieve the desired power density is in progress. Fig. F2 shows the preliminary results for this comparative study.

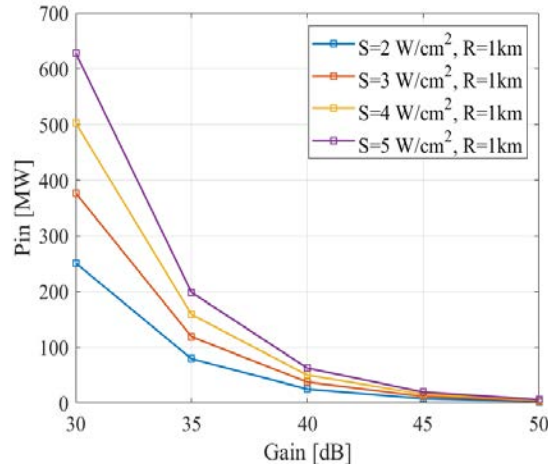


Figure F2. Peak input power to the array as a function of Array Gain

Reference:

- [1] M. Gustafsson, C. Sohl, and G. Kristensson, "Physical limitations on antennas of arbitrary shape," *Proceedings of the Royal Society A: Mathematical, Physical and Engineering Sciences*, vol. 463, issue 2086, 2007, pp. 2589-2607.
- [2] A. D. Yaghjian and S. R. Best, "Impedance, Bandwidth, and Q of Antennas," *IEEE Trans Antennas and Propag.*, AP-53, 4, April 2005, pp. 1298- 1324.
- [3] J. L. Volakis, C. C. Chen, and K. Fujimoto, *Small Antenna Miniaturization Techniques and Applications*. New York, NY, USA: McGraw-Hill, 2010.

G. Antennas - Log Periodic (Hassan)

- Designed three helical antennas and iterated on their designs to meet program requirements.
- Performed an exhaustive review of high-power microwave antennas and are currently preparing a summary of this review.

IV. Research and Development Support (Smith)

H. Device and Materials Modelling (Eifler)

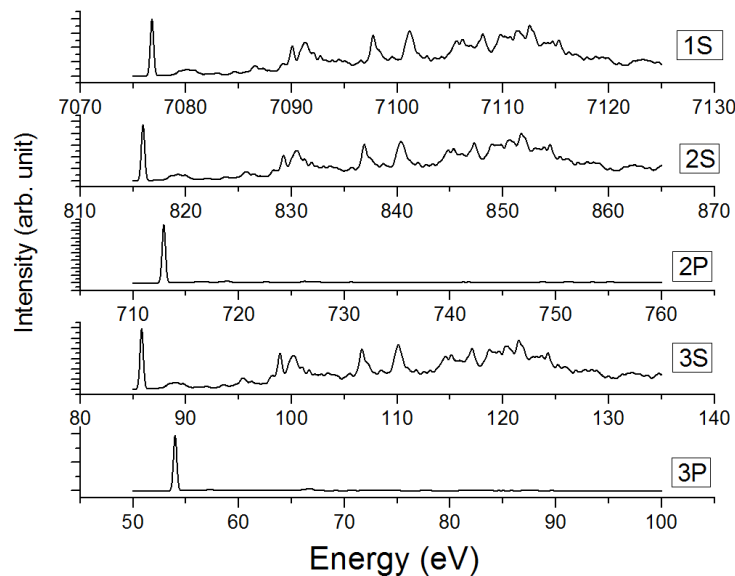
- TCAD simulation of the PCSS to full recovery of on-state resistance (R_{on}) to initial R_{on} was performed. Earlier simulations did not recover to R_{on} since a DC voltage was continually placed across the PCSS. Therefore, we disconnected the power supply and used transient simulation to charge and re-charge the transmission lines within the PCSS external circuit to accomplish the recovery. These

simulations enable future simulations for two pulse testing which yields pre- and post-pulse characteristics.

- Impedance spectroscopy experiments are compared to corresponding TCAD simulations with good agreement. Simulations allow a connection to the underlying device properties for a more detailed understanding of the device characterization for which impedance spectroscopy is used. New experiments with modification to the experimental test fixture and more detailed simulation data output will be performed to characterize PCSS.
- TCAD simulations for varying defect profiles were performed. Simulations showed tailoring of the charge storage time and forward and initial reverse voltages occurred in relation to the simulation defect profile. With a better understanding of the defect profiles present in actual devices, future simulations will be quantitatively compared to experiments for the effect of the defect profile.

I. GaN:C Calculation (Rulis)

- Students continued the use of VASP+PyCDT to produce a range of defect models using Fe compensated GaN: (substitutions, interstitials, different charge states, and vacancies in standard hexagonal and also cubic GaN). For many of the models the students were able to compute core level spectra (ELNES spectra). Charge defect models were created, but computing ELNES spectra of such models is not straightforward requiring a significant amount of effort going forward.



- Students worked with other team members (Eifler) to evaluate the capabilities of Silvaco TCAD and determine its capability for calculating device properties with the list of parameters that we will generate from *ab initio* methods. For C and Fe doped GaN, new scripts will be developed to compute the dielectric constant, thermal conductivity and breakdown voltage for C-GaN and Fe-GaN, and will be explored in the next round of simulations.

Issues:*[1] Scope – N/A**[2] Budget – N/A**[3] Schedule – N/A***Interactions with Others:**

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

Major Procurement Actions: N/A**Travel:****Travel Plans**

Please note that all papers or presentations on any aspect of the ONR HPM program must be reviewed by the ONR Program Officer (or in some cases ONR Public Release Process) regardless of classification level or audience.

Planned Travel Details	Purpose	Attendees	Estimated Costs

Pending or Upcoming Public Release Requests:

N/A

ONR HPM 2019 Program - Monthly Status Report (MSR) – DECEMBER2019
Anthony Caruso – Univ. Missouri – Kansas City
N00014-17-1-3016 [OSPRES Grant]
ONR Short Pulse Research, Evaluation and non-SWaP Demonstration for C-sUAV
Study

01/15/2020

Distribution Statement ~A: Distribution authorized to all OSPRES Grant and Contract Participants, and any uncompensated collaborators. Further dissemination per Ryan Hoffman (ryan.hoffman@navy.mil).

OSPRES Project Description: Measure, analyze and summarize the optimal high power microwave waveform (pulse width, pulse repetition frequency, center frequency, polarization, bandwidth) that simultaneously yields the lowest power density AND lowest possible size, weight, monetary cost and cooling high power microwave system (i.e., driver and radiator) that can produce such waveforms. Modularity and scalability are also critical requirements of the overall design/downselect criteria, that are considered as part of the tradespace study analysis (i.e., within the context of the optimal waveform).

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Executive Summary:

A new cooling system for 1-cm Si-PCSS devices was designed and is being fabricated to facilitate testing of Si-Behlke based transmission line charging systems. Lack of adequate cooling systems caused the Si switch to heat up while performing low voltage tests. Transition to the new SiC-Behlke switches with higher operating voltages will be aided by the new cooling designs. The Khan group also investigated three alternative topologies for Behlke based designs to operate at voltages in excess of 2-kV and above 50-kHz using SiC MOSFETs. Further development and critical analysis of these designs to come in the next few months.

The Hassan group continued to investigate coupling of RF waveforms to UAV electronic speed controllers and microcontrollers; this month studying four disparate UAV frame materials including carbon fiber, polyethylene, and Plexiglas in reference to near perfect conductors. It was found that the RF coupling response of carbon fiber and the polyethylene were similar and that the dielectric frames have little effect on the RF coupling to the electronic components. It was also validated that perfectly conducting frame materials result in reduced RF coupling to electronic components as the conducting materials “shield” the components through induction of some of the RF energy. The RF coupling group also varied the conductivity of dielectric materials used in the UAV frames, while holding the dielectric permittivity of the materials constant, to investigate trends in the shielding of the electronic components. Increased conductivities of the shields resulted in a reduction of RF coupling to components as expected. While trivial individually, this convolved information helps to understand the relation between UAV frame materials, RF coupling current in components, and helps to inform the incident waveforms and powers necessary to induce coupling.

Dr. Chatterjee’s antenna design group used circular and rectangular substrate geometries to evaluate the bandwidth and gain of two different dielectric constant substrates. It was found that the TMM-10i based material (i.e., higher permittivity) had higher bandwidth (~53%) versus the TMM-6 (~40%) while maintaining similar VSWR’s. Further investigation will be needed to determine if there is an advantage to moving to higher real permittivity materials.

Current Month Progress:**I. Source Development (Myers)**

The source research and development efforts currently under OSPRES have focused on (1) preparing for delivery of a SiC-Behlke for the next pulse charger prototype, (2) conducting a literature review for SiC-Behlke alternatives with greater performance, reliability, and efficiency, (3) continued testing of the GaN:C PCSS devices delivered by NRL, (4) continuing the cost-analysis of GaN:C manufacturing, and (5) improving and expanding the high-power figure of merit developed for linear PCSS materials. The highlights of these efforts are included in Table I.

Table I. Progress tracker for critical tasks and status: (GREEN) low risk and/or no major delays, (YELLOW) medium risk and/or minor delays, (RED) major risk and or delays.

Critical Effort	Critical Task	Status
Power Conditioning	HV switch-based charger for SiPCSS charging	The new SiC-Behlke-based charger being designed by UMKC will serve as a compact, capacitor charging solution for continued PCSS research. Initial prototypes based on Si-Behlke switches from prior art require less than 1 ft ³ and enable 10-kV charge at tens of kHz (burst mode) for tens of shots. The new SiC-based charger will enable hundreds of shots per burst at hundreds of kHz in a package of similar size, and will have built in thermal management. UMKC is also exploring HV switch topologies as alternatives to the Behlke HV switches and identified three possible topologies for continued HV switch development.
Switches – GaN:C PCSS	Cost estimation and evaluation	GaN:C devices were tested at UMKC using the PCB pulser circuit with reduced capacitance, resulting in reduced pulse voltage. While testing continues, the primary effort is shifting toward a cost-analysis manuscript and concluding our evaluation of the GaN:C PCSS system under the OSPRES program.
Switches – Figure of merit	High power PCSS figure of merit (FoM)	For source evaluation efforts, UMKC has calculated both turn-on and turn-off power loss and applied them to select linear PCSS materials. The results indicate diamond and SiC for WBG and UWBG materials, respectively, however, the constraints imposed on the recombination time, electric field (operating voltage) to simplify the model are not yet representative of high power switching. Effort continues on the FoM manuscript.

A. PCSS Power Conditioning - IGBT (Khan)

During the month of December, we continued to evaluate the Si-Belkhe based HV switch for SiPCSS transmission line charging, and conducted a literature and modeling study on high-voltage solid-state switch topologies to select suitable architectures for generating high-voltage (>1 kV) pulses at high switching frequency (> 50 kHz).

Testing has been slowed due to the thermal limits of the 1-cm SiPCSS devices being used for characterization, which are not currently fitted with a cooling system. While the average power of the Si-BEHLKE HV switch is similar to that of the 1-cm SiPCSS, without cooling the SiC-BEHLKE HV switch being delivered to UMKC will far exceed that of the SiPCSS, requiring cooling for continued testing. To overcome this, the mechanical design and development group at UMKC has modified the most recent SiPCSS cooling system to fit the 1-cm devices. The design was sent for fabrication and results using this cooling system will be reported in the subsequent report(s).

For designing an alternative to BEHLKE HV switches, we have primarily focused on the design of a string of series-connected SiC switches driven by a single gate driver. We have performed simulations of the initial models of the HV switch to determine the limitations and scope of improvements in these designs. We have chosen a HV SiC switch from BEHLKE with additional cooling features to operate at a voltage greater than 2 kV at a switching frequency of 50 kHz and higher.

SiC switching technologies are under rapid development in recent years. That said, the maximum voltage rating of the existing commercial SiC MOSFET is limited to 4.5 kV (from IXYS). For high voltage switching applications, a HV switch built from multiple MOSFETs or IGBTs connected in series and parallel is becoming an appealing option. In a series configuration, a major constraint is the voltage unbalance across the MOSFETs or IGBTs during both steady-state (the device is OFF) and switching transitions that can lead to significant damage of the switches. Ensuring equal voltage stress across devices is a challenging task because of a slight deviation in the gate signal or switch rise/fall time or manufacturing mismatch can result in unbalanced voltage distribution across the switches. Therefore, one of the primary challenges in HV switch design is to ensure voltage balancing techniques for the series string to prevent overvoltage stress and excessive losses followed by significant damage to the switches.

Existing voltage balancing (static and dynamic) techniques for series-connected switches can be primarily categorized into two types, namely, drain-source side and gate side techniques. Static voltage balancing in a series combination is generally achieved by connecting large value resistors in parallel with the switches. The drain-source side method of dynamic voltage balancing during switching transients uses snubber circuits on the drain-source side of the switches. The gate-side control method involves auxiliary circuits or feedback loops to control timing delays of different gate signals depending on the overvoltage level of the switching devices. However, the major drawback of this method is the additional complexity and cost of the auxiliary circuits.

We are presently exploring the architecture proposed in [1] which utilized the capacitive coupling method of the internal capacitances of two discrete MOSFETs using a single gate driver. However, this method is designed for two series-connected switches only. We have introduced modifications to the design to achieve static and dynamic voltage balancing across a three series-connected MOSFET based HV switch. The initial design of this switch is illustrated in Figure A1.

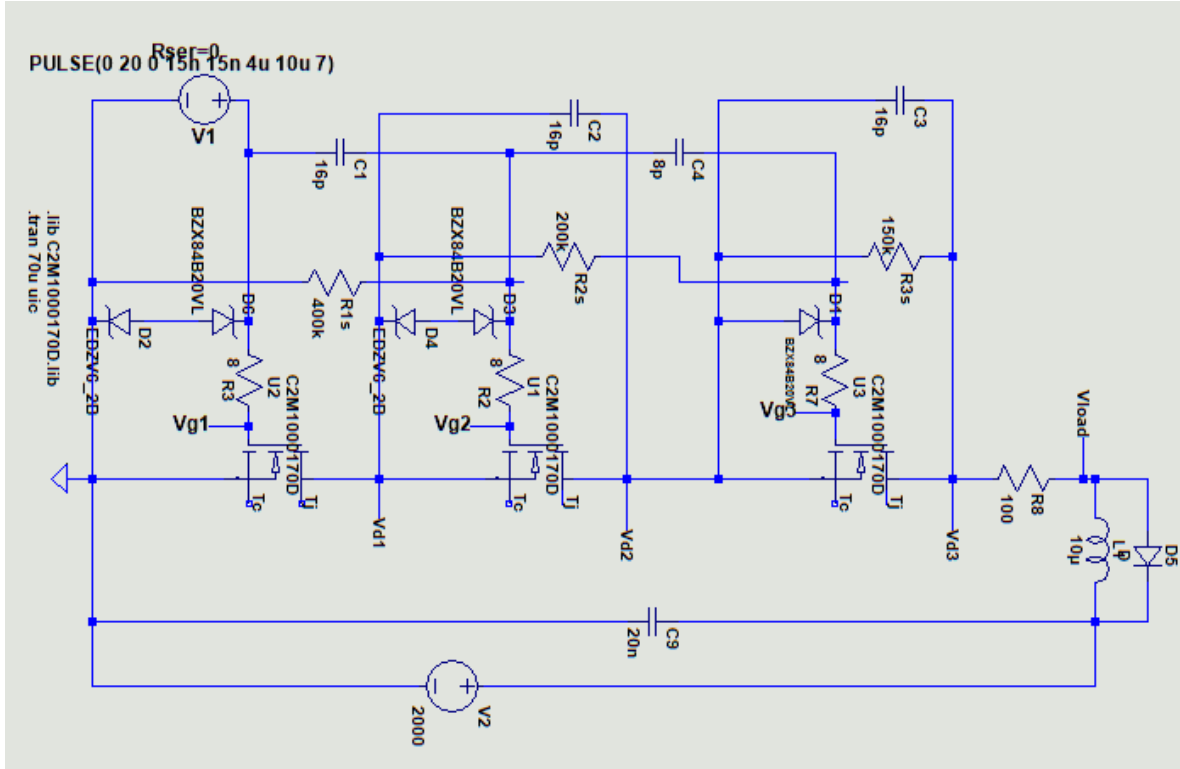


Figure A1. Schematic of series-stacked MOSFET based HV switch

Large resistors are connected across the drain-source terminals of each MOSFET to establish static voltage balancing during the OFF state. The chosen values of the resistors are lower than the channel resistance of the SiC MOSFET during its OFF period to provide a path for the leakage current. Dynamic voltage balancing is realized by the sequential turn-ON of the MOSFETs controlled by the capacitors (C1, C2, C3). However, the static voltage balancing among the stacked MOSFETs is not equal in the present design. We are identifying all the possible current paths in the proposed design during the ON and OFF states of the MOSFETs. We will modify the design based on the findings to establish equal static and dynamic voltage balancing among the MOSFETs.

The second HV switch design is based on a single gate driver as in the first design. However, the arrangement of the R-C blocks to carry out the sequential turn-ON process of the series-connected MOSFETs differs from the first design. Capacitors are connected between the gates of the consecutive MOSFETs. An additional capacitor is connected across the drain-gate terminal of the top MOSFET. In addition, two zener diodes are placed between the capacitors and the power supply to allow a level shift to prevent static

conduction from the gate circuit to the drain circuit. The schematic of this HV switch is illustrated in Figure A2. The simulated static voltage balancing across the MOSFETs in this design is illustrated in Figure A3.

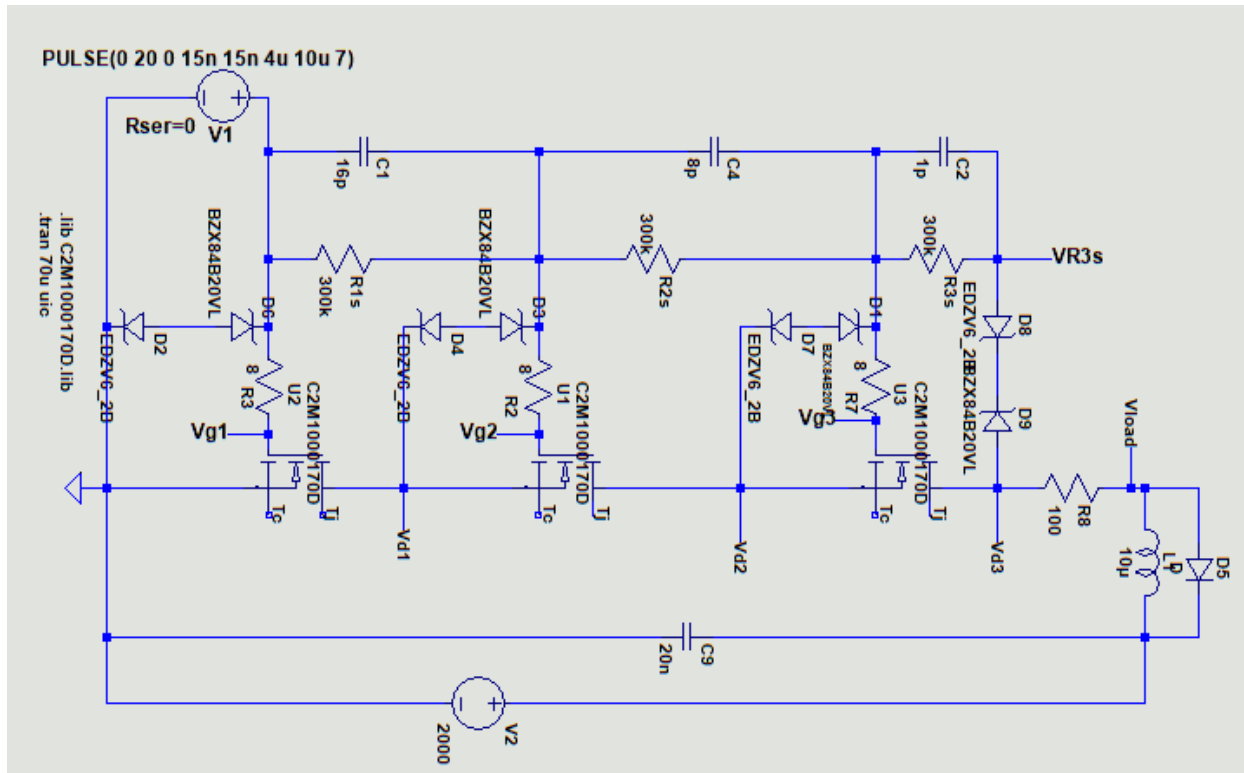


Figure A2. Schematic of series-stacked MOSFET based HV switch.

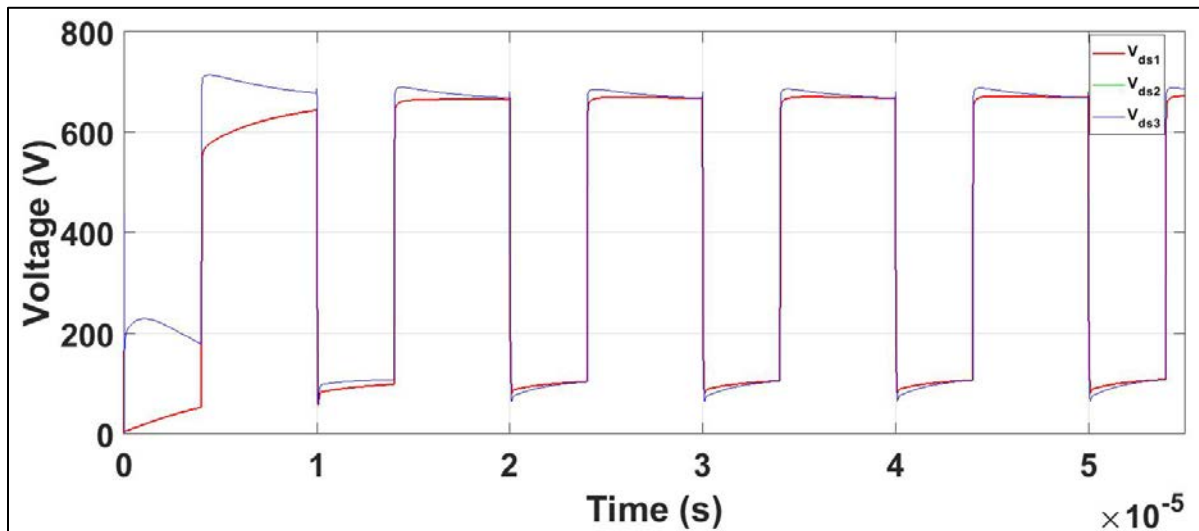


Figure A3. Simulated voltages across the drain-source terminals of the series-connected MOSFETs.

The static voltage levels across the stacked MOSFETs differ by 50 V for an input of 2 kV during the OFF period. However, the voltages across the switches do not go to zero as intended during the ON state of the switch. Determining the precise values of the buffer capacitors can be one of the ways to address the issue.

The third HV switch configuration consists of strings constructed from series switches, which are then connected in parallel [3]. The drain of a switch is connected to the source of the subsequent switch. Capacitors are connected to the common gates of two switches connected in parallel. The lower switch is driven by a gate driver without any external capacitor. The schematic of the switch is shown in Figure A4.

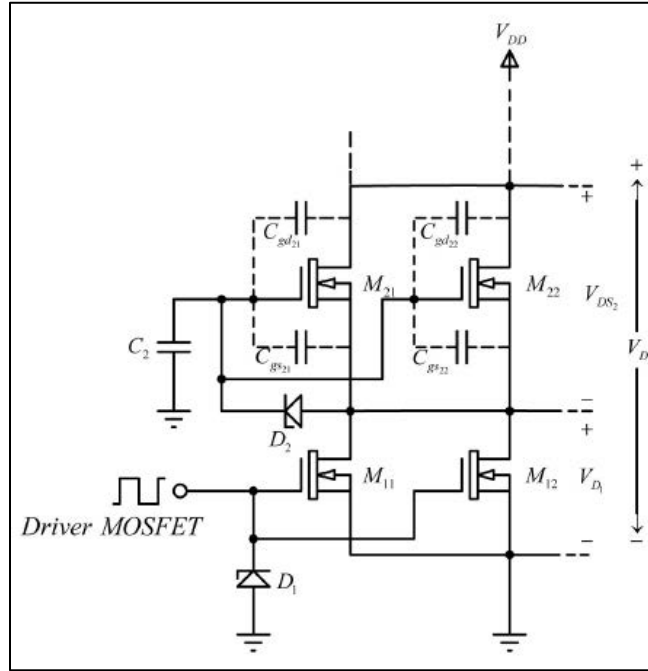


Figure A4. Schematic of series-parallel stacked MOSFET based HV switch [3].

For a string of n series switches, the value of the capacitor connected to the common gate of the n^{th} switch is given by [3]:

$$C_n = C_2 * \frac{1}{(n - 1)}$$

At present, we are conducting HV experiments (≥ 2 kV) at a high switching frequency (≥ 50 kHz) using the HV Si MOSFET based power conditioning board. At this operating condition, the cooling requirement for both the PCSS and the power conditioning board gets elevated. During the experiment, the existing air cooling method has failed to sufficiently cool the PCSS, resulting in the failure of the PCSS. Three cooling blocks have been designed to aid the cooling process of the Si PCSS at HV environments at a high switching frequency. Potential coolant leaking issues have been addressed in these designs. We will perform the HV experiments with the cooling blocks attached to the PCSS. We are in the process of procuring a SiC-based 15 kV rated commercial MOSFET switch from BEHLKE. It can handle a turn-ON peak current in the range of (duty cycle

<1%). The switch can be equipped with a direct liquid cooling (DLC) option which can facilitate a maximum continuous power dissipation of 1600 W and a maximum switching frequency of 100 kHz. The low ON-resistance of SiC BEHLKE switch can minimize the conduction loss of the switch, and increase the output pulse count at higher switching frequency (100 kHz). The maximum burst frequency of the switch is specified as 2 MHz in the datasheet.

Immediate Tasks for the Power Conditioning Subgroup:

- Designing a new power conditioning board based on the HV SiC switch from BEHLKE.
- Continued Si-BEHLKE power conditioning circuit testing with SiPCSS (with a cooling system) at a lower voltage (<3 kV) and at a higher repetition frequency (>50 kHz).
- Designing an HV SiC MOSFET based power conditioning circuit.

Expected Products of Power Conditioning Subgroup:

- Efficient thermal management system with the power conditioning circuit.
- Single gate driver-based HV series-connected switch for capacitor charging.

References:

- [1] Y. Ren *et al.*, "A Compact Gate Control and Voltage-Balancing Circuit for Series-Connected SiC MOSFETs and Its Application in a DC Breaker," in *IEEE Transactions on Industrial Electronics*, vol. 64, no. 10, pp. 8299-8309, Oct. 2017.
- [2] N. Teerakawanich and C. M. Johnson, "Design Optimization of Quasi-Active Gate Control for Series-Connected Power Devices," in *IEEE Transactions on Power Electronics*, vol. 29, no. 6, pp. 2705-2714, June 2014.
- [3] A. F. Guerrero-Guerrero, A. J. Ustariz-Farfan, H. E. Tacca and E. A. Cano-Plata, "High Voltage Switch with MOSFET Series/Parallel Connection," *2018 IEEE ANDESCON*, Santiago de Cali, 2018, pp. 1-6.

B. Switches - GaN:C (Thompson)

The GaN:C PCSS team is continuing to empirically test and evaluate GaN:C PCSS capabilities and performance, while composing a manuscript evaluating the cost of GaN manufacturing to determine feasibility for pulsed power and/or RF applications.

The focus of the manuscript is the current and projected costs of GaN manufacturing, which varies widely depending on the material specifications required by application, giving the material a high-risk in terms of cost. If new, more mainstream uses for GaN:C result in more efficient manufacturing methods exist, this cost of GaN could be greatly reduced. The second focus is to compare GaN manufacturing to other wide bandgap and ultra-wide bandgap materials, such as SiC, and their manufacturing histories, to predict the possible future cost trends of GaN:C.

Testing of the GaN:C provided by Naval Research Laboratories (NRL) continues using the PCB pulsing circuit with a reduced charge storage capacitance in an attempt to reduce

the pulse width and shape. Although not ideal, the capacitors were chosen based on readily available high-voltage multi-layer ceramic capacitors. Per Figure B1, the peak pulse voltage was reduced by approximately 14%, and the shape of the recovery profile was virtually unaffected. In hindsight, the capacitance chosen (0.78 nF) was far too large to create an appreciable decrease in pulse width; a capacitance of 40 pF would be necessary to achieve full discharge with a 50 Ω line (approximating 2 ns FWHM and 5RC as fully discharged). However, this low capacitance, thus energy storage capacity, would further decrease the peak pulse voltage to an unacceptable level. Instead, using a standard pulse-forming network (PFN) or Blumlein PFN, with a propagation time of less than half the recovery time of the GaN:C PCSS, will decrease the fall time of the output pulse, similar to how SiPCSS pulse width is controlled via the transmission line length. For future testing, a PFN with a 1-ns propagation time will be used to evaluate GaN:C PCSS.

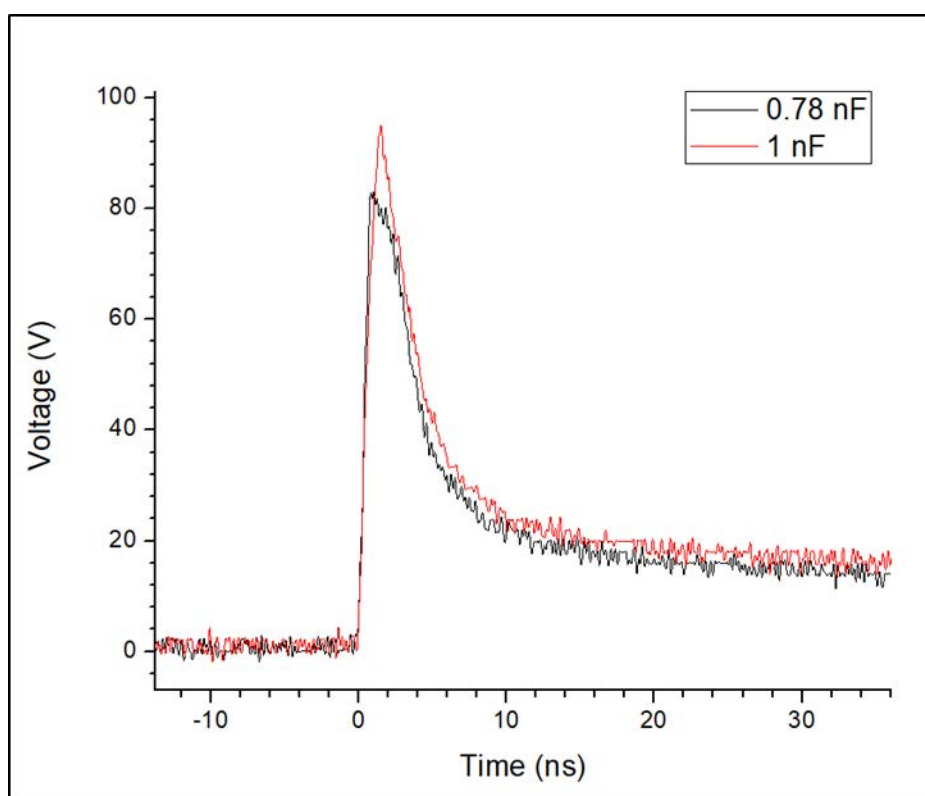


Figure B1. Comparison of output pulses of 500- μm GaN:C device using a PCB with an effective capacitance of 0.78 nF and 1.0 nF.

Immediate Tasks of GaN:C PCSS Subgroup:

- On-state resistance comparison of Carbon concentrations of $8\text{E}16$ vs. $5\text{E}17$ at low voltages for devices with 50 μm and 500 μm gap lengths.
- Measurements using a PFN with a 1 ns propagation time for reduced fall time proof of concept.
- Continue efforts on the manuscript.

Expected Products of GaN:C PCSS Subgroup:

- Summary report of GaN:C PCSS evaluation efforts including cost analysis and performance data.

C. Switches - Figure of Merit (Bhamidipati)

The FoM team continued its focus on improving the high power figure of merit (HP FoM) for uncompensated/intrinsic material high power PCSS. During the reporting period, the power dissipated in the material was calculated during turn-on and turn-off (per equations reported in the October 2019 report) for devices with varying lengths, electric fields, recombination times, and optical intensities. The material properties used in the analysis are provided in Table C1 for reference.

Table C1. Material properties used in HP FoM calculations.

Property	Si	GaN	SiC 4H	Diamond	GaAs	Ga2O3	BN	AlN
E _g (eV)	1.12	3.45	3.26	5.45	1.43	5.5	6.5	6.2
Electron mobility (cm ² /V-s)	1400	1000	1000	2200	8500	400	825	500
Hole mobility (cm ² /V-s)	450	350	115	850	400	100	500	100
E _{crit} (V/cm)	3E+5	3.3E+6	2.2E+6	1.0E+7	4.0E+5	1.3E+7	1.8E+7	1.5E+7
Dielectric constant	11.7	5.35	9.66	5.7	10.89	9.9	4.5	4.77
Thermal Cond (W/m-K)	1.5	1.3	4.9	22	0.5	0.2	7.6	3
Photon Energy (J)	1.8E-19	5.52E-19	5.17E-19	8.72E-19	2.23E-19	8.8E-19	10.4E-19	9.92E-19
Saturation drift velocity (m/s)	1E+05	2.5E+05	2.2E+5	2.7E+5	1.2E+05	1.6E+05	2.7E+05	1.6E+05

To simplify the calculations, operating conditions/restrictions and assumptions were applied. Many of these assumptions will be improved and expanded in future calculations. The assumptions/conditions are as follows:

Optical pulse Assumptions:

- A temporal Gaussian profile with $T_{FWHM} = 2$ ns and pulse energy to create 10^{17} charge carriers (corresponding to an average current of 1.6×10^{-2} A).
- Uniform spatial illumination of devices with 100% absorption.
- 100% contact injection efficiency.

Device conditions:

- Devices are held at electric fields of 6 kV/mm—*This will be material-specific in future calculations.*

- Charge carrier transit time is calculated based on the material's electron saturation velocity and device length.
- A theoretical recombination time of 50 ns is assumed across all the materials—in real-world calculations, the recombination time would vary across the materials between sub-ns and ones to tens of μ s for uncompensated and indirect bandgap materials. This will be material specific in future calculations.
- Turn-on loss is calculated between 0.1 ns and 1.2 ns (10%-90% of the optical pulse rise time) and turn-off loss from 1.5 ns-3.5 ns (90%-10% of the optical pulse fall time)—Turn-off time is buffered with approximately 1 ns to account for the recovery time greater than optical pulse tail.

For comparison, the power dissipated in each material when operated at 50 kHz pulse repetition frequency (PRF) is calculated for the device lengths ranging from 50–950 μ m. The results of these calculations are broken into turn-on and turn-off power dissipation in Figure C1 (top) and (bottom), respectively.

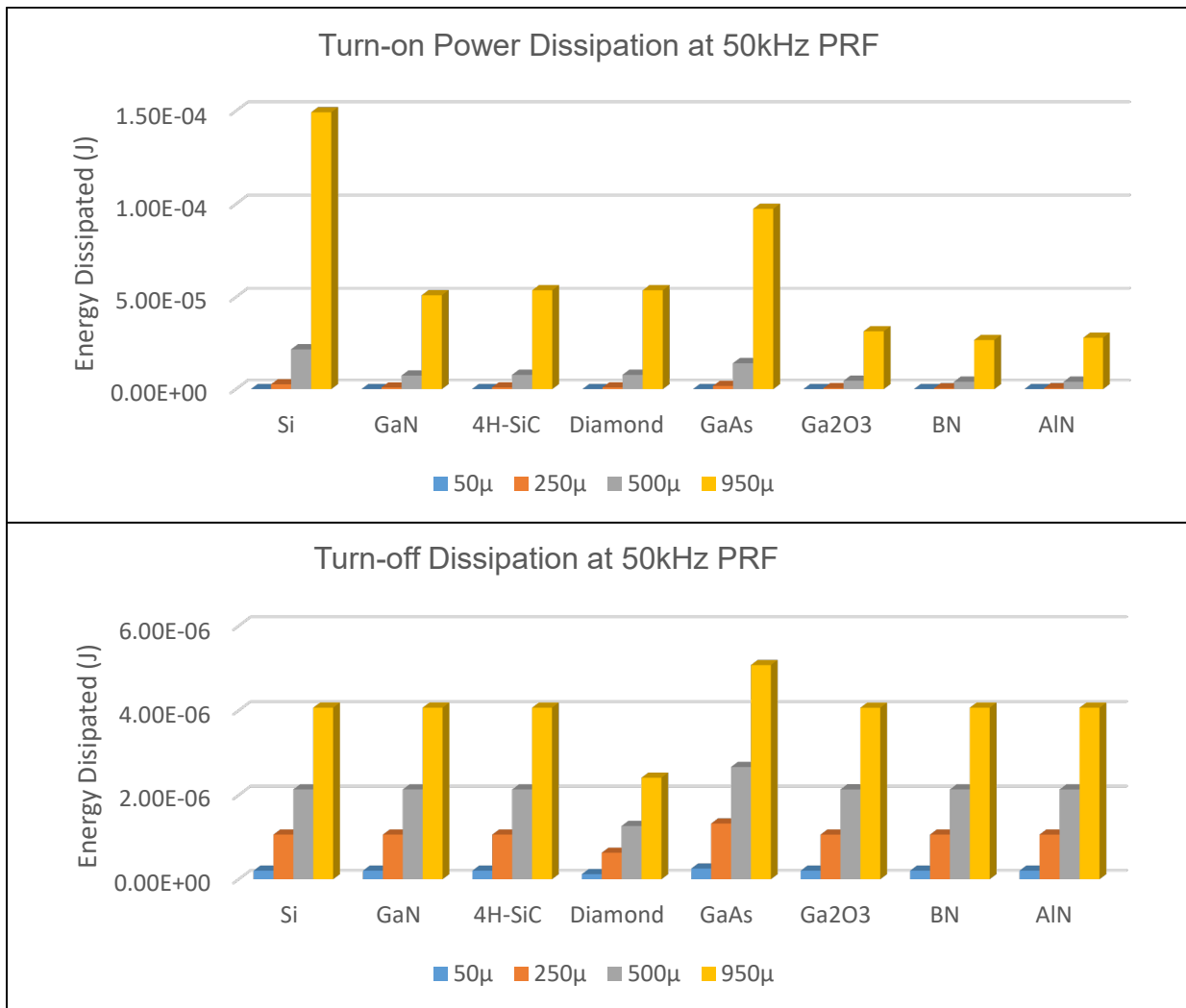


Figure C1. Power dissipation during (top) turn-on and (bottom) turn-off per second in 50-, 250-, 500-, and 950- μ m devices when operated at PRF of 50 kHz.

Following calculation of the energy dissipated in the device during switching, we calculated the rate of temperature rise in each material. To more accurately portray the FoM and avoid ambiguity, the inverse of the rate of temperature rise was taken and normalized to that of Si, shown in Figure C2. Table C1 shows the material properties used to calculate the device power dissipation and the temperature rise in the material.

