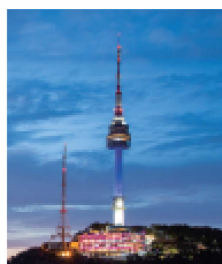




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Dual-frequency A8 Magnetron with ACAE-based TE41-TM01 Mode Converter

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Abstract– Magnetrons’ applications are limited by their single-frequency output characteristic. To address this limitation, this paper proposes a dual-frequency A8 magnetron system with a TM01 mode output, based on the All-cavity Axial Extraction (ACAE) structure, employing two magnetrons. A deliberate impedance mismatching structure is applied to ensure high efficiency for each magnetron. The performance measured by CST-PS demonstrates that the magnetrons operate at 4.8 GHz and 5.8 GHz, achieving peak efficiencies of 61% and 59%, respectively.

Keywords– All-cavity Axial Extraction (ACAE), CST Particle Studio, dual-frequency, high-power microwave, magnetron

I. INTRODUCTION

Due to their exceptional power-conversion efficiency and low-profile characteristics, magnetrons have established themselves as a premier choice for high-power microwave (HPM) sources. However, their single frequency output characteristic limits their application in systems that require various frequency bands. Therefore, we propose a dual-frequency system that generates the TM01 mode output by connecting magnetrons in series. To achieve high system efficiency, the concept of impedance mismatching is applied to the ACAE structure[1]. The target frequencies of the magnetrons are set to 4.8 GHz and 5.8 GHz, and the proposed design was validated using CST Particle Studio.

II. DESIGN OF DUAL-FREQUENCY MAGNETRON SYSTEM

The system consists of two magnetrons which have eight cavities. Although increasing the number of cavities improves efficiency, it also intensifies mode competition. Therefore, the A8 magnetron was selected over the A6 or A10 models. Mag1 operates at 4.8 GHz and Mag2 operates at 5.8 GHz. To ensure that both magnetrons have identical input and output powers, their cathodes were designed with the same surface area. Furthermore, because a single operating point must be applied, the interaction regions were designed to be identical in size to ensure the alignment of their operating regions.

In the proposed extraction structure, the power couples to the slots and propagates axially. The slot length, which significantly affects the efficiency and resonant frequency of the magnetron, was tuned to $\lambda/4$. The separator and end-cap inserted between the two magnetrons decouple each magnetron and prevent axial leakage of RF waves. The distance from the slot center of Mag1 to the waveguide short (D_1) was designed to be $\lambda_{g1}/4$, and the distance from the slot center of Mag2 to the waveguide short was designed to be $5\lambda_{g2}/4$. Because the waveguide is composed of PEC, the impedance of the waveguide short observed from the slot center of each magnetron becomes an open circuit. Consequently, the λ_{g1} , λ_{g2} components are totally reflected at the waveguide short and

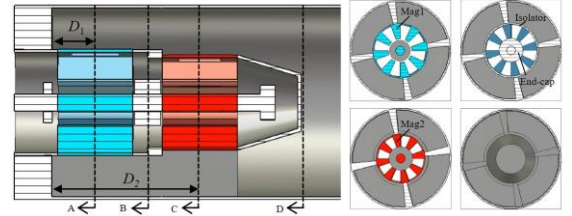


Fig. 1. Proposed dual-frequency magnetron system. (a) Axial cross-section. (b) Radial cross-section.

propagate toward the output waveguide. Here, λ_{g1} , λ_{g2} denote the wavelengths of the 4.8 GHz and 5.8 GHz components, respectively. Fig. 1 illustrates a cross-sectional view of the designed magnetrons.

III. SIMULATION RESULT

Simulation results obtained via CST-PS verified that the magnetrons are tuned to 4.8 GHz and 5.8 GHz. Furthermore, it was confirmed that the maximum efficiencies were achieved when D_1 and D_2 are $\lambda_{g1}/4$ and $5\lambda_{g2}/4$, respectively, resulting in efficiencies of 61% and 59% for Mag1 and Mag2. Table I compares the performance of the proposed structure with previously reported dual-frequency structures, demonstrating that the efficiency and rise time of the proposed structure are superior to previous studies.

TABLE I. SYSTEM PERFORMANCE COMPARISON

SYSTEM	RISE TIME OF MAG1	RISE TIME OF MAG2	EFFICIENCY OF MAG1	EFFICIENCY OF MAG2
[2]	13 ns	18 ns	50%	50%
PROPOSED SYSTEM	10 ns (-23%)	12 ns (-33%)	61% (+11%p)	59% (+9%p)

IV. CONCLUSION

In this paper, a dual-frequency magnetron system utilizing the ACAE-based extraction structure is proposed. To achieve high efficiency, a deliberate impedance mismatching structure was employed. Simulation results demonstrated that the magnetrons operating at 4.8 GHz and 5.8 GHz achieved efficiencies of 61% and 59%, respectively. We anticipate that the proposed design method can be extended to realize a multi-frequency system by axially cascading multiple magnetrons.