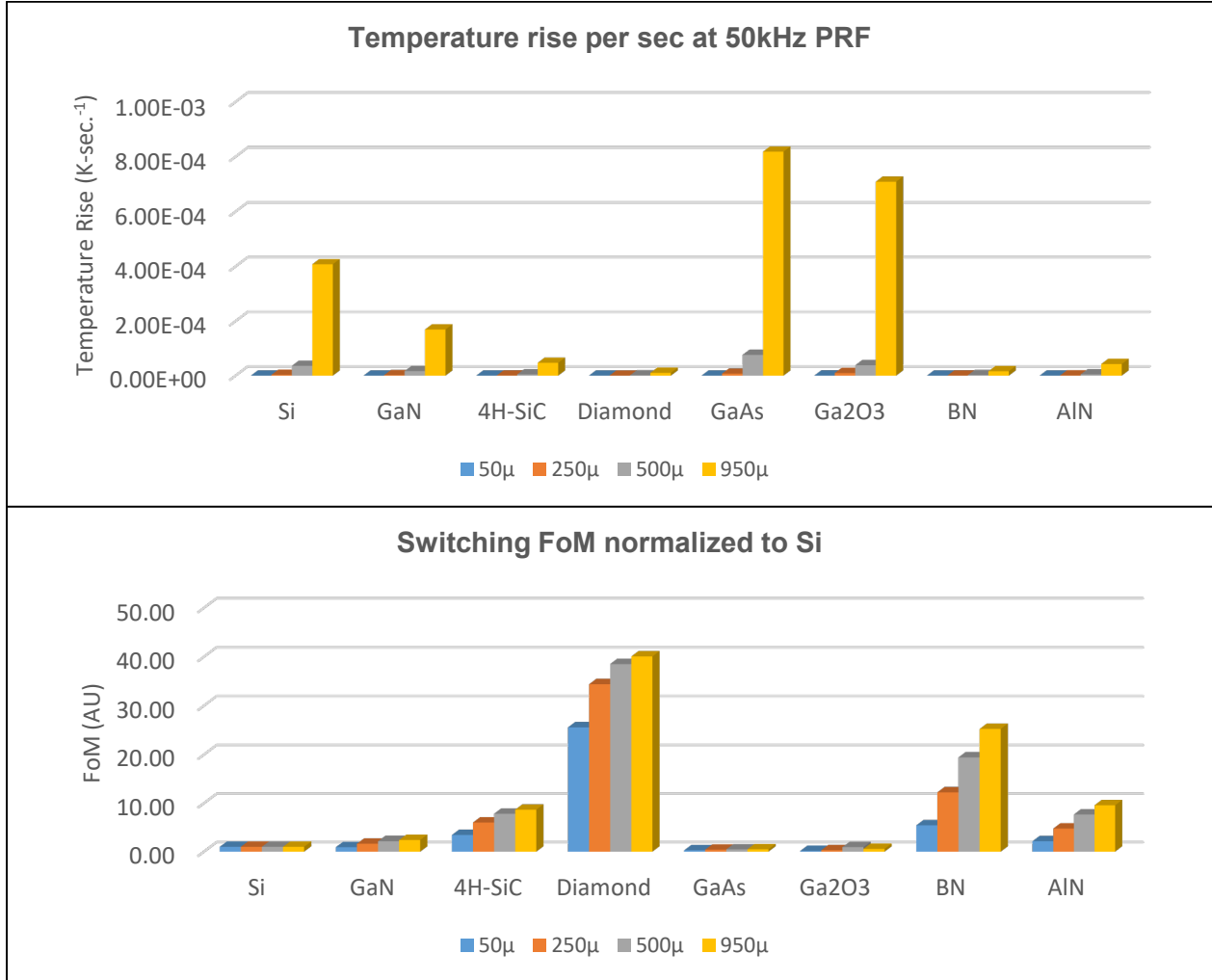


Figure C2. (Top) Rate of temperature increase in devices with lengths of 50-, 250-, 500-, and 950-μm devices and (bottom) high power material figure of merit for devices operating at a PRF of 50 kHz. All results are normalized to Si.

Using this figure of merit and limiting the analysis to the 950-μm devices, of the ultra-wide bandgap materials, diamond ranks highest owing to its thermal conductivity and charge carrier mobility, followed by BN, AlN, and Ga₂O₃. For the wide band gap materials, 4H-SiC ranks highest, followed by GaN, Si and GaAs. The resulting figure of merit is of these results are provided in Figure C3.

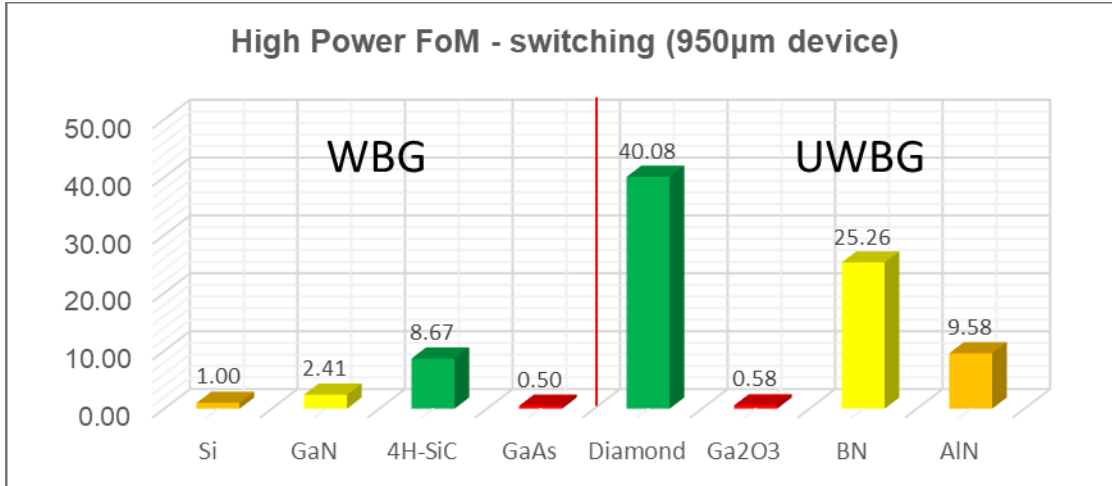


Figure C3. Calculated HP FoM for wide band gap (WBG) and ultra-wide band gap (UWBG) materials switched at a PRF of 50 kHz and normalized to Si.

Immediate tasks for FoM subgroup:

- Complete cost-effectiveness analysis to determine optimal material for PCSS implementation in a given application using material FoM in comparison with SWaP-C² requirements.
- Completion of the manuscript “Figures of Merit for High Power Linear Mode PCSS.”

II. Effects, Coupling and UAS Developments (Currie)

D. RF Coupling (Hassan)

Effect of frame material and shape on EMC of UAV’s wires/loads:

A simple wire configuration that emulates the connections between the Electronic Speed Controller (ESC) and the UAV microcontroller above frames is shown in Figure D1. The ESCs are placed at the midway between the center part of the drone and the motors at the edge of the arms. In this report, the effect of the frame shape and material on the coupling to ESCs is presented. UAV frames composed of carbon fibers (CF), Plexiglas (PG), Polyethylene (PE), and perfect electric conducting frames were investigated. Carbon fibers were designed as 3 layers of isotropic materials. Each layer is rotated by 45° relative to the other layer as modeled in [1]. The relative permittivity of the Plexiglas (PG) is 2.6 with 0.0057 loss tangent and the relative permittivity of the Polyethylene (PE) is 2.25 with 0.0004 loss tangent [2].

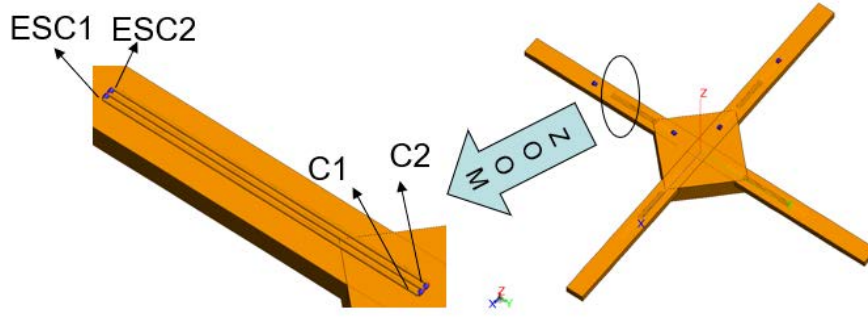


Figure D1. Sketch of the UAV frame with the wires connecting the center controller 'C' with the electronic speed controllers 'ESC' at the 4 arms of the frame.

A $50 \, \Omega$ impedance is assigned for the controller impedances, $2.5 \, \text{k}\Omega$, and $35 \, \text{M}\Omega$ is assigned for the electronic speed controllers. Thus, $\text{ESC1} = 2.5 \, \text{k}\Omega$, $\text{ESC2} = 35 \, \text{M}\Omega$, and $\text{C1} = \text{C2} = 50 \, \Omega$. The previous values were obtained from DC measurements on the components using a multimeter. The wires are placed 10 mm above the frame and are excited using 625 different electric field directions, $0^\circ < \theta < 180^\circ$, 7.5° increment, $0^\circ < \Phi < 360^\circ$, 15° increment. All the currents shown in this report are the average over all the 625 angles.

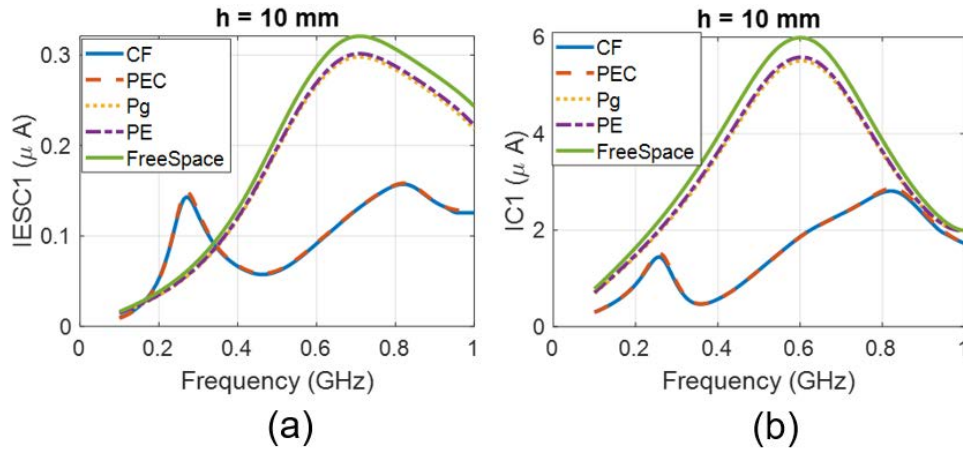


Figure D2. Average coupled current over 625 incidence angles on: (a) Electronic speed controller, (b) controller at the center part of the frame.

Figure D2 shows the comparison between the average coupled current to the controller (IC) and the speed controller (IESC). Figure D2 shows that the ESC loads are significantly affected by the modes of the UAV frame. The peak at 0.25 GHz in the average coupling is due to the excitation of Mode 1 of the frame. The observations of Figure. D2 can be summarized as follow:

- On average, at low-frequency range, the response of the carbon fiber is similar to the PEC frame regardless of the non-symmetric layers design of the carbon fiber.
- The dielectric frames have a small effect on the coupling to the electronic components and can be approximated by a red-shift in the resonance frequency

and a small reduction in the peak amplitude of the coupled current in comparison to coupling to loads in air without a frame.

- The current coupled to the ESC at the resonance frequency of the conducting frame can reach 300% of the current coupled to the ESC in the air or above a dielectric frame. However, in general, conducting frames provide a significant reduction in the coupled current over a wide range of frequencies due to the shielding effect.

To further highlight the dependency of the coupled current on the frame's electrical properties UAV frames with different conductivities were tested. The relative dielectric permittivity was fixed at a value of 2 whereas the conductivity was valued from 10 S/m to 1000 S/m. Again, the frame and the wires are illuminated by an electric field from 625 angles and the average induced current on ESC, the sensitive component, is calculated. Figure D3 shows the comparison between average coupling to the ESC for different frame conductivity. Figure D3 shows that as the conductivity of the frame increases the shielding effectiveness of the frame increases most frequencies except at the resonance frequencies of the modes of the frame (Mode 1 resonates at 0.25 GHz). The effect of the conductivity on the frame modes start to saturate at $\sigma = 100$.

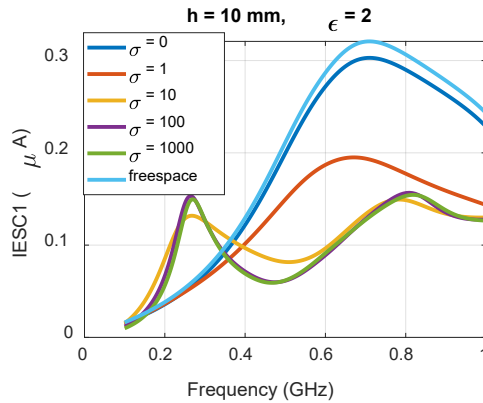


Figure D3. Average coupled current over 625 incidence angles on the Electronic speed controller for different frame materials.

In the next set of computational experiments, we studied the effect of the permittivity as well as the effect of the shape of the frame. To eliminate the effect of the frame shape, we simulated the same wire configuration above an infinite planer dielectric layer and plotted the coupled current of the loads in Figure. D4a. The thickness of the layer is like the thickness of the frame used in the previous analysis.

Figure D4b shows the coupling to the ESC loads for wires above the actual frame composed of the same permittivity values shown in Figure D4a. Clearly, the difference between Figure D4a and Figure D4b shows that the frame shape significantly affects the coupling for high permittivity values. That is, comparing the green curve in Figure D4a and Figure D4b, show additional resonances for the wire above the frame that do not exist for the wires above the infinite dielectric layer. However, for low permittivity frames, the exact shape of the frame does not have a significant effect on the coupled current

and the frame can be simplified by an infinite dielectric slab of the same thickness and the same relative dielectric permittivity with a significant saving in the computational time.

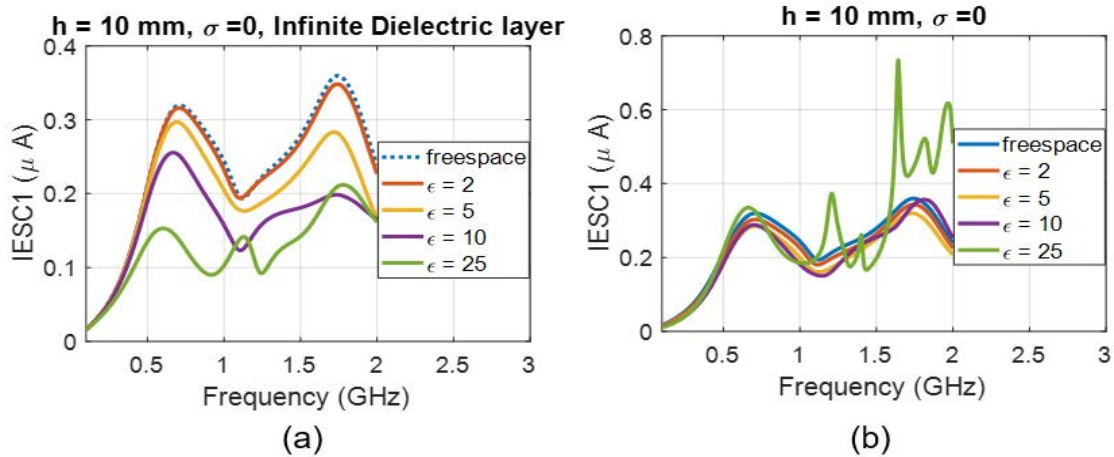


Figure D4. Average coupled current over 625 incidence angles on Electronic speed controller for different permittivity of (a) infinite layer, (b) frame as shown in Figure D1.

Immediate Tasks for RF Coupling Subgroup:

- Study different scanned UAV frames (different shapes) with different dielectric properties.
- Perform an exhaustive statistical analysis of coupling to randomly shaped loops above the UAV frames with dielectric properties.

References:

- [1] C. L. Holloway, M. S. Sarto, and M. Johansson, "Analyzing carbon-fiber composite materials with equivalent-Layer models," *IEEE Trans. Electromagn. Compat.*, vol. 47, no. 4, pp. 833–844, Nov. 2005.
- [2] "Plexiglas® acrylic sheet and acrylic resin products - Plexiglas.com." [Online]. Available: <https://www.plexiglas.com/en/>. [Accessed: 21-Dec-2019].

III. Antenna Development (Crow)

E. Antennas - Electrically Small Antennas (Chatterjee/Barman)

Coaxial probe fed microstrip patch antennas, which can be used to generate > 30% 2:1 VSWR bandwidth by strategically placing the probe along the patch diagonal, was previously reported. In this report, validation of the simulated data with the measured ones has been presented. Measurement of the antenna parameters are performed using GTEM.

A simple rectangular patch can yield > 30% 2:1 VSWR bandwidth and a reasonable gain if the probe is placed along the diagonal at 2/3rd distance from the center. The geometry of the patch is shown below in Figure E1.

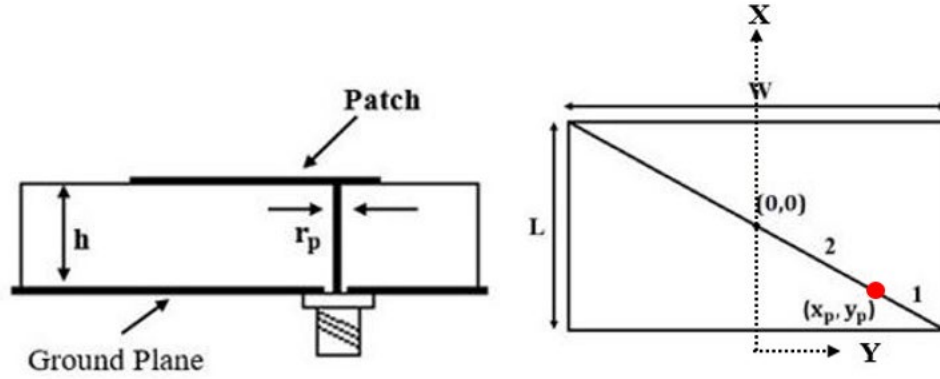


Figure E1. Geometry of the designed coaxially fed rectangular microstrip patch antenna.

To verify the phenomenon, two different types of substrates, TMM-6 ($\epsilon_r = 6, \tan \delta = 0.0023$) and TMM-10i ($\epsilon_r = 9.8, \tan \delta = 0.0020$), were used. The antenna dimensions and the corresponding results are demonstrated in Tables E1, 2, and 3.

Table E1: Optimized antenna dimensions.

Substrate	h (mm)	$h\sqrt{\epsilon_r}/\lambda$	L (mm)	W (mm)	r_p (mm)	(x_p, y_p)
TMM-6	6.35	0.1868	12.77	19.16	0.6	(4.26, 6.38)
TMM-10i	6.35	0.2387	9.61	14.42	0.6	(3.20, 4.80)

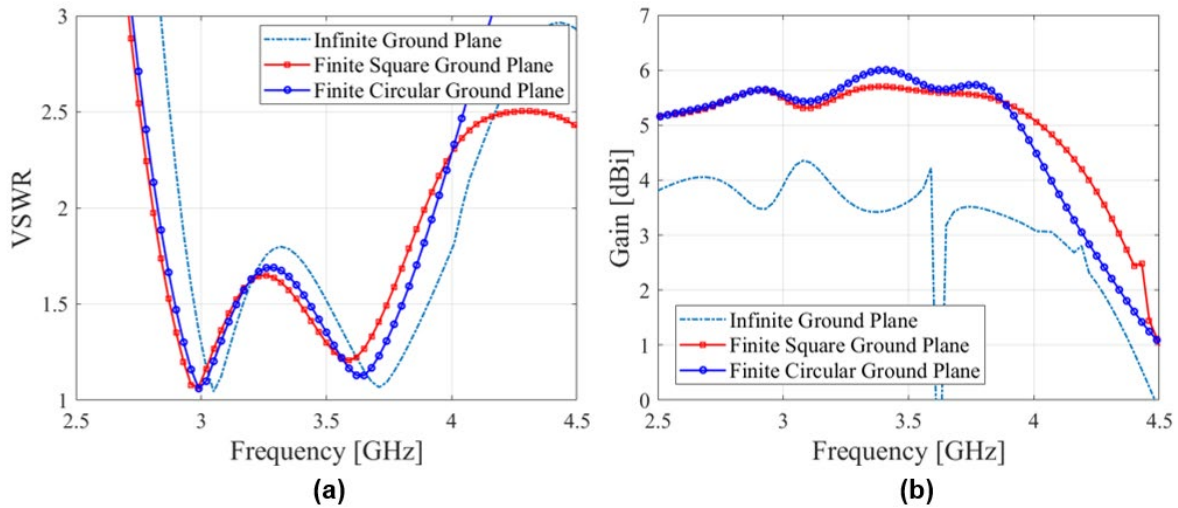
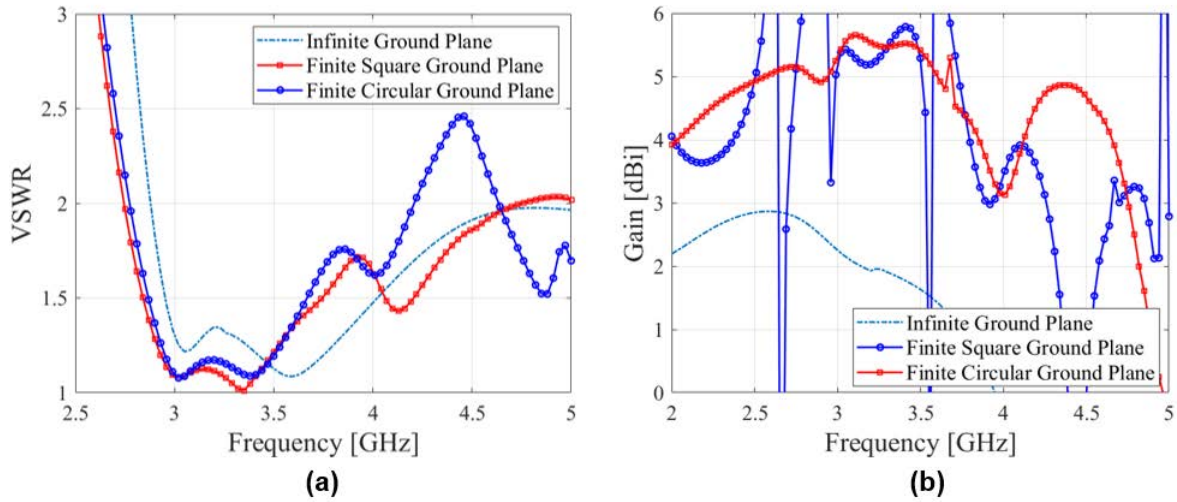


Figure E2. (a) VSWR and (b) Gain vs. frequency of the antenna designed on TMM-6 substrate.

Table E2. Summary of the results for the antenna designed on TMM-6 substrate.

Ground Plane	f_L (GHz)	f_c (GHz)	f_U (GHz)	% BW	Max. Gain (dB)	ka
Infinite	2.917	3.48	4.042	32.3	4.35	0.8398
Finite Square ($L_g=41.64$ mm)	2.826	3.38	3.935	32.8	5.7	0.8156
Finite Circular ($r_g=23.49$ mm)	2.806	3.35	3.894	32.5	6.00	0.8084

**Figure E3.** (a) VSWR and (b) Gain vs. frequency of the antenna designed on TMM-10i substrate.**Table E3.** Summary of the results for the antenna designed on TMM-6 substrate.

Ground Plane	f_L (GHz)	f_c (GHz)	f_U (GHz)	% BW	Max. Gain (dB)	ka
Finite Square ($L_g=41.64$ mm)	2.773	3.491	4.209	41.13	unstable	0.6339
Finite Circular ($r_g=23.49$ mm)	2.745	3.7475	4.75	53.50	5.65	0.6805

The antennas are initially simulated using the EM solver FEKO. For validation of the simulated results, the antennas are fabricated and measured in the transmit mode in GTEM. The test setup for the antennas in transmit mode is shown in Figure E4.

Figure E5 illustrates a comparison between the simulated and measured VSWR of the antennas on finite circular ground planes.

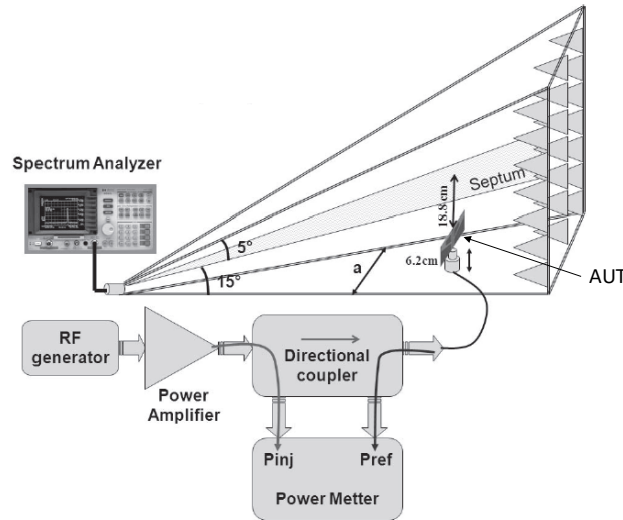
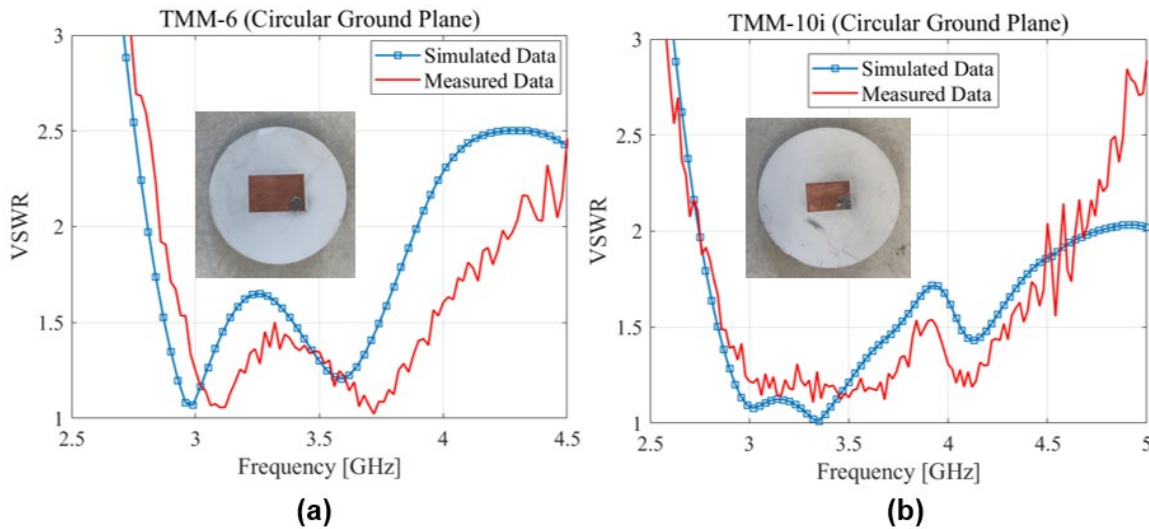


Figure E4. Test setup for the antennas in the transmit mode.



	F _c (GHz)	%BW
Simulated	3.35	32.5
Measured	3.58	40.22

	F _c (GHz)	%BW
Simulated	3.75	53.50
Measured	3.68	53.26

Figure E5. Comparison between the simulated and measured VSWR of the antennas designed on (a) TMM-6 and (b) TMM-10i substrates. (Fabricated prototypes are shown in inset).

Immediate Tasks:

- Validation of the simulated results in terms of antenna gain and radiation pattern is being looked into.
- Optimization of the antenna in terms of its near and far-field parameters (D/Q ratio) is being investigated.

F. Antennas - Log Periodic (Hassan)

Last month's MSR, we presented our progress in the helical antenna and array design. This month, we discuss the parameters characterizing the time domain response of the helical antenna. In Ultra-Wideband (UWB) systems, antenna characteristics play a significant role compared to narrowband antennas used for typical RF applications. The performance of narrowband antennas can be described by its center frequency response since many applications are constrained to the small predefined channel frequency. On the other hand, the performance of UWB antennas can vary significantly over the bandwidth, and an antenna may modify its input signal in several ways such as differentiation or dispersion. Therefore, both the time domain and frequency domain properties must be considered during the design process. In this report, we present a summary of UWB time-domain performance parameters and then we apply time-domain analysis to our design of the 1 GHz helical antenna.

Frequency Domain Performance:

Figure F1 shows the frequency domain representation for a simple two-antenna link where for a given antenna pair $G(f, \theta, \psi)$ is the gain; r_{TxRx} is the Tx-Rx distance, $E_{Tx}(f, r, \theta_{Tx}, \psi_{Tx})$ is the radiated electric field due to the transmit antenna at location r . Then for the transmit and receive antennas respectively, $U_{Tx}(f)$ and $U_{Rx}(f)$ are the voltage amplitudes; $Z_{C,Tx}(f)$ and $Z_{C,Rx}(f)$ are the antenna characteristic impedance; and $H_{Tx}(f, \theta_{Tx}, \psi_{Tx})$ and $H_{Rx}(f, \theta_{Rx}, \psi_{Rx})$ are the antenna transfer functions in meters, which can be also represented as the frequency-dependent effective length of the antennas [1]. Since antennas are passive components, linear time-invariant theory can be used to describe the two-antenna system in Figure F1 [2], with two transfer functions, one for the Tx antenna and another for the Rx antenna. Equation (1a) describes the output field at a point r due to the input voltage $U_{Tx}(f)$. The equation can be read as a cascade of the effects the signal undergoes from right to left, which can be rearranged into the transfer function form in equation (1b) relating the output to the input. Equation (2) describes the input voltage at the Rx antenna's port due to the field by Tx, $E_{Tx}(f, r)$, which is simply $H_{Rx}^T(f, \theta_{Rx}, \psi_{Rx})$ cascaded with equation (1a). This simple description can be used in the frequency domain to fully characterize the system and solve for the relevant parameters.

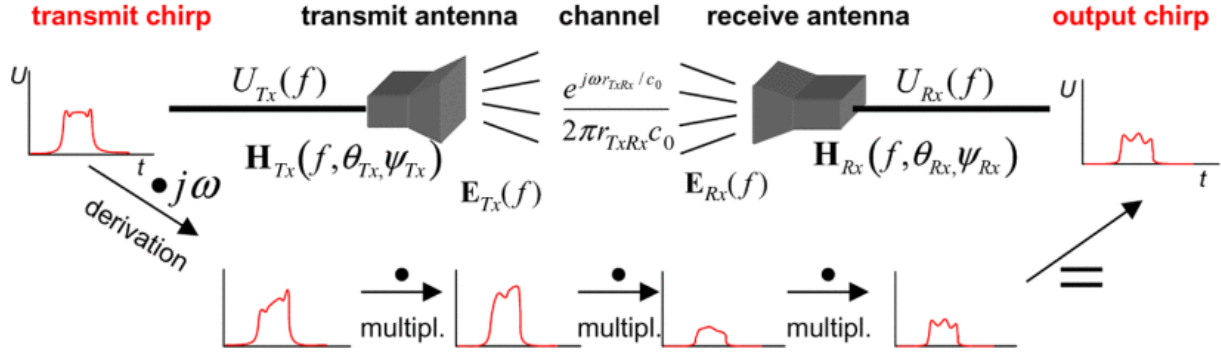


Figure F1. Frequency-domain system link-level characterization [3]

$$\frac{E_{Tx}(f, r)}{\sqrt{Z_0}} = \frac{e^{j\omega r_{TxRx}/c_0}}{2\pi r_{TxRx}c_0} H_{Tx}(f, \theta_{Tx}, \psi_{Tx}) \cdot j\omega \frac{U_{Tx}(f)}{\sqrt{Z_{C,Tx}}} \quad (1a)$$

$$H_{Tx}(f, \theta_{Tx}, \psi_{Tx}) = \frac{\text{Output}}{\text{Input}} = \frac{E_{Tx}(f, r) \cdot 2\pi r_{TxRx}c_0 \cdot \sqrt{Z_{C,Tx}}}{j\omega U_{Tx}(f) \cdot e^{j\omega r_{TxRx}/c_0} \cdot \sqrt{Z_0}} \quad (1b)$$

$$\frac{U_{Rx}(f)}{\sqrt{Z_{C,Rx}}} = H_{Rx}^T(f, \theta_{Rx}, \psi_{Rx}) \cdot \frac{e^{j\omega r_{TxRx}/c_0}}{2\pi r_{TxRx}c_0} \cdot H_{Tx}(f, \theta_{Tx}, \psi_{Tx}) \cdot j\omega \frac{U_{Tx}(f)}{\sqrt{Z_{C,Tx}}} \quad (2)$$

Group Delay:

Arguably the most critical frequency domain performance parameter of UWB antennas is the group delay, τ_g , represented by the levelness of the phase of the radiated signal. Group delay is a measure of time delay across the frequency bandwidth. Mathematically it's the negative derivative of the phase with respect to frequency,

$$\tau_g(\omega) = -\frac{d\varphi(\omega)}{d\omega} \quad (3)$$

Group delay is sometimes expressed as the mean group delay, $\overline{\tau_g}$, over the antenna bandwidth as a single number spec for comparing antennas.

$$\overline{\tau_g} = \frac{1}{\omega_2 - \omega_1} \int_{\omega_1}^{\omega_2} \tau_g(\omega) d\omega \quad (4)$$

Time Domain Performance:

Similar to the frequency domain, the impulse response of an antenna by convolution can describe its time-domain performance [3]. Figure F2 represents an equivalent antenna link in the time domain with the typical mathematical operations a Gaussian pulse undergoes in the system shown in the small plots. A similar approach will be used to simulate the helical antenna excite with a modulated Gaussian pulse.

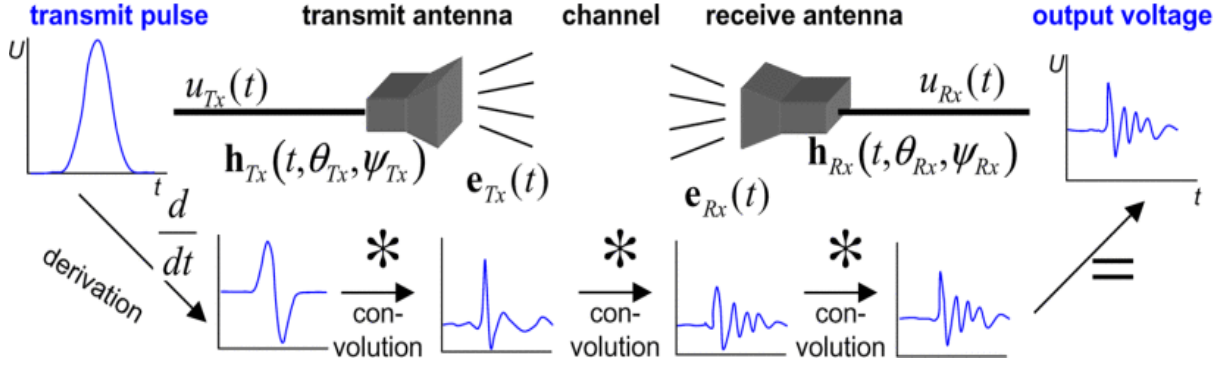


Figure F2. UWB system link-level characterization in the time domain [3].

In Figure F2, the distance between Tx and Rx is r_{TxRx} ; the radiated field due to Tx is $e_{Tx}(t, r, \theta_{Tx}, \psi_{Tx})$; for the transmit and receive antennas respectively $u_{Tx}(t)$ and $u_{Rx}(t)$ are the exciting waveform; while $h_{Tx}(t, \theta_{Tx}, \psi_{Tx})$ and $h_{Rx}(t, \theta_{Rx}, \psi_{Rx})$ are the impulse response in m/ns [1]. The time-domain impulse response for the two antennas in equations (3) and (4) can be arrived at by transforming equations (1) and (2) to the time domain [3], resulting in:

$$\frac{e_{Tx}(t, r)}{\sqrt{Z_0}} = \frac{1}{2\pi r_{TxRx} c_0} \delta\left(t - \frac{r_{TxRx}}{c_0}\right) * h_{Tx}(t, \theta_{Tx}, \psi_{Tx}) * \frac{\partial u_{Tx}(t)}{\partial t \sqrt{Z_{C,Tx}}} \quad (5)$$

$$\frac{u_{Rx}(t)}{\sqrt{Z_{C,Rx}}} = h_{Rx}^T(t, \theta_{Rx}, \psi_{Rx}) * \frac{1}{2\pi r_{TxRx} c_0} \delta\left(t - \frac{r_{TxRx}}{c_0}\right) * h_{Tx}(t, \theta_{Tx}, \psi_{Tx}) * \frac{\partial u_{Tx}(t)}{\partial t \sqrt{Z_{C,Tx}}} \quad (6)$$

Time-domain parameters of an antenna can be analyzed by looking at the analytic impulse response, which is calculated by the Hilbert transform H such that the analytic impulse is:

$$h^+(t) = h(t) + jH\{h(t)\} \quad (7)$$

The envelope of the analytic function, $|h^+(t)|$ provides a measure of antenna dispersion, the two main parameters extracted out are peak value and width at half peak [3].

Peak Value of the Envelope:

The peak value of the envelope of the analytic response is inversely related to dispersion because a higher envelope peak value would mean localization of the energy in the time domain.

$$p(\theta, \psi) = \max|h^+(t, \theta, \psi)| \quad (8)$$

Full width at Half maximum time:

The width at half maximum describes the widening of the input pulse due to dispersion in the antenna; lower FWHM time is very much desired in a UWB antenna.

$$\tau_{FWHM} = t_{1| |h^+(t_1)|=p/2} - t_{2| |h^+(t_2)|=p/2} \quad (9)$$

Ringing time measures oscillations in the radiated pulse beyond the max peak value, which is another critical parameter in the time-domain characterization of UWB antennas, a short ringing time is typically desired. Ringing is defined for a given lower bound ratio, α , with respect to the envelope peak value for a given antenna, such that the lower bound is $\alpha \cdot p(\theta, \psi)$ and ringing time τ_r is:

$$\tau_{r=\alpha} = t_{1|_{h^+(t_1)=\alpha p}} - t_{2|_{t_2 < t_1 \wedge h^+(t_2)=p}} \quad (10)$$

Helical Antenna Time-Domain Performance:

In order to compare the time-domain performance of the helical antenna, two antennas are simulated in CST Microwave Studio, with the design parameters are kept constant except for the number of turns and helix length. Both structures are then excited by an impeding plane wave in the azimuth direction ($e_i(t)$), which is equivalent to the field intensity due to the Tx antenna $e_{Tx}(t)$ that was discussed earlier. The plane wave excitation is a modulated Gaussian pulse with spectral components covering the 1 GHz antenna operating bandwidth. The normalized pulse where the absolute peak value is 1 and its envelope are shown in Figure F3 (a).

Voltage ($u_{Rx}(t)$) is then measured as the output at each antenna port, where the port is not being excited by a signal; instead, it is used to monitor the incoming signal. It follows that the voltage transfer function in equation (6) discussed earlier is a complete description of the simulated link in the time domain. In this report, we consider the ideal case, where only the Rx antenna is excited by a plane wave, meaning r_{TxRx} is large. More can be found about the impulse response measurement, and the analysis of transmitting and receiving modes can be found in [4], [5].

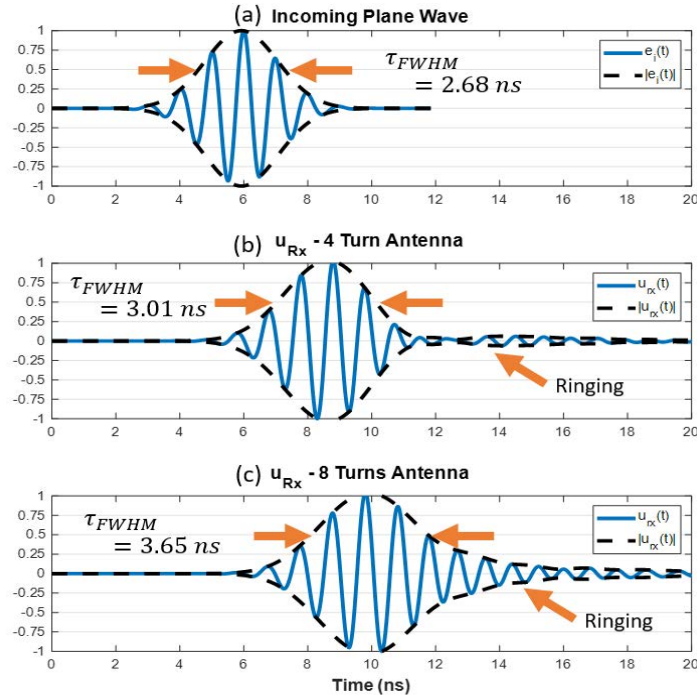


Figure F3. CST Simulation of (a) Impeding wave, and port received voltage at the (b) 4 turn, and (c) 8 turn Helical Antenna.

From Figure F3, it's evident that the received normalized pulse at the 8-turn helical in Figure F3 (c) is more dispersed than that of the 4-turn helical in Figure F3 (b). The pulse received by the 8-turn helical antenna is around 20% wider than the 4-turn case, because $\tau_{FWHM}^{8TURN} = 3.65 \text{ ns}$ and $\tau_{FWHM}^{4TURN} = 3.01 \text{ ns}$. As for ringing, the threshold α is chosen to be .2 of the peak value for the two antennas. Calculating ringing using equation (9) leads to $\tau_{r=2}^{4TURN} = 4.53 \text{ ns}$, and for the 8 turn $\tau_{r=2}^{8TURN} = 6.29 \text{ ns}$. Meaning dispersion and ringing are reduced in the antenna with less turns at the cost of gain. Since maintaining a low dispersion and achieving a high antenna gain are both critically favored in our design, further investigation is needed to optimize both parameters and achieve dispersion levels that can be tolerated or at a satisfactory gain.