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Compensation Inductor Optimization for Reducing Third-Harmonic Leakage Magnetic Field in WPT

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Abstract– This paper proposes a method to reduce third-harmonic current by optimizing the transmitter-side compensation inductor, without additional shielding coils or filters. At the third-harmonic, the output-to-input voltage ratio is utilized to induce cancellation of the leakage magnetic field. Simulation and experimental results verify that the proposed method effectively reduces third-harmonic current and leakage magnetic fields.

Keywords– Wireless power transfer, third-harmonic leakage magnetic field, electromagnetic interference

I. INTRODUCTION

Wireless power transfer (WPT) is widely used for charging electric vehicles (EVs) and automated guided vehicles (AGVs). Because power is transferred via magnetic coupling between transmitter (TX) and receiver (RX) coils, WPT generates leakage magnetic fields that can cause Electromagnetic Interference (EMI) and interfere with nearby electronics [1]. Prior mitigation approaches such as shielding coils and coil-structure optimization have been proposed, but they increase system volume and weight and mainly target fundamental-frequency performance. Therefore, this study reduces third-harmonic current and leakage magnetic fields by optimizing the TX-side compensation inductor in an LCC-S topology.

II. METHODOLOGY

To reduce harmonics, the WPT circuit topology was selected as LCC-S, as shown in Fig. 1(a). At the fundamental frequency, the TX and RX currents have a 90° phase difference. In the harmonic frequency, under the assumption that the harmonic induced voltages on the TX and RX sides can be neglected, Kirchhoff's voltage law indicates that the TX and RX currents have a 180° phase difference and that the input and output voltages are in phase. By applying the condition that the magnitude of the sum of the TX and RX currents becomes zero, the input-output voltage ratio in (1) can be obtained. Although L_{TX} and L_{RX} are already fixed by design, V_{out}/V_{in} can be adjusted using L_s . In addition, the input-output voltage relationship can be derived from the induced voltage, which leads to (2).

$$\frac{V_{out}}{V_{in}} = \frac{(n^2-1)L_{RX}}{(n^2-1)^2 L_{TX} - L_s} \quad (1)$$

$$\frac{V_{out}}{V_{in}} = \omega_o^2 M C_p = \frac{M}{L_s}, \quad C_p = \frac{1}{\omega_o^2 L_s} \quad (2)$$

$$L_s = \frac{M(n^2-1)^2 L_{TX}}{M + (n^2-1)L_{RX}} \quad (3)$$

Using (1) and (2), the expression for the optimized L_s in the harmonic domain is given in (3). To verify the proposed technique, circuit simulations were performed under identical input-output conditions to extract the magnitude and phase of the third-harmonic current. These values were then used as inputs to the finite-element method simulation shown in Fig. 1(b) to confirm the reduction effect of the leakage magnetic field in the WPT coils.

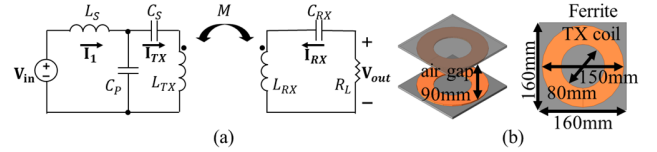


Fig. 1. WPT coil system. (a) LCC-S topology. (b) Simulation setup.

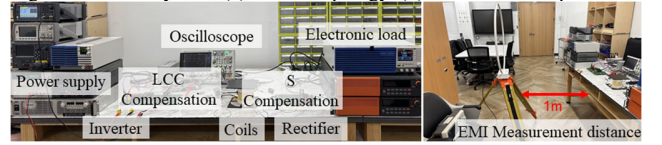


Fig. 2. Experimental setup used to verify the effectiveness of the proposed L_s for third-harmonic reduction.

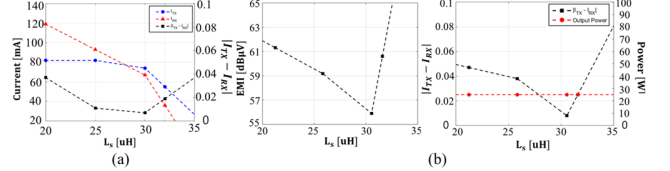


Fig. 3. Simulation and experimental results as a function of L_s . (a) Simulated third-harmonic currents of the TX and RX coils and their magnitude difference. (b) Measured EMI, the magnitude of the TX and RX coil currents at the third harmonic, and output power.

III. RESULTS AND DISCUSSION

By substituting the fixed inductance values of the TX and RX coils into the L_s optimized for third-harmonic reduction, as in (3), a value of 30.54 μH is obtained. After satisfying the resonance condition of the circuit with these values, the third-harmonic current obtained through circuit simulation is applied to the FEM simulation to compare the proposed value with each case and to confirm whether effective reduction is achieved. The experiment was conducted as shown in Fig. 2. As shown in Fig. 3, it was confirmed that, at the third harmonic, compared with the comparison cases, the proposed value reduced the magnitude of the TX and RX coil currents by 20 mA and EMI by 3.27 dB μV .