Summary:

In this MSR, we showed how we plan to quantify both the frequency domain and the time domain response of our helical antenna design. Work is underway to fully characterize the antenna impulse response of the helical antenna and analyze any adverse effects the antenna can have on the realistic pulses of interest to OSPRES.

Immediate Tasks for the Antenna subgroup:

- Quantify the adverse effect of the helical antenna on the time-domain pulse used in the UMKC measurement setup.
- Explore ideas for full antenna characterization at UMKC facilities using the impulse response method outlined in [4].
- Design a transition to bring down the input impedance to 50 ohms.

References:

- [1] A. A. Smith, "Received Voltage Versus Antenna Height," *IEEE Trans. Electromagn. Compat.*, vol. EMC-11, no. 3, pp. 104–111, 1969.
- [2] IEEE Standards Association, *145-2013 - IEEE standard for definitions of terms for antennas*, vol. 2013. IEEE, 2013.
- [3] W. Wiesbeck, G. Adamiuk, and C. Sturm, "Basic Properties and Design Principles of UWB Antennas," *Proc. IEEE*, vol. 97, no. 2, pp. 372–385, Feb. 2009.
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IV. Research and Development Support (Smith)

G. Device and Materials Modelling (Eifler)

Decreasing Si-PCSS recovery times enable higher pulse repetition rates and therefore higher average power and more efficient waveforms to be coupled to UAS. Device recovery time has been increased by a factor of 5–10 via electron irradiation of devices which introduced defects into the silicon material. To inform and optimize this approach,

simulations are needed expose the transient charge carrier profile within the device and correlate it with imperial recovery time measurements. In this report, we combine visualization of the charge profile with modeling of defects to quantify decreases in charge concentration in the Si-PCSS, as observed for rectifying diodes used in signal conversion [1].

Irradiating a 1.3 mm thick Si-PCSS with 3 MeV electrons at sufficient doses result in approximately uniform defect profiles that significantly reduce device recovery time. In the previous report, we compared simulated and empirical measurement of the charge storage time during electrical reverse recovery. Alternatively, if electrons of a low enough energy are used (hundreds of keV), the absorption depth within the Si results in a decaying defect profiles along the axis of irradiation. These profiles can be combined with the uniform profiles (e.g., from 3 MeV irradiation) to produce a variety of defect profiles, including a profile in which defects are concentrated in the n^- base near the P^+-n^- junction. In theory, such a profile will improve recovery in a way uniform profiles cannot [1].

To evaluate the irradiation profiles, simulations were conducted using two defect profiles: no defects and step defect profile, the latter to coarsely represent a sloped defect profile as a first attempt. Both simulations used a 500- μm thick Si-PCSSs which was photo-illuminated with a 1064-nm light pulse. The stepped profile incorporated 7.8×10^{12} defects/ cm^3 in the region from 0–225 μm into the device, and zero defects from 225–500 μm . Adaptive time-steps were used, hence non-uniform time-steps. The resulting transient carrier concentration profiles from these simulations are shown in Figure G1 and G2 for the zero-defect case and the step defect profile, respectively.

The excess carrier concentration was obtained from the difference of the total carrier concentration and equilibrium carrier concentration (obtained from open-circuited simulations). Crosses and open circles are used to denote excess electrons and holes, respectively, and color correlates to time following photo-illumination.

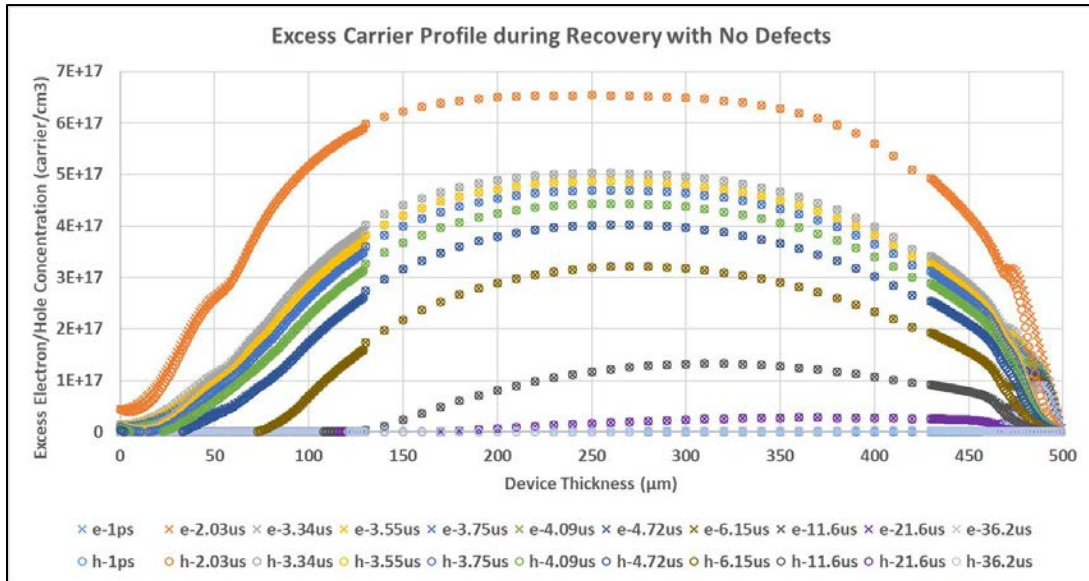


Figure G1. Excess electron and hole concentrations vs depth over various time-steps during photo-illuminated recovery for no defects.

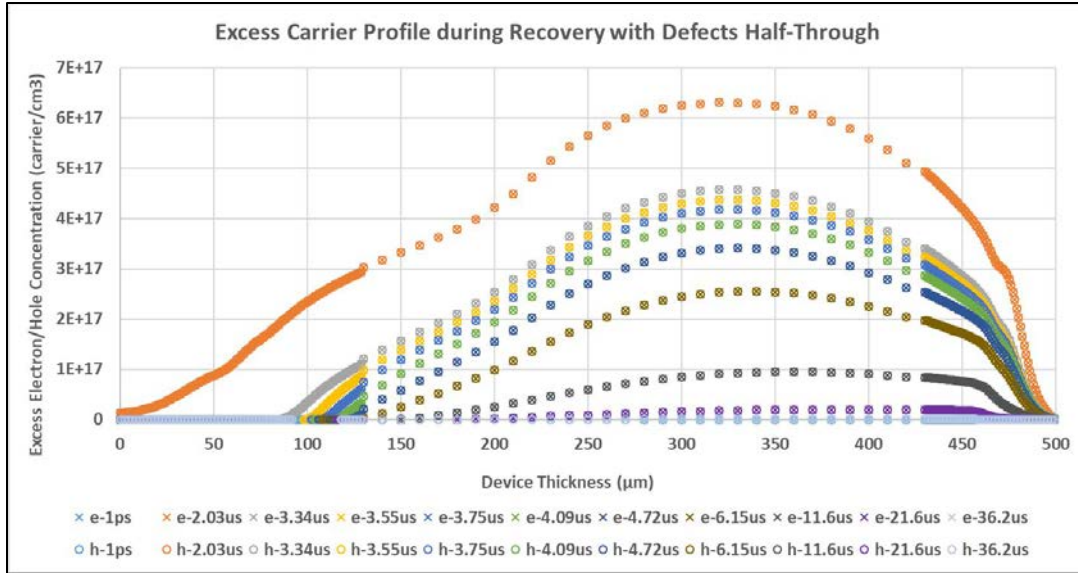


Figure G2. Excess electron and hole concentrations vs depth over various time-steps during photo-illuminated recovery for 7.8×10^{12} defects/cm³ half-way through depth of device.

In both Figures G1 and G2, at 1 ps (medium blue), the initial off-state resistance of the switch is high and excess carrier concentration is low. Immediately after illumination, a large number of charge carriers are produced and the low on-state resistance is achieved. The results from the simulations indicate, as expected, that the excess carrier concentrations (both electrons and holes) are reduced in regions where defects are present, which agrees with [1]. At 21.6 μs, the charge profile has returned to the initial off-state resistance of a recovered device. Another feature, less apparent from the figures, is that the carrier concentration reduces with depth as the optical pulse decreases in intensity due to absorption. Ignoring this, the excess carrier profile is uniform in the intrinsic region and declines only near the device edges.

These results demonstrate that sloped defect profiles are promising for further optimizing recovery time and losses in Si-PCSS as in the rectifying diodes. Moving forward, we will continue to correlate the defect levels produced from electron irradiation of specific energies to better correlate experiment with simulation. Future simulations will also incorporate: (1) more realistic sloped defect profiles by increasing the number of steps to match that of the anticipated defect profiles for specific electron energies, and (2) additional defect types, and validating experimentally.

Critical Tasks for Device Modeling and Simulation subgroup:

- Determine the amounts of defects in simulation vs electron irradiation.
- Use Gaussian slopes for defect profiles.
- Study of laser window width effect on current density within device.

Reference:

- [1] Grekhov, I. V., L. S. Kostina, V. V. Kozlovskii, V. N. Lomasov, and A. V. Rozhkov. "Electron irradiation controlled profile of recombination center concentration in silicon." *Technical Physics Letters* 37, no. 5 (2011): 442-444.

H. GaN:C Calculation (Rulis)

The main activity to report over the past month is that TCAD simulations of GaN transport properties were performed using the MOCASIM simulator. MOCASIM is a device simulator tool based on Monte Carlo methods that is integrated with Silvaco TCAD. The student determined the size of the band gap and the set of Brillouin zone k-points to define the valleys and the scattering phenomena between different energy states under an applied electrical field.

Because GaN requires certain parameters we have modified the input script parameters such as band gap, DC dielectric constant, etc.

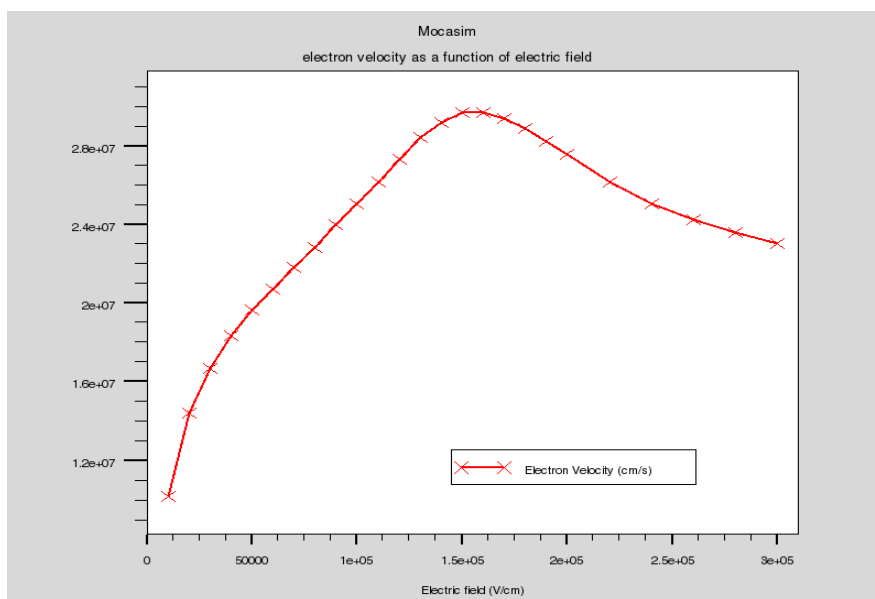
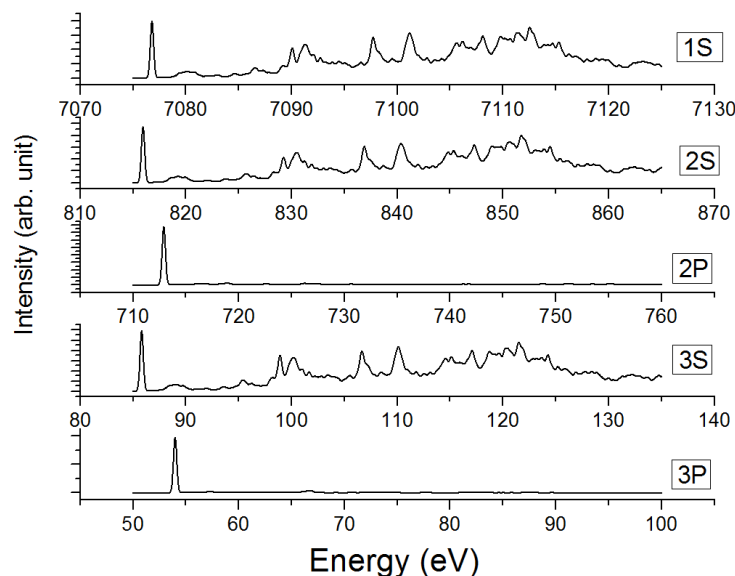


Figure H1. The electron velocity, as a function of electric field, increases into 2.9×10^7 (cm/s) when the applied electrical fields is 1.5×10^5 (V/cm) the velocity of electron starts to decrease slowly.

We did an evaluation on Silvaco TCAD code to connect its capability of calculating device properties to our list of parameters that we will generate from *ab initio* methods. For C and Fe doped GaN, new scripts need to be developed to compute the dielectric constant, thermal conductivity and breakdown voltage for C-GaN and Fe-GaN based on the new generated parameters.

The main activity to report over the past month is that the student continued using VASP+PyCDT to produce a range of defect models. This time of Fe compensated GaN: (substitutions, interstitials, different charge states, and vacancies in standard hexagonal

and also cubic GaN). For many of the models, the student was able to compute core level spectra (ELNES spectra). Charge defect models were created, but computing ELNES spectra of such models is not straightforward.



Additionally, the student discussed with other team members (Eifler) methods to evaluate the capabilities of Silvaco TCAD and determine its capability for calculating device properties making use of the list of parameters that we will generate from *ab initio* methods. For C and Fe doped GaN, new scripts need to be developed to compute the dielectric constant, thermal conductivity and breakdown voltage for C-GaN and Fe-GaN.

Issues:

[1] **Scope** – N/A

[2] **Budget** – N/A

[3] **Schedule** – N/A

Interactions with Others:

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

Major Procurement Actions: N/A

Travel:

Travel Plans

Please note that all papers or presentations on any aspect of the ONR HPM program must be reviewed by the ONR Program Officer (or in some cases ONR Public Release Process) regardless of classification level or audience.

Planned Travel Details	Purpose	Attendees	Estimated Costs

Pending or Upcoming Public Release Requests:

N/A

ONR HPM 2019 Program - Monthly Status Report (MSR) – JANUARY2020
Anthony Caruso – Univ. Missouri – Kansas City
N00014-17-1-3016 [OSPRES Grant]
ONR Short Pulse Research, Evaluation and non-SWaP Demonstration for C-sUAV
Study

02/15/2020

Distribution Statement ~A: Distribution authorized to all OSPRES Grant and Contract Participants, and any uncompensated collaborators. Further dissemination per Ryan Hoffman (ryan.hoffman@navy.mil).

OSPRES Project Description: Measure, analyze and summarize the optimal high power microwave waveform (pulse width, pulse repetition frequency, center frequency, polarization, bandwidth) that simultaneously yields the lowest power density AND lowest possible size, weight, monetary cost and cooling high power microwave system (i.e., driver and radiator) that can produce such waveforms. Modularity and scalability are also critical requirements of the overall design/downselect criteria, that are considered as part of the tradespace study analysis (i.e., within the context of the optimal waveform).

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Executive Summary:

Notable achievements during the month of January include:

- A new 4-level HV switch using SiC MOSFETs has been designed by Dr. Khan's group. Simulation results are extremely favorable for giving the program a low cost solution for initial testing of the SiPCSS system. Fabrication and testing will begin shortly to verify the simulation results.
- The RF coupling group began investigating the efficiency with which varying RF pulse shapes (pulse duration, peak, rise and fall times, etc..) couple to particular wire geometries. This information, when combined with experimental RF coupling data, will inform the most ideal pulse/waveform shape to achieve optimal coupling and induced current within a specific wire system.
- By operating microstrip patch antenna prototypes in transmit mode within the GTEM, improved agreement with simulations has been achieved. Previous attempts to characterize the antenna prototypes in receive mode showed substantial amounts of noise in the acquired datasets causing concern over the validity of the simulated data. The transmit mode eliminated the noise and improved overall confidence in both the experimental and simulated data. The results can now inform further research into arrays of the microstrip patch antenna designs.

Current Month Progress:

I. Source Development (Myers)

A. PCSS Power Conditioning - IGBT (Khan)

- Designed a 4-level HV switch using SiC MOSFETs. Initial simulation results are favorable, and we are in the process of fabricating this circuit.
- Ordered SiC BEHLKE switches, and expect to have them delivered by March.
- Performed Si-BEHLKE power conditioning circuit testing with SiPCSS (with a cooling system) at a lower voltage (<3 kV) and at a higher repetition frequency (>50 kHz).

B. Switches - GaN:C (Thompson)

- Continued work on manuscript for cost-analysis of GaN:C manufacturing for HPM and pulsed power applications.

C. Switches - Figure of Merit (Bhamidipati)

- Efforts have been focused on refining content for publication, to include the need-for-cooling, and a cost-effectiveness perspective.
- Data analysis has been completed as a part of material classification based on heat dissipation with varied recombination time.
- To quantify device lifetime and the need for cooling at the rate of heat dissipation, temperature-rise rate analysis has been completed for the following scenarios:
 - o Considering heat transfer based on thermal conductivity
 - o Ignoring heat transfer theoretically

II. Effects, Coupling and UAS Developments (Currie)

D. RF Coupling (Hassan)

- Studied several simple pulse shapes to see how the Characteristic Mode Analysis can predict which pulse can maximize coupling to a particular wire system.
- Worked on establishing a framework to link the RF coupling analysis and predictions with experimental results.

E. UAS Developments (Fields)

- Creating (in progress) UAV platform manuals and building procedures for standardizing the vehicles.

III. Antenna Development (Crow)

F. Antennas - Electrically Small Antennas (Chatterjee/Barman)

- A single element antenna designed on Rogers TMM-10i substrate at a center frequency of 3.6 GHz, which yielded 53.5% 2:1 VSWR bandwidth and a peak gain of 5.65 dBi, was first measured using GTEM with the antenna in the receive mode.
- Due to the jagged nature of the measured data (as shown in Figure F1(a)), the measurement was repeated in the open space as well as using GTEM with the antenna in Transmit mode.
- Measurements from the repeated tests showed improved alignment with the simulated data (Figure F1(b)).

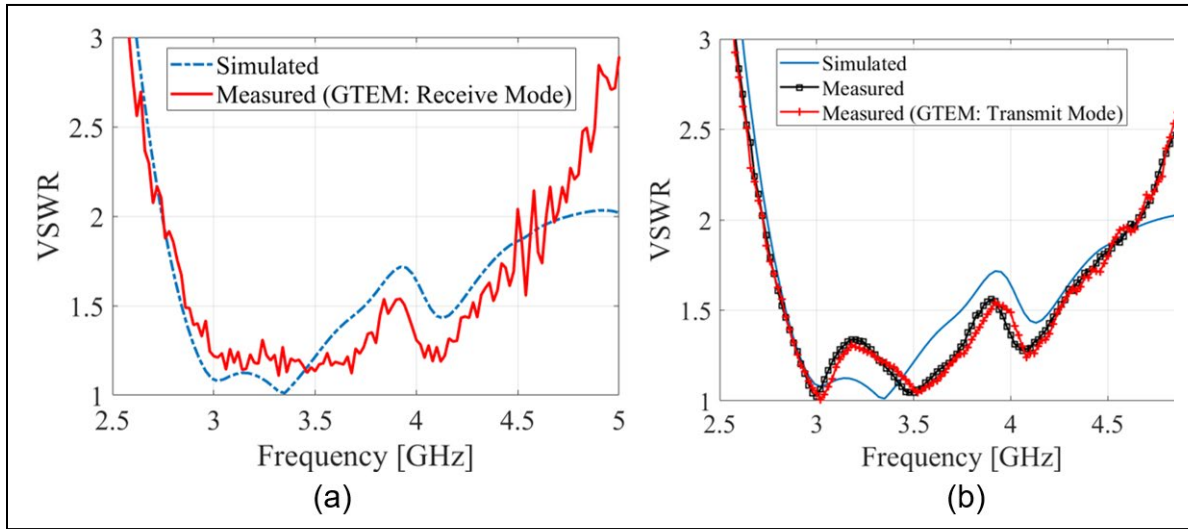


Figure F1. Simulated data for an antenna in (a) receive mode and (b) transmit mode compared with simulation.

- Investigations into array radiation patterns for different types of antenna elements and for various amplitude distributions were performed.
- MATLAB code to accomplish the above task is in progress.

G. Antennas - Log Periodic (Hassan)

- We worked on an impedance taper design for the 1 GHz helical antenna to smoothly transition the input impedance from 50 to 160 ohms. Two impedance transformer designs of interest are the exponential and Klopfenstein tapers. The design parameters were calculated in MATLAB.
- The next step is CST simulation and design verification. An exponential taper provides a smooth and continuous transition with low ripples, while the Klopfenstein taper is the most optimal for a given length. Both designs will be useful in future optimizations to minimize size and maintaining performance.
- The impedance taper acts as a bandpass filter. Inserting a filter at the feed point can affect the input pulse in several ways. So, time-domain analysis and optimization going forward will include the taper.

IV. Research and Development Support (Smith)

H. Device and Materials Modelling (Eifler)

- TCAD simulations were conducted using uniform defect profiles over a larger range of defect concentrations (1×10^{12} to 1×10^{15} defects/cm³). The recovery process was analyzed via the diode output voltage, bias, resistance and charge-carrier profile. The simulations suggested that recovery time can be reduced to lower levels than seen in experiment with larger defect concentrations (produced by longer experimental electron irradiation times). For a defect concentration of 1×10^{15} defects/cm³, a recovery time of 141 ns was observed in simulation. This lowered recovery time is not obvious nor have we found experimental data that shows such decreases. Experimental irradiation on devices of a dose of 40 MRad are being performed to determine if recovery time can be reduced further. If recovery time can be significantly reduced with more irradiation time, as modeled in simulation by the defect concentration, then a major improvement in device performance will be achieved.
- Gaussian defect profiles based on published data for electron irradiation of devices found in the literature were generated comprising defect depths of 100, 200, 300 and 400 μm . Ten simulations were conducted, eight comprised of single-side (n- or p-side) and two comprised of first a 100 μm depth on the p and n-sides and another of 200 μm depth on both sides. These Gaussian defect profiles can be combined with uniform defect profiles to give higher overall defect concentrations. We have not analyzed the data yet and will do so in the next report.

I. GaN:C Calculation (Rulis)

- TCAD simulations of GaN transport properties were completed using MOCASIM simulator: MOCASIM is a device simulator tool based on Monte Carlo Methods.
- We determined the size of the band gap and the k -points of the Brillouin zone to reproduce a three-valley model using the three lowest conduction band minima.

Table I1. The valley parameters corresponding to wurtzite GaN (El-Ela, F. & Mohamed, A.. (2013). *Electron Transport Characteristics of Wurtzite GaN*. ISRN Condensed Matter Physics. 2013. 10.1155/2013/654752.).

Valley	$\Gamma 1$	$\Gamma 2$	L-M
Valley degeneracy	1	1	6
Effective mass	$0.2 m_e$	m_e	m_e
Intervalley energy separation (eV)	---	1.9	2.1
Energy gap (eV)	3.39	5.29	5.49
Nonparabolicity (eV ⁻¹)	0.189	0.0	0.0

- Transport properties of GaN were reproduced using the three-valley model of conduction band using published parameters that were theoretically calculated.

Issues:*[1] Scope – N/A**[2] Budget – N/A**[3] Schedule – N/A***Interactions with Others:**

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

Major Procurement Actions: N/A**Travel:**

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Planned Travel Details	Purpose	Attendees	Estimated Costs
DEPS Spring	Conference	12 members of OSPRES	TBD

Pending or Upcoming Public Release Requests:

N/A

ONR HPM Program - Monthly Status Report (MSR) – February 2020

Anthony Caruso – Univ. Missouri – Kansas City

N00014-17-1-3016 [OSPRES Grant]

ONR Short Pulse Research, Evaluation and non-SWaP Demonstration for C-sUAS

03/15/2020

Distribution Statement ~A: Distribution authorized to all OSPRES Grant and Contract Participants, and any uncompensated collaborators. Further dissemination per Ryan Hoffman (ryan.hoffman@navy.mil).

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2.0 Executive Summary

During February, the UMKC team focused on: (1) finalizing defect-controlled recovery time studies on Si-PCSS via TCAD simulations, (2) continued testing and simulation of GaN:C devices, (3), refining, collecting, and analyzing results from high power, figure of merit calculations for linear PCSS, (4) HV switch design using capacitively coupled single-gate driver topologies, (5) transformer design, fabrication, and testing for helical antennas, (6), trade-space studies of electrically small antennas (ESAs), and (7) RF coupling between loaded wire loops as a function of angle of incidence, loop curvature, and frame material.

The SiPCSS modeling and design team has previously demonstrated the benefits of electron irradiation for reducing recovery time to single microseconds, and recent efforts indicate that Gaussian defect profiles (corresponding to irradiation at lower electron energy) show little benefit for photo-illuminated recovery time (increasing the Gaussian profile depth converges with the uniform case). Over the coming weeks, efforts will be shifting toward optical coupling to minimize the approximately 30% losses due to contact metallization.

A series of tests were conducted on GaN:C PCSSs achieving a peak power of 10 kW, but ultimately limited by the encapsulation of the PCSS and pulse forming network. While testing will continue, efforts will be shifted toward evaluation of GaN-based PCSS in terms of cost and future performance, relying on modeling and experimental validation to quantify optimal modes of operation. Special cases of the figure of merit for high power, linear PCSS will be generated to aid in these answers.

The pulse charger team has simulated, designed, and sent for fabrication the first prototype of the single gate driver HV switch for charging of nF-scale pulse forming transmission lines. This approach has the potential to drive the cost of pulse chargers down by exploiting relatively low-cost, medium voltage SiC-MOSFETs in a series configuration driven by only a single gate driver. Design validation continues using a surrogate load, which will be transitioned to a SiPCSS-PFTL for final testing and optimization. In parallel, the SiC-BEHLKE switch topology will be completed and used as a COTs solution for cross-comparison.

The helical antenna team focused on designing, simulating, fabricating, and testing impedance transformers for the helical antenna based on Binomial and Klopfenstein tapers. The initial results with the Klopfenstein show reasonable agreement with experiment (i.e., within a 3 dB of simulation). The team will continue with prototyping and validating the Binomial transformer, and explore options for integrating the impedance transformer directly with the helical antenna feed.

The ESA team conducted simulations of single-element and 11×11 microstrip patch ESA arrays and characterized VSWR and beam patterns in the horizontal and vertical planes for finite and infinite ground plane under uniform and single-element excitation conditions. While the infinite case show a gain of 25 dBi, the performance degraded substantially in the finite ground plane case. Single-element excitation is also being explored via simulation to determine the behavior of single elements embedded within an array.

3.0 Switch Evaluation and Design

3.1 Si-PCSS Modeling and Evaluation

3.1.1 Overview

At high pulse repetition rates (PRRs), Si-PCSS based pulsed power and RF sources can be improved by reducing recovery time to prevent unnecessary leakage current during the off-state. Specifically, for PRR of 1 MHz or more, recovery time should be reduced to 1 μ s or less. Reducing the recovery time is generally achieved by introducing defect states using charged particles (e.g., electrons, protons), a standard process in the semiconductor industry. Our current focus is to quantify the improved recovery time of Si-PCSS devices via simulation by comparing sloped vs uniform defect profiles at various concentrations, and subsequently quantify improvements in efficiency at high PRR.

3.1.2 Recovery-time optimization

TCAD simulations of Si-PCSS recovery time were performed by modeling a capacitor discharging via a photo-illuminated SiPCSS using the circuit model in the Aug 2019 Grant Report, comprised of a 32- μ F capacitor, SiPCSS TCAD model, and 50- Ω load. The large capacitance is used to reduce the depth of discharge on the capacitor during pulsing. The laser energy used is 326 μ J, and we defined recovery time as 10% of the maximum of the diode output voltage.

In device processing, defects are introduced in the semiconductor devices using electron irradiation, largely due to its relatively low cost (not requiring a vacuum) compared to other techniques. In simulation the defect profiles representative of irradiation are modeled directly, rather than modeling the electron irradiation. We split the defect profiles into two categories: (1) uniform, representative of 3-MeV electrons and (2) Gaussian sloped, representative of 270–700-keV electrons. The defect state having the greatest impact on carrier lifetime—and thus recovery time—is the Vacancy-Oxygen (VO) acceptor state at 0.165 eV, per Figure 3.1.1.

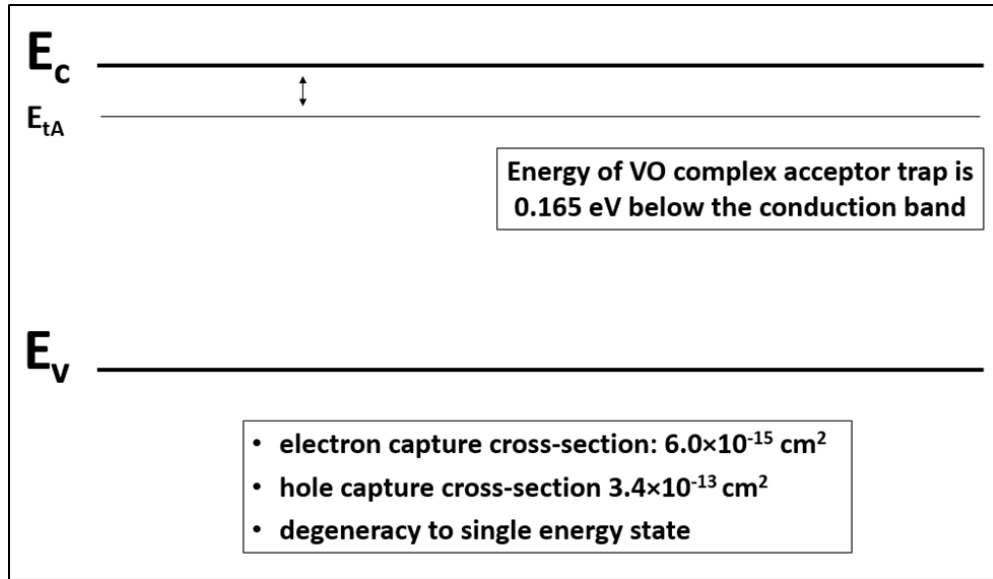


Figure 3.1.1. Band-gap diagram and defect parameters.

The Gaussian slope is generated by an analytical function and adjusted the Gaussian profile to approximate profiles due to varying electron energy. The mean of the Gaussian profile was varied to emulate irradiation from the p-side, n-side, and from both sides of the device. All Gaussian defect profiles peak at of 1×10^{13} defects/cm³ and fall off to 1×10^{12} defects/cm³ at the maximum depth. The Gaussian defect profiles simulated are shown in Figure 3.1.2.

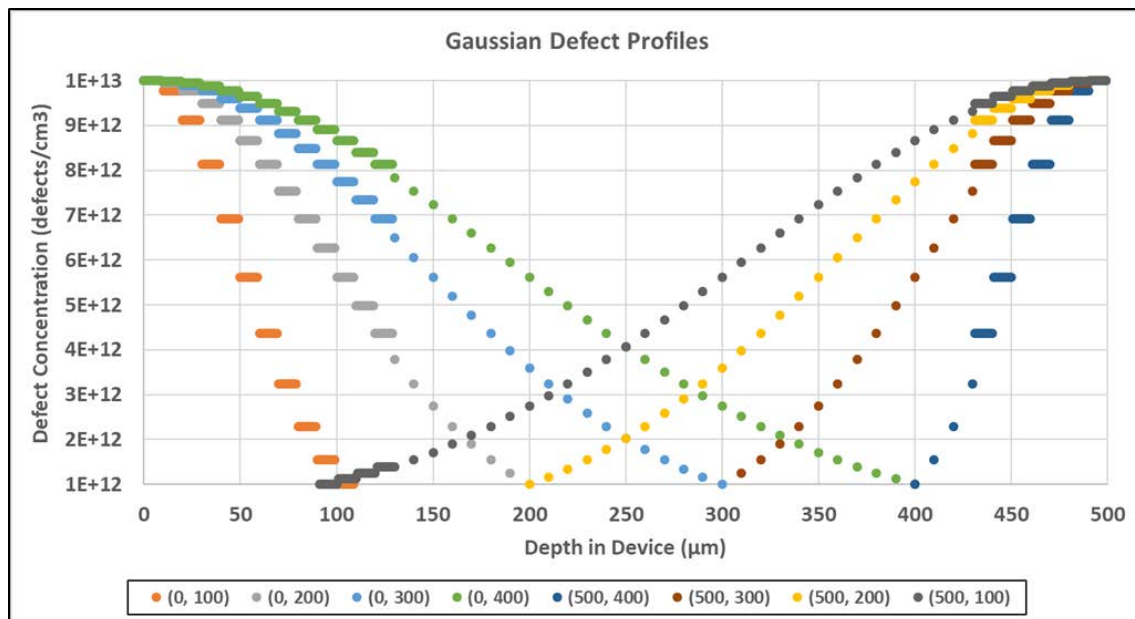


Figure 3.1.2. Gaussian defect profiles from p or n-side used in simulation for a 500- μm device. The depth of the Gaussian defect profiles is notated as (peak-position, depth) such that p-side profiles are (0, depth) and n-side are (500, 500 - depth).

Unlike the Gaussian defect profiles, the uniform profiles used a fixed defect concentration throughout the entire device model. A total of seventeen profiles were studied, listed in Table 3.1.1.

Table 3.1.1. Simulation Uniform and Gaussian Defect Profiles.

Uniform Defect Profile Concentration (defects/cm ³)	Gaussian Defect Profile Depths (μm)
No Defects	(0,100)
1.0×10^{12}	(0,200)
3.9×10^{12}	(0,300)
7.8×10^{12}	(0,400)
1.0×10^{13}	(500,400)
1.0×10^{14}	(500,300)
1.0×10^{15}	(500,200)
	(500,100)
	(0,100) (500,400)
	(0,200) (500,300)

The recovery time was calculated from the output pulse and compared to the total defects, as shown in Figure 3.1.3. Note, a SRH lifetime of 10-μs is also included in cases with defects as a base material lifetime before irradiation. As predicted, the “No Defects” case (not shown) had the longest recovery time (17 μs). First, for uniform defect profiles, increasing defect concentration decreased the recovery time, making it theoretically possible to achieve sub-μs recovery time. However, VO complex defect concentrations have a practical limit corresponding to the concentration of O atoms. Second, a fall-off of reduced recovery time is seen in experiments for increasing irradiation doses which may be independent of the O atom concentration. To quantify these defect concentration limits, Si-PCSS devices are currently being irradiated at greater dose following which their recovery time will be characterized.

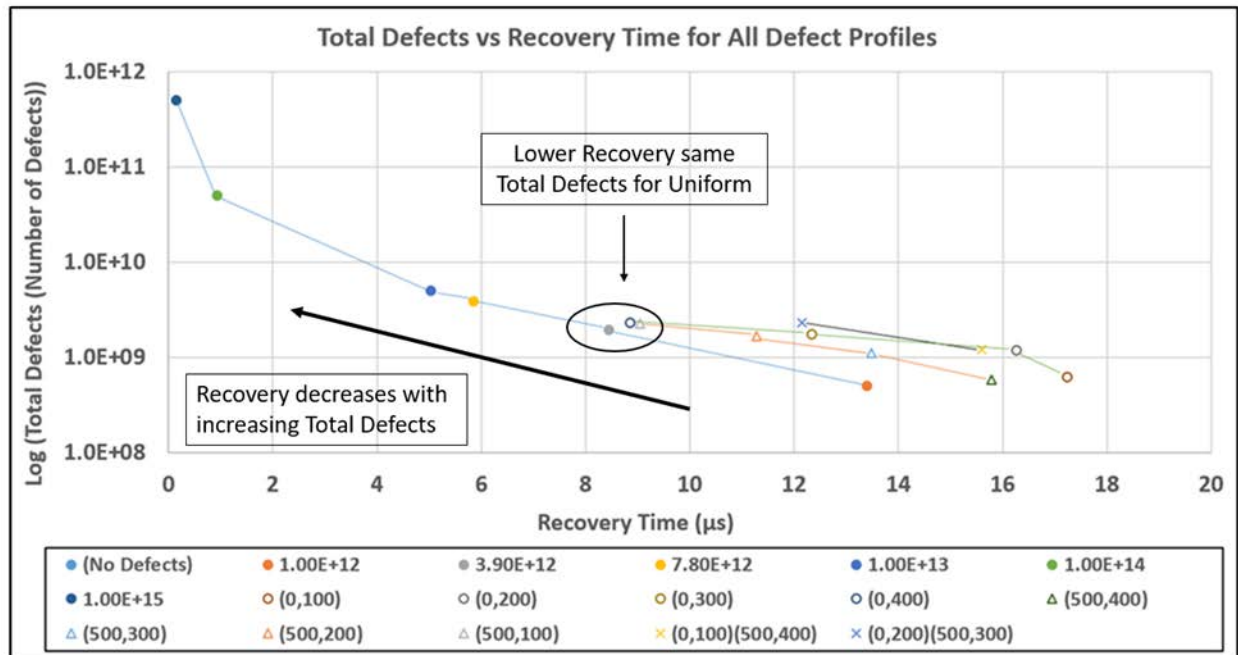


Figure 3.1.3. Total defect concentration vs recovery time for all simulated defect profiles: (circles, filled and blue line) uniform defect profiles, (circles, open, and green line) p-side defect profiles, (triangles, open, and orange line) n-side doping profiles, and (crosses, black line), p- and n-side defect profiles.

Regarding the Gaussian defect profiles, no significant improvement in recovery time was achieved over the uniform profiles at the same total concentrations. In fact, for the p-side only (open circles) and n-side only (open triangles) at a 400- μm depth, the recovery time appears to be converging with to the uniform case, as indicated by the circled region on Figure 3.1.3.

The original interest in sloped defect profiles was motivated for improvements in switching efficiency, independent from recovery time, applied to rectifying diodes rather than laser-triggered pulsed power. In those applications, defects are included near the P⁺-n⁻ junction and even the N⁺-n⁻ junction to improve switching efficiency. For photo-conductive switches, switching efficiency in electrical reverse recovery (i.e., the analog to rectification) and photo-illuminated recovery in a discrete (capacitor discharge) or transmission line (PCSS pulsed power) pulse forming network are not identical nor quantitatively comparable. Thus, it was not feasible to directly quantify improvements in efficiency in all circuits using only the discrete (capacitor discharge) circuit model. Instead, the energy dissipated in the switch vs the load due to Joule heating was calculated and for n-side only defects at a 200- μm depth, the “efficiency” was somewhat improved, apart from recovery time. It is not well understood how such defects quantitatively influence switching efficiency for a TL discharging through the switch to the load. Future simulations can help to address these questions.

3.1.3 Next Steps for Si-PCSS

- i. Determination of minimum laser requirements from 2D parameter study in TCAD varying metal contact window spacing and laser energy

3.2 GaN:C

3.2.1 GaN:C Testing and Evaluation

To push the performance of GaN:C PCSS and measure peak output power, a lateral GaN:C PCSS with a 500- μm gap length was pulse tested from 100–1010 V charge voltage, with fixture failure occurring at 1010 V. The device was vacuum encapsulated using Sylgard™ 184 (dielectric strength: 19 kV/mm) to prevent arcing between contacts. Attenuation of the incident 355-nm laser pulse was increased due to an additional layer of encapsulant, resulting in a 20- Ω on-state resistance. While differences in doping or defect density between devices could also cause this difference, this is less likely as the devices were diced from the same wafer.