IV. CONCLUSION

Optimizing L_s in an LCC-S WPT system provides a compact circuit-level solution to suppress third-harmonic current and leakage magnetic field without additional shielding or filters, while maintaining fundamental-frequency power-transfer performance.

ACKNOWLEDGMENT

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Experimental setup for High Power Microwave effect on a target using ground reflection effect

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Abstract– This paper proposes an experimental setup for analyzing the effects of high-power microwave exposure on a target by leveraging ground reflection effect of electromagnetic waves to control the electric field strength. The proposed setup is simple to construct using a high-power microwave system and sensors, and it enables systematic variation of the electric field strength from 11.52 kV/m to 50.19 kV/m by adjusting the distance to the target at a fixed height within the test range.

Keywords– High power microwave (HPM), ground reflection effect

I. INTRODUCTION

Accurate test facilities are required to analyze the effects of high-power microwave (HPM) exposure on targets by controlling the electric field (E-field) strength as a function of frequency, as demonstrated by systems such as the Swedish MTF, the British Orion, the French Hyperion, and the German Supra [1]. However, these facilities are costly and time-consuming to construct.

This paper proposes a simple and cost-effective experimental setup for analyzing HPM effects on a target, consisting of an HPM system, a D-dot sensor, and an open-ended waveguide (OEG) antenna. The proposed approach leverages ground reflection effect of electromagnetic waves to characterize E-field variations with respect to distance under both vertical and horizontal polarizations. Based on this characterization, an appropriate E-field range for HPM testing is established, and the target response is systematically evaluated by varying the distance from the antenna aperture.

II. EXPERIMENTAL SETUP

Fig. 1 illustrates the simulated variation of the E-field strength as a function of distance when 500 MW of input power (P_{in}) is applied to an antenna positioned 1.62m above a ground (GND) plane modeled as a perfect electric conductor (PEC). Due to ground reflection effect, the E-field variation above the ground is more pronounced than in free space. Furthermore, beyond a distance of 26.5 m, the horizontally polarized field exhibits a more gradual decay and a wider dynamic range compared to the vertically polarized field. Based on this analysis, the test range is selected from 26.5 m to 38 m, corresponding to an E-field strength range of 11.52 kV/m to 50.19 kV/m.

Fig. 2 presents the proposed experimental setup for evaluating HPM effects on a drone target. A 90 dB waveguide directional coupler (WGDC) is inserted between the HPM source and the transmitting antenna to monitor the input power. An OEG antenna, positioned 40 m from the transmitting antenna aperture, is used as a reference sensor. The E-field distribution is measured by translating a D-dot sensor along the test range. The measured data are cross-validated with the WGDC output and the OEG antenna measurements to ensure consistency in the E-field characterization.

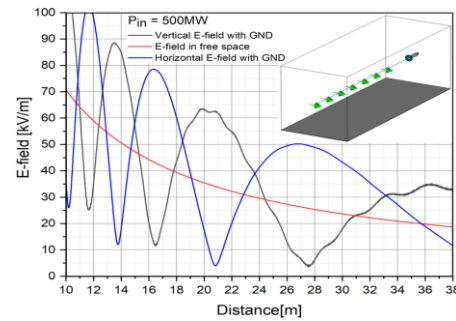


Fig. 1. The simulated E-field strength variation with the distance



Fig. 2. Experimental setup for HPM effect on a target

III. CONCLUSION

This paper presents a simple experimental setup for analyzing the effects of high-power microwave exposure on a target by exploiting ground reflection effect of the electromagnetic waves to control the E-field strength. The proposed approach enables systematic measurement of target responses by varying the distance from the antenna aperture at a fixed height. Furthermore, flexible E-field control can be achieved by adjusting the target height while maintaining a constant distance. The proposed setup provides a practical and cost-effective alternative to conventional large-scale HPM test facilities, enabling controlled E-field variation for target-level evaluation.

ACKNOWLEDGMENT

This work was supported by the Agency for Defense Development.

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Investigation of Ringing Suppression in Inverters Using Capacitors Connected to MOSFET Terminals

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Abstract– This paper analyzes ringing and electromagnetic interference (EMI) in a full-bridge inverter caused by PCB parasitic components. The impact of parasitic elements on switching behavior is investigated, and the application of additional capacitance is proposed to suppress voltage overshoot and ringing. Simulation results validate the effectiveness of the proposed approach.

Keywords– Ringing, Electromagnetic interference (EMI), Full-bridge inverter, Switching noise, Parasitic components

I. INTRODUCTION

As power switching devices continue to evolve toward higher power and higher switching frequency operation, switching noise, including electromagnetic interference (EMI), becomes increasingly significant, as it may lead to device failure, excessive thermal stress, and acoustic noise. Therefore, effective suppression of ringing in full-bridge inverters is of critical importance [1, 2]. In practical implementations, printed circuit board (PCB) layouts inherently include unintended parasitic inductances and capacitances, which can introduce ringing phenomena.

In this paper, a detailed analysis of the PCB circuit is conducted, and component placement strategies aimed at reducing parasitic effects for EMI suppression are proposed. The effectiveness of the proposed approach is verified through experimental results.

II. METHODOLOGY

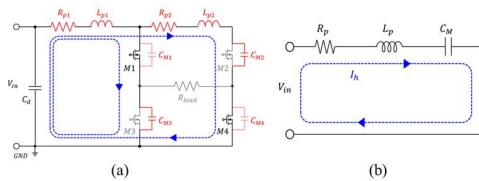


Fig. 1. Circuit of full bridge inverter. (a) Current path of full bridge inverter. (b) Equivalent circuit.

To analyze the full-bridge inverter, circuit analysis is conducted. Fig. 1(a) illustrates the current paths according to the operating mode. When the FETs are turned off, the current flows through the parasitic capacitances of the FETs. Furthermore, assuming that the FETs in the on state behave as short circuits, the simplified circuit representing a single current path can be described as shown in Fig. 1(b). As a result, the relationship can be expressed as in (1).

$$\frac{d^2 v_c}{dt^2} + \frac{R}{L} \frac{dv_c}{dt} + \frac{1}{LC} v_c(t) = \frac{1}{LC} V_{in} \quad (1)$$

, where the V_{in} , v_c , R , L , C represent the input voltage, the capacitor voltage resistance, inductance and capacitance respectively. Furthermore, considering that $\omega_0 = 1/\sqrt{LC}$, and $\zeta = R/2L\omega_0$, (1) can be rewritten as (2).

$$\ddot{v}_c + 2\zeta\omega_0 \dot{v}_c + \omega_0^2 v_c = \omega_0^2 V_{in} \quad (2)$$

Since (2) is a second order of differential equation, ringing occurs when the damping ratio ζ is lower than 1, corresponding to underdamped condition. Considering that the $\zeta = \frac{R}{2} \sqrt{\frac{C}{L}}$, increasing capacitance results in a higher damping ratio. Therefore, additional capacitance is applied near the MOSFET to increase the ζ , thereby reducing ringing and voltage overshoot.

III. RESULTS AND DISCUSSION

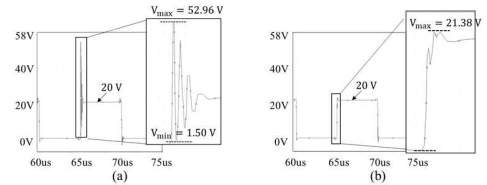


Fig. 2. Simulation results depending on the additionally applied capacitor. (a) Control case and (b) two 100 μ F are applied in parallel to the MOSFETs.

For the simulation, the parasitic resistance and inductance of the PCB traces are extracted using the EMCoS PEEC Solver, resulting in values of 5.5 m Ω and 66 nH, respectively. In addition, to consider practical electric device characteristics, SPICE model of the SPW47N60CFD is employed. In this research, two cases are considered. One is used as a control condition, and the other case applied two 100 μ F capacitors are additionally connected in parallel between the power and ground nodes. Fig. 2 shows the transient response for the different inductance conditions. As illustrated in Fig. 2(a), the overshoot voltage reaches nearly twice the desired voltage. In contrast, in Fig. 2(b), where the two additional capacitors applied case, the overshoot voltage is significantly reduced. As a result, it is confirmed that ringing can be effectively suppressed by applying additional capacitance.

IV. CONCLUSION

In this paper, the ringing phenomenon in an inverter is theoretically analyzed using a ringed equivalent circuit model. To investigate the suppression of voltage overshoot, additional capacitance is examined through simulation. The results demonstrate that reducing parasitic inductance effectively suppresses ringing at the FET terminals, thereby improving the reliability of inverter operation.

ACKNOWLEDGMENT

The authors would like to thank Huwin for their technical support and helpful discussions.

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Electrode-Geometry Optimization of ZnO Varistor Elements for EMP Protection Filters

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Abstract– This paper presents a compact design study of a ZnO varistor element for EMP protection filters, focusing on electrode-to-edge clearance as a key factor governing edge overstress. Coupled electro-thermal analysis was performed for silver-electrode clearances of 0.5 mm, 1.0 mm, and 1.5 mm under a 50 kA, 10/350 μ s impulse. Although the peak surface current density remained nearly unchanged among the three cases, thermal and electric-field concentrations showed clear differences. The 0.5 mm design exhibited the lowest hotspot temperature and peak electric field, while the 1.5 mm case showed the highest local overstress, implying a greater puncture risk. Impulse tests supported the analysis: improper geometry caused puncture near the terminal edge, whereas zero clearance led to flashover. The optimized 0.5 mm structure maintained stable operation up to 50 kA with a limiting voltage below 2 kV. These results show that a controlled minimum clearance, together with terminal-edge shaping, is an effective design strategy for robust varistor integration in EMP protection filters.