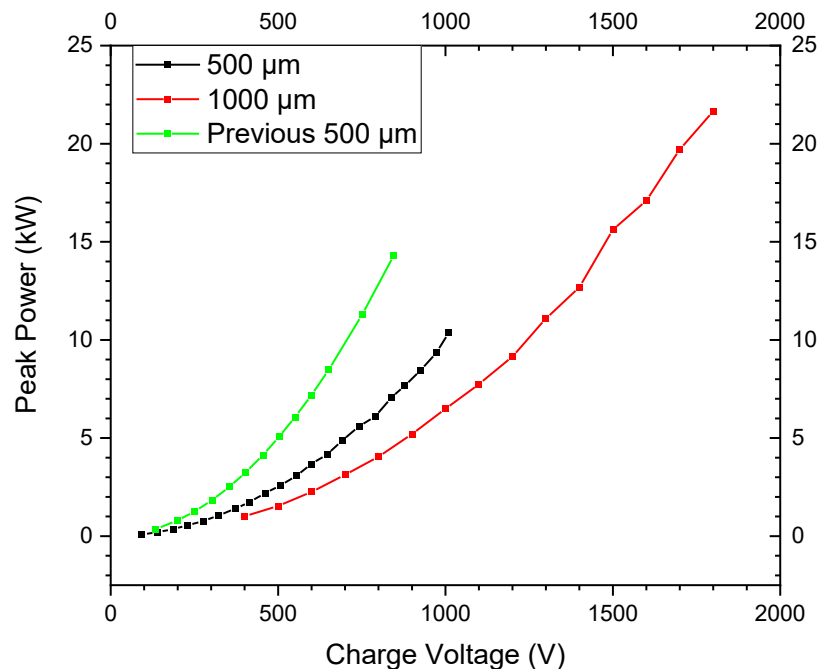


Figure 3.2.1. Comparison of calculated peak power from testing on three GaN:C-PCSSs: (black) recent tests on 500- μm device, (red) recent tests on 1000- μm device, and (green) tests from previous reports on 500- μm device.

The calculated peak power output (V^2/Z with $Z = 50 \Omega$) was approximately 10.3 kW at 1010-V charge voltage, resulting in peak pulse voltage of 720 V (Figure 3.2.1). Improving the application of encapsulant will be a priority for future testing. In order to improve light coupling and decrease on-state resistance, it is expected that—with sub-ohm on-state resistance—20-kW peak power can be achieved. Testing will be performed with multiple

1000- μm and 100- μm gap length devices to determine device consistency, improve encapsulation, and peak power.

Device failure occurred due to arcing between the resistors and copper on the PCB. Future testing, better cleaning/encapsulation techniques will be used and the entire board will be covered in encapsulant to prevent arcing between the charge resistors.

3.2.2 GaN:C Calculations

Different transport properties of GaN were obtained using the three-valley model of the conduction band using published parameters that were theoretically calculated.

We have reproduced results from the published literature (1,2) with generic n-type doping and are now working to reproduce breakdown voltage results and to extend to include specific C and Fe doping effects from first principles. The results from these simulations are shown in Figure 3.2.2 through 3.2.5 over an electric field from 1 kV to 300 kV at a doping concentration of $1\text{e}14\text{ N/cm}^3$, and the temperature of 300 K.

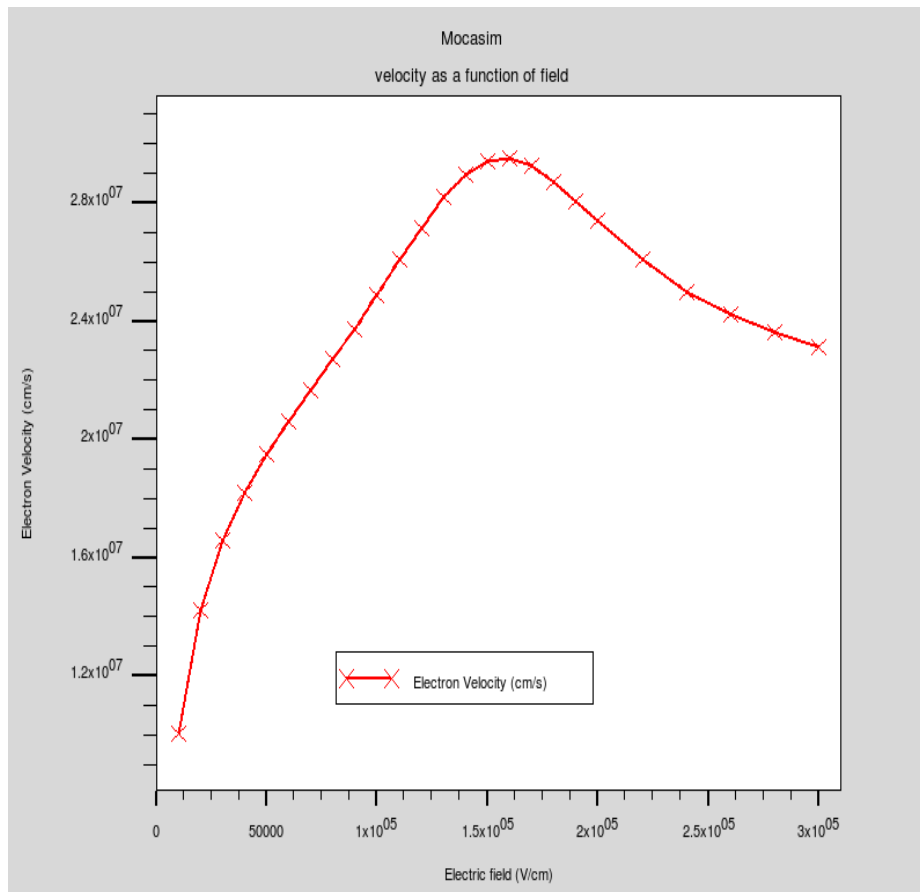


Figure 3.2.2. Electron velocity vs electric field.

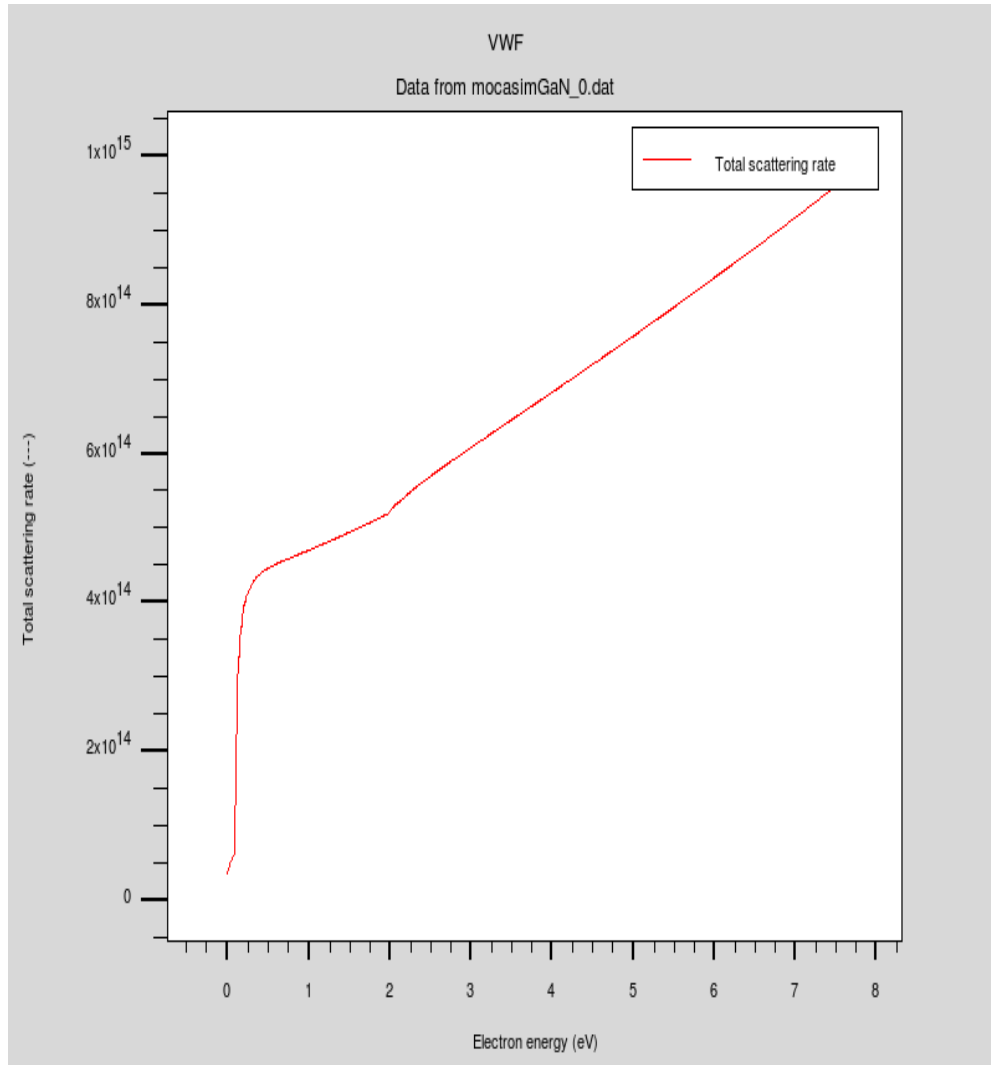


Figure 3.2.3. Total scattering rate.

From Figure 3.2.2 and Figure 3.2.3, when the applied electric fields less than 1.6×10^5 V/cm, the scattering processes are dominated by acoustic deformation potential, piezoelectric phonon scattering, and impurity scattering. Thus, at low fields, where the electron temperature is less than the polar optical phonon temperature, Polar optical phonon emissions are a rare event and the free carriers remain a majority in the Γ_1 valley where their effective mass is weak. Their speed increases with the electric field to reach a maximum value of 2.85×10^5 m/s at 1.6×10^5 V/cm.

At higher electric field, electrons transfer to the upper valleys where they become heavier and their velocity decreases toward its steady-state value due to the increasing of their effective masses.

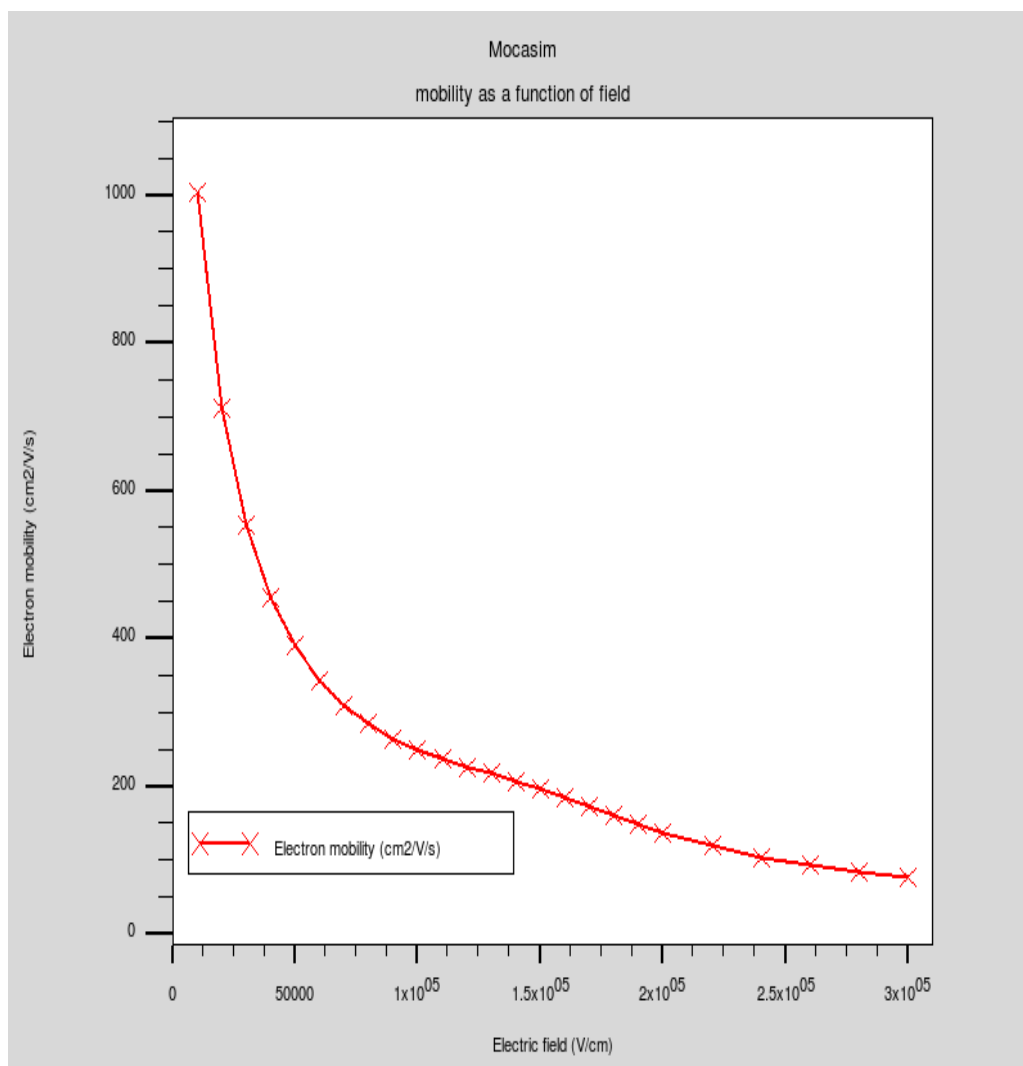


Figure 3.2.4. *Electron mobility vs electric field.*

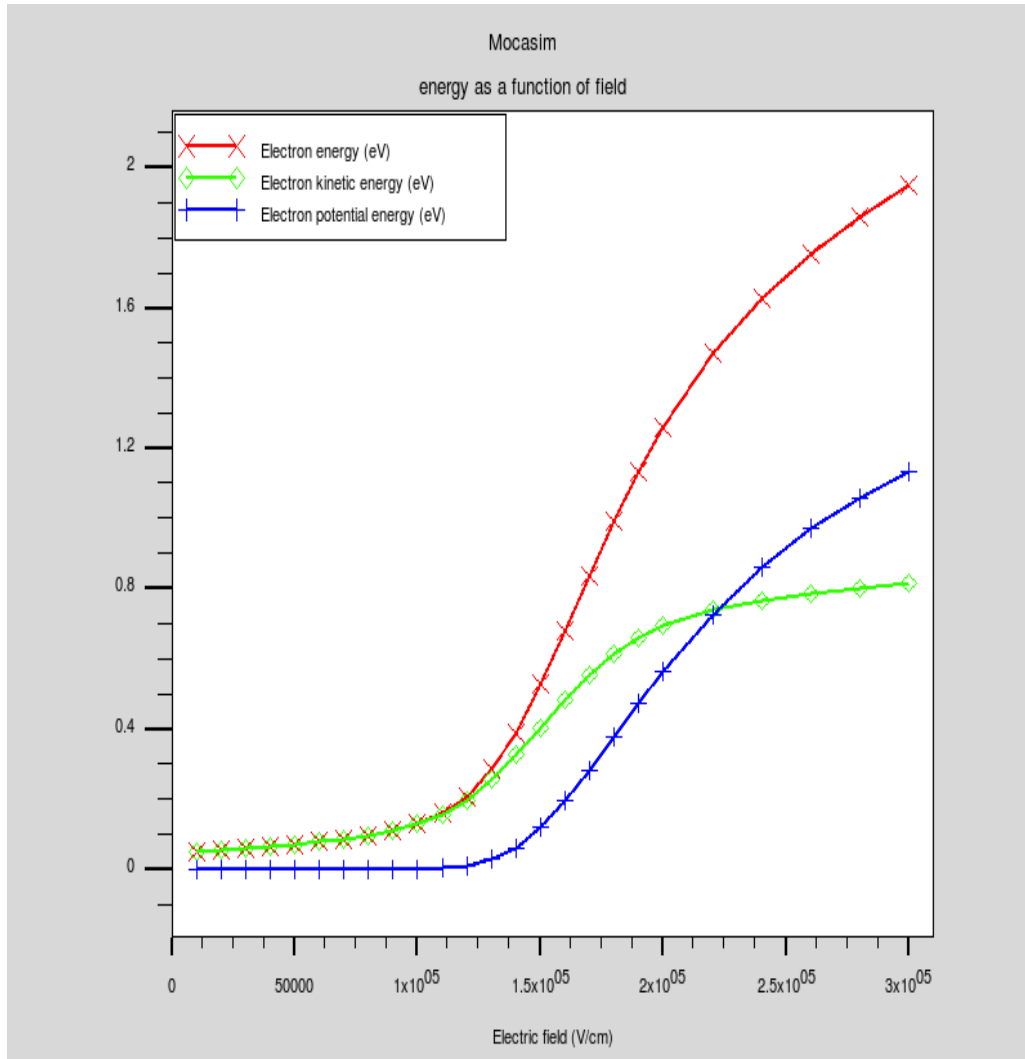


Figure 3.2.5. Kinetic, potential, and electron energy.

From Figure 3.2.4, electron drift mobility has its maximum value at very low values of the electric field due to the smallest effective mass for the electrons in the central valley. The negative differential mobility NDM is observed by increasing the electric field using the inter-valley mechanism. Linear low-field electrons drift velocity depends on the applied electric field, $V_d = \mu F$ in this regime. At high electric field, the average electron drift mobility reaches low values due to the increase of electron effective mass in higher valleys.

From Figure 3.2.5, Kinetic energy, Potential energy, and electron velocity are maximized with the increasing of the applied electric field.

References:

- [1] El-Ela, F. & Mohamed, A.. (2013). Electron Transport Characteristics of Wurtzite GaN. ISRN Condensed Matter Physics. 2013. 10.1155/2013/654752.

[2] Karmalkar, S., & Mishra, U. K. (2001). Enhancement of breakdown voltage in AlGaIn/GaN high electron mobility transistors using a field plate. IEEE Transactions on Electron Devices. <https://doi.org/10.1109/16.936500>

3.2.3 Next Steps for GaN:C

- i. On-state resistance comparison of devices with varying carbon concentrations of $8E16$ vs. $5E17$.
- ii. Determination of pulsed peak power limits for 1000-, 500-, and 100- μm devices by improving encapsulation techniques and applying to entire pulse forming network.

3.3 FOM

3.3.1 High Power FoM for Linear PCSS

We have concluded initial efforts to evaluate uncompensated photoconductive materials using a high-power figure of merit (HPFoM) based on power dissipation. Continuing from the December 2019 period, we focused on quantifying turn-off and turn-on losses, and temperature rise rate at varied- lengths and recombination times at $3E+5$ V/cm electric fields, using the material properties reported previously. For reference, the conditions/restrictions and assumptions applied are re-iterated below:

Optical pulse assumptions (carrier generation):

- Gaussian temporal profile with $t_{FWHM} = 2$ ns
- Uniform optical power density throughout switch volume
- 100% absorption (no reflection or transmittance)
- 100% contact injection efficiency
- 10^{17} photons per pulse (i.e., 10^{17} photo-generated carriers)

Carrier transport:

- Transit time calculated from electron saturation velocity and carrier path length.
- Recombination times varied from 5 ns to 100 ns for all materials, irrespective of bandgap and type.

The temperature rise and the device lifetime were calculated for two different cases: (1), No heat transfer (adiabatic heating)—all power dissipated contributes to lattice temperature rise based on mass and specific heat of device and (2) maximum heat transfer, based on device's thermal conductivity and geometry, assuming a T_{COLD} of 298K. These cases are considered to help evaluate materials based on their cooling requirements at different heat transfer rates.

Figure 3.3.1 shows the trends of heat dissipation rate with varied carrier path length (device length) quantitatively. With increased carrier transit time, adiabatic dissipation increases by a scaling factor unique to the material. At 500 μm carrier path length, the dissipation range is in the order of mJ/sec, with GaAs has the highest dissipation at an average of 1.1mJ/sec and Ga_2O_3 at lowest with 0.1 mJ/sec. These trends are better understood when studied under constant carrier density with varied device lengths.

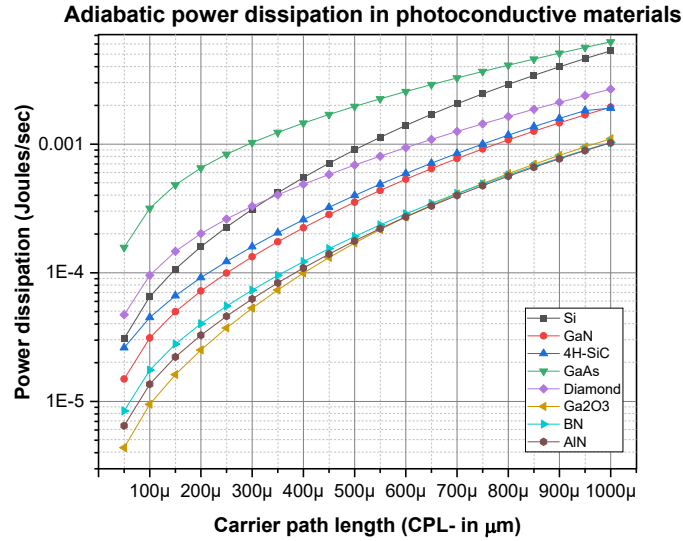


Figure 3.3.1. Power dissipation rate (Joules/sec) calculated with electric field $3E+5$ V/cm field, $1E+17$ carriers per pulse and 50 kHz pulse repetition frequency and 50ns recombination time.

The power dissipated during turn-off is dependent on recombination- time and events, directly affecting the turn-off transient time increasing heat dissipation, as shown in Figure 3.3.2. From the calculations so far, recombination time dependence of power dissipation is found to be in the order of μJ ; but theoretically, is expected to be in the order of 100's of μJ . The realistic trend is not reflected in the plots as the heat released per recombination event has not yet been included in the evaluation model, and will be investigated in the near future. To coherently represent the adiabatic dissipation in different materials and to accommodate wide range of data, only two materials are plotted per figure. The temperature rise rate calculations for adiabatic and non-adiabatic lattice heating for different lengths are shown in Figure 3.3.3 and Figure 3.3.4. These trends are used to calculate device lifetimes (not reported here).

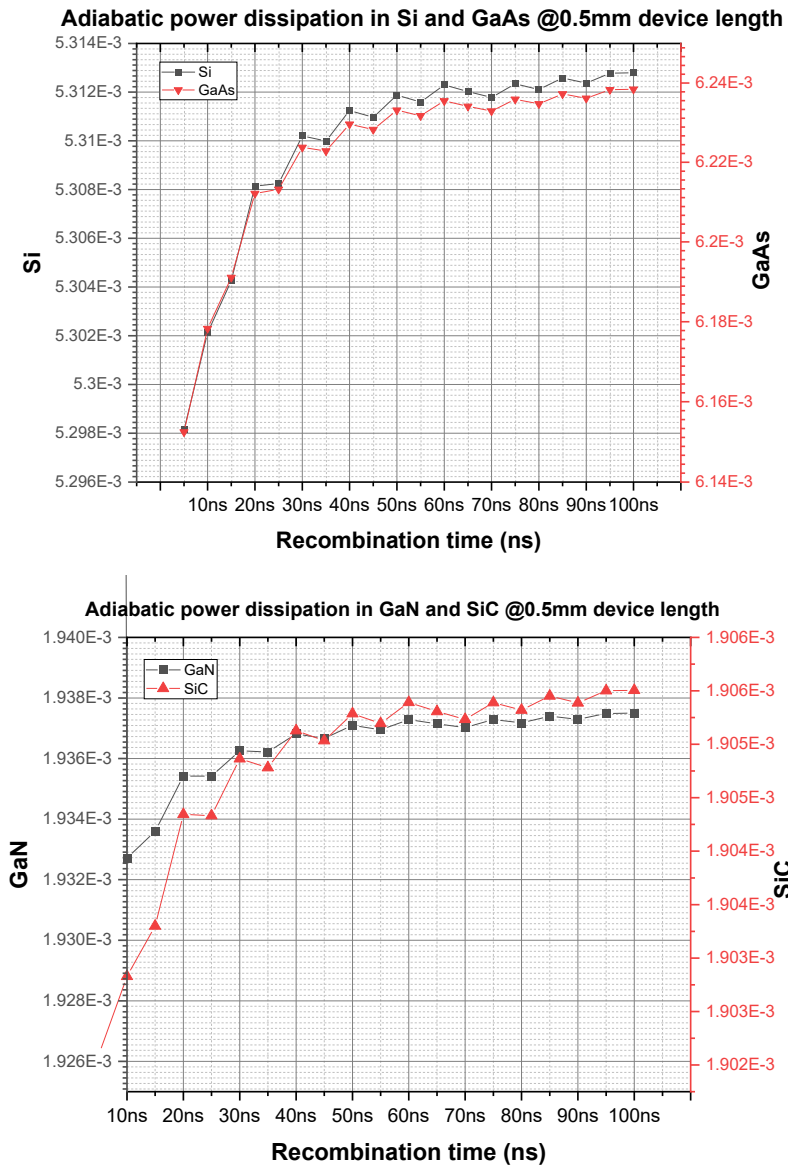


Figure 3.3.2. Heat dissipation rate (Joules/sec) calculated for 0.5mm device with electric field $3E+5$ V/cm field, $1E+17$ carriers per pulse and 50 kHz pulse repetition frequency at varied recombination times. Two materials are plotted per figure to coherently accommodate data in different scales.

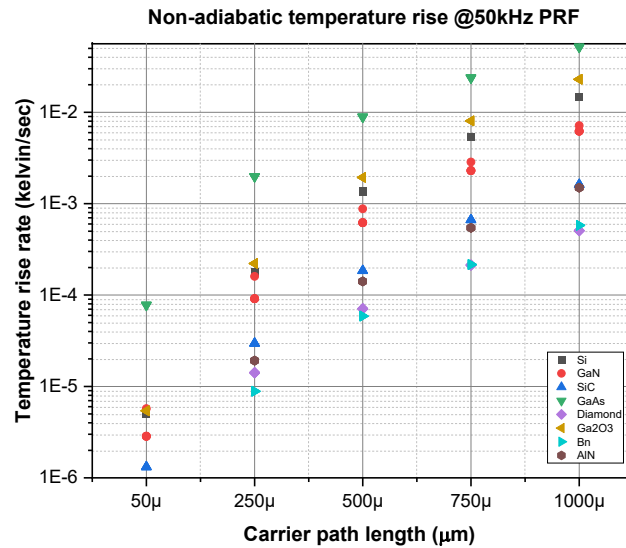


Figure 3.3.3. Temperature rise rate (Kelvin/sec) across materials at 50 kHz PRF and 3×10^5 electric field assuming heat transfer based on thermal conductivity with one end at 298K.

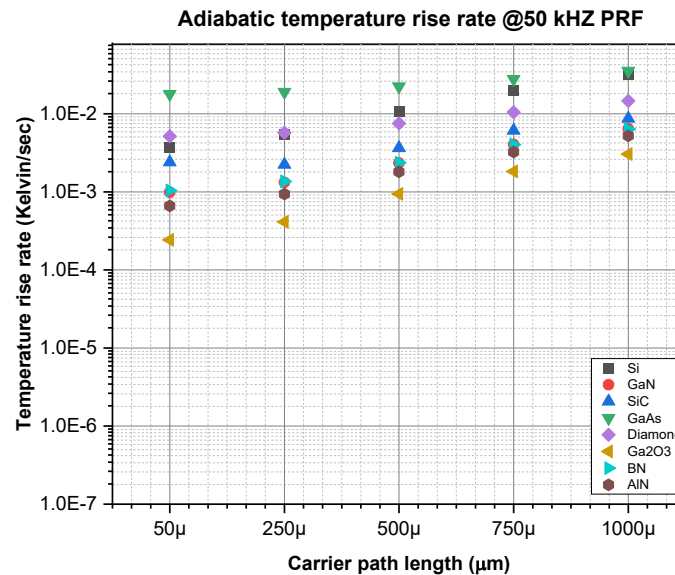


Figure 3.3.4. Temperature rise rate (Kelvin/sec) across materials at 50 kHz PRF and $3E+5$ V/cm field assuming no heat transfer theoretically. Calculations are based on specific heat, density and mass of the material.

In adiabatic heating, the temperature rise rate is in m-kelvin/sec order and is dependent on materials' density and specific heat, which is higher compared with non-adiabatic temperature rise (~ 10 's of μ -kelvin/sec), dependent on thermal conductivity and heat sink.

UWBG materials BN and AlN perform better than diamond during adiabatic heating; but when heat transfer is considered, diamond's thermal conductivity compensates for the losses incurred, resulting in low temperature rise rate for high PRF operation.

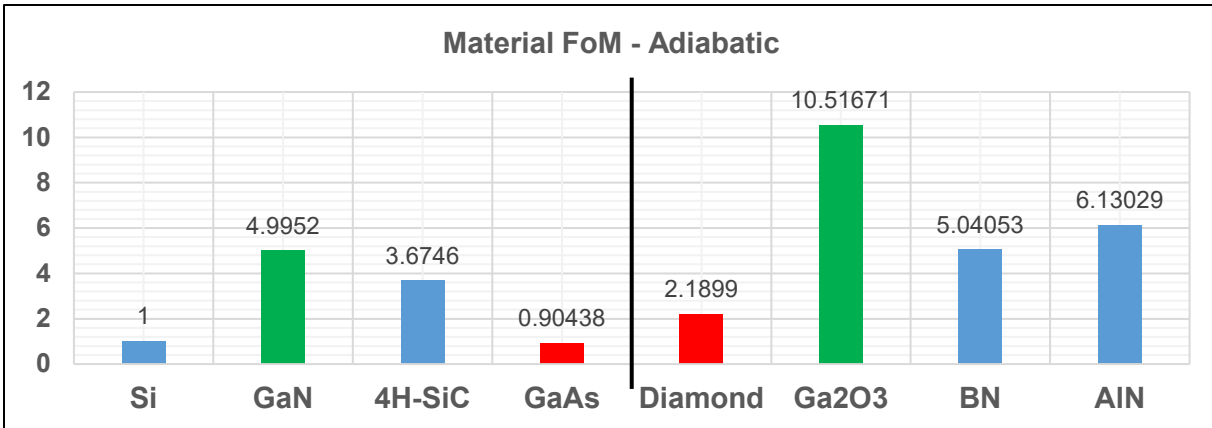


Figure 3.3.5. Calculated HP FoM for wide band gap (WBG) and ultra-wide band gap (UWBG) materials switched at a PRF of 50 kHz and normalized to Si, when no heat transfer is considered.

Figure 3.3.5 shows normalized material FoM when adiabatic heating of materials is considered. Adiabatic implementation of source driver is not ideal for high power applications with higher kHz-MHz repetition rates, but helps evaluate source driver materials for applications requiring packaging that may induce adiabatic processes.

Initial evaluation of uncompensated materials has included semiconducting/semi-insulating materials with narrow bandgap (NBG) to ultra-wide bandgap (UWBG). The calculations performed until now have reflected the superiority of UWBG materials performance in terms of power handling and higher temperature operational capability over the NBG and WBG materials. However, as we transition into switch evaluation at device level involving PCSS and non-PCSS technologies, we are considering Si-, SiC- and GaN-PCSS (planar and lateral devices), Si-DSRDs and AlGaIn/GaN HEMTs considering the limitations in technological maturity to obtain high quality UWBG substrates.

3.3.2 Next Steps for FoM

- i. Complete calculations with constant carrier density across all device lengths to wrap up the uncompensated material evaluation
- ii. Transition to next phase of evaluation involving TCAD and COMSOL simulations for device level evaluation.

4.0 Pulse Charging

4.1 Single Gate Driver HV Switch

4.1.1 Overview

During the months of January and February 2020, we have explored multiple single gate driver-based high-voltage (HV) switch designs. We have selected two reference HV switch architectures from literature to design our own HV switch using four MOSFETs as alternatives to expensive COTs HV switches. We have simulated both topologies in LTSpice to evaluate their static (OFF-state) and dynamic (switching transients) performance. The simulation demonstrated equal voltage distribution among the four switches during the OFF-state and proper gate turn-ON scenario to drive a 50 Ω resistive load in one of the switches. The second switch is designed to drive a capacitive load to emulate the characteristic of a transmission line, which is used in a PCSS based pulse forming network. The second switch has shown 4.2% voltage unbalance during OFF-state at a supply voltage of 3 kV and a switching frequency of 50 kHz. A PCB of the second switch has been designed in Altium and sent out for fabrication.

4.1.2 Parallel-Series HV Switch Design

Designing a fast HV switch with relatively low cost and reduced complexity while achieving voltage balancing during the static and dynamic conditions of the switches has been the key objective in our approach. SiC-based semiconductor switches have a higher electric breakdown field, improved thermal conductivity leading to superior switching frequency compared to Si-based switches. Therefore, SiC MOSFET has been chosen as the building block of an HV switch in our design.

The first type of HV switch consists of two strings connected in parallel for higher current carrying capability, each constructed from four series-connected switches. Buffer capacitors are connected between the common gates of the switch groups, DC bus and the system ground. These capacitors supply the required gate charge to drive the MOSFETs in the HV switch which are not connected to the external gate driver. The capacitors, being connected to the DC bus, are quickly replenished during each switching cycle, thereby mitigating the gate turn-ON delay issue for the four series-connected switches. Minimizing delay between gate drive signals ensures none of the four switches is subjected to overvoltage stress for a period long enough to cause significant damage to any of them. Out of the four MOSFETs, the low side (very bottom) MOSFET is driven by an external gate driver. Once the gate drive signal turns the bottom MOSFET ON, the upper MOSFETs in the string turns ON sequentially following the bottom MOSFET. Large resistors are used to establish static voltage balancing by providing a path for the leakage current to flow to the ground during the OFF state of the switches. Zener diodes are placed between the gate and source terminals of the switches to mitigate unexpected voltage overshoot in the gate drive signals. The schematic of the designed HV switch is shown in Figure 4.1.1.

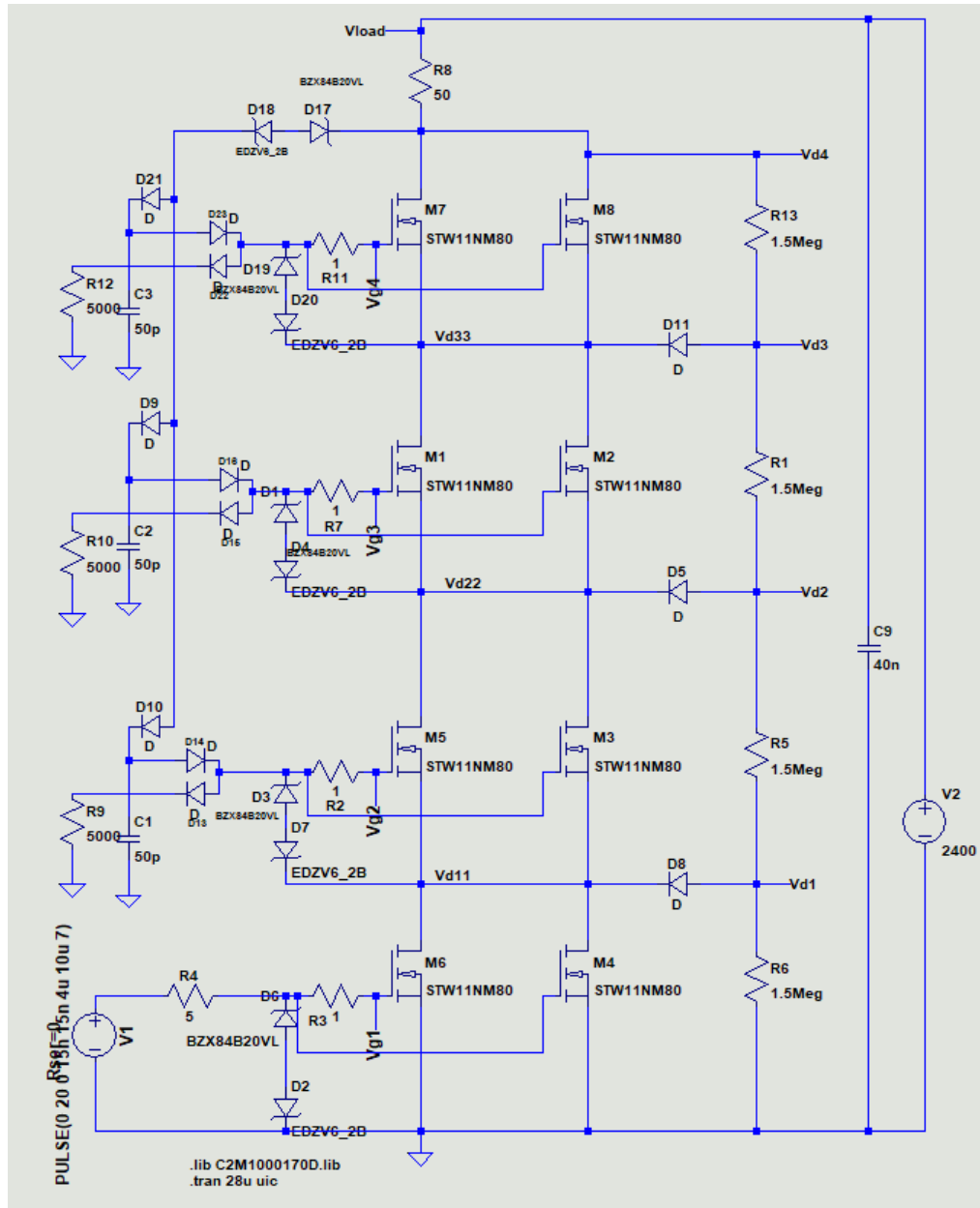


Figure 4.1.1. Schematic of a single gate driver-based 3.2 kV HV switch.

The 3.2-kV rated HV switch has been modeled and simulated at a switching frequency of 50 kHz in LTSpice using the spice model of STW11NM80 MOSFET developed by STMicroelectronics. Simulation has demonstrated 5.1% voltage variation among the series-connected switches during the OFF-state. The simulated OFF-state voltage distribution across the switches' drain and gate, and gate and source are shown in Figure 4.1.2 and Figure 4.1.3, respectively.

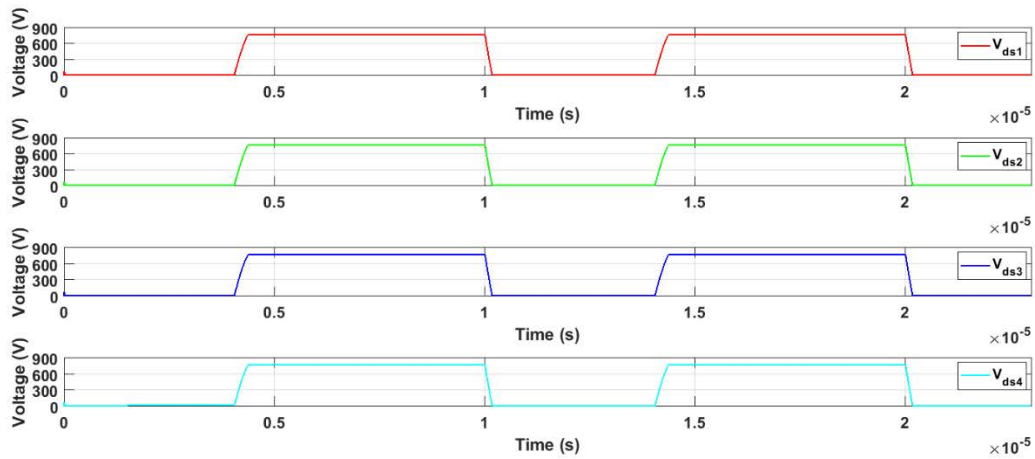


Figure 4.1.2. Simulated voltage across the drain and source terminals of series connected MOSFETs.

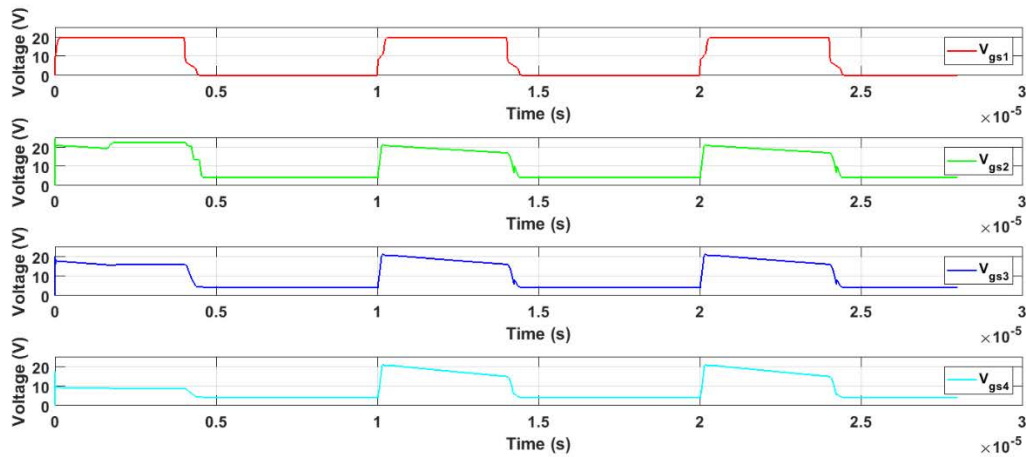


Figure 4.1.3. Simulated voltage across the gate and source terminals of the series connected MOSFETs.