Keywords– EMP protection filter, varistor, ZnO, electrode geometry, puncture, surge protection

I. INTRODUCTION

EMP protection filters require nonlinear clamping elements that can survive high-energy transients while maintaining stable limiting performance. In practical ZnO varistors, failure is often initiated not in the ceramic bulk but near the terminal edge, where current spreading becomes nonuniform and local thermal and electrical stress are intensified. This issue is especially important in compact filters with limited geometric margin. In this work, electrode-to-edge clearance is treated as a primary design parameter, and its effect on current density, hotspot formation, electric-field concentration, and surge endurance is examined.

II. DESIGN CONCEPT AND ANALYSIS METHOD

A ZnO varistor with screen-printed silver electrodes and attached copper terminals was analyzed as given in Fig. 1. The main variable was the clearance between the electrode edge and ceramic sidewall: 0.5 mm, 1.0 mm, and 1.5 mm. A coupled electro-thermal simulation was performed under a 50 kA, 10/350 μ s impulse to evaluate current density, temperature rise, and electric-field stress near the terminal edge. Impulse experiments were also conducted to verify failure behavior and limiting performance.

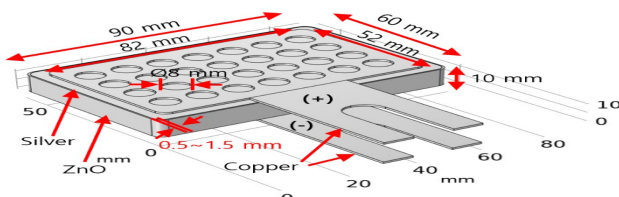


Fig. 1. Schematic view of simulated ZnO Varistor

III. RESULTS AND DISCUSSION

In all cases, the terminal edge was the most vulnerable region because broadband impulse components caused current crowding near the outer terminal boundary as given in Fig. 2. The peak surface current density was similar in all cases, indicating that current-density magnitude alone did not determine design margin. In contrast, thermal and electric-field stress strongly depended on clearance. The 1.5 mm case showed the highest hotspot temperature and electric field, whereas the 0.5 mm case gave the lowest values, demonstrating improved current spreading and reduced local overstress. Experimental results supported this trend: improper electrode geometry led to puncture near the terminal edge, while zero clearance caused flashover between the upper and lower electrodes. The 0.5 mm structure showed stable operation up to 50 kA with a limiting voltage below 2 kV.

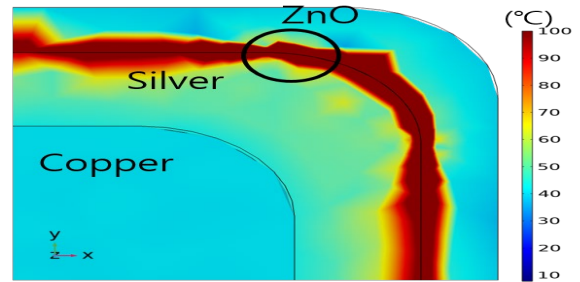


Fig. 2. Surface temperature distribution of ZnO Varistor with clearance area of 1.5 mm

IV. CONCLUSION

This study showed that electrode geometry significantly affects edge-initiated failure in ZnO varistors for EMP protection filters. Among the examined cases, a 0.5 mm clearance provided the best balance between current spreading and insulation margin, minimizing thermal and electrical overstress and enabling stable high-current operation. The results indicate that electrode clearance and terminal-edge treatment should be considered key design parameters for improving the surge reliability of varistor-based EMP protection filters.

ACKNOWLEDGMENT

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Development of a -3 kV IEMI Pulse Generator with a 20 kHz Pulse Repetition Rate

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Abstract We designed and fabricated a 20 kHz MOSFET Marx generator, driven by a single-gate driver architecture, which proved to work successfully at the level of -3 kV for IEMI test application.

Keywords– IEMI, SGMG, SSMG, MOSFET, PRR, Single gate driver.

I. INTRODUCTION

Securing reliability verification technologies against intentional electromagnetic interference (IEMI) is essential in future mobility environments. Conventional spark-gap based marx generator (SGMG) are inherently constrained in high-repetition pulse generation, as their pulse repetition rate (PRR) is limited to a few kilohertz or less due to the plasma deionization time [1], [2]. We designed a MOSFET-based solid-state based marx generator (SSMG) capable of generating a PRR of 20 kHz and -3 kV to be utilized for verifying the impact of IEMI. The proposed circuit introduced a topology that controlled multi-stage switches through a single gate driver [3].

II. OPERATING PRINCIPLE

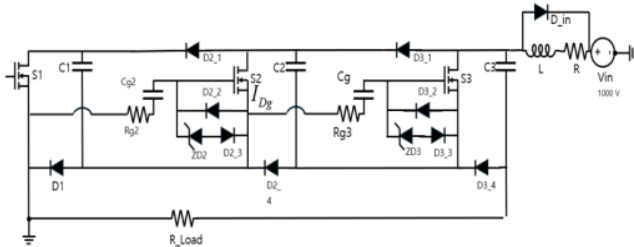


Fig. 1. Schematic of the 3-stage MARX generator

The circuit operation is categorized into two primary phases. All switches are in the OFF state. The upper and lower diodes are forward-biased, and the capacitors at each stage are charged to the input voltage. When the switches are turned ON, the potential of each stage capacitor shifts, causing the upper and lower diodes to become reverse-biased. All capacitors are connected in series to generate the output voltage. Where k represents the voltage utilization factor, and n denotes the number of marx stages.

$$V_{out} = knV_{in} \quad (1)$$

The circuit employs a single active gate driver combined with stage-wise passive clamping networks. When a trigger pulse is applied to S1, the Zener diode in the subsequent stages exploits the transient voltage drop to passively trigger the switches, instantly clamping the gate-source voltage (V_{GS}) to a safe level to protect the gate oxide. This self-synchronized architecture enables the stable driving of multi-stage MOSFETs strictly through passive components, completely eliminating the propagation delays and bulky isolated power supplies associated with conventional active drivers, thereby facilitating seamless stage extension.

III. EXPERIMENTAL RESULTS

To verify the performance of the proposed three-stage MOSFET Marx generator, a prototype was fabricated and tested. The experimental measurements were compared with SPICE simulation results to evaluate the accuracy of the design model in achieving a target output of -3 kV at PRR of 20 kHz ($T=50\mu s$).

Figure 2 presents a comparative analysis of the simulated and experimentally measured output voltage waveforms. The experimental results validate that the proposed generator produced a pulse of -3 kV, corresponding to triple the applied input voltage. The rise time is approximately 110 ns. The transient timing characteristics of the pulse including the pulse width and the slopes of the rising and falling edges demonstrated consistency with the simulation model. This correlation substantiated that the adopted single-gate driving mechanism achieved synchronization in actuating the MOSFET switches across all stages. Future studies will focus on tuning critical circuit parameters to further optimize the quality of the final output waveforms.

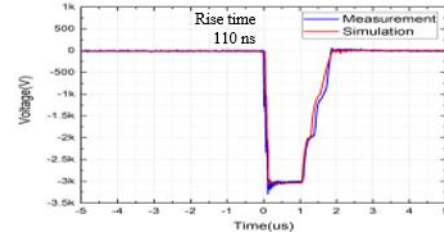


Fig. 2. Output pulse waveform

IV. CONCLUSION

A three-stage MOSFET Marx generator, driven by a single-gate architecture, was developed for IEMI pulse generation. Experimental results confirmed the design's effectiveness, showing consistency between simulation and measured data. Through the selection of optimized component values, a voltage output of -3 kV was demonstrated at a PRR of 20 kHz. These results suggest that the proposed architecture provides an IEMI test solution for high-voltage, high-repetition rate IEMI susceptibility and reliability verification fields. Future research will focus on increasing the number of stages to achieve higher voltage outputs at a PRR of 20 kHz.

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Four-Layered Electromagnetic Defense Framework for Critical Military Facilities Against Drone Swarm Attacks

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The proliferation of low-cost commercial drones and coordinated swarm tactics poses unprecedented asymmetric threats to military critical facilities worldwide. Recent conflicts in Ukraine and the Middle East have demonstrated that conventional single-layer defense systems are insufficient against mass drone attacks employing saturation tactics. This paper proposes an innovative four-layered electromagnetic (EM) defense framework integrating: (1) Detection and Deception using GPS spoofing to redirect hostile drones into designated kill zones, (2) Soft-Kill through selective inter-drone communication link jamming to disintegrate swarm coordination, (3) EM Fence establishing persistent electromagnetic barriers around facility perimeters, and (4) Hard-Kill employing High-Power Microwave (HPM) and directed energy weapons for final neutralization. This progressive layered approach maximizes cost-effectiveness by filtering threats at each stage. International case studies from Israel's multi-layered defense system, Russia's GPS spoofing operations, and UK's SkyFence validate the proposed concepts.

Keywords– Counter-UAS, drone swarm, GPS spoofing, soft-kill, hard-kill, EM fence, HPM, critical infrastructure protection, electromagnetic warfare

I. INTRODUCTION

The rapid advancement of drone technology and proliferation of low-cost commercial UAVs have created new asymmetric threats to military critical facilities. In Ukraine, Russian forces have been neutralizing over 10,000 drones monthly through electronic warfare jamming systems. During the 2025 Iran-Israel conflict, Israel successfully intercepted 98% of more than 1,000 drones using its multi-layered defense architecture combining Iron Dome, electronic warfare units (5114th Spectrum Warfare Battalion), and the newly deployed Iron Beam laser system [1]. These real-world cases conclusively demonstrate that single countermeasure approaches cannot effectively defeat coordinated drone swarm saturation attacks, necessitating an integrated electromagnetic protection framework with cost-effective staged response capabilities.

II. FOUR-LAYERED DEFENSE FRAMEWORK

A. Layer 1 - Detection and Deception: This layer combines RF/Radar-based detection with GPS spoofing technology to manipulate drone navigation systems. By broadcasting counterfeit GNSS signals, hostile drones are redirected to designated kill zones where they can be efficiently neutralized. Russia's Khmeimim Air Base in Syria successfully employed this technique against multiple U.S. and Israeli surveillance drone incursions, as documented by C4ADS and UT Austin researchers [2]. The estimated cost per engagement is approximately \$1.