Simulation demonstrates output voltage across a 50 Ω resistive load alternates between 3.17 kV and 60 V following the triggering signal from the gate driver. The simulation result is shown in Figure 4.1.4.

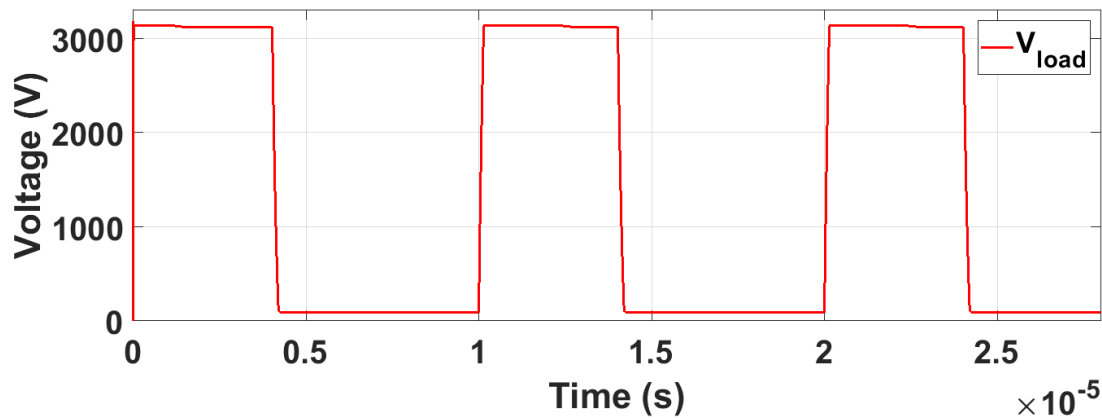


Figure 4.1.4. Simulated voltage across the 50 Ω resistive load.

The topology shown Figure 3.1.1 was eventually not considered for prototyping. The reason is this circuit was not originally designed for SiC switches. Therefore, the gate driving signals reaching to the gate of the MOSFETs would not go to negative values during turn OFF, and this is critical for SiC switches. The next design overcomes this limitation, and the gate signals can reach to -5V when the MOSFETs need to be turned OFF. We need to keep in mind that all circuits found in the literature was done for two MOSFETs in series, and we have extended the designs and modified them to accommodate four MOSFETs so that operating voltage can be increased.

4.1.3 Series-only HV Switch Design

The second HV switch designed by our group is based on a single gate driver with four 1200 V rated SiC switches from CREE (C2M0160120D). However, the gate driving mechanism of the upper switches differs from the first (very bottom) switch. Coupling capacitors that are connected between the gate terminals and ground act as the driving sources of the upper MOSFETs. Snubber capacitors control the slew rate of the drain-source voltages to ensure voltage balancing during switching transients, and balancing resistors control the OFF-state voltage across the four MOSFETs. The switch has been designed specifically to drive a capacitive load (≈ 5 nF) to emulate the characteristic of a transmission line. A bleeding resistor is connected in parallel with the load capacitor to emulate the discharge operation while the PCSS is activated. A low-side switch, connected in series with the resistor, is turned ON for a short time period to discharge the capacitor after the main HV switch is turned off. The schematic of the switch is illustrated in Figure 4.1.5.

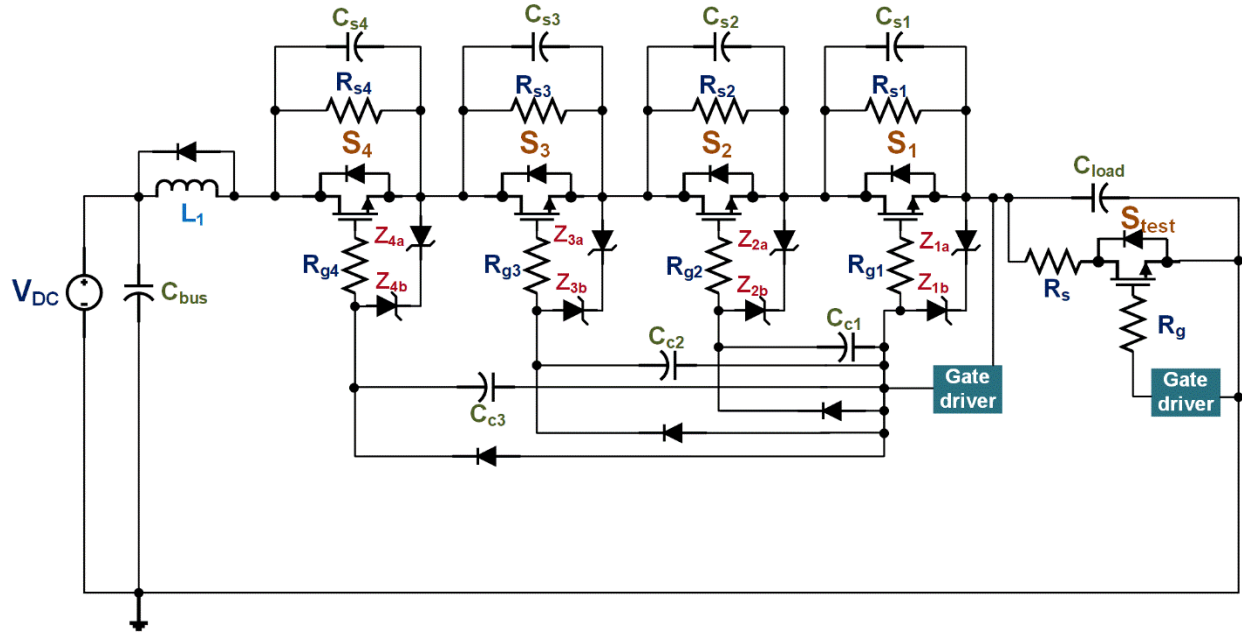


Figure 4.1.5. Schematic of a 4.8-kV HV switch based on single gate driver.

The 4.8 kV rated HV switch has been modeled and simulated at a switching frequency of 50 kHz and a DC supply voltage of 3 kV in LTSpice using the spice model of C2M0160120D developed by CREE. Simulation has demonstrated a maximum voltage unbalance of 4.2% during the OFF-state among the series-connected MOSFETs. The simulated OFF-state voltage distribution across the switches and the output voltage across the capacitive load are shown in Figure 4.1.6 and Figure 4.1.7 respectively.

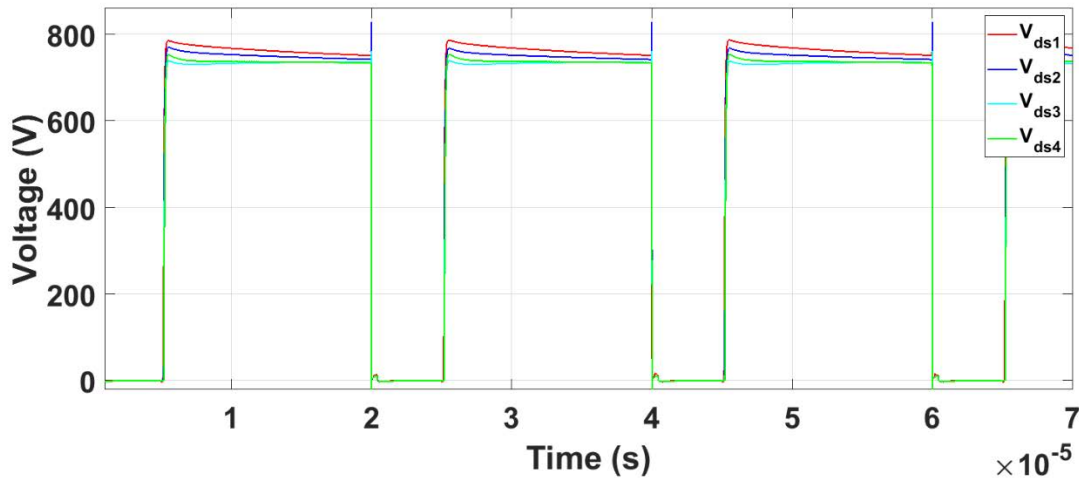


Figure 4.1.6. Simulated voltage across the drain and source terminals of the series connected MOSFETs.

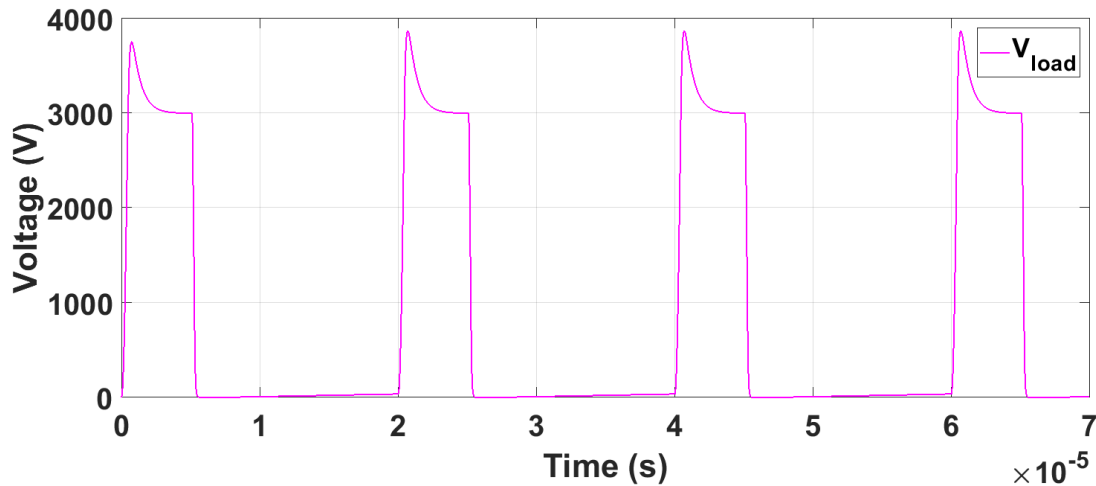


Figure 4.1.7. Simulated voltage across a 5-nF capacitive load.

A prototype of this high-voltage switch has been designed in Altium. The PCB design of the switch is shown in Figure 4.1.8.

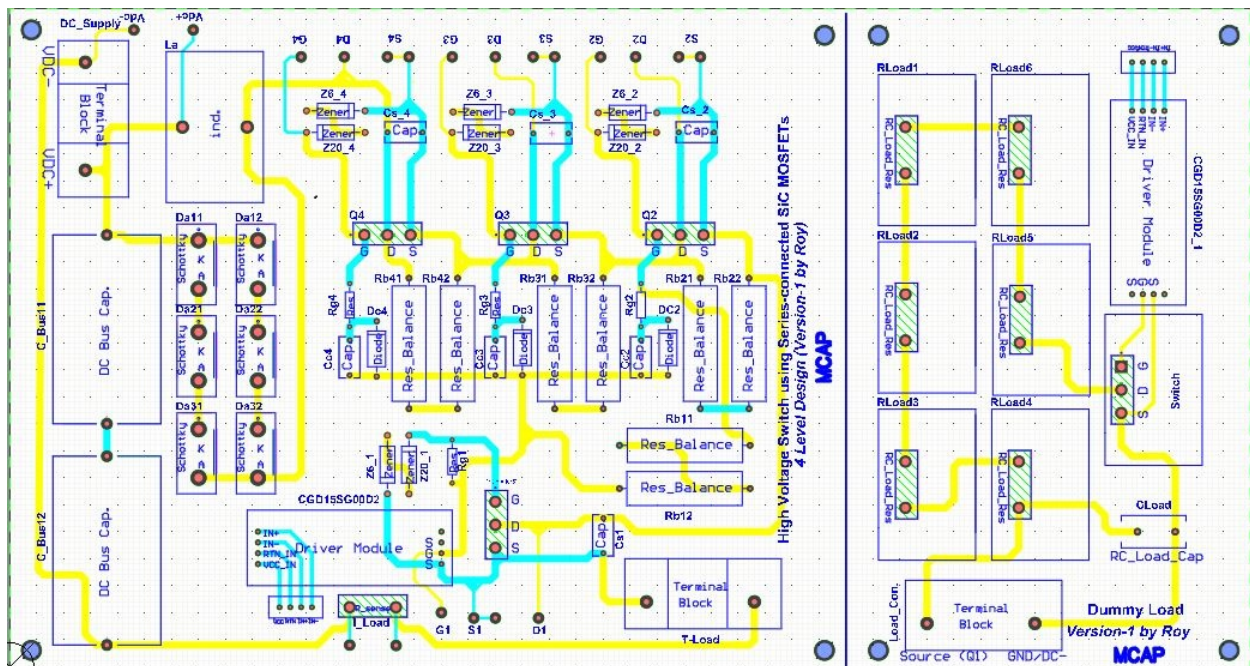


Figure 4.1.8. PCB layout of the series-only HV switch design.

We placed an order for fabrication of the pictured PCB. Upon receipt, we will populate the board and start testing the HV switch PCB to characterize its behavior for different loading conditions at different input voltages and switching frequencies.

We are in the process of procuring a SiC-based 15 kV rated commercial MOSFET switch from BEHLKE. We will conduct testing with a modified power conditioning board fitted with the new SiC switch from BEHLKE.

4.1.4 Next Steps for HV Switch

- i. Conducting experiments with the in-house HV switch
- ii. Designing a new power conditioning board based on the HV SiC switch from BEHLKE.
- iii. Continued Si-BEHLKE power conditioning circuit testing with SiPCSS (with a cooling system) at a lower voltage (<3 kV) and at a higher repetition frequency (>50 kHz).
- iv. Efficient thermal management system with the power conditioning circuit.
- v. Single gate driver-based HV series-connected switch for capacitor charging.

5.0 Antenna

5.1 Helical antennas

5.1.1 Introduction

An Impedance mismatch at the input of an antenna causes a power rejection. We have shown in a previous MSR that a challenge with the helical antenna is that the input impedance of an axial mode element is approximately a function of the circumference (C), and wavelength λ_0 at the design frequency f_0 [1], [2],

$$R \approx 140 \left(\frac{C}{\lambda_0} \right)$$

The input impedance, R , is typically a real value in the range of $100 - 300 \Omega$ depending on the frequency and geometry, which must be accounted for in the design of the antenna to match it to the characteristic impedance of the input port. In this MSR, we discuss the impedance matching theory and present a couple of techniques for matching the impedance to a standard 50Ω .

5.1.2 Impedance Matching

In general, it's always possible to design a matching network for any load, if Z_L has a positive real part. The input impedance of helical antennas can always be made approximately positive and real without a significant imaginary component. The challenge becomes finding the optimal impedance matching network based on several considerations such as complexity of the design, operating bandwidth, the medium of implementation (i.e., waveguide vs. PCB), adjustability to post-fabrication tuning, and time-domain performance [3].

5.1.3 Small Reflections: Single Section

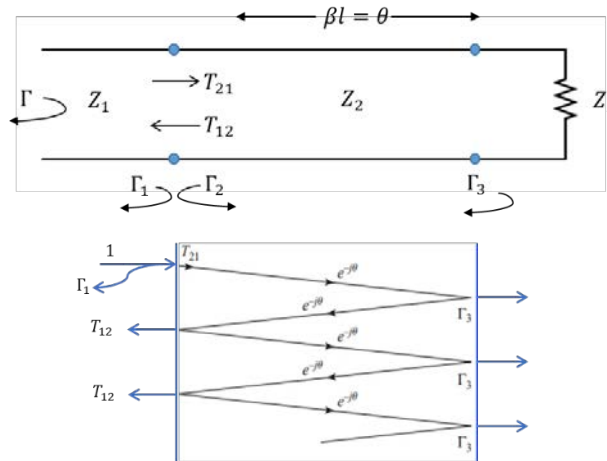


Figure 5.1.1. Partial reflections and transmissions on a single-section matching transformer [3].

The configuration in Figure 5.1.1 shows a transmission line with characteristic impedance Z_1 feeding a small transmission line section with a characteristic impedance Z_2 terminated

with the load to be matched Z_L . The general expression for the reflection coefficient of this configuration can be found by a series summation. Given $\Gamma_1 = \frac{Z_2 - Z_1}{Z_2 + Z_1}$, $\Gamma_2 = -\Gamma_1$, and $\Gamma_3 = \frac{Z_L - Z_2}{Z_L + Z_2}$, meaning $T_{21} = 1 + \Gamma_1 = \frac{2Z_2}{Z_1 + Z_2}$ and $T_{12} = 1 + \Gamma_2 = \frac{2Z_1}{Z_1 + Z_2}$. Extending this to an infinite sum of the small reflections for the single section gives the total reflection Γ as:

$$\Gamma = \Gamma_1 + T_{12}T_{21}\Gamma_3 e^{-2j\theta} \sum_{n=0}^{\infty} \Gamma_2^n \Gamma_3^n e^{-2jn\theta}$$

The above is simplified as a converging geometric series into the form:

$$\Gamma = \Gamma_1 + \frac{T_{12}T_{21}\Gamma_3 e^{-2j\theta}}{1 - \Gamma_2\Gamma_3 e^{-2jn\theta}} = \frac{\Gamma_1 + \Gamma_3 e^{-2j\theta}}{1 - \Gamma_1\Gamma_3 e^{-2jn\theta}}$$

This can be further simplified when the discontinuities between Z_1, Z_2 and Z_2, Z_L are small, so $|\Gamma_1\Gamma_3|$ is then negligible. So, the total reflection can be approximated by the numerator, $\Gamma = \Gamma_1 + \Gamma_3 e^{-2j\theta}$. Another way to say the same is the total reflection is dominated by the first respective reflection of each section's interface.

5.1.4 Small Reflections: Multi-Section

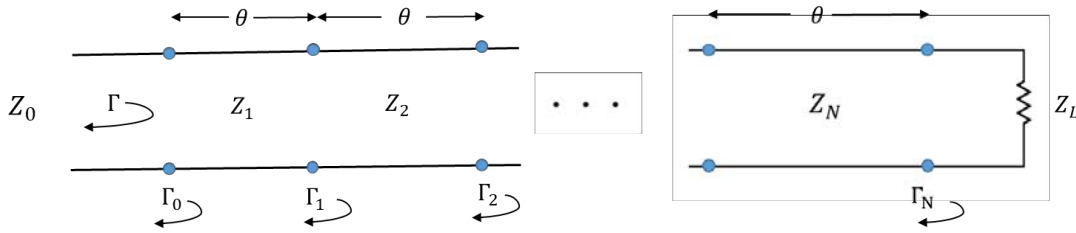


Figure 5.1.2. Partial reflections for a multi-section matching transformer [3].

The small reflections approximation can be extended to a multi-section transmission line with N equal-length sections shown in Figure 5.1.2. The total reflection coefficient by the small reflections of all the sections of a TL with monotonically varying impedance and a real load at the end is:

$$\Gamma(\theta) = \Gamma_0 + \Gamma_1 e^{-2j\theta} + \Gamma_2 e^{-4j\theta} + \dots + \Gamma_N e^{-2jN\theta}$$

Using the equation above, a device with the desired reflection coefficient as a function of frequency (θ) can be synthesized by using enough sections (N). The challenge then becomes, what number of sections, length, and bandwidth are acceptable for the application [3]. Another problem specific to this project is the impact of a matching network on the time domain response of the antenna.

5.1.5 Binomial and Klopfenstein Transformer Design

Table 5.1.1. 2D Binomial and Klopfenstein Transformers' Design Parameter.

PARAMETER	BINOMIAL	KLOPFENSTEIN
f_{min}	800 MHz	800 MHz
R_L	160 Ω	160 Ω
ϵ_r	4.35	4.35
Transformer length	14.86 cm	11.76 cm
Input line Width	4.5 mm	2.9 mm
Output line Width	205.6 μm	130.5 μm
Substrate thickness	2.36 mm	1.5 mm
N	4	10

There are many possible implantations for a multi-section transformer, below are Binomial and Klopfenstein implementations, they were selected as a starting point due to their suitability to the application at hand. Two design validation prototypes for the 1 GHz band with the parameters in Table 5.1.1, were initially designed via Antenna Magus, then optimized in MATLAB and simulated in CST on an FR4 substrate, one of the two devices was measured showing good agreement with the simulation values. The MATLAB code was written with the goal of designing an optimized 3D taper capable of meeting the antenna requirement.

The advantage of the binomial transform is the ability to provide a maximally flat response in over the working bandwidth starting from f_0 , where ripples and impedance variations are minimized. The flat and low Γ response is at the expense of bandwidth and length of transition, for example, the S_{11} the response shown in Figure 5.1.3, based on the parameters in Table 5.1.1, shows that even though the Klopfenstein TL is shorter in length, it's still operational over a much wider bandwidth, with the disadvantage of high ripples and slightly degraded performance. On the other hand, the binomial transformer gives significantly better matching at f_0 compared to the Klopfenstein.

The main advantage of a Klopfenstein transition is the guarantee of an optimal design; that is, for a given length the best match bandwidth is achieved, or for a given bandwidth, the shortest possible length of TL is achieved [4].

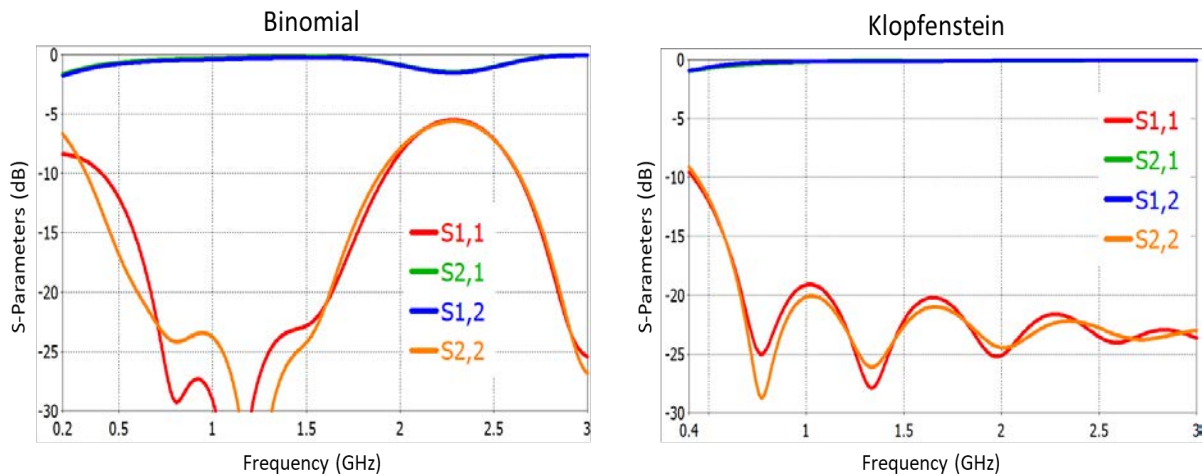


Figure 5.1.3. CST Performance of Klopfenstein and Binomial Transformers (50Ω at Port 1 and 160Ω at Port 2).

For the helical antenna, the binomial transition may suffice in terms of bandwidth and impedance performance. That being said, size and exciting waveform considerations can be limiting, that is, if there are frequency components of the exciting pulse outside of the matching bandwidth, it would be better to use the Klopfenstein transition.

In Table 5.1.1, the output line's width for the two tappers is $205.6\ \mu\text{m}$ and $130.5\ \mu\text{m}$ respectively, due to the FR4 material used and prototype substrate high limitations. A small trace would not be able to handle a significant amount of power. That's why we are improving the design by using a 3D taper instead of the two 2D examples shown above. That would also allow for an additional degree of freedom, which would make it easier to optimize for small size, reasonable input and output terminal size, better bandwidth, and acceptable time-domain performance.

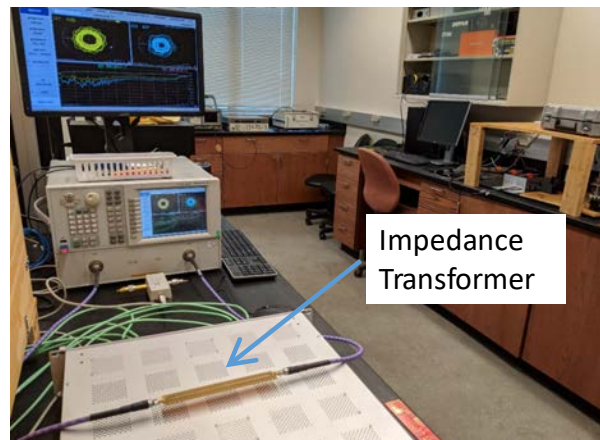


Figure 5.1.4. Impedance transformer measurement setup.

5.1.6 Klopfenstein Measurement

The two designs above were sent to the UMKC fabrication shop, the first of which (Klopfenstein) was fabricated. The binomial will be fabricated in the coming days. VNA

measurements comparing the fabricated impedance tapper with the CST design simulation were conducted. In the simulation, both ports are terminated with $50\ \Omega$ loads, to compare quantitatively with the VNA measurement given that the connectors used, and the VNA test cables are terminated with a characteristic impedance of $50\ \Omega$. A Keysight eCal was used to calibrate the system, the IFBW used was 1 kHz, and the power level was set to 0 dBm. The measurement setup is shown in Figure 5.1.4. Figure 5.2.4.

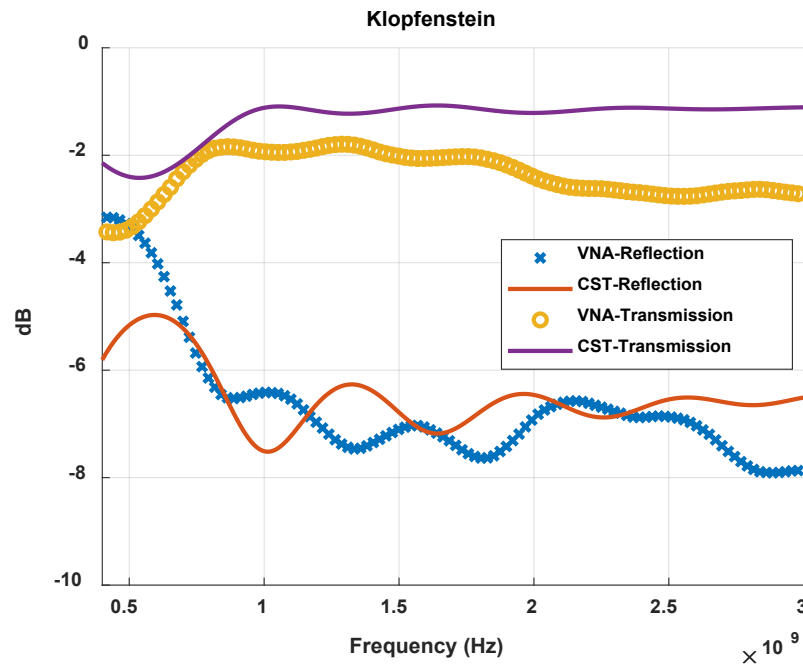


Figure 5.1.5. Simulation and $50\ \Omega$ Reflection and Transmission Measurement.

The reflection and transmission magnitude values are shown in Figure 5.1.5. The measurement is sufficient to show good agreement with the CST simulation as a validation to move forward with more complicated designs. Both the transmitted and reflected signals seem to match the expected simulation values within a few dB. Any slight mismatch can be due to reasonable loss and uncertainties in the form of dielectric uncertainty, soldering, connectors, and fabrication tolerance.

5.1.7 Summary

This month, we designed several impedance tappers and were able to fabricate and test a validation design. We intend to expand by designing a tapper capable of high-power levels, acceptable time domain response, and small footprint. The same theory discussed here can be applied to variation of the feeding of the helical antenna to optimize operation further.

5.1.8 Next Steps

- i. Measure the impedance transformer prototypes and validate against the CST and MATLAB models

- ii. Extend the design shown here to a 3D or 2D continuous taper to decrease footprint, optimize performance, and better integrate with the helical antenna.
- iii. Investigate using the helical antenna's feedline/conductor as an integrated impedance transformer without the need for a separate impedance transition.

References

- [1] H. E. King, J. L. Wong, and E. H. Newman, "Helical Antennas," pp. 1–26.
- [2] J. D. Kraus, "The Helical Antenna," *Proc. IRE*, vol. 37, no. 3, pp. 263–272, 1949.
- [3] D. Pozar, "Microwave Engineering," 2012.
- [4] R. W. Klopfenstein, "A Transmission Line Taper of Improved Design," *Proc. IRE*, vol. 44, no. 1, pp. 31–35, 1956.

5.2 Trade Space Analysis of ESAs

5.2.1 Overview

The primary goal for the antenna array, in the current framework of the project, is to achieve a specific power density S at a certain distance R from the array and to enable steering of short pulses. Power density can be related to the realized gain G^R and the input peak power P_{in} by the well-known equation [1]:

$$S(\theta, \phi) = \frac{1}{4\pi} \frac{P_{in}}{R^2} G^R(\theta, \phi)$$

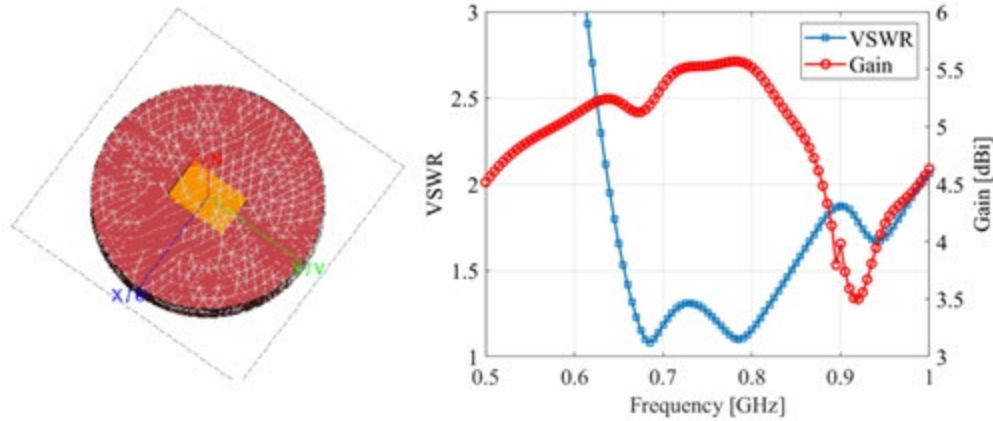
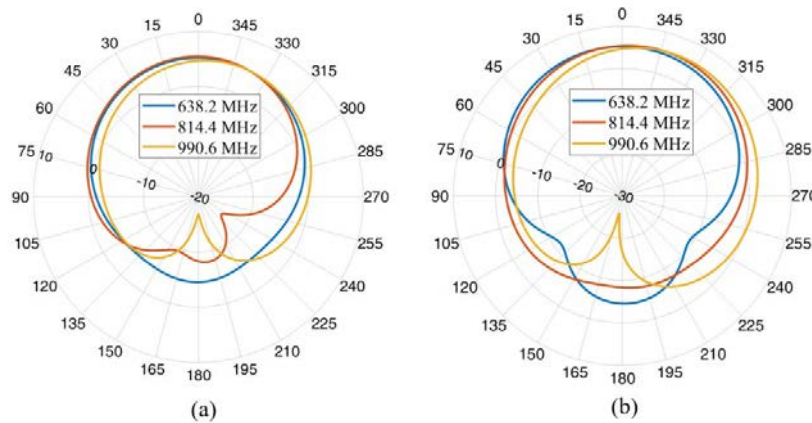
Clearly, the required power density can be achieved either by having a very high gain antenna or by supplying high peak power to the array. However, generating high power is undesirable from economic and environmental point of view. This leads us to the investigation of high gain antenna elements for the array. However, high gain antennas can limit the steer-ability of the array. So, the ultimate goal is to achieve the right gain-steering-balance at reduced element/aperture size. This report focuses on the investigations into a single element design in the UHF region and hence its integration in a planar array to achieve the required power density at the specific distance.

5.2.2 Element Design

An electrically small ($ka = 0.67$) coaxial probe fed rectangular microstrip patch antenna is designed on a 2.67 cm thick TMM-10i ($\epsilon_r = 9.8$, $\tan \delta = 0.0020$) substrate. The antenna achieves 43.27% 2:1 VSWR bandwidth and a maximum gain of 5.67 dBi, when the probe is strategically placed along 2/3rd of the patch diagonal. Table 5.2.1 enlists the optimized antenna parameters. The antenna geometry and its simulated VSWR and gain plots are shown in Figure 5.2.1 and Figure 5.2.2. Full wave simulations have been performed via commercially available EM Solver FEKO.

Table 5.2.1. Optimized Antenna Parameters.

h (mm)	L (mm)	W (mm)	r_p (mm)	(x_p, y_p)	f_L (MHz)	f_C (MHz)	f_U (MHz)	% BW	Max. Gain (dBi)	ka
26.67	41.73	62.59	2	(13.91, 20.86)	638.2	814.4	990.6	43.27	5.67	0.6

**Figure 5.2.1.** Simulated VSWR and Gain of the isolated antenna.**Figure 5.2.2.** Isolated radiation pattern (in dBi) at the edge and center frequencies in (a) XZ plane ($\phi=0^\circ$) and (b) YZ plane ($\phi=90^\circ$).

5.2.3 Trade Space Analysis of the Phased Array

An 11x11 planar array is modeled, in FEKO, using the designed antenna elements, first on an infinite ground plane and later on a finite circular ground plane. In this report, we have presented the case, where all the array elements have been uniformly excited. The radiation patterns of the array on infinite and finite ground planes are shown in Figure 5.2.3 and Figure 5.2.4 respectively.

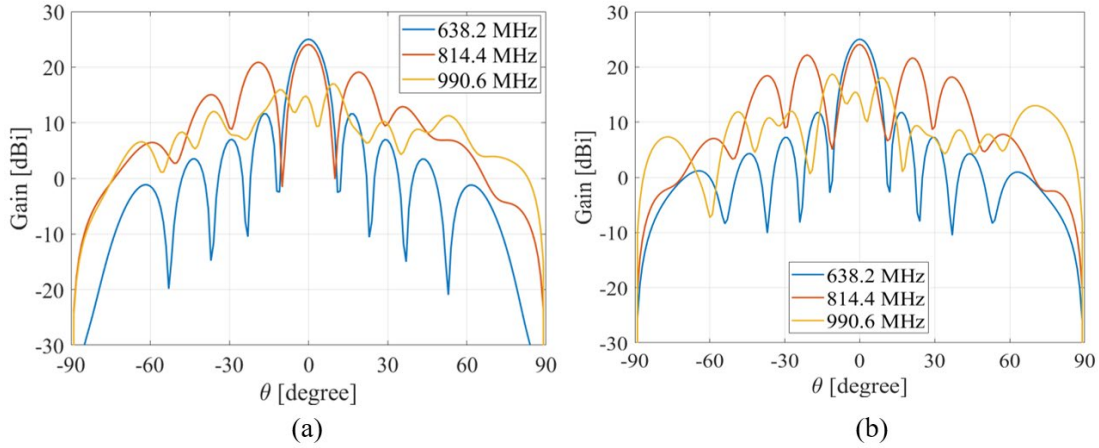


Figure 5.2.3. Radiation pattern (in dBi) of a 11×11 array on infinite ground plane at the edge and center frequencies in (a) XZ plane ($\phi = 0^\circ$) and (b) YZ plane ($\phi = 90^\circ$).

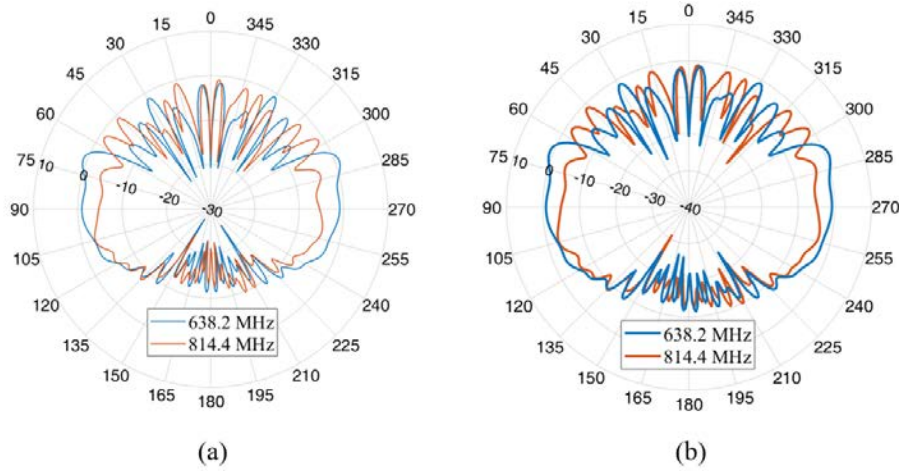


Figure 5.2.4. Radiation pattern (in dBi) of an 11×11 array on finite circular ground plane (radius= $5 \times \lambda_c$) at the edge and center frequencies in (a) XZ plane ($\phi = 0^\circ$) and (b) YZ plane ($\phi = 90^\circ$).

On infinite ground plane, the array has a maximum gain of 25.01 dBi. However, the radiation pattern of the array completely degrades on finite circular ground plane. This can be attributed to the effect of diffraction from the edges of the ground plane.

The investigations have been furthered by studying the characteristics of the embedded elements. To characterize an embedded element in the array, only a specific element is excited, with all other elements being terminated in matched loads. In this case, we have considered two specific elements. These include the center most element in the array (6, 6), which views an infinite array environment around it, and also the one at the edge (1, 1). The radiation pattern for both the cases have been shown in Figure 5.2.6 and Figure 5.2.7. Clearly, the embedded elements patterns on finite ground plane include the effect of edge diffractions, which in turn leads to the degradation of the overall array behavior.

Further investigations into the embedded element characteristics, including their VSWR and gain will be furnished in the future reports.

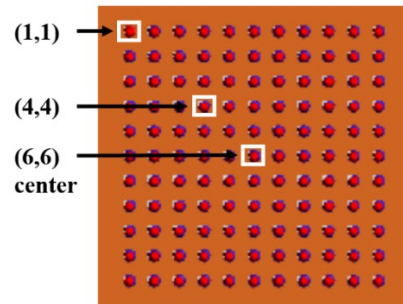


Figure 5.2.5. Geometry of the 11×11 array.

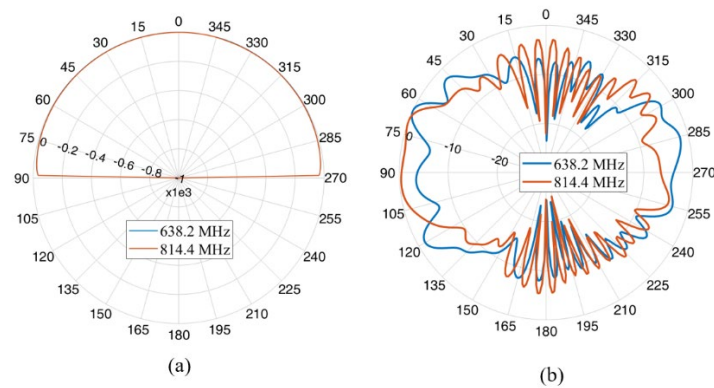


Figure 5.2.6. Normalized embedded element (with only the edge element (1, 1) being excited) pattern in XZ plane ($\phi=0^\circ$) with the 11×11 array on (a) infinite ground plane and (b) finite circular ground plane.

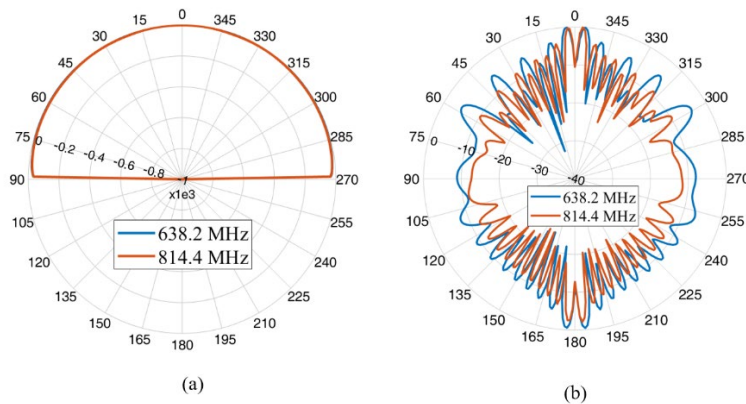


Figure 5.2.7. Normalized embedded element (with only the center element (6, 6) being excited) pattern in XZ plane ($\phi=0^\circ$) with the 11×11 array on (a) infinite ground plane and (b) finite circular ground plane.