B. Layer 2 - Soft-Kill: This layer employs selective jamming of C2 (command and control) links and inter-drone mesh network communications. Unlike conventional broadband jamming, targeting specific mesh network frequencies (typically 2.4/5.8 GHz) effectively disintegrates swarm coordination capabilities while minimizing interference to friendly systems. Ukraine's Doodle Labs Mesh Rider radios with anti-jam "Sense" technology have demonstrated the vulnerability of swarm communications to selective disruption [3].

C. Layer 3 - EM Fence: This innovative concept establishes persistent electromagnetic protection zones around critical facility perimeters. Any drone entering this zone is automatically neutralized through continuous EM field exposure. UK's SkyFence system, deployed for prison protection, provides a validated operational example of this approach [4].

D. Layer 4 - Hard-Kill: The final layer employs HPM (High-Power Microwave) systems and directed energy weapons (DEW) for complete drone destruction. Israel's Iron Beam 100kW laser system, operationally deployed in December 2025, achieves interception at only \$0.03 per shot compared to \$50,000 for conventional missile interceptors [5]. This revolutionary cost reduction makes sustained defense against mass drone attacks economically viable.

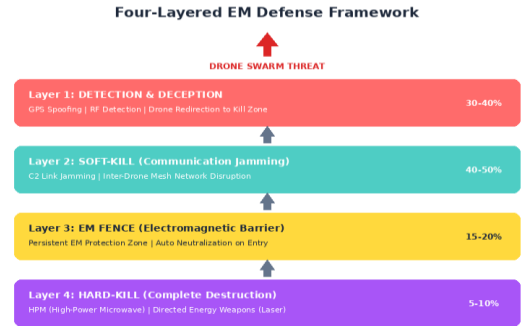


Fig. 1. Four-Layered EM Defense Framework Architecture

III. INTERNATIONAL CASE STUDIES AND VALIDATION

TABLE I. INTERNATIONAL COUNTER-UAS IMPLEMENTATIONS

Layer	Country/System	Cost	Neutralization Rate	Key Technology
1. Deception	Russia/Khmeimim	~\$1	30-40%	GNSS Spoofing
2. Soft-Kill	Ukraine/Mesh Rider	~\$10	40-50%	Mesh Net Jamming
3. EM Fence	UK/SkyFence	~\$100	15-20%	Perimeter EM Zone
4. Hard-Kill	Israel/Iron Beam	\$0.03	5-10%	100kW Laser DEW

IV. LEGAL AND OPERATIONAL FRAMEWORK

Peacetime deployment of this framework requires careful consideration of radio frequency regulations, rules of engagement (ROE), and friendly system interference minimization. Layers 1-2 (non-destructive measures) can potentially operate under automated authorization protocols. Layer 3 (EM Fence) requires geographically-limited operation zones with proper coordination. Layer 4 (Hard-Kill) necessitates explicit commander approval under current ROE frameworks. The framework design inherently aligns with international humanitarian law principles of proportionality and distinction by employing graduated response escalation.

V. CONCLUSION

This paper proposes a comprehensive four-layered electromagnetic defense framework: Detection/Deception → Soft-Kill → EM Fence → Hard-Kill. The progressive filtering approach achieves cumulative neutralization rates exceeding 95% while reducing operational costs by approximately 80% compared to conventional kinetic-only solutions. The framework integrates combat-proven concepts validated through recent operational experiences in Israel, Russia, Ukraine, and the United Kingdom. Future research will focus on simulation-based validation, system integration optimization, and field testing for Republic of Korea military facility applications.

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Online Impedance Measurement of Automotive Sensor Lines for Crosstalk Prediction

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Abstract– This paper presents an online impedance measurement method for automotive sensor power and signal lines and its application to crosstalk analysis. By introducing an external attenuator at the input of an impedance analyzer, the method enables measurements under DC-bias conditions without instrument saturation, allowing characterization under actual operating conditions. Experiments on various automotive sensor lines evaluated both power and signal line impedances, which were then used to analyze crosstalk. Comparison between predicted and measured results demonstrates that the proposed approach provides practical insight into crosstalk characteristics in complex automotive wiring systems.

Keywords– online impedance measurement, crosstalk, automotive sensors, sensor power line, signal line

I. INTRODUCTION

With the increasing complexity of automotive electronic systems, electromagnetic coupling between sensor and control signal lines has become an important EMC concern [1]. Previous online impedance measurement approaches mainly rely on current-probe-based non-intrusive techniques, which are limited in directly characterizing interface impedance [2]. In this study, an online impedance measurement method using an external attenuator is applied to determine the near-end and far-end impedances of 5-V automotive sensor power and signal lines under DC-bias conditions. The measured impedances are then used to assess and validate crosstalk behavior between threat and victim lines.