5.2.4 Next Steps for ESA trade space

- i. The edge diffractions encountered in the array on finite ground planes can be reduced by changing the ground plane size or by putting absorbers around the array. Methods to reduce edge diffractions will be investigated in the future.
- ii. To obtain a power density of $X \text{ W}/(\text{cm})^2$ at $Y \text{ km}$ from the source (array), the required array gain is $\approx 30 \text{ dBi}$, which can be obtained via $\approx 15 \times 15$ array. Simulation of such large arrays on infinite/finite ground planes is a computational bottleneck even on supercomputers. A method to enhance the computational speed is being looked into currently and will be reported in the upcoming months.
- iii. Characterization of the designed antennas subject to short pulse input signal will be investigated.

References:

[1] R. J. Mailloux, *Phased Array Antenna Handbook*, 3rd ed. Boston, USA: Artech House, 2018.

6.0 Program Management

6.1 Issues

- i. Scope – N/A
- ii. Budget – N/A
- iii. Schedule – N/A

6.2 Interactions with Others

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

6.3 Major Procurement Actions

- i. N/A

6.4 Travel

Destination	Purpose	Attendees	Estimated Costs

6.5 Pending or Upcoming Public Release Requests

- i. N/A

ONR HPM Program - Monthly Status Report (MSR) – March 2020

Anthony Caruso – Univ. Missouri – Kansas City

N00014-17-1-3016 [OSPRES Grant]

ONR Short Pulse Research, Evaluation and non-SWaP Demonstration for C-sUAS

04/15/2020

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2.0 Executive Summary

In March, the OSPRES team continued research and evaluation on short-pulse RF technologies and ancillary systems, high power electrically small antennas, and mechanisms of RF coupling to arbitrary wire configurations: Specific areas of focus included:

- i. modeling and evaluation of Si-PCSS, GaN:C-PCSS, and figures of merit (FOMs) for linear PCSS,
- ii. single-gate driver HV-switch (pulse charger) simulation, development, and validation,
- iii. experimental validation of the jet-impingement thermal management system prototype (JI-TMS),
- iv. genetic-algorithm optimization of helical antenna designs and adding ring elements to microstrip-path ESAs for improved radiation patterns,
- v. deploying equivalent circuit models for reducing computational demands of time-domain RF coupling to arbitrary wire systems.

3.0 Switch Evaluation and Design

3.1 Si-PCSS Modeling and Evaluation

- i. Using TCAD simulations, we investigated current density uniformity and magnitude as a function of Si-PCSS metallization pattern spacing. Results indicated greater uniformity and lower current levels for larger window spacing.
- ii. Simulated R_{on} and recovery time as a function of incident laser energy for 1-mm 2D Si-PCSS devices using adaptive time steps. The results agreed with experimental results.
- iii. Further simulations will increase the device size to 1-cm for further validation.

3.2 GaN:C

- i. PyCDT was used to generate a set of models with C-doped and Fe-doped GaN devices that had different charged defects (-1,0,1).
- ii. We have tested a range of different charge defect states and concentrations. Our preliminary calculations showed a bandgap of 3.06 eV, which is less than expected. If we are not able to improve bandgap calculations in the near future, we will proceed with current models in order to quantify additional trends.
- iii. Continued work on manuscript for cost-analysis of GaN:C manufacturing for HPM and pulsed power applications along with analysis of additional applications for the commercialization of GaN:C and how the historical price trends of SiC can be used to predict the future price trends of GaN:C.

3.3 FOM

- i. We have expanded our focus from uncompensated materials to device level evaluation.
- ii. Began generating 2D device models using COMSOL and Silvaco TCAD for advancing FOM, beginning with linear Si-PCSS
- iii. Temperature dependent physical models for material parameters and carrier dynamics are being used in simulations to study the PCSS' response at different lattice temperatures.

4.0 Pulse Charging

4.1 Single Gate Driver HV Switch

- i. A 4.8 kV rated single gate driver-based high-voltage switch has been designed and fabricated using four SiC MOSFETs.
- ii. The in-house high-voltage switch has been tested up to a DC supply voltage of 600 V at a switching frequency of 50 kHz.
- iii. Further testing needs to be done in April 2020, and other circuit topologies are being investigated in parallel.

5.0 Thermal Management

5.1 Jet Impingement

- i. Simulations have been performed using the computational fluid dynamics software ANSYS Fluent which include refined mass flow rate parameters to model those observed during the first experimental iteration with the jet impingement thermal management system (JI-TMS) prototype.
- ii. The jet impingement TMS was tested using pure silicon fluid PSF-20cSt and the upgraded chiller for increased flow-rate. However, the measured flow-rate (value) was significantly less than the expected 240 gph indicating excessive head-loss in the coolant loop and JI-TMS, resulting in decreased cooling capacity (i.e., higher PCSS temperatures) with respect to calculations.
- iii. We are investigating coolant-loop restrictions and will make modifications to eliminate problematic fittings and connections.

6.0 Antenna

6.1 Helical antennas

- i. This month, we conducted a genetic algorithm optimization on the helical antenna input feed to reduce its input impedance to $50\ \Omega$ for better matching to the $50\ \Omega$ feeding network. The parameters to be optimized included the conductor diameter, feed pitch angle, and feed height. So far, the optimization improved our original design by reducing the impedance of the helical antenna from 168 ohms to 122 ohms.
- ii. The gain was improved from 10.9 dBi to 11.7 dBi in a single helical element while keeping the same element footprint.
- iii. In the next month, we will continue tuning the optimization process to further improve the antenna's frequency domain and time domain response.

6.2 Trade Space Analysis of ESAs

- i. Earlier investigations into a 7×7 array of microstrip patch antennas, on infinite ground plane, at inter-element spacing, $d=0.5\lambda$ (λ calculated at 814.4 MHz), showed excessive pattern degradation at 990.6 MHz (as reported in the MSR for February, 2020).
- ii. Improvement in the radiation pattern, especially at higher frequencies, is achieved by choosing $d=0.485\lambda_1$ (λ_1 calculated at 990.6 MHz) and by addition of a ring of dummy elements around the 7×7 array.
- iii. All full wave simulated results, obtained via FEKO, involve the effect of mutual coupling.
- iv. Mutual coupling analysis between array elements and investigations into beam steering, particularly in case of larger arrays (such as, 11×11 and 15×15), which strongly affect the power density and SWAP requirements, are being looked into.
- v. The majority of the trade space analysis will be completed by the end of summer 2020.

7.0 RF Coupling

7.1 Time-domain Coupling using Equivalent-circuit Models

- i. In March, we applied an equivalent circuit approach to predict RF coupling to practical wiring systems with nonlinear loads.
- ii. The approach is based on replacing the whole structure (frame + wires) with an equivalent circuit and using a SPICE circuit simulator to estimate the time domain response.
- iii. Using this approach, we show that RF coupling to wire systems with nonlinear loads can be calculated accurately in a computationally efficient manner.

8.0 Program Management

8.1 Issues

- i. Scope – N/A
- ii. Budget – N/A
- iii. Schedule – N/A

8.2 Interactions with Others

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

8.3 Major Procurement Actions

- i. N/A

8.4 Travel

Destination	Purpose	Attendees	Estimated Costs

8.5 Pending or Upcoming Public Release Requests

- i. *Berg, Jordan N., and Sarvenaz Sobhansarbandi. "CFD Modeling of Thermal Management Systems for a Silicon Semiconductor Switches". ASME 2020 Summer Heat Transfer Conference. American Society of Mechanical Engineers Digital Collection, In Press – July 12–15, 2020.*

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Anthony Caruso – Univ. Missouri – Kansas City

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ONR Short Pulse Research, Evaluation and non-SWaP Demonstration for C-sUAS

05/15/2020

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2.0 Executive Summary

In April, the OSPRES team: (1) conducted a series of simulations on Si-PCSS, GaN:C-PCSS, and GaN:C atomic structure; (2) continued development of single-gate-driver HV switch and SiC-Behlke based pulsed chargers; (3) tested jet-impingement cooling and compared with experiment; (4) continued trade-space analysis of ESAs and log-periodic antenna; and (5) continued development of RF-coupling time-domain analysis methods using equivalent circuit models.

Specific highlights from the reporting period include:

- i. Developed a preliminary Si- PCSS diode model compatible with SPICE simulations to improve the fidelity of pulse-forming network simulations.
- ii. Began TCAD simulations of SiPCSS specifically aimed at switch figure-of-merit and evaluation beginning with comparative analysis of carrier lifetime methods, which confirmed electron-beam irradiation as the preferred method.
- iii. Developed preliminary models of GaN-PCSS in COMSOL based on examples of LED and photodiodes.
- iv. Reproduced voltage-breakdown of GaN HEMTs.
- v. Demonstrated 2-kV switching at 20 kHz using the custom single-gate driven HV switch topology and designed a PCB for the SiC-Behlke (COTs) HV switch.
- vi. Achieved 160 W of cooling with the jet-impingement thermal management system and identified inlet/outlet geometry issues, which, once resolved, we expect to improve cooling to 200 W or more.
- vii. Incorporated fidelity measurements into antenna analysis that allow for quantitative measures of short-pulse dispersion on antennas in the time domain.
- viii. Simulations of a 15x15 element array of microstrip-patch ESAs demonstrated an average gain of over 26 dB over a 400 MHz bandwidth.

3.0 Switch Evaluation and Design

3.1 Si-PCSS Modeling

- i. Began development of Si-PCSS TCAD device models specifically for incorporation into pulse forming network models in COMSOL.
- ii. Compared TCAD simulations to experiment for R_{on} and recovery time as a function of laser energy using a full-sized device model, adaptive time-steps, and a capacitor discharging topology. Despite minor differences in capacitor voltage, the recovery time results matched when we adjusted the defect concentration accordingly.
- iii. Developed a simple PCSS SPICE model from standard diode models to evaluate against variable resistance models currently used in SPICE modeling efforts. The dependence of switching efficiency and pulse-repetition rate comprise the basis for evaluating this new model, which is still under development.
- iv. Began implementing a user-defined C program of diode equations to develop a custom PCSS SPICE model for implementing laser triggering and photogeneration models into.

3.2 GaN:C

3.2.1 *Testing and Evaluation*

- i. Conducted preliminary COMSOL simulations to model of GaN devices using example models of a GaAs photodiode and an InGaN/AlGaIn LED as a basis for defining device geometry, material properties, and incident optical pulse properties. Results of these simulations will be validated against experimental measurements of GaN:C on-state resistance.
- ii. Continued work the on commercialization analysis of semi-insulating GaN (manuscript in progress) by analyzing price and application through the lens of SiC commercialization to estimate the trajectory of GaN commercialization.
- iii. Experimental testing of devices has been limited during recent months; we plan to resume GaN:C-PCSS pulse testing in May.

3.2.2 *Atomic Structure Calculations*

- i. In the past month we focused our efforts on relaxing all of the atomic structures that were obtained from PyCDT and with the DFT parameters that were determined during the last month. The relaxation process is slow and time consuming because the systems contain charge defects.
- ii. While the DFT relaxations have been running we also spent time studying results we obtained for pure GaN HEMT breakdown voltages from TCAD. We were able to reproduce the breakdowns over a range of GaN thicknesses from approximately 1 to 2.25 microns. However, after that thickness we started observing reduced breakdowns for even thicker sample simulations, which we are continuing to understand and improve upon.

3.3 FOM

We began TCAD simulations to improve the fidelity of our FoM and switch evaluation, beginning with Si-PCSS simulations including:

- i. Temperature dependence of band parameters, carrier dynamics and recombination rates of silicon to evaluate switch's performance as a function of temperature.
- ii. Photo-illuminated recovery of a 2-D, 1000×500-μm Si-PCSS device with a p-i-n doping profile $(p^+, n^-, n^+) = (10^{19}, 10^{12}, 10^{19})$ and a 500-μm optical window to map recovery time as a function of defect/trap-type, -density, -level and –cross sectional area.
- iii. Effects of defect-producing processes such as Au diffusion, Pt diffusion, electron-beam irradiation, proton-bombardment, γ -irradiation and neutron transmutation doping (NTD) on carrier lifetime control. Initial results from point defects from Au diffusion yielded ns-order recovery time reduction with increased defect densities between 10^8 and 10^{13} cm^{-3} .

4.0 Pulse Charging

4.1 Single Gate Driver HV Switch

- i. The in-house high-voltage switch (HVS) has been operated reliably up to a DC supply voltage of 2 kV at a switching frequency of 20 kHz with a dummy PCSS load. This switch is designed to switch up to 4.5 kV; increased voltage will require higher-voltage MOSFETs and/or additional series switches.
- ii. Laboratory tests showed a maximum mismatch of 11.5% among the individual blocking voltages of all four SiC MOSFETs at a supply voltage of 1 kV, thus limiting the maximum charge voltage to (theoretically) ~88.5% of the maximum rated voltage.
- iii. We have received the high voltage SiC-BEHLKE switch and designed a PCB to use it in the charger circuit, and are currently working on an enclosure for use with Si-PCSS testing.

5.0 Thermal Management

5.1 Jet Impingement

- i. We compared simulations performed in ANSYS™ to experimental data with the jet-impingement thermal management system (JI-TMS). Both agreed within ± 2 °C when heated at 100–200 W with flow-rate of a 1 gpm.
 - a. At 100 W, the maximum temperature was 54 °C and 58 °C in the simulation and experiment, respectively.
 - b. At 200 W, the maximum temperature was 97 °C and 99 °C in the simulation and experiment, respectively.
- ii. Simulations indicated that the staggered jet array geometry limits cooling to ~160 W when using low-viscosity silicon dielectric fluid (PSF-20cSt) at 1 gpm and maintaining PCSS temperature below 80 °C. While we have continued simulations of DI-water and water-glycol mixtures that demonstrate improved flow-rate, the high-voltage environment may preclude their use.
- iii. We determined inlet/outlet of the JI-TMS were causing unnecessary flow-restrictions leading to excessive pressure to achieve sufficient flow rate. We are revisiting the design to eliminate these restrictions.

6.0 Antenna

6.1 Log-periodic (Helical) Antennas

- i. This month, we investigated the Fidelity Factor (FF) as one metric to quantify the time-domain performance of the helical antenna. The FF can be calculated as follows:

$$FF = \max_t \frac{\left| \int_{-\infty}^{\infty} \mathbf{e}_{Tx}(t' + t, \mathbf{r}) u_{Tx}(t') dt' \right|}{\sqrt{\int_{-\infty}^{\infty} |\mathbf{e}_{Tx}(t', \mathbf{r})|^2 dt'} \sqrt{\int_{-\infty}^{\infty} |u_{Tx}(t')|^2 dt'}}$$

where $\mathbf{e}_{Tx}$ is the radiated electric field by the antenna, u_{Tx} is the input voltage pulse to the antenna. Therefore, the FF measures the capability of the antenna to preserve the shape of the input pulse. If the shape of the radiated electric field pulse matches exactly the shape of the input voltage, the FF factor will be unity, whereas if the antenna severely distorts the shape of the pulse the FF will drop to zero. Hence, the FF provides a quantitative metric to quantify the time-domain response of the antenna.

- ii. Over this month, we showed that increasing the number of turns of the helical antenna increases the gain of the antenna, and therefore increases the amount of the radiated power. At the same time, increasing the number of turns leads to lower values of the FF, and hence more distortion to the shape of the radiated pulse. In the next month, we will continue with the multi-objective optimization to investigate enhancing both the gain and FF of the antenna at the same time.

6.2 Trade Space Analysis of ESAs

- i. A 15x15 planar array of microstrip patch antennas on infinite ground plane, with inter-element spacing $d=0.485\lambda_1$ (λ_1 calculated at 990.6 MHz) and with a ring of dummy elements, is modeled and investigated via full-wave simulations in FEKO.
- ii. An average gain of ≈ 26 dBi is obtained over the frequency range 0.6 -1 GHz.
- iii. As the main beam is scanned from the zenith to $\pm 60^\circ$, the array gain is found to drop at some of the scan angles, especially at higher frequencies (> 850 MHz). The dips are more prominently observed in the E-plane ($\phi=0^\circ$).
- iv. This can be due to the excessive surface wave loss which eventually leads to scan blindness in the array.
- v. A method to predict the blind scan angles has been looked up in the literature and is being implemented to address scan blindness reduction/compensation issues.

7.0 RF Coupling

7.1 Time-domain Coupling using Equivalent-circuit Models

- i. Developed equivalent circuits to represent full-wave RF coupling to wire systems with nonlinear loads. This approach replaces the wire system with an open circuit voltage and an input impedance represented by a series of RLC sub-circuits. This equivalent circuit approach is computationally more efficient, and it allows us to explain the variations in the response with different excitation conditions.
- ii. Continued our investigation of developing equivalent circuits for wire systems placed at different locations. We showed that the equivalent circuit could change considerably, based on the location of the wire system relevant to the frame. Based on the changes in the equivalent circuit, we can explain the variations in the coupling with respect to pulses of different properties.
- iii. In our future work, we will show how the equivalent circuit approach will facilitate the training of machine learning networks to predict the variations in coupling with the relative location of the wire system.

8.0 Program Management

8.1 Issues

- i. Scope – N/A
- ii. Budget – N/A
- iii. Schedule – N/A

8.2 Interactions with Others

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

8.3 Major Procurement Actions

- i. N/A

8.4 Travel

Destination	Purpose	Attendees	Estimated Costs

8.5 Pending or Upcoming Public Release Requests

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Anthony Caruso – Univ. Missouri – Kansas City

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2.0 Executive Summary

In May, the OSPRES team: (1) continued implementation and validation of Si-PCSS diode models for improving simulations of PCSS-PFN configurations in SPICE, (2) conducted a comparative study of the methods for introducing defects into Si-PCSS, (3) conducted ab-initio calculations of polarization, lattice constants, and band-gap of GaN, GaN:C, and GaN:Fe using multiple density functional theory models, (4) began testing new GaN:C-PCSS devices from NRL, (5) tested both the in-house and the SiC-Belkhe HV switch topologies at a repetition rate of 50-kHz, (6) modified the inlet and outlet jet impingement thermal management system to decrease the head loss and maximize flow rate, (7) continued investigation of ESA trade space analysis.

Specific highlights from the reporting period include:

- i. Implemented improved and reverse current and recovery physics in Si-PCSS SPICE model.
- ii. Proton-irradiation simulations exhibited five times less recovery time in Si-PCSS devices at the cost of a hundred-fold increase in leakage current.
- iii. Demonstrated 2-kV charging at 50-kHz using the SiC-Belkhe topology, and 600-V charging at 50-kHz using the single-gate driver HV switch topology.

3.0 Switch Evaluation and Design

3.1 Si-PCSS Modeling

- i. SPICE Si-PCSS models allow for less computationally expensive simulations when compared to TCAD simulations for repetitive pulse generation in TL-based PFN. The Spice models can be connected to experimental measurements and TCAD models through spice modeling equations and their device parameters which allows models indicative of a custom, non-COTS device.
- ii. Continuing improvement of Si-PCSS SPICE models, we implemented a PN diode model for improved forward and reverse current physics, and another for recovery time. The recovery time model (refer to April 2020 MSR) was modified to work with adaptive time-steps in SPICE simulation. The model included current contributions to the Shockley ideal diode equation from non-ideal forward- and reverse-biased trap-assisted tunneling, Shockley-Read-Hall generation, band-to-band tunneling, and avalanche multiplication.
- iii. We validated the diode models via IV and electrical reverse recovery simulations, which demonstrated the characteristic curves expected for default device parameters. Then a diode-based spice Si-PCSS model was implemented from the recovery time diode model.
- iv. To move towards quantifying the improvements of the new models, we compared simulations of Si-PCSS on a simple PFTL ($Z_{\text{chg}} = Z_{\text{load}}$) using the (old) time-varying resistance model (for the PCSS on-state resistance) and the improved diode-based SPICE model. Pulse trains of a few pulses generated had similar pulse height and width and distortions due to data points selected using adaptive time-stepping.

3.1.1 Next Steps

- i. To quantify improvements of the SPICE Si-PCSS model, longer pulse trains and comparable model parameters will next be used. Efficiency vs pulse repetition rate and recovery time as well as added computational cost will be used to assess improvements between time-varying resistance and diode-based SPICE Si-PCSS models.

3.2 GaN:C

3.2.1 GaN:C PCSS Test and Analysis

- i. Preliminary simulations in COMSOL to model GaN:C PCSS devices continued using Si solar panel and bipolar transistor models as reference models. Challenges arose in modeling GaN:C devices due to not having the wave optics COMSOL package that allows for coupling of light sources to devices to model how changes in incident energy and wavelengths affect device on-state resistance and efficiency. Instead, user-defined carrier generation is required, leading to a model (in the Si solar cell example) reliant on approximating photo-carrier generation using Shockley-Read-Hall recombination combined with the Beer-Lambert Law which does not account for certain parameters of incident light such

as scattering and reflection in the material or how material defects affect optical absorption. Other generation methods included in the RF package will be explored as possible ways to more accurately model photo-carrier generation in GaN:C PCSS.

- ii. Testing begins this week to characterize the new NRL devices with differing carbon concentrations (8×10^{16} and $5 \times 10^{17} \text{ cm}^{-3}$) and gap lengths (1000 and 50 μm). Initial pulsing will be done at low voltages (up to low ones-of-hundreds of volts) to explore the relationship that carbon concentration has on the on-state resistance and efficiency (P_{in} vs P_{out} , where P_{in} includes electric power along with incident optical power) at as a function of laser energy and voltage. High voltage pulse testing will be done after all low voltage is completed due to the small number of devices available.

3.2.2 GaN:C Atomic Structure Calculations

- i. In the past month the main activity was to set up and perform ab initio calculations to obtain the polarization constants of GaN, carbon compensated GaN (GaN:C), and iron doped GaN (GaN:Fe). The values obtained are in the table below.

Parameter	GaN	GaN-C	GaN-Fe
Lattice constant (supercell)	9.649	9.65	9.669
Elastic constant (C_{33} , C_{13}) GPa	346.623 , 73.369	338.426 , 75.681	321.574 , 62.212

- ii. The next activity that was performed in the past month was to set up and perform ab initio calculations of the relaxed lattice constants and elastic constants of GaN, GaN:C, and GaN:Fe. The values obtained are in the table below.

Parameter	GaN	GaN-C	GaN-Fe
Piezoelectric constant (z) (e_{33})	0.88	0.47	- 0.517
Piezoelectric constant (x,z) (e_{31})	-0.27	-0.25	- 0.052

- iii. The final activity was to test a series of calculations on GaN, GaN:C, and GaN:Fe to compute the band gap using different density functions. The key goal is to obtain a band gap close to experimental values of ~ 3.4 . Initial results looking only at the density of states were misleading and showed some reasonable agreement. Later results (using various GGA and hybrid DFT functions) of the band structure showed consistently smaller gaps on the order of slightly more than 2 eV. We are continuing to investigate the causes of the discrepancy.

3.3 FOM

- i. We continued the efforts on switch evaluation using TCAD simulations for $1000 \times 500\text{-}\mu\text{m}$ Si-PCSS devices with a doping profile of $[(p^+, n^-, n^+) = (10^{19}\text{-}, 10^{12}\text{-}, 10^{19}\text{-cm}^{-3})]$, illuminated by a trapezoidal 1064 nm optical pulse with 120 ps FWHM, and an intensity of $4 \times 10^9 \text{ Watts/cm}^2$, with 48 μJ incident energy. Photo-illuminated recovery is being studied as a function of defect level, density and cross section to compare effectiveness of electron-beam irradiation, proton-irradiation, and a combination thereof.

- ii. Electron-beam irradiation: We conducted a study of the dominant acceptor defect level ($E = E_C - 0.4$ eV), due to electron-beam irradiation [1], distributed uniformly throughout the device as a function of defect density. The recovery time at defect densities of 10^{13} cm^{-3} (~ 35 ns) was reduced to ~ 5 ns at $5 \times 10^{17} \text{ cm}^{-3}$. However, this reduced recovery time comes with an increase in leakage current with increasing defects, which we are continuing to quantify.
- iii. Proton-irradiation: Two dominant acceptor defect levels exist at $E_t = 0.165$ eV and 0.42 eV [2], corresponding to high- and low-injection levels, respectively, with two orders of magnitude larger capture cross sections when compared with electron-beam irradiation. We studied the proton-irradiation defects as a function of defect density and non-uniform, local distributions within the device. Simulations so far have shown 5x reduced recovery time at defect densities of 10^{16} cm^{-3} , but also increased leakage current by two orders of magnitude higher when compared with electron beam irradiation at similar densities.

3.3.1 Next Steps

- i. A relationship will be quantified for recovery time vs. defect concentration for electron-beam irradiation and recovery time vs. proton-irradiation induced defect relationship in terms of defect concentration and region.
- ii. A combination of electron and proton irradiation induced defects (both acceptor- and donor-type) will be simulated to explore leakage current mitigation at higher defect densities by implementing clusters of localized defects across the device at varied densities.

3.3.2 References

- [1] Eywaraye, A. O., and B. J. Baliga. 1977. "The Dominant Recombination Centers in Electron-Irradiated Semiconductors Devices." *Journal of Electrochemical Society* 124 (6).
- [2] Anders Hallén, Niclas Keskitalo, Ferenc Masszi, and Viliam Nág, 1996. "Lifetime in Proton Irradiated Silicon." *Journal of Applied Physics* 79, 3906. <https://doi.org/https://doi.org/10.1063/1.361816>.

4.0 Pulse Charging

4.1 Single Gate Driver HV Switch

- i. We have conducted experiments using the in-house high-voltage switch coupled with the PCSS at a repetition frequency of 50 kHz and a positive supply voltage up to 600 V obtaining similar results compared to a commercial Si-based BEHLKE switch.
- ii. We have tested a modified in-house high-voltage switch coupled with the PCSS at a repetition frequency of 50 kHz and a negative supply voltage up to 120 V as a proof of concept obtaining similar results compared to a commercial Si-based BEHLKE switch.
- iii. We have designed a power conditioner circuit using a SiC-based BEHLKE switch with 15 kV rating and operated the circuit with the PCSS in continuous mode at a repetition frequency of 50 kHz and up to a negative DC supply of 1 kV as initial test.

5.0 Thermal Management

5.1 Jet Impingement (JI-TMS)

- i. Continuing on last month's progress, we are investigating techniques to further reduce the pressure drop across the JI-TMS's inlet and outlet while maintaining the Si-PCSS temperature at or below 80 °C while inducing 200 W average heat dissipation. Simulations indicate that the current staggered jet array geometry limits cooling to ~160 W when using low-viscosity silicone fluid (PSF-20cSt) flowing at 1 gpm and a loop pressure below 80 psi.
- ii. Techniques which could achieve these constraints are being evaluated using ANSYS™ CFD simulations. Their results are summarized as follows:
 - a. Serving as the baseline simulation- when flowing PSF-20cSt at 1 gpm, the maximum Si-PCSS temperature and pressure drop achieved is 98.8 °C and 44.6 psi, respectively. This indicates that a greater flow rate is required to reduce the temperature towards the allowable operating temperature regime.
 - b. Increasing the PSF-20cSt flow rate to 2 gpm results in a decrease of the maximum Si-PCSS temperature achieved to 80 °C, thus meeting the operating temperature performance constraint. In doing so however, this resulted in an undesirable increase in pressure to 171.2 psi. This indicates that to achieve greater flow rates we need to minimize the pressure drop within the JI-TMS while not exceeding the maximum allowable temperature.
 - c. Therefore we increased the number of fluid outlets to two, instead of one, where the second outlet was positioned half way between the original inlet and outlet. Now with two outlets and flowing PSF-20cSt at 1 gpm into the singular inlet, the maximum temperature the SiPCSS achieved was 97.39 °C with a TMS pressure drop of 10.85 psi. Thus the pressure drop was improved from the single outlet case (a), however the PCSS still exceeded the maximum allowable temperature.
 - d. To mitigate this we next simulated the two-outlet condition with an increased flow rate of 3 gpm. As a result, the maximum SiPCSS temperature achieved was 83 °C with a pressure drop of 74.35 psi across the TMS inlet and outlets.
- iii. The resulting simulations performed indicates that just over 3 gpm of flow rate is necessary to achieve the desired PCSS operating conditions when inducing 200 W of average power input. This is well within the operating limitations of the upgraded pump and chiller thus revisions to the JI-TMS's CAD model will be made to ensure manufacturability of the design simulated and followed by experimental testing thereafter.

6.0 Antenna

6.1 Log-periodic (Helical) Antennas

- i. This month, we investigated the multi-physics electromagnetics-thermal analysis of antennas to predict their maximum power handling capabilities for continuous and pulsed wave excitations.

6.2 Trade Space Analysis of ESAs

- i. Earlier investigations have shown that an array of microstrip patch antennas designed on a finite ground plane results in excessive pattern degradation/spoilage. This is apparently due to excessive surface wave excitation, a commonly observed phenomenon in microstrip antennas.
- ii. Methods to suppress surface waves, which would in turn reduce edge diffraction and mutual coupling, are being looked into in detail. Detection of blind scan angles by locating surface wave poles is being done using the method suggested in [1]. Further investigations into the trade space analysis of the array will be meaningful once the surface wave suppression issues are resolved.
- iii. Performance validation of a small 4x4 array via fabrication and measurement will be carried out by the end of summer, 2020.
- iv. On a separate study, a MATLAB code has been developed for the rapid estimation of the radiation pattern of a linear array. The code will be extended for a planar array in the near future.

Reference:

- [1] K. C. Durbhakula, D. Chatterjee and A. M. Hassan, "An Efficient Algorithm for Locating TE and TM Poles for a Class of Multiscale Inhomogeneous Media Problems," *IEEE Journal on Multiscale and Multiphysics Computational Techniques*, Vol. 4, pp. 364-373, 2019.

7.0 RF Coupling

7.1 Time-domain Coupling using Equivalent-circuit Models

- i. Recently, we started to develop equivalent circuits to represent full-wave RF coupling to wire systems with nonlinear loads. This approach replaces the wire system with an open circuit voltage and an input impedance represented by a series of RLC sub-circuits. Over the last month, we explored how the equivalent circuit varied with the environment of the wire system. That is a wire system was placed inside a perforated cavity or a cavity with a slot and the changes in the RF coupling equivalent circuit were quantified.

8.0 Program Management

8.1 Issues

- i. Scope – N/A
- ii. Budget – N/A
- iii. Schedule – N/A

8.2 Interactions with Others

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

8.3 Major Procurement Actions

- i. N/A

8.4 Travel

Destination	Purpose	Attendees	Estimated Costs

8.5 Pending or Upcoming Public Release Requests

- i. *Berg, Jordan N., and Sarvenaz Sobhansarbandi. "CFD Modeling of Thermal Management Systems for a Silicon Semiconductor Switches". ASME 2020 Summer Heat Transfer Conference. American Society of Mechanical Engineers Digital Collection, In Press – July 12–15, 2020.*

ONR HPM Program - Monthly Status Report (MSR) – June 2020

Anthony Caruso – Univ. Missouri – Kansas City

N00014-17-1-3016 [OSPRES Grant]

ONR Short Pulse Research, Evaluation and non-SWaP Demonstration for C-sUAS

07/15/2020

Distribution Statement ~A: Distribution authorized to all OSPRES Grant and Contract Participants, and any uncompensated collaborators. Further dissemination per Ryan Hoffman (ryan.hoffman@navy.mil).

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2.0 Executive Summary

In June the OSPRES team: (1) conducted simulations aimed at optimizing the Si-PCSS contact geometry to minimize laser requirements, (2), conducted studies on non-uniform defect profiles created via proton irradiation in Si-PCSS, (3) began testing new GaN:C PCSS devices from NRL, (4) setup and applied methods for improving *ab initio* band-gap calculations for GaN, (5) redesigned the single-gate-driver HV switch using surface mount devices for increased voltage and rise time, (6) conducted a preliminary analysis on the use of nanoparticle fluids for improving the low viscosity silicone oil (PSF-20cst) used in the jet-impingement thermal management system, (7) validated the causes of scan-blindness in microstrip-path ESA arrays and began time-domain characterization, and (8) studied the rf coupling dynamics of a wire loop within a metal cavity. Specific highlights from those efforts include:

- 1) Simulations indicate that further increasing the spacing of the metal filament pattern on the Si-PCSS optical window may decrease the laser energy required to achieve the same R_{on} by a factor of 1.5-2.0.
- 2) The use of proton irradiation to produce localized (non-uniform) defect densities has the potential to achieve the same recovery time while reducing leakage current by up to two orders of magnitude.
- 3) Photoconductivity measurements on GaN:C PCSS demonstrated peak photoresponse at 365 nm, and decreased rapidly beyond 375 nm. Thus, the 408-nm laser diodes will be insufficient for GaN:C PCSS activation.
- 4) Improved the *ab initio* modeled bandgap to 2.97 eV, however this is still well below the experimentally measured bandgap of 3.5 eV.
- 5) Designed (ready for fabrication) a new single-gate-driver HV-switch for 6.8 kV operation.
- 6) Calculations of aluminum oxide and silicon carbide nanoparticle additives indicate a potential increase of 17% and 22% in heat-removal rate, respectively.
- 7) Microstrip-patch antenna simulation show scan blindness at $\approx 10^\circ$ is due to Surface Wave (SW) excitations at ≈ 990 MHz.
- 8) Modeling indicated that placing a wire within a metal cavity creates additional resonance in the wire's input impedance which results in additional coupling paths to the loop.

3.0 Switch Evaluation and Design

3.1 Si-PCSS Modeling

The Si-PCSS modeling team conducted simulations to optimize the contact patterns of Si-PCSS devices to minimize R_{on} by tailoring the metal filament spacing and geometry to the peak doping concentration. These results, although still early, indicate that by further increasing the spacing of the contacts, it may be possible to reduce the R_{on} by a factor of 1.5–2.0, corresponding to roughly halving the required laser energy to achieve the same R_{on} . The team is finalizing a selection of new metal filament spacing's to be fabricated and characterized. The literature review and modeling approach to these findings is detailed in the following two sections:

3.1.1 Top Contact Equation-based Modeling from Literature (Serreze and Heizer)

The basic design of the two (regular grid and fishbone) Si-PCSS top contact patterns in use can be found in Serreze [1]. Only the increased metal resistance (from decreased metal) vs decreased shadowing loss (from increased coverage of metallization area) is considered in Serreze. From direction of current flow into and out of the top contact, metal resistive losses are calculated in python modeling and implemented as a resistor network in SPICE. This method can be used to determine an optimal width of the metallization. The optimal spacing or window width is dependent on the losses in the silicon top layer where most of the contact resistance losses occur. The optimal width calculated was 0.2 μm compared to the 10 μm metal line width used. This is in agreement with optimal width usually being less than metallization line width limits set by manufacturability.

Using python modeling, the relevant literature resistive losses due to the top layer of the semiconductor were added to the metal resistivity. This follows the method of Heizer [2]. The Heizer method can also be used in SPICE simulation using an equivalent circuit. Inclusion of the top layer losses along with the metal losses allows an optimal window spacing (S) to be considered. Nine different device parameters (height, metal line width, metal resistivity, top-layer sheet resistivity, laser power, available photocurrent or quantum efficiency, diode ideality, diode reverse saturation current, and diode output voltage) are used to determine a relative efficiency and an optimal S . Results for an optimal S using the Heizer method are still inconclusive.

3.1.2 Top Contact TCAD Modeling

Several attempts were made to use the Serreze-style resistor network with TCAD. Modeling of the resistor network as metal regions and/or electrodes was also attempted. TCAD is too costly to model an entire top contact and this is usually done in some SPICE implementation [3] but sections of the top contact can be modeled in TCAD.

Other 3D TCAD simulations for different window spacing were performed. The device model used separate electrodes and electrode resistances for (3 or 5) metal filaments with (50 μm and 20 μm) spacing, respectively. The device was simulated in a discharging capacitor circuit for a single output pulse for which on-state resistance, recovery time, switching losses (for the turn-on, on-state, turn-off and off-state) and efficiency were

calculated. Increased filament spacing decreased R_{on} and increased recovery time, switching losses, and efficiency.

Next Steps:

- i. TCAD simulations to determine an optimal window spacing S with reduced laser energy at fixed on-state resistance will be performed. As S is increased the resistive losses in the top-layer will begin to offset reduction in on-state resistance from increased available photocurrent.
- ii. The Heizer method will continue to be studied using an equivalent circuit (or lumped) model of the Si-PCSS for determination of on-state resistance to compare to TCAD results.
- iii. Free-carrier absorption will be included in TCAD simulations. The spreading resistance profile of the p-doped region can be used to calculate top-layer sheet resistivity and free-carrier absorption. Adding free-carrier absorption improves the modeling for required laser energy.

References:

- [1] Serreze, H. B. "Optimizing solar cell performance by simultaneous consideration of grid pattern design and interconnect configuration." *13th IEEE PVSC*, 1978 609 (1978).
- [2] Heizer, K. W., and T. L. Chu. "Solar cell conducting grid structure." *Solid-State Electronics* 19, no. 6 (1976): 471-472.
- [3] Wong, Johnson. "Griddler: Intelligent computer aided design of complex solar cell metallization patterns." In *2013 IEEE 39th Photovoltaic Specialists Conference (PVSC)*, pp. 093

3.2 GaN:C

3.2.1 Switches

Progress towards a COMSOL simulation for GaN:C PCSS devices continued by comparing recombination and generation types used in Silvaco TCAD simulation to types in COMSOL. The goal for next month is to have a GaN PCSS simulation device model built replacing the altered Si solar cell model used.

On-state resistance and efficiency measurements were delayed due to being unable to solder to the contacts on the newer NRL devices made of Ti, Al, Ni, and Au (deposited respectively and annealed at 800°C for 30 s) with little or no Au remaining. A fixture using gold probes will be designed and built to allow for low-voltage characterization of on-state resistance and efficiency.

Using a monochromator and Keithley 2450 source-meter, we measured the photoconductivity of two NRL GaN:C PCSS devices (with 30 μm gap length and 2.5 mm long with carbon concentration of $5 \times 10^{17} \text{cm}^{-3}$) and a similar older NRL device by plotting I-V curves (up to 200 V) as a function of wavelength. Once normalized using the incident

power density, 365 nm achieved the highest photocurrent in both devices at $\sim 15 \mu\text{A}/(\mu\text{W}/\text{cm}^2)$. Photocurrent fell to approximately 25% (of $\sim 15 \mu\text{A}/(\mu\text{W}/\text{cm}^2)$) at 375 nm for the new device (further dropping at subsequently higher wavelengths) and to approximately zero in the older device (Figure 3.1). This shows efficient optical excitation using a cheaper violet laser ($\sim 408 \text{ nm}$) is not possible, without changes to trap states and/or geometry, leading to the need to continue with a more costly UV laser with current GaN:C devices.

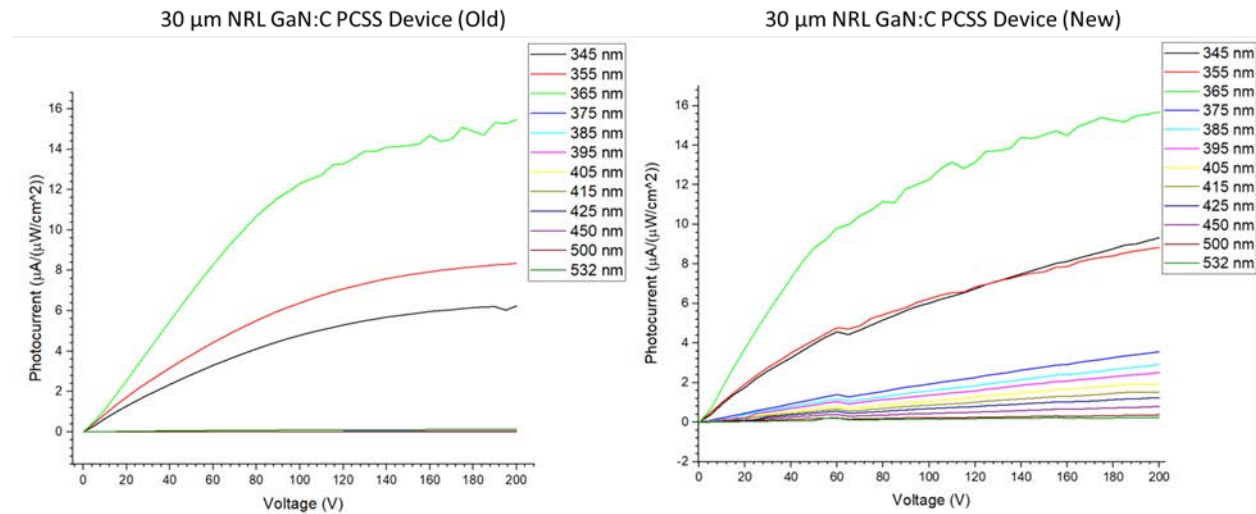


Figure 3.1. Photocurrent measurements of an old vs new NRL device as a function of incident wavelength normalized by the incident power density.

Next Steps:

- i. Complete building a GaN PCSS device in COMSOL.
- ii. Further quantify the effect of carbon-concentration on the on-state resistance and efficiency, low-voltage measurements will be performed on the new NRL devices using a gold-probe fixture.
- iii. Continue pushing the performance limits of GaN:C PCSS, the PCB test fixture will be fully encapsulated to prevent PCB component failure from arcing and allow for higher operating voltages, which previously reached $\sim 1900 \text{ V}$ using a 1 mm gap length device.

3.2.2 Calculations

In the past month the main activity was to set up and perform ab initio band structure calculations to obtain the band gap and effective masses (electron, and heavy hole, and light hole) of GaN. These parameters are necessary inputs for TCAD simulations.

We applied different density functionality, basis sets, and other methods to try to improve the value of the band gap of GaN to more closely match experiment. Unfortunately, our bandgap is still slightly different from the experimental result (in our work we improved the bandgap to 2.97 eV whereas the gap size is 3.5 eV experimentally).

Next Steps:

- i. Apply the same methodologies to carbon compensated GaN and Fe doped GaN to try to get useful parameters for TCAD simulations of these materials.
- ii. Also, we will work to obtain the shear deformation potential for GaN, which will be used to control the deformation of the material when applying force, temperature, pressure or all three at once during TCAD simulation.

3.3 FOM

We have continued our switch evaluation efforts using TCAD simulations for recovery time optimization in Si-PCSS to study and improve device average power handling capability with low-to-no increase in reverse leakage current. We have studied an electron-beam irradiated 1000×500-μm Si-PCSS model with a 1000-μm width, having a photo-illuminated area of 1000×600-μm with a doping profile of [(p+, n-, n+) = (1019-, 1012-, 1019-cm-3)] and illuminated using a 1064 nm optical pulse with 20 ps rise- and fall-times, 120 ps FWHM and 180 μJ optical energy for a single pulse.

3.3.1 Electron-beam irradiation

Devices simulated at 5 kV bias attained a space charge region (SCR) electric field of 2.83×10^5 V/cm and 15 mΩ on-state resistance at 0.5 MW peak power. We have studied the effect of the dominant acceptor-type defect level ($E_{\text{acceptor}} = E_c - 0.4$ eV) at high injection and varied uniformly distributed defect densities and observed an off-state (dark) resistance range of 15–19.9 MΩ for the studied defect density range of $[0, 3 \times 10^{17} \text{ cm}^{-3}]$. The models showed a decrease in off-state resistance and breakdown voltage with increase in defect densities $> 1 \times 10^{16} \text{ cm}^{-3}$, due to an exponential increase in the leakage current resulting from thermal carrier generation and tunneling. This has resulted in a decreased SCR, limiting the device power handling. Devices with no defects showed recovery times of $\sim 1 \mu\text{s}$ and there was no significant impact on recovery-time reduction at defect densities $< 1 \times 10^{14} \text{ cm}^{-3}$, but for densities $> 1 \times 10^{14} \text{ cm}^{-3}$, hundreds-of-ns reduction in recovery time was observed, with a trade-off in the leakage current trends reaching the order of mA as shown in Figure 3.2.

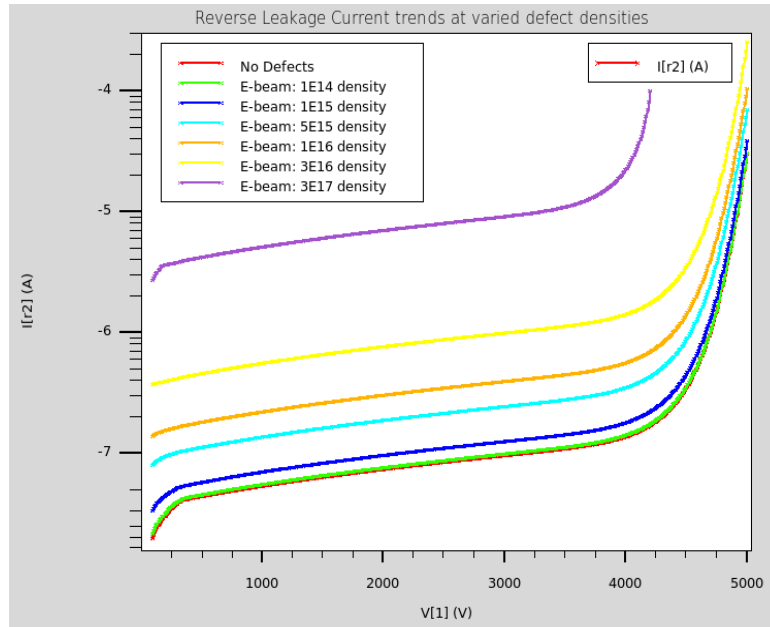


Figure 3.2. Reverse leakage current trends in electron-beam irradiated Si-PCSS at varied defect densities. Voltage hold-off diminished at $>3E16 \text{ cm}^{-3}$ defect density alongside exponentially increasing leakage currents.

3.3.2 Proton irradiation

We have established in the earlier report that proton irradiation with geometrically localized crystal defects and larger capture cross-sections was more effective for carrier lifetime control and recovery time optimization than electron irradiation devices with a uniform defect distribution, which contributed to higher leakage currents. We continued our study to mitigate the high leakage currents with proton irradiated device simulations and modeled devices with high injection dominant acceptor defect level ($E_{\text{acceptor}} = E_c - 0.165 \text{ eV}$) localized in high carrier concentration regions. We have simulated donor-type ($E_{\text{donor}} = E_v + 0.195 \text{ eV}$), acceptor-type and a combination of both defects at varied geometric locations at a defect density of $1 \times 10^{16} \text{ cm}^{-3}$ in a 2-D device of geometry $1000 \times 500 \text{ } \mu\text{m}$ with $1000 \text{ } \mu\text{m}$ width at 5kV bias with proton irradiation defect parameters. This has effectively reduced the leakage currents by three orders of magnitude (shown in Figure 3.3) along with $\sim 50 \text{ ns}$ faster recovery compared to electron beam irradiated models.

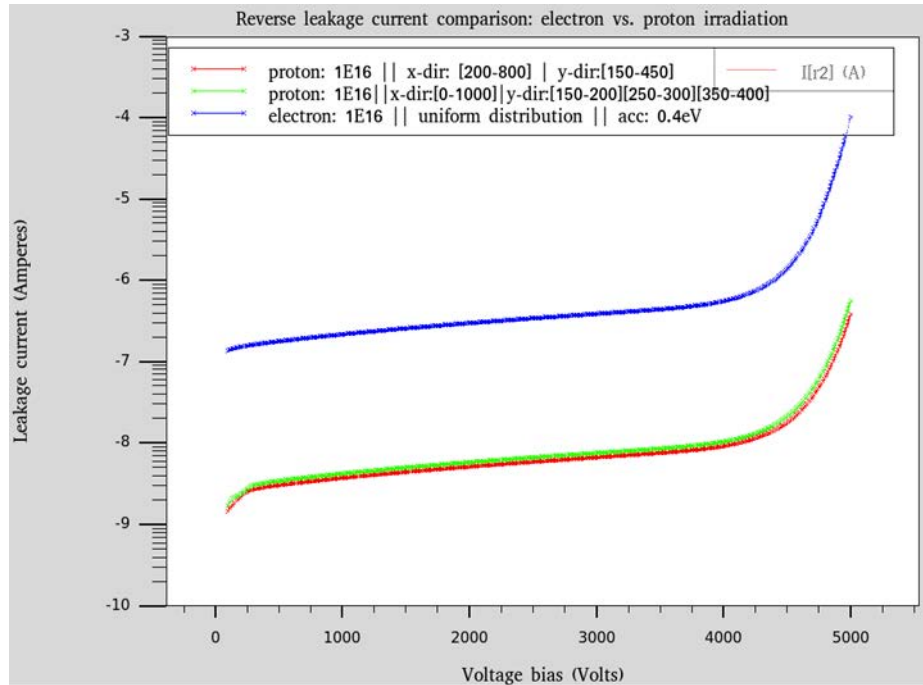


Figure 3.3. Comparison of reverse leakage currents in the electron irradiated model with uniformly distributed defects and the proton-irradiated model with geometrically localized defect clusters at $1E16 \text{ cm}^{-3}$ defect density.

3.3.3 Next Steps

- A Si-PCSS study with proton irradiation defect parameters using $1000 \times 500\text{-}\mu\text{m}$ device model with $1000\text{-}\mu\text{m}$ width will be completed and recovery time vs. defect concentration for electron-beam and localized proton/alpha particle irradiation induced defects will be quantified.
- Pulse forming network (PFN) limitations in pulse shaping and charging/discharging times will be studied to determine the preferred PFN and lifetime control technique for given average power requirements.

4.0 Pulse Charging

4.1 Single Gate Driver HV Switch

- i. A 4.8 kV high-voltage switch, customized for power conditioning of PCSS operation, has been designed, fabricated, and tested to operate with a negative voltage bus.
- ii. A new 6.8 kV HV switch has been designed and simulated in-house using SMD components and reduced trace length for continuous PCSS operation using a negative power supply. The PCB layout has been completed.
- iii. A nonlinear capacitance-assisted single gate driver based HV switch with improved dynamic voltage balancing capability is now in the design phase.

5.0 Thermal Management

5.1 Jet Impingement (JI-TMS)

- i. This June, the thermal management team focused our efforts on exploring viable nanoparticle additives which could be entrained in the thermal management fluid to improve the fluid's capability to remove heat from a HPRF source. Specifically, using low viscosity silicon as the dielectric base fluid (PSF-20cSt) and adding nano-sized particles comprised of aluminum oxide particles or silicon carbide, the following numerical results obtained are provided:
 - a. When the nanofluid is comprised of a 0.2% volume fraction of nanoparticles, aluminum oxide or silicon carbide nanoparticle additives are able to increase the maximum heat removal rate by ~17% and ~23%, respectively.
 - b. The required volumetric flow to produce the turbulent flow characteristics for impinging jets decreases by ~15% when using 0.2% of aluminum oxide particles, or ~5% if using 0.2% of silicon carbide nanoparticles in comparison to the PSF-20cSt base fluid.
 - c. Between these two findings, a clear, definitive answer as to which of the two is the optimal choice for improving the thermal management fluid's performance is not achieved. Although aluminum oxide nanoparticles, when added to PSF-20cSt, reduce the required flow rate more than a silicon carbide nanoparticles, it is able to increase the cooling capability (heat removal rate) of the nanofluid the least between the two.
 - d. Both of these two competing thermodynamic effects affect the overall operation and performance of the JI-TMS and require further analysis, namely by simulations, to occur.
- ii. These heat removal rate and flow rate performance metrics will be used in the continuing ANSYS FLUENT CFD simulations on the jet impingement TMS design. Ultimately in an effort to determine which of the two options produces the optimal heat removal rate capability at the lowest flow rate and corresponding pressure drop. These results will be provided in next month's Grant MSR.

6.0 Antenna

6.1 Trade Space Analysis of ESAs

- i. Scan Blindness Issues
 - a. Earlier a drop in the array gain at $\approx 10^\circ$ from boresight on scanning the 15×15 array at 990.6 MHz was reported.
 - b. We were able to validate our assumption that the appearance of scan blindness at $\approx 10^\circ$ is due to Surface Wave (SW) excitations.
 - c. Methods to suppress SW in the array are being looked into currently.
- ii. Pulse Distortion
 - a. Preliminary investigations into pulse dispersion/distortions have been initiated via simulations in CST.
 - b. However, investigations are more focused on power delivery - directed energy at this stage to meet the OSPRES requirements.
- iii. Fabrication and Measurement
 - a. For the validation of design approach, it has been decided to fabricate a smaller (4×4) array with an external ring of dummy elements but at a higher frequency (3.75 GHz).
 - b. The array is being characterized via different EM solvers (FEKO, CST and HFSS) and the results obtained so far show good agreement.

7.0 RF Coupling

7.1 Time-domain Coupling using Equivalent-circuit Models

During this month, we studied a loop inside a metallic cavity. The summary of this study is flowing:

- i. Placing the wire-loop inside the cavity causes the first minimum of the pristine wire loop in free-space to shift to a lower frequency and decrease significantly its magnitude.
- ii. The minimum in the input admittance of the loop becomes narrower in bandwidth when the wire loop is placed inside the cavity, indicating a resonance of a higher quality factor.
- iii. The cavity creates additional resonance in the wire's input impedance which results in additional coupling paths to the loop.

The open-circuit voltage at the diode port plays the other most important role in defining the coupling to the nonlinear load. It is well known that the cavity will shield the loop resulting in a low induced voltage across the linear/nonlinear components of the loop. However, the authors in [1] showed that the shielding effectiveness of an aluminum cylindrical cavity with a slot can reach more than 0 dB, indicating that the field inside the shield is higher than that outside it. In other words, the voltage across a load attached to a loop inside the cavity can be higher than a load attached to the loop outside a cavity. This phenomenon is still under investigation and more details will be indicated in the next MSR.

Reference:

1. S. Campione et al., "Modeling and Experiments of High-Quality Factor Cavity Shielding Effectiveness," 2019 International Applied Computational Electromagnetics Society Symposium (ACES), Miami, FL, USA, 2019, pp. 1-2.

8.0 Program Management

8.1 Issues

- i. Scope – N/A
- ii. Budget – N/A
- iii. Schedule – N/A

8.2 Interactions with Others

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

8.3 Major Procurement Actions

- i. N/A

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2.0 Executive Summary

In July, the OSPRES team:

- i. Focused optimization of Si-PCSS metallization patterns and proton-irradiated recovery time control.
- ii. Tested non-encapsulated GaN:C PCSS devices and exploration of suitable approaches for RF generation.
- iii. Continued modeling of GaN:C material parameters.
- iv. Continued development of single-gate driver HV switch for pulsed charging, time-domain analysis of pulse dispersion in microstrip-patch antennas.
- v. Modeled diode capacitance and resistance effects on RF coupling to wire loops.

3.0 Switch Evaluation and Design

3.1 Si-PCSS Modeling

3.1.1 Si-PCSS Top Contact Design Optimization

- i. Last month, theoretical methods and SPICE simulation were used to try to determine the optimal spacing for the top contact of 500 μm thick Si-PCSS. Initial TCAD simulations for adapting resistor networks, developed from the theoretical methods for TCAD, led to exploring optimal spacing without the need for the top contact metallization model.
- ii. It was reasoned that as the spacing was increased, the on-state resistance (R_{on}) would reach a minimum and then begin to increase—the minimum determining the optimal spacing. This was confirmed by an initial TCAD simulation of a small 3D device. However, the small device did not allow for a wide range of spacing (or number of current filaments across the top contact).
- iii. 2D 1 cm sized device models were simulated in TCAD in both a discrete capacitor discharge circuit and a simple pulse forming transmission line (described in previous reports). For the discharge of the PFTL, we simulated three laser energies (14 μJ , 28 μJ and 56 μJ) and 25 spacing's from 10 μm (including no filaments, window of 8000 μm).
 - a. The resulting R_{on} reached a minimum at a spacing of 400 μm and increased only slightly when spacing approach 200 and 800 μm . The percent reduction in laser energy was interpolated from the results over a range in laser energy range from 14 μJ to 56 μJ , showing a reduction of 10–40% depending on the on-state resistance.
 - b. At 400 μm (minimum R_{on}), the required laser energy to achieve a R_{on} of 0.12 Ω was reduced by 37.9% (28 μJ) when compared to a spacing of 20 μm .
- iv. Simulations of 1.8 cm device were also conducted, but these results did not differ significantly from 1 cm device results.
- v. For the capacitor discharge circuit, we extended the simulation to vary dopant profiles and increased voltage. Doping profiles from previous Si-PCSS devices were compared to the most recent doping profiles. The previous profile was shallower with a greater peak concentration (same background doping used in both), resulting in greater R_{on} .
 - a. While we expect higher doping of p-side to decrease resistive losses, this is complicated by role of free-carrier absorption, which has not been yet been simulated. This will be explored by varying only the doping concentration and reported in subsequent reports.
 - b. High voltage greatly increased the on-state resistance, with all else constant. For the same voltage and doping profile, the discharged capacitor simulations yielded the same results for optimal spacing except the optimal spacing was at 500 μm .

Next Steps:

- i. Explore options to account for free-carrier absorption, such as the Drude model in Silvaco TCAD. The optimization of the doping concentration on the p-side can be considered once an accurate absorption model is found and would include resistive losses in the top-layer of the p-side due to the spacing used.
- ii. Simulate at higher voltages within an in-line pulse forming transmission line to determine the optimal spacing and laser energy reduction.
- iii. Repeat simulations In addition to the previously studied 500 μm thick, 1 cm-sized device, a 1300 μm thick, $\frac{1}{2}$ cm-sized Si-PCSS device model will be used to allow for higher voltage simulations.

3.2 GaN:C**3.2.1 Pre-encapsulation Pulsed Testing**

To understand the effects of encapsulation on GaN:C PCSS performance, the on-state resistance of a 5 mm device at low voltage (50 V – 200 V) was measured before encapsulation, which will be compared to results measured after encapsulation with Sylgard-184. The on-state resistance was minimized ($\sim 1 \Omega$) at a laser energy density of $\sim 16 \mu\text{J}/\text{mm}^2$, matching previous results for a 10 mm device. Use of a smaller device conserved the remaining 10 mm devices and provided initial results for smaller GaN:C devices.

Next Steps:

- i. Measure on-state resistance and efficiency as a function of carbon concentration by performing low-voltage measurements on the new NRL devices with the pogo pin fixture. If the pogo pin fixture is not suitable for these measurements, a new gold-probe fixture will be fabricated.
- ii. Evaluate the performance limits of the Sylgard-184 encapsulated device and PCB by measuring the on-state resistance, efficiency, and peak output power at voltages up to device failure.
- iii. Demonstrate RF capability of GaN:C PCSS by using a circuit, based on the *pulse-switch-out generator* used for GaAs BOSS devices [1], to show bipolar modulation.

Reference:

1. Stoudt, D. C., Richardson, M. A., & Peterkin, F. E. (1997). "Bistable optically controlled semiconductor switches in a frequency-agile RF source." *IEEE Transactions on Plasma Science*, 25(2), 131–137. <https://doi.org/10.1109/27.602483>

3.2.2 Ab initio Calculations

Ab initio band structure calculations were continued to obtain the band gap and effective masses (electron, heavy hole, and light hole) of GaN.

We computed elastic constants and lattice constants for all 3 models (GaN, C-doped GaN and Fe-doped GaN). When we combine the polarization constant and mechanical constant, the total polarization of the material is obtained. We are currently computing the

shear deformation potential. These calculated parameters will be used as inputs for TCAD simulations.

Next Steps:

- i. Complete the heat deformation potential calculations.
- ii. Continue to improve the band gap calculations.
- iii. Apply the same methodologies to GaN:C and GaN:Fe to extract useful parameters for TCAD simulations.
- iv. Continue to explore the effects of mesoscale structure (i.e., extended defects such as dislocations and interfaces) on the bulk properties of GaN that are relevant for TCAD simulations.

3.3 FOM

3.3.1 Si-PCSS Defect and Recovery Optimization

TCAD simulations were utilized to evaluate Si-PCSS designs that incorporated proton irradiation induced defects. We modeled a 1000- μm wide, 1000 $\times$ 500- μm Si-PCSS with a photo-illuminated area of 600 $\times$ 1000- μm and a Gaussian doping profile of [(p+, n-, n+) = (10^{19} -, 5×10^{12} -, 10^{19} - cm^{-3})]. Defects from proton irradiation with crystal defect parameters for acceptor- and donor-type defects at ($E_c-0.165$ eV), and ($E_v+0.195$ eV) energy levels were included; the simulated device was illuminated by a 180 μJ , 120 ps FWHM, 1064 nm optical pulse, with a 20 ps rise/fall time.

Penetration depth, fluence rates, and defect distribution required to achieve localized defect densities of 1×10^{16} cm^{-3} and 1×10^{17} cm^{-3} were calculated for H^+ and He^{2+} ions using SRIM (Stopping and Range of Ions in Matter) and TRIM (Transport of Ions in Matter). Sample devices will be irradiated using the calculated values and tested to experimentally verify the leakage current and recovery time reduction predicted by simulation; pre-irradiation reverse recovery measurements of 7 cm Si-PCSS devices with no electron irradiation were completed before sending out for proton irradiation tests.

Temperature dependent simulations for Si-PCSS at 5 kV bias are being performed to quantify total heat dissipation and device temperature-rise rate per pulse caused collectively by: carrier-carrier and carrier-lattice collisions, carrier generation/recombination, Peltier-Thomson effect, and joule heating. This is also done as a function of recombination rates based on varied crystal defect densities and distributions induced by electron and proton irradiation.

Next Steps:

- i. Quantify heat dissipation at varied recombination rates via non-isothermal simulations to compare Si-PCSS with other switching technologies for device level evaluation.
- ii. Experimentally verify simulated results for proton irradiated defect parameters and distribution profiles to compare with electron beam irradiated devices at identical or lower defect densities.

- iii. Determine the preferred PFN and lifetime control technique for given average power requirements by studying pulse forming network (PFN) limitations in pulse shaping and charge/discharge times.

4.0 Pulse Charging

4.1 Single Gate Driver HV Switch

- i. We have simulated a modular HV switch design rated for 10 kV in LTspice using three modules each containing three 1200 V rated MOSFETs. The research effort also includes following tasks.
 - 1. Design an isolated HV gate driver with isolation rating close to 10 kV
 - 2. Limit voltage imbalance among the MOSFETs in the HV switch below 5-10%.
- ii. The BEHLKE based power conditioner board has been tested in the GTEM using a negative supply voltage up to 2.5kV; the in-house HV switch for negative power supply had been tested up to a 1 kV supply voltage.

5.0 Antenna

5.1 Trade Space Analysis of ESAs

- i. Time Analysis
 - a. A single microstrip patch antenna, which was designed at 3.75 GHz, is subjected to input Gaussian pulse via FEKO and the output is observed at 1 and 2 km from the antenna.
 - b. The output pulse has shown much less distortion but an expected decay in amplitude [1].
 - c. The output of the antenna subjected to other types of pulsed inputs is being investigated.
- ii. Fabrication and Measurement
 - a. For the validation of the design approach, a smaller 6×6 array (including an external ring of dummy elements), designed to operate at 3.75 GHz, has been characterized via different EM solvers (CST, HFSS, and FEKO).
 - b. The model will be sent for fabrication and measurement in the upcoming weeks.
- iii. Trade Space Analysis
 - a. A code is being developed for faster computation of array parameters from user inputs such as: engagement range, target effect levels, antenna element type, lattice type, and amplitude distribution.

Reference:

1. D. V. Giri, High-Power Electromagnetic Radiators, Harvard University Press, Cambridge MA, 2004, p 102.

6.0 RF Coupling

6.1 Time-domain Coupling using Equivalent-circuit Models

July, we studied the effect of the diode junction capacitance and reverse resistance on the coupling to a loop loaded with the diode. The analysis illustrated that the voltage across the diode is exponentially proportional to the diode junction capacitance and its reverse impedance.

We also have studied the shielding effectiveness of a cavity with one slot on one of the cavity's faces.

Summary:

- i. The shielding effectiveness of the cavity with a slot can reach more than 0 dB at all the cavity resonance frequencies, which also creates additional resonances in the input impedance of the circuits inside the cavity, adding coupling paths to the circuits.
- ii. The slot size determines the bandwidth and magnitude of the fields inside the cavity. The smaller the slot size the higher the power content of the signal at the cavity resonances is.

Next Steps:

- i. Continue to investigate the effect of slot size, shape, and location on shielding effectiveness.

7.0 Program Management

7.1 Issues

- i. Scope – N/A
- ii. Budget – N/A
- iii. Schedule – N/A

7.2 Interactions with Others

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

7.3 Major Procurement Actions

- i. N/A

7.4 Travel

Destination	Purpose	Attendees	Estimated Costs

7.5 Pending or Upcoming Public Release Requests

- i. N/A

ONR HPM Program - Monthly Status Report (MSR) – August 2020

Anthony Caruso – Univ. Missouri – Kansas City

N00014-17-1-3016 [OSPRES Grant]

ONR Short Pulse Research, Evaluation and non-SWaP Demonstration for C-sUAS

09/15/2020

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2.0 Executive Summary

In August, the OSPRES team focused on (3.1.1) optimization of Si-PCSS metallization patterns to minimize laser energy, (3.1.2) continued evaluation of proton-irradiation for recovery time-enhancement, (3.2.1) testing GaN:C-PCSS and designing a new PCB for based on prior are on pulse-switch-out-generators (PSOGs), (3.2.2) obtaining the total polarization constants of GaN for improving TCAD simulations, (4.1) designing a new gate-driver for and testing the negative-polarity single-gate driver HV switch for pulse charging, (5.1) experimental characterization of the microstrip-ESA array performance including scan blindness array at Honeywell, (6.1) continued modeling of rf-coupling dynamics of slotted cavities, and (7.1) optimization of the jet-impingement cooling system to maximize flow rate and minimize head-loss, and characterize silicone-oil contamination effects on cooling performance. Highlights from those efforts include:

- i. Simulations indicate that increasing the Si-PCSS window spacing to 350 μm could reduce laser requirements by up to 31%.
- ii. Si-PCSS devices and quotes are being obtained for studies of 3 proton irradiation profiles.
- iii. A new PCB has been designed based on pulse-switch-out-generators (PSOGs) from prior art in GaN BOSS switches.
- iv. In-house HV-switch tested at 2.5 kV with a switching frequency of 40 kHz and 50% duty cycle.
- v. No scan blindness is observed in the microstrip ESAs hen steered over $\pm 60^\circ$ from boresight at frequencies of 2.7 GHz to 3.7 GHz), but severe degradation occurred at 4.75 GHz.
- vi. Reduced the jet-impingement cooling system head-loss (pressure drop) to 10.9 psi at 1 gpm, and 37.4 psi at 2 gpm while maintaining equivalent or better cooling rates.

3.0 Switch Evaluation and Design

3.1 Si-PCSS Modeling and Optimization

3.1.1 Contact Optimization

- i. Last month, we modeled Si-PCSS photoconductivity to reduce laser requirements and determined that the optimal top contact spacing was ~400 μm at 100 V to minimize on-state resistance. The objective was to reduce required laser energy by at least 10% and achieve the same or reduced on-state resistance. We defined the baseline required laser energy for devices with a window spacing of 20 μm (E_{20}) and defined *percent laser energy reduction* (E_{per}) as:

$$E_{\text{per}} = \frac{E_{20} - E_{\text{wid}}}{E_{20}} * 100\%$$

where the wider window laser energy is E_{wid} .

- i. TCAD studies were completed for high voltage (2000 V) in an inline transmission line (TL) based pulse forming network (PFN) for optimal spacing and the required laser energy reduction. Because the energies simulated previously were too sparse for accurate interpolation, the new simulations used a larger number of laser energies of 10, 20, 30, 50, 70, 90, and 112, resulting in on-state resistances ranging from 0.018 to 0.332 Ω .
- ii. The percent laser energy reduction was relatively constant over the range of on-state resistances achieved for the (interpolated) laser energies used. Constant percent laser energy reduction vs on-state resistance allowed for determining percent laser energy reduction vs spacing used (Figure 3.1). At the optimal spacing of 350 μm , the laser energy required was reduced by 31%.

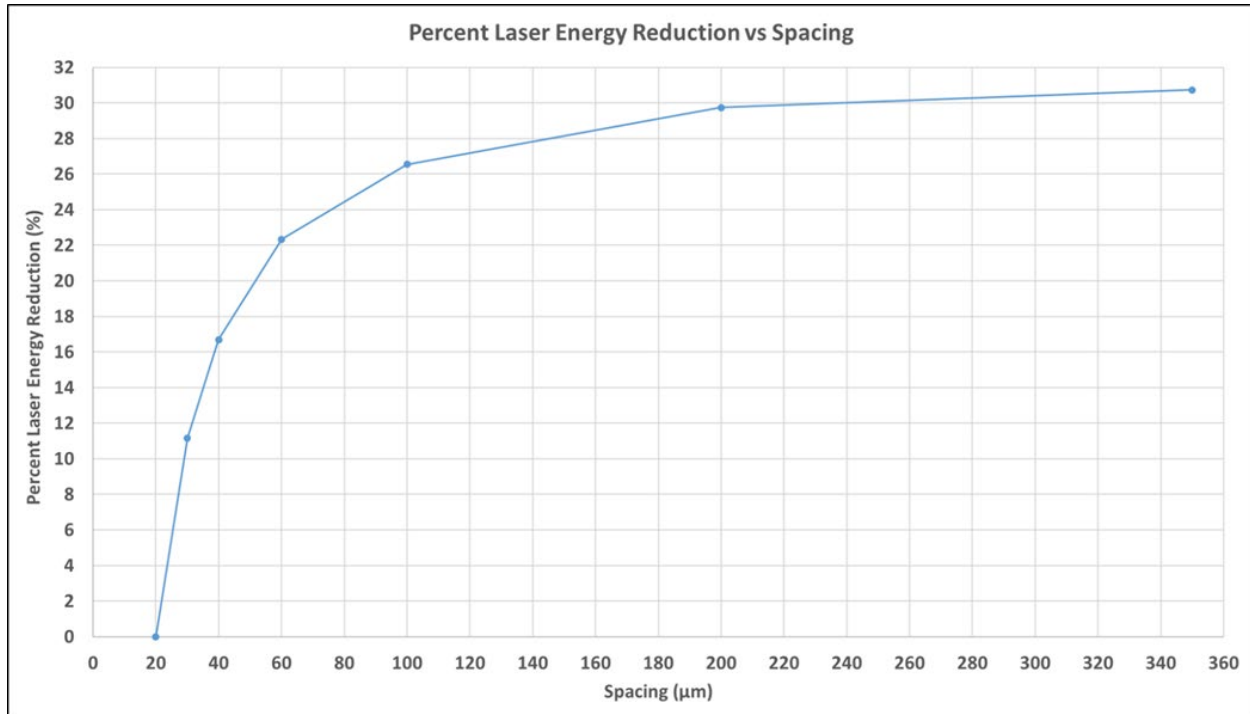


Figure 3.1 Laser energy reduction up to the optimal 350 μm spacing. The 20 μm spacing is an extrapolation and serves as the reference spacing, and thus corresponds to zero laser energy reduction.

- i. After this relationship between percent laser energy reduction and spacing was established, we initiated an experimental metallization study for 1 cm and 7 cm Si-PCSS. A series of wafer patterns were requested for Radiation Detection Technologies to fabricate for validation:
 - a. For 1 cm Si-PCSSs, window spacing's were 20, 40, 60, 100, 200, and 350 μm with 6 devices per wafer;
 - b. For 7 cm Si-PCSSs, spacing's were 60 μm , 350 μm with 5 devices per wafer.
- ii. We expect a ~2 month turnaround time for the above wafers, after which validation of the modeling results will begin. Hence, we expect to report results on this study in the November, 2020 report.
- iii. Because free-carrier absorption was previously omitted from TCAD models, which we believe to negatively effect on-state resistance, we began incorporating free-carrier photogeneration to quantify its effect and improve model accuracy. By implementing a user-defined photogeneration function, we accounted for dopant dependent free carrier absorption in the model. As expected, results confirm that dopant dependent free carrier absorption increases the on-state resistance for laser energy used. Further analysis of the result are still in progress.

Next Steps

- iv. Determine the dopant dependent laser energy required for on-state resistance that more closely matches experimental results.

- v. Optionally (pending results from analysis) include carrier concentration dependent free carrier absorption to better match experimental results for the magnitude of the required laser energy for some on-state resistance (after including additional contact resistance contributions to the on-state resistance).
- vi. Develop a TCAD GaN breakdown model for the effort in custom material parameters for carbon compensated GaN (see Rulis).

3.1.2 Si-PCSS Defect and Recovery Optimization

- i. Based on (a) simulated TCAD models of recovery time and leakage current optimization and (b) SRIM/TRIM calculations of penetration depth and fluence rates, we have designed experiments to obtain 3 defect profiles via irradiation with proton beam energies capable of achieving up to 250 μm penetration depth and fluence to create $1\text{E}16\text{ cm}^{-3}$, $5\text{E}16\text{ cm}^{-3}$ and $1\text{E}17\text{ cm}^{-3}$ defect densities.
- ii. We conducted pre-irradiation reverse recovery and on-state resistance testing on $7\text{ cm} \times 0.5\text{ cm}$ devices (500 μm and 1300 μm thick) and $1\text{ cm} \times 1\text{ cm}$ devices (500 μm thick). The devices will be sent for proton irradiation for experimental validation.
- iii. Vertical and lateral geometry WBG-PCSS are being modeled in TCAD for vanadium-compensated 4H-SiC and carbon-compensated GaN to simulate transient and steady-state response. Response characteristics to be studied include rise- and fall-time, voltage hold-off, and peak- and average powers for performance evaluation against Si-PCSS models.

3.2 GaN:C Modeling and Optimization

3.2.1 GaN:C Device Testing

- i. Optics were received and setup for the demonstration of bipolar RF capability of GaN:C PCSS using a circuit, based on the *pulse-switch-out generator* (PSOG) used for GaAs BOSS devices. PCBs were designed and fabricated for use with the PSOG-based setup.
- ii. On-state resistance of Sylgard-185 encapsulated devices was measured to determine the effect of encapsulation on device performance and breakdown. The minimum on-state resistance achieved using Sylgard was $1.4\ \Omega$ at $\sim 16\ \mu\text{J}/\text{mm}^2$, compared to $<1\ \Omega$ without encapsulation at the same incident energy density. Pushing the device (with full PCB encapsulation) to higher voltages resulted in failure of the charge resistors at $\sim 800\text{ V}$ due to exceeding the power rating. One could conclude that encapsulation of the entire PCB does not achieve higher voltages with this setup due to pulse-repetition and power limitations of the resistors in the PCB circuit. To overcome these limitations, higher voltage operation will be demonstrated by eliminating the charge resistors by either using a pulse charger or flyback circuit, to maintain isolation and reduced capacitance from the DC power supply.

Next Steps:

- iii. Measure the on-state resistance and efficiency as a function of carbon concentration by performing pulsed, low-voltage measurements, using the wedge

laser, on the new NRL devices with the pogo pin fixture. If the pogo pin fixture is not suitable for these measurements, a new gold-probe fixture will be fabricated.

- iv. Demonstrate RF bipolar modulation by using the PSOG based circuit and accompanying setup with GaN:C PCSS.
- v. Demonstrate high-voltage operation of GaN:C PCSS by using a pulse charger and/or flyback/inductor circuit with the current PCB setup.

3.2.2 *GaN:C Calculations*

- i. We continued computing the shear deformation potentials, but had to put calculations on hold for machine maintenance impeded progress. When combining the polarization constant and other mechanical constants, we can obtain the total polarization of the material. These parameters are necessary inputs for TCAD simulations.

Next Steps:

- ii. Continue improving the band gap calculations.
- iii. Apply the same methodologies (used for regular GaN) to carbon compensated GaN and Fe doped GaN, to obtain useful parameters for TCAD simulations of these materials.
- iv. Continue to explore the role that mesoscale structure (i.e., extended defects such as dislocations and interfaces) plays on the bulk properties of GaN that are relevant for TCAD simulations.

4.0 Pulse Charging

4.1 Single Gate Driver HV Switch

- i. We have tested the in-house HV SiC switch using a negative power supply using following parameters and recoded a maximum imbalance close to 50 V among the series connected MOSFETs.
 - a. Input voltage - 2.5 kV
 - b. Switching frequency - 40 kHz
 - c. Resistive load- 3.2 k Ω
 - d. Duty cycle- 50%
- ii. We have initiated the design of an isolated HV gate driver (~ 10 kV rated) required to build a modular in-house HV switch.
 - a. Initiated the design of an isolation transformer with close to 10 kV isolation and minimal coupling capacitance to safely power the gate driver IC.
 - b. Selected suitable core type, core geometry, required turns-ratio and probable methods of insulation between primary and secondary sides to build the isolation transformer.
- iii. We have tested a series stack of six avalanche transistors (ATs) to evaluate its feasibility to use as Si PCSS surrogates.
 - a. The triggering voltage of the stack of ATs depends on the load connected at the end of a transmission line which may vary.
 - b. The limited peak pulse current withstanding capability of ATs makes it difficult to use it in a pulse forming network where the peak current exceeds the maximum rating of the ATs

5.0 Antenna

5.1 Trade Space Analysis of ESAs

5.1.1 *Single Antenna Design:*

- i. Radiation pattern of the electrically small microstrip patch antenna (center frequency = 3.75 GHz), which was fabricated earlier, has been measured at Honeywell Research Lab.
- ii. Experimental data shows good agreement with simulated data obtained via FEKO, CST and HFSS.

5.1.2 *Array Design:*

- iii. To validate the array design approach, a small 4×4 array has been modeled and characterized via FEKO and CST at a frequency of 3.75 GHz.
- iv. No scan-blindness is observed as the array main beam is steered from boresight to $\pm 60^\circ$ at the lower frequencies (2.745 and 3.7475 GHz).
- v. Severe pattern degradation has been seen at 4.75 GHz – due to surface wave (SW) excitations.
- vi. Reduction in SW excitations is necessary to reduce: edge diffraction from the ground plane/substrate, back radiation (harmful for r.f circuitry behind the ground plane – EMI/EMC issues for HPM applications), and mutual coupling between antenna elements.
- vii. An effort to design RSW (Reduced Surface Wave) antennas has been initiated.

References:

[1] V. R. Komanduri, D. R. Jackson, J. T. Williams and A. R. Mehrotra, "A General Method for Designing Reduced Surface Wave Microstrip Antennas," *IEEE Trans. on Antennas and Propag.*, Vol. 61, No. 6, pp. 2887-2894, Jun 2013.

6.0 RF Coupling

6.1 Slotted Cavity Coupling

- i. A metallic enclosure, or a cavity with a slot, has a high-quality factor causing the penetrating waves to bounce back and forth, creating a standing wave. This standing wave can create large fields at the resonance frequencies of the cavity, i.e. the field inside the cavity can be higher than that outside it [1]. We showed in the last MSR that as the slot size increases the time it takes the cavity internal fields to bounce back and forth and leave the cavity decreases resulting in a low-quality factor response hence high bandwidth response with low magnitude. Over the last month, we investigated different UAV structures to identify possible cavity configurations and the slots where electromagnetic waves can penetrate. Different ways to enhance/mitigate RF coupling to these cavities are also under investigation.

7.0 Thermal Management

7.1 Jet Impingement Optimization

- i. This August, the thermal management team focused on finalizing ANSYS computational fluid dynamics simulations for the second iteration of the single-phase jet impingement TMS for Si-PCSS. This included ensuring consistency among each simulation by fixing parameters such as turbulence model, Courant number, and mesh resolution to produce directly comparable results.
 - a. The following summarizes the results of simulations using the second iteration with reduced head loss (pressure):
 - b. Flowing silicone oil at 1 gpm while inducing 88.4 W of average heat generation rate through the Si-PCSS yields a pressure drop of 10.9 psi and a max Si-PCSS temperature of 51 °C.
 - c. At 2 gpm under the same conditions yields a pressure drop of 37.4 psi and a maximum temperature of 47.4 °C.
 - d. Increasing the average power input to 200 W while maintain flowrate at 1 gpm resulted in a head-loss of 10.9 psi and maximum Si-PCSS temperature of 96 °C.
- ii. At 2 gpm under the same conditions yields a pressure drop and max Si-PCSS temperature of 37.4 psi and 94.8 °C, respectively.
- iii. Moving forward, we will utilize what we have learned from both of the JI-TMS iterations and produce a third JI-TMS design iteration which will aim to incorporate additional head loss reduction techniques (i.e., increase the fluid inlet/outlet's cross-sectional area) while reducing the number of fluid outlets back to one and continue to improve on the maximum heat generation rate handling capabilities while limiting the Si-PCSS temperature to 80 °C or less. In addition, incorporate cyclic heating simulations to emulate burst profiles and laser shot-to-shot transient simulations for HPRF-based systems.
- iv. Experimental evaluation of the silicone dielectric fluid, PSF-20cSt, is ongoing to obtain its thermophysical properties such as specific heat as well as including entrained nanoparticle additives (e.g., Alumina) to increase the thermal performance of the thermal management fluid itself. Initial results on PSF-20cSt indicate that unwanted contaminants found within the fluid yield a significant drop of specific heat than that specified by the manufacturer. Specifically, at 20 °C, the measured specific heat was found to be ~1.1 [J/g·K]. When compared to the value of 1.6 [J/g·K] reported by the manufacturer, Clearco, lead to the conclusion that the silicone oil was contaminated which resulted in a 31.25% decrease in specific heat and therefore the heat removal rate capabilities of the JI-TMS results. Moving forward, we will filter the fluid and perform the experiments again. Followed thereafter by including volumetric fractions of nanoparticle additives to complete the nanofluid experimental study. These results will be provided in next month's Grant MSR.

8.0 Program Management

8.1 Issues

- i. Scope – N/A
- ii. Budget – N/A
- iii. Schedule – N/A

8.2 Interactions with Others

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

8.3 Major Procurement Actions

- i. N/A

8.4 Travel

Destination	Purpose	Attendees	Estimated Costs

8.5 Pending or Upcoming Public Release Requests

- i. N/A

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2.0 Executive Summary

In September, the OSPRES team continued modeling and analysis of the mechanisms by which RF waveforms couple to traces within shielded enclosures, and research and development of source technologies spanning GaN-PCSS, Si-PCSS, SiC-PCSS, diode-based NLTs, high voltage switching for power conditioning, and analysis of microstrip-path electronically small antenna (ESA) arrays—all aimed at (1) uncovering innovations that enable more compact and efficient effector trade-space and future technological capabilities, (2) identifying and evaluating present and future capabilities of each technology and class, and (3) providing key insights on the underlying coupling mechanisms to the advance the HPM community knowledge base.

Recent focus of RF coupling effort is aimed at understanding how metalized enclosures having slots and/or feedthroughs potentially decrease and/or amplify incident RF waveforms, thus enhancing coupled currents to UAVs under specific waveform parameters, and attenuate coupled currents at others—critical to extending the simplified RF coupling models more representative of actual UAV electronics.

The efforts to evaluate and optimize PCSS switching technologies has continued surveying the trade-space within each material set, including optimizing contact patterns to reduce laser energy demands for Si-PCSS by as much as 30%, which translates directly to reduced laser SWaP-C². The GaN-PCSS-focused efforts have had a recent breakthrough in ab-initio modeling, accurately modeling the materials bandgap within 0.01 eV. The ability to model this fundamental properties shows promise for accessing a deeper understanding of the fundamental properties that may be exploited to increase the photoresponse, carrier lifetime, and hold-off voltage. Moreover, the GaN-PCSS testing and evaluation have begun evaluating methods of direct RF-modulation of GaN pushing toward the ability to generate monopolar and bipolar pulses at arbitrary frequencies and/or multi-tone waveforms, a crucial step toward advancing GaN-PCSS based RF generation capability.

Research and development in compact, high-voltage switch and their enabling components (i.e. isolated gate-drivers), are aimed at advancing critical power conditioning technologies for PCSS-based pulse forming networks and their alternatives. Currently, the supply of similar technologies is extremely limited and costly—developing this technology using COTS components increases the availability and decreases overall system SWaP-C² for numerous technologies relying on capacitive energy storage.

3.0 Switch Evaluation and Innovation

3.1 Si-PCSS Modeling and Optimization

For all high power PCSS-based RF sources, the activating laser system represents a significant cost—upwards of an order of magnitude greater than any other ancillary or auxiliary system. For an electronically steered short-pulse HPM array, this cost is further exacerbated by the requirement for multiple extremely low-jitter (< 50 ps) lasers to drive each PCSS element independently to enable true-time-delay steering. To minimize these costs, the OSPRES team has conducted a series of studies aimed at reducing the required laser energy by optimizing the Si-PCSS metallization pattern.

3.1.1 Photo-response Optimization

Results over recent months indicate that a laser energy reduction of up to 30% is possible—we are awaiting device to validate this experimentally.

- i. Doping dependent free carrier absorption (DDFA) was improved by refining the meshing used in TCAD simulation to produce no photogeneration underneath top contact filaments. The top contact filaments are $10\text{ }\mu\text{m}$ wide and the mesh had to be refined to at least $5\text{ }\mu\text{m}$ with finer meshing down to $1\text{ }\mu\text{m}$ producing better results. Next, the TCAD model was implemented for 1 cm^2 device models as opposed to 1 mm^2 . DDFA raised the required laser energy for an on-state resistance by 10% showing it was non-negligible.
- ii. Experimental on-state resistance versus laser energy results were almost exclusively conducted on $1300\text{-}\mu\text{m}$ thick Si-PCSS while TCAD simulations made use of thinner devices to reduce mesh points. A new DDFA TCAD study is being completed for percent laser energy reduction vs spacing by first producing on-state resistance vs laser energy curves with a comparison to experimental data for $1300\text{ }\mu\text{m}$ devices. The inclusion of the DDFA photogeneration model greatly increases simulation time for the larger models. Due to mesh point limitations a $1300\text{ }\mu\text{m}$ thick, 0.8 cm wide 2D device model was used. Initial results show a slightly reduced percent laser energy reduction vs spacing compared to TCAD simulations last month without DDFA and for $500\text{ }\mu\text{m}$ devices.

3.1.2 Next Steps

- i. The percent laser energy reduction vs spacing for $1300\text{ }\mu\text{m}$ thick devices with DDFA will be completed. These results from TCAD can then be compared to experimental tests of new RDT Si-PCSS devices with the top contact using wider window spacings than previous devices. These new devices are expected to arrive mid to late October.
- ii. Optimization of the doping profile of the Si-PCSS in conjunction with the top contact spacing using DDFA photogeneration models will be completed in TCAD.

3.2 Si- and Si-C Modeling and Evaluation

Source driver evaluation is focused on studying and comparing the fundamental operation of Si-, SiC- and GaN-PCSS through performance-based, application-specific figure(s) of

merit (FOM) for HPM systems based on switching speeds, power dissipation, PFNs and cooling needs, and explore possibilities to push the practical limitations of each technology by proposing novel fabrication techniques using TCAD modeling and simulation. This study involves a comprehensive study of the effects of device geometry, doping-levels and distribution, crystal defect types, energy levels, and concentration/densities on transient photo carrier- generation, mobility, scattering and recombination.

Device technologies' optical-, electrical- and thermal-properties are used to compare optical absorption profile, laser requirements, power density, peak and average power, on-state resistance, transient and DC response, and switching and on-state losses. This effort also involves quantifying the cooling requirements for each technology based on the heat dissipation and temperature-rise rates as a function of aforementioned device parameters. This month, our efforts have been focused on SiC-PCSS baseline modeling to obtain photo-absorption profile, and Si-PCSS non-isothermal simulations to quantify heat dissipation as described below:

- i. SiC-PCSS models with 900 x 360- μm lateral, and 1000 x 200- μm vertical geometries with gold contacts and Si_3N_4 anti-reflective coating have been modeled in TCAD with trap models to account for vanadium compensation. Optical absorption and carrier generation with wavelengths 265, 355, 385 and 532 nm and optical energies between 250 μJ and 2.5 mJ have been performed at 10 kV DC bias. And we continue to study both the DC and transient performance to compare with Si-PCSS.
- ii. Si-PCSS non-isothermal simulations to quantify heat dissipation and temperature rise per pulse due to joule heating, generation-recombination, Peltier-Thomson effect have been performed. However, the results failed to precisely quantify heat dissipation from each process. We are working on a 3D model to compute the power dissipation more accurately.
- iii. Experimental validation of proton irradiation study with varied defect profile localization for Si-PCSS carrier lifetime control is on hold for the vendor.

3.2.1 *Next steps:*

- i. Continue to study optical and electrical performance of P-N doped and undoped semi-insulating vanadium compensated SiC-PCSS including optical absorption profile characterization, geometric (vertical and lateral) and doping profile optimization, and their effects on carrier dynamics.
- ii. Simulate 3D Si-PCSS models at varied defect density and localization and optical energy to quantify their effects on power throughput, output pulse width, heat dissipation and temperature rise.
- iii. Model carbon- and iron- compensated GaN-PCSS with vertical and lateral geometries to simulate transient response, including rise- and fall-time, voltage hold-off, and peak- and average powers for performance evaluation against Si- and SiC-PCSS models.

3.3 GaN:C Modeling and Optimization

3.3.1 GaN:C Testing and Characterization

For the development of high-power RF source technologies, overall system SWaP may be reduced by using a material that allows for smaller PCSS devices with less cooling requirements and also allows for direct RF modulation using the laser/photon source driving the device—resulting in a more compact and mobile system. To work towards a more compact system, GaN:C PCSS is being tested to demonstrate unipolar and bipolar RF modulation in the L-band via direct modulation of the laser using an optical delay stage to generate unipolar and bipolar pulses. Unipolar pulses require a single GaN:C PCSS with variable delay between the initial and subsequent laser pulse. For bipolar pulse generation, a second GaN:C device will be used with a negative power supply with the same optical delay setup. This work is based off the GaAs bulk optically controlled semiconductor switch (BOSS) developed in the 1990's as a wide-band frequency-agile RF source. The immediate challenges in applying this technique to GaN is the slow recovery time (fast relative to Si, ones of nanoseconds) and power limitations of the PCB circuit components at present (charge resistors). To overcome charge resistor failure on the PCB, a pulse charger has been obtained from the HV switch development team that will allow for high-voltage operation, which will begin in the ones-of-kV range. To mitigate thermal challenges, losses must be minimized in the system, so the on-state resistance and recovery of newer NRL devices with differing carbon concentrations and buffer-layer annealing temperatures will be measured for comparison with prior devices. Measurements on these devices will be taken at lower voltages due to changes in the contacts that disallow direct soldering of switch leads. As demonstrated in previous reports, on-state resistance scales linearly with higher voltages, so low-voltage measurements can be used to predict performance at higher voltages.

- i. Initial energy measurements of the optical delay for the *pulse-switch-out generator* PSOG setup (Figure 3.1) were taken, and the average energy output was 3.7 μJ to 49.5 μJ for the transmitted beam and 2.2 μJ to 45.8 μJ for the reflected (delayed) beam.
- ii. Initial waveforms were captured using a single GaN:C PCSS to show that there is sufficient optical energy in the delay setup to trigger the device (Figure 3.2). The spacing of the optics will be further refined as testing begins.
- iii. In collaboration with the HV switch development team, incorporation of a pulse charger into the PSOG/GaN:C PCSS setup is being done to eliminate the charge resistors that failed in previous testing and allow for high-voltage operation.
- iv. We started measuring on-state resistance as a function of incident energy at low voltages for the newer NRL devices with carbon concentrations of $8\text{E}16$ and $5\text{E}17$ cm^{-2} to quantify carbon concentration effects on on-state resistance and recovery time.

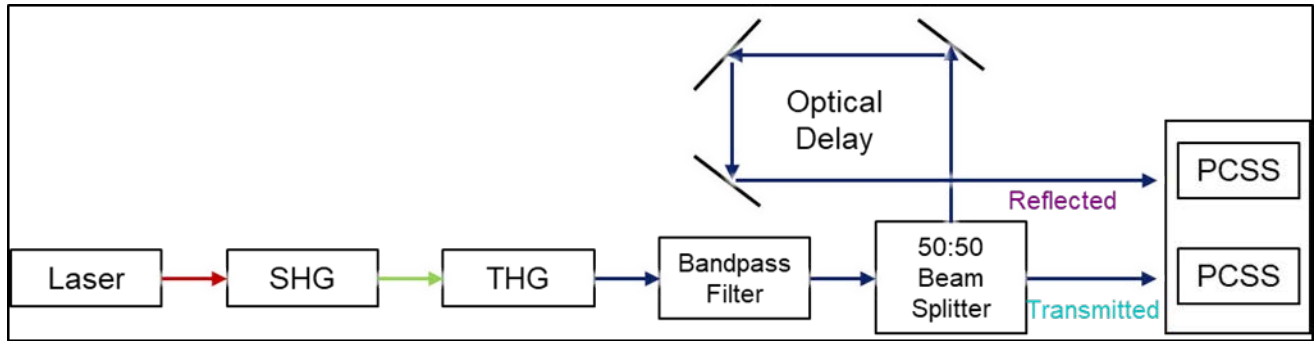


Figure 3.1: Schematic of the PSOG, optical delay setup.

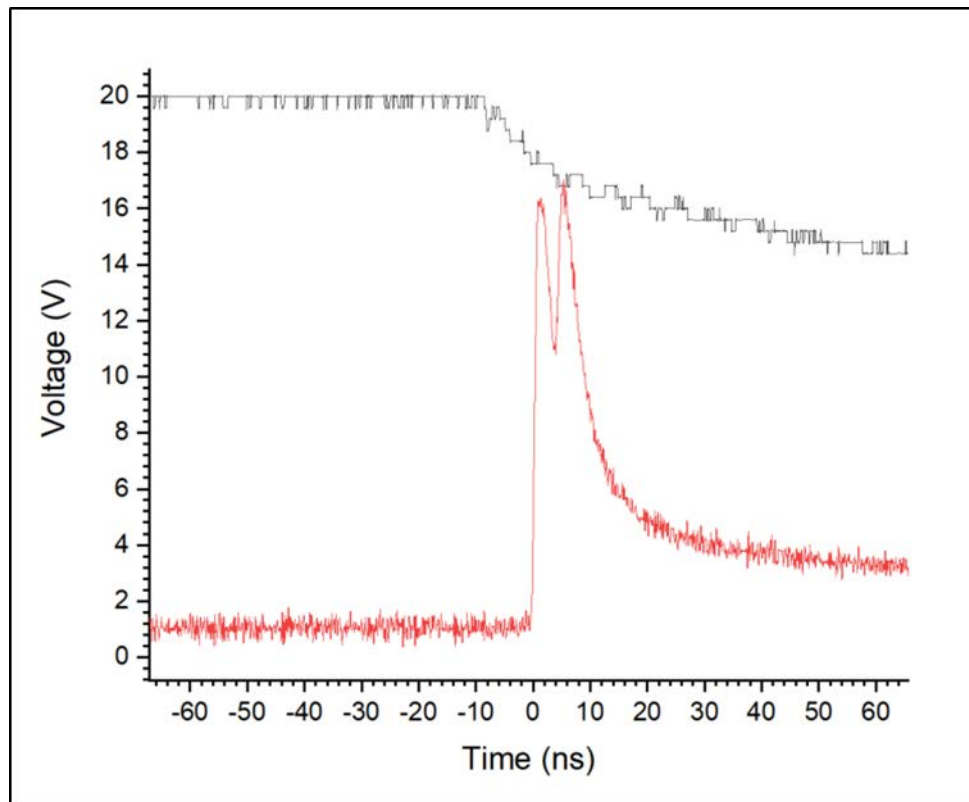


Figure 3.2: Waveforms from initial trials of direct unipolar RF modulation of GaN:C PCSS using an optical delay. The modulation frequency is approximately 200 MHz based on the peak-to-peak period.

3.1.2 Next Steps:

- i. Collection and analysis of on-state resistance measurements as a function of incident laser energy newer NRL devices.
- ii. Demonstrate RF unipolar modulation in the GHz regime with current PCB setup and bipolar modulation by using the PSOG based circuit and setup with GaN:C PCSS. Bipolar modulation will be demonstrated by incorporating a negative power supply and a second GaN:C-PCSS.

- iii. Show GaN:C PCSS high-voltage operation (ones of kV to start) using the pulse charger obtained from the HV switch development team.

3.3.2 GaN:C *Ab-initio* Calculations

Technical Computer Aided Design (TCAD) simulations are essential for the design of GaN (or alternative material) based high power photoconductive semiconductor switch (PCSS) devices. However, for the cases of carbon, iron, or other dopants in GaN the bulk material properties (conductivities, breakdown voltages, quantum efficiencies, etc.) needed for accurate TCAD simulation are not known. Similarly, the effects associated with mesoscale defects (e.g. dislocations, interfaces, etc.) derived from variations in fabrication of bulk materials and device integration are not typically incorporated into TCAD simulation. The Rulis group has been and continues to compute material properties via *ab initio* methods that can be used to provide necessary TCAD parameters and also to help verify the atomic structure of GaN as fabricated by collaborating partners. (Example parameters include the electron and hole (light and heavy) effective masses, the band gap, polarization parameters, bulk mechanical properties, etc.)

- i. At long last, we have now finally computed a band gap for pure GaN that is in agreement with experiment (3.49 eV computed vs. 3.5 eV experimental). Therefore, we will have confidence when applying the same computational method to C-, Fe-doped models of GaN for the computation of the band gap and effective masses. (To be done in the next month.)
- ii. As mentioned in the mini-summary above, there is some concern that the bulk GaN used in TCAD simulations does not account for realistic concentrations of mesoscale defects, nor for their influence on key properties such as electrical and thermal conductivity. Hence a student has spent some time to search for and begin evaluating the usefulness of mesoscale modeling tools that bridge the atomic and continuum scales. Specifically, atomman from NIST and a collection of tools exhibited on Nanohub. The student has also sought out classical interatomic potentials for GaN.
- iii. The parameters in the first bullet point are necessary inputs for TCAD simulations. The purpose of the activity in the second bullet point is to pursue a suspicion that bulk GaN as represented in the TCAD simulations and bulk GaN as fabricated may not be entirely consistent (particularly with respect to the interfacial properties and the influence of extended dislocation defects).

3.3.2.1 Next Steps

- i. We will continue now proceed to computation of the band gaps and band structures of C and Fe doped GaN.
- ii. We will continue to explore the role that mesoscale structure (i.e., extended defects such as dislocations and interfaces) plays on the bulk properties of GaN that are relevant for TCAD simulations.

3.4 Diode-based NLTLs

D-NLTLs garner interest as an HPM source due to their SWAP-C² capabilities. Currently three challenges present barriers to HPM production: (1) impedance mismatches due to a dynamic characteristic impedance (2) frequency generation limited to hundreds of MHz, and (3) material limitations on peak power generation. The D-NLTL development and optimization effort is focused on characterizing and overcoming each of these challenges by designing solutions to each.

3.4.1 Impedance matching

The amplitude dependent capacitance of a D-NLTL induces a signal dependent characteristic impedance ($Z(V)$) creating issues in energy insertion and extraction from the line. For D-NLTLs to be a viable HPM source, the issues presented by $Z(V)$ must be characterized and the effects mitigated.

- i. Three 20-C lumped-element D-NLTLs consisting of 100 nH, 47 nH, and 22 nH cell inductances were tested with resistive termination impedances of; 5, 10, 15, 20, 24.9, 30.1, and 50 Ω . It was expected that a line terminated with an impedance equal to Z_{avg} would experience a maximal amplitude with minimal reflections at the line-load interface.
- ii. Z_{avg} for each of the 3 lines were 25 Ω (L=100 nH), 16 Ω (L=47 nH), and 11 Ω (L=22nH). Results indicated success in impedance matching with minimal reflections occurring at $Z=24.9 \Omega$ for L=100 nH, $Z=16 \Omega$ for L = 47 nH, and $Z=11 \Omega$ for L=22 nH.

3.4.2 Frequency Generation

To be considered an HPM source D-NLTLs need to generate frequencies in the UHF range and at sufficient power when arrayed. Current frequency generation of D-NLTLs have been limited to tenths of a GHz in functioning prototypes. Frequency generation efforts are aimed at identifying and overcoming the limitations of D-NLTL frequency generation.

- i. Cell inductances were decreased to to 5 nH and 2.5 nH. In each case lines were terminated with a 5 Ω resistive load for reflection supression. Center frequency estimations of 125 MHz for the 5nH line and 128 MHz for the 2.5 nH line were measured.

3.4.3 Next Steps

- i. Search for new LV diode candidates capable of greater frequency generation to the UHF range required of a HPM source. Good diode candidates consist of junction capacitances of 200 pF or less and a decrease in capacitance of 1 order of magnitude over a voltage range of 5-10 V.

4.0 Prime Power and Conditioning

4.1 HV Switch Pulsed Chargers

The major objective of this initiative is to develop a low-cost, 10 kV, high-speed SiC MOSFET to be used with high-power microwave (HPM) generation circuit. Within the scope of this project, the pulsed power team at UMKC has designed a new switch architecture using multiple low-voltage MOSFETs to achieve 10 kV breakdown capacity. As of today, a commercially available but very expensive 15 kV switch is being used to generate high-voltage pulses. In contrast, the UMKC developed switch will take advantage of low-cost but commercially available SiC MOSFETs and configure them in a modular, multilevel structure to achieve greater than 10 kV voltage withstanding ability. Once completed, this part of the project will result in significant cost savings.

- This effort will achieve the following milestones pertinent to OSPRES grant mission:
- A new multilevel, modular architecture will be developed to achieve more than 10 kV breakdown voltage using multiple low-voltage MOSFETs
- A PhD student will be trained to design and fabricate high-frequency dc-dc converter circuit with 10 kV isolation, data TX/RX with optical fiber link and modular MOSFET architectures

4.1.1 *Design of a low-cost 10 kV gate driver circuit to fabricate a 10 kV SiC switch*

- i. We have fabricated multiple custom 10 kV rated but low-power (~ 5W) isolation transformers to facilitate galvanic isolation in the order of 10 kV for the power supply of the in-house gate driver.
 - a. Ungapped E-type ferrite core has been used (made in house) as the magnetic material of the transformer.
 - b. Silicone potting has been used to fill up the intentional airgap between two cores to increase the isolation rating of the transformer.
 - c. The secondary side of the transformer has been designed to accommodate three voltage rails (-20V, +5V, -5V) required to drive multiple ICs in the gate driver circuits.
- ii. We have designed a complete architecture of a custom in-house HV (~ 10 kV) gate driver in ALTIUM and fabricated it.
 - a. Optical transmitter and receiver has been used in the design to provide the required isolation through an optical link that is immune to interference and high-voltage surge for the PWM signal of the gate driver IC.
 - b. SMD components has been predominantly used in the design to minimize parasitic inductances.
- iii. Preliminary simulation of parallel transmission line coupled with PCSS and power-conditioning circuit has been performed with the addition of decoupling diode.
 - a. Initial results indicate decoupling between parallel transmission lines (microstrip type line) can be achieved using diodes with sufficient voltage rating.

- b. Literature study on Blumlein type transmission type line has been initiated to model this type of line along with PCSS and power-conditioning circuit to see the impact of triggering multiple PCSS, each coupled with one Blumlein line, in the output.

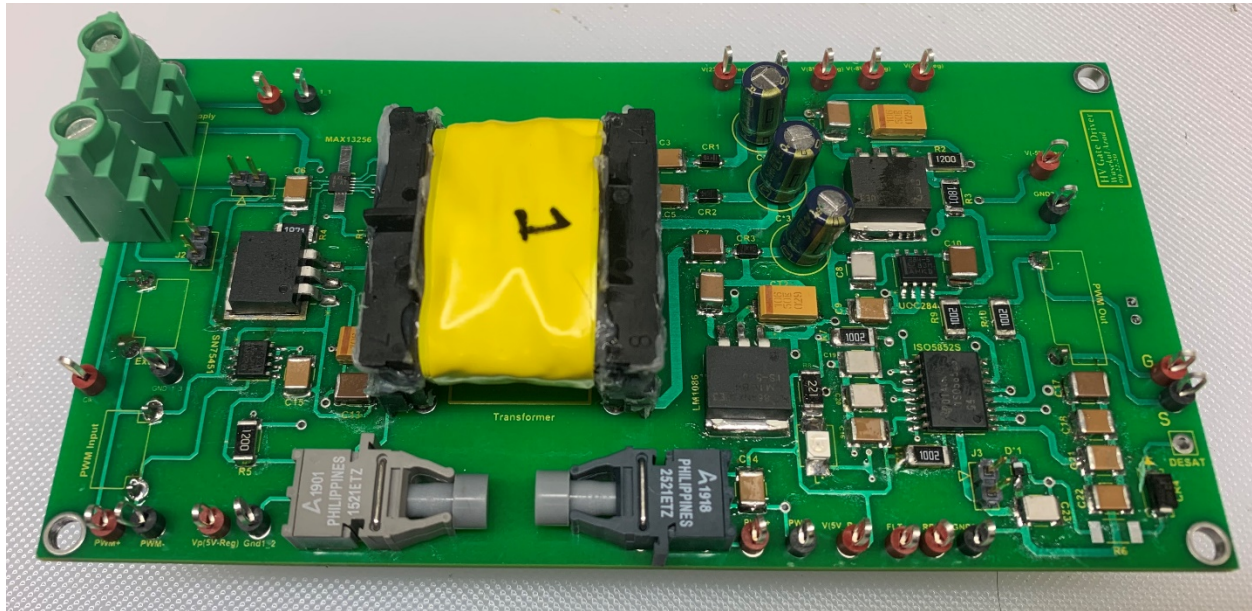


Figure 3.1. Prototype of the 10 kV gate driver circuit

4.1.2 Next Steps

The next step involves the testing of this gate driver circuit to achieve the necessary 10 kV isolation.

5.0 Antenna Trade Space

5.1 Trade Space Analysis of ESAs

The objective of this work is to achieve a specific power density (3 Watts/cm²) at a distance of 2 km using a planar array of Electrically Small Antenna (ESA) elements in the UHF region (0.6 – 1 GHz) and to enable electronic beam steering of short pulses. The total peak input power to the array is taken to be roughly around 800 MW.

5.1.1 Single Antenna Design:

- i. A *new* technique, to design wideband (> 30% 2:1 VSWR bandwidth) coaxial probe fed rectangular microstrip patch antennas, is used to model an antenna in the targeted UHF region (acquired bandwidth: 638.2 – 990.6 MHz). At this frequency we did not get any free samples from the vendor Rogers Corp., and the custom-made orders were of prohibitively high cost.
- ii. However, the proposed new design methodology was validated at higher frequencies 3.75 GHz (due to availability of free sample TMM-10i substrate panels), via measurements at Honeywell's Research Lab and using GTEM. The measured and simulated 2:1 VSWR bandwidths agreed very closely (~53.5%). The single element gains and radiation patterns were also in good agreement.

5.1.2 Array Design:

- iii. A 15 × 15 array is modeled in FEKO using the designed antenna at the UHF range – maximum boresight gain ≈ 27.5 dBi.
- iv. As reported earlier, steering of the main beam from boresight to ±60° resulted in scan blindness at a higher frequency of 990.6 MHz.

5.1.3 Reduced Surface Wave (RSW) Antennas:

- v. Investigations show pattern degradation and scan blindness at the higher frequency end occur due to excitation of TM₀ and TE₁ surface wave (SW) modes.
- vi. A method to choose substrate parameters (permittivity ϵ_r , thickness d) that minimizes SW excitations, at any specified frequency range, is currently being investigated.
- vii. Arrays designed on such substrates will have significantly reduced pattern degradation (at high frequencies) and scan blindness (which otherwise limits steerability).

5.1.4 Progress towards validation of array design technique:

- i. To validate basic array parameters viz., boresight gain, and radiation pattern, a small 3×3 array is modeled and characterized via FEKO at 3.75 GHz (due to availability of substrate of required dimension) - design sent for fabrication.

5.1.5 Next Steps:

- i. For a more thorough trade-space analysis, polyethylene ($\epsilon_r = 2.25$ and $\tan \delta = 0.00025$) and ceramic ($\epsilon_r = 9.6$ and $\tan \delta = 0.0001$) based substrates (for high power applications) are being explored for design antenna arrays at the UHF range to achieve better performance for pulsed HPM transmissions.

6.0 RF Coupling

The goal of this task is to improve our understanding of the mechanisms by which various RF waveforms couple to wires and electronics. Wires and electronics are typically enclosed by a dielectric plastic casing/enclosure, a metallic casing/enclosure, or something in between similar to casings/enclosures made of carbon fiber composites. These enclosures or casings, with the absence of any slots or openings, typically shield the enclosed electronics. However, the introduction of slots or gaps in the casings, for a variety of practical reasons, compromise this shielding and might even amplify the fields sensed by the enclosed electronics. Knowing the exact shielding/amplification provided by the casing/enclosure will help us better determine the desired power levels from the HPM sources to create a desired coupled voltage/current to the enclosed electronics.

In particular, a metallic enclosure, or a cavity with a slot, has a high-quality factor causing the penetrating waves to bounce back and forth, creating a standing wave. This standing wave can create large fields at the resonance frequencies of the cavity, i.e. the field inside the cavity can be higher than that outside it. Over the last month, we upgraded our analysis to UAV casing/frames with realistic shapes instead of the rectangular cavity studied in the previous reports. The 21 cm × 12 cm UAV shown in Fig. 1a is studied where different materials were assigned to the dielectric casing/enclosure. Four slots were added to the casing/enclosure where the motors typically go and these slots will be the optimum pathway for electromagnetic fields to penetrate to the enclosed electronics. We focus on the shielding effectiveness (SE) as defined in [1]:

$$SE = \frac{|E_{cavity}(r)|}{E_0}$$

where $E_{cavity}(r)$ is the field inside the frame at a probe position r and E_0 is the external incident field magnitude which will be fixed in this analysis as $E_0 = 1 \text{ V/m}$. Carbon fibers (CF) and Plexiglass were assigned as common materials to fabricate those UAVs and their response is compared to that of a perfect electric conductor, PEC, casing/enclosure. Fig 1(b) shows the SE in the center of the UAV and the following conclusions can be observed:

- i. PEC enclosure/casing shows the best SE as it acts as a faraday cage and shields the electronics inside the UAV at most frequencies. However, due to the slots at the motor locations, the fields can move through the hollow arms to the center of the casing/enclosure. Therefore, the PEC enclosure/casing provides the worst SE at the cavity resonance frequencies, e.g. 4.75 GHz, where the field can be amplified by 22 dB more than the incident field. The SE in Fig. 1 tells us that at 4.75 GHz we can lower the source requirement by 22 dB to excite a desired voltage/current in the enclosed electronics compared to electronics in free space. At 4.5 GHz, the source requirements need to be increased by 50 dB because the enclosure/casing will attenuate the incident field.
- ii. At frequencies other than the resonance frequency of 4.75 GHz, the studied CF enclosure/casing shows ~ 20 dB higher SE than PEC casings/enclosure. This means that it will attenuate the incident field by 20 dB less than a PEC

casing/enclosure due to its lower conductivity. At the resonance frequency of the cavity, CF shows a maximum SE of 2 dB. The SE analysis in Fig. 1 tells us that we can lower the source requirement by 2 dB at 4.75 GHz because the enclosure/casing will amplify the incident wave. At any other frequency, e.g. 4.5 GHz, the source requirements need to be increased by 30 dB, compared to electronics in free space, because the enclosure/casing will attenuate the incident field.

- iii. The dielectric casing/enclosure, plexiglass ($\epsilon_r = 2.6, \tan \delta = 0.0057$), amplify the incident over a wide frequency bandwidth. To validate this amplification, we simulated the SE using two independent computational techniques, FEKO and CST as shown in Fig. 1b. At 4.5 GHz, the SE of the plexiglass is 1.5 dB and it increases to 2.5 dB at 4.9 GHz. The SE analysis in Fig. 1 tells us that we can lower the source requirement by ~ 2-3 dB because the enclosure/casing will amplify the incident wave.

In the coming reports, we will focus more on studying: (i) the statistical variation of the fields at different locations inside the UAV enclosure/casing, (ii) how variations in the UAV enclosure/casings shape/size affect the SE, and (iii) different carbon fiber UAV enclosure/casings with different material properties over a wider frequency range. This analysis will help us update the source requirements to excite a certain voltage/current in enclosed electronics versus electronics in free space.

Reference:

- [1] S. Campione et al., "Modeling and Experiments of High-Quality Factor Cavity Shielding Effectiveness," 2019 International Applied Computational Electromagnetics Society Symposium (ACES), Miami, FL, USA, 2s019, pp. 1-2.

7.0 Thermal Management

7.1 Jet Impingement Optimization

This September, the Thermal Management team focused efforts towards a third iteration of the jet impingement thermal management system (JI-TMS). Before doing so, we took a step back to broadly analyze what our team's objective was as part of the overall OSPRES mission, what progress we have made since the last OSPRES Annual Review to meet that objective, and really the alignment and misalignment between the two. Our objective is simple, develop the novel cooling technology, jet impingement, from a SWaP-C perspective that facilitates the maximal and optimal operation of technologies enabling modular, scalable, electronically-steered HPM arrays.

Until now, we have devoted our efforts on (1) the original JI-TMS's design, (2) simulating and experimentally evaluating its SWaP-C characteristics, and (3) pinpointing specific design limitations that otherwise inhibit the system from achieving its full potential as a viable option to increase the cooling capabilities of OSPRES HPM technologies beyond that of traditional methods. The last of which included utilizing the PCSS-based HPM source's cross-flow thermal management system as a reference and baseline for SWaP-C comparison.

When analyzing the progress and performance of the current JI-TMS design through a multitude of simulations and experiments, we realized that no small change in the current design's geometry would ultimately unlock the full potential of jet impingement as a novel cooling technology utilized in HPM-based sources; PCSS-based or otherwise. This was attributed to inadequacies within the current design to provide sufficient coolant flowrates necessary to induce turbulence (a fundamental aspect of jet impingement cooling) without generating a large pressure drop across the fluid's inlet and outlet of the JI-TMS. Within past MSRs, we have portrayed the efforts and steps we have taken to overcome this issue by investigating multi-outlet design revisions to improve the flowrate capacity and the accompanying head loss (pressure drop) reduction in doing so. As viable as this approach may have seemed to us at the time— we still were not truly achieving the conditions required to (1) induce turbulence, (2) generate a large convective heat transfer coefficient between the coolant and the heat source (device under load), and (3) capitalize on jet impingement as a cooling technology to further HPM-based Source's capabilities (e.g., higher average powers) beyond that of traditional cooling methods.

Building on the lessons we have learned from the simulation- and experimental-based efforts on the previous JI-TMS design and iterations thereof, we are in the process of finalizing a complete overhaul of the JI-TMS's design which is discussed in the following section.

7.1.1 Optimal JI-TMS Design Review

This subsection provides a brief on the ongoing JI-TMS design overhaul effort. The following bullet points outline the steps taken, lessons learned and employed, and pinpoints the JI-TMS's redesign requirements to fulfil the OSPRES mission. As a visual aid to accompany the following bullet point, has been generated.

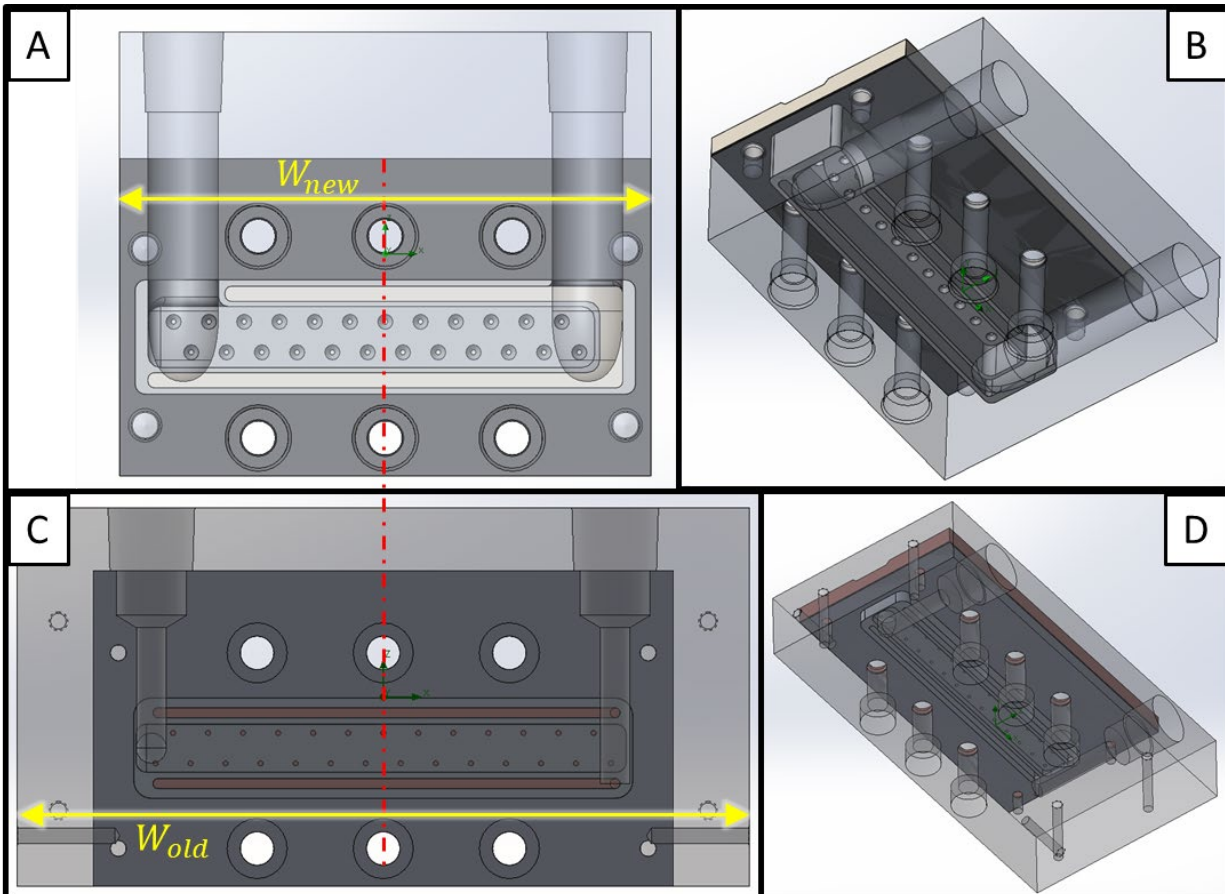


Figure 6.1 CAD model visualization aid to compare the new JI-TMS design's geometry to the original version with (A) a bottom view of the new JI-TMS where the reduced width and increased inlet and outlet diameters are evident, (B) an isometric view of the new design highlighting the jet manifold's increased height, (C) the original design having a greater overall width and smaller inlet and outlet diameters, and (D) the isometric view of the original JI-TMS with much smaller jet manifold geometry.

- i. Utilizing the PCSS-based HPM Source as a reference point, decrease the overall footprint of the JI-TMS aluminum cooling block (i.e., the cradle) which houses the flow channel that supplies fluid into and out of the jet impingement manifold.
 - a. From a space (SWaP-C) perspective– the original design possessed an overall width of 116 mm. The revised cradle's width has been reduced to 83.65 mm; a nearly 28% decrease from the previous design.
 - b. This is visualized in Figure 6.1. when comparing (A) vs. (C).

- ii. Increase flowrate and reduce pressure drop to achieve a turbulent flow regime between the jet-nozzle's exits and the surface above from which heat is generate.
 - a. By increasing the coolant's inlet and outlet diameter, a larger flowrate can be supplied to the jet impingement manifold while reducing the pressure drop across the flow path.
 - b. The geometry accommodations should:
 - i. Achieve the desired minimum fluid flowrate of 7.8 gpm
 - ii. Result in the production of turbulent vorticities
 - iii. Yield a convective heat transfer of $\sim 40 \text{ kW}/(\text{m}\cdot\text{K})$
 - iv. Allow for $>1 \text{ kW}$ of average power handling capabilities while keeping the heat source (e.g., PCSS, GaN, DSRD) below 80°C . *Note, these values are derived from theory and are ideal as well as based on the revised JI-TMS design's geometry nor account for the pressure drop associated with these flow-rates.*
 - c. This is visualized in Figure 6.1

7.1.2 Next Steps

- i. Utilize the optimal turbulence models and other simulation parameters deduced from past iteration's analysis to numerically analyze the redesigned JI-TMS parametrically.
- ii. Experimentally verify the simulation results by generating a new prototype of the JI-TMS and testing its performance capabilities.
- iii. Generate an operational-space model to understand and predict how specific HPM-based Source's waveform parameters effect the transient time-to-overheat (time in which a device-under-load exceeds its maximum operating temperature).

8.0 Program Management

8.1 Issues

- i. Scope – N/A
- ii. Budget – N/A
- iii. Schedule – N/A

8.2 Interactions with Others

Agency/Org	Performer	Project Name	Purpose of Research/ Collaboration

8.3 Major Procurement Actions

- i. N/A

8.4 Travel

Destination	Purpose	Attendees	Estimated Costs

8.5 Pending or Upcoming Public Release Requests

- i. N/A