II. METHODOLOGY

Fig. 1. illustrates the online impedance measurement setup incorporating a 10-dB attenuator, which enables accurate characterization under DC-bias conditions. Impedances at both the controller (near-end) and load (far-end) sides were measured separately for power and signal lines.

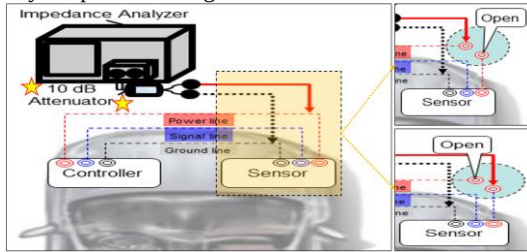


Fig. 1. Measurement setup for online impedance measurement using an external attenuator.

Fig. 2. illustrates the threat–victim crosstalk measurement configuration. An ignition coil was employed as the threat source, while sensor signal lines served as victims with a crosstalk length of 0.7 m. Voltages and currents of the threat source, as well as crosstalk voltages at the victim near-end (NE) and far-end (FE), were measured using appropriate voltage and current probes.

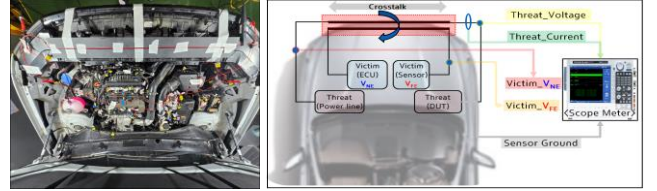


Fig. 2. Measurement setup for threat-victim crosstalk characterization.

III. RESULTS AND DISCUSSION

Fig. 3. presents the measured power-line and signal-line impedances, along with the resulting threat–victim crosstalk, obtained from the configurations shown in Fig. 1. and 2. Under powered conditions, the measured impedances exhibit lower values in the low-frequency region compared with the power-off state, reflecting the operational characteristics of the controller and sensor interfaces.

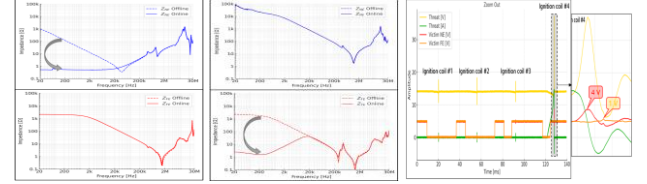


Fig. 3. Measured power-line impedance (left), signal-line impedance (center), and threat–victim crosstalk (right).

During the crosstalk experiment, the victim near-end voltage increased to approximately 4 V, while the far-end voltage increased to around 1 V, reflecting the relatively higher impedance at the controller side. The predicted crosstalk voltage, calculated using (1) with the measured impedance, showed good agreement with the experimental results [3].

$$V_{NE(FE)}(t) = \frac{Z_{NE(FE)}}{Z_{NE} + Z_{FE}} I_m L \frac{dI_{Max}(t)}{dt} + \frac{Z_{NE} Z_{FE}}{Z_{NE} + Z_{FE}} C_m L \frac{dV_{Max}(t)}{dt} \quad (1)$$

IV. CONCLUSION

An online impedance measurement approach employing an external attenuator was applied to characterize automotive sensor power and signal lines under DC-bias conditions. The measured near-end and far-end impedances were used to analyze threat–victim crosstalk, and the predicted crosstalk demonstrated good agreement with experimental results. The proposed approach provides a practical method for evaluating crosstalk characteristics in automotive sensor interfaces.

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Measurement of Shielding Effectiveness of Wire Mesh

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Abstract—This paper investigates the shielding effectiveness of wire meshes and wire mesh cages to develop low-cost electromagnetic pulse (EMP) protection measures for major facilities. Three wire mesh sizes (10 cm, 7.7 cm, and 5 cm) were tested in a shielded room and a 10m semi-anechoic chamber across a 10 kHz to 1 GHz frequency range. The measured results demonstrate Shielding Effectiveness (SE) values of 15–20 dB in the 10–150 MHz band, with higher effectiveness observed in the frequency band where nuclear EMP energy is concentrated.

Keywords—EMP, Shielding Effectiveness, Wire Mesh,

I. INTRODUCTION

High-power electromagnetic environments, such as Nuclear EMP (NEMP), can cause severe upset or damage to modern electronic systems. Despite the increasing threat of high-power electromagnetic waves due to global conflicts and technological advancements, most major facilities lack adequate EMP protection due to the heavy financial burden of traditional, high-cost shielding measures. The objective of this study is to analyze the shielding performance of wire meshes with various grid sizes to verify their shielding effectiveness and to propose a low-cost shielding measure.

II. METHODOLOGY

To compare the shielding effectiveness of the wire mesh, three grid sizes of 10 cm, 7.7 cm, and 5 cm were prepared. The shielding effectiveness of two different types, a flat wire mesh and a wire mesh cage, were respectively compared and measured. The wire mesh was installed and measured at the penetration panel of a shielded room, while the wire mesh cage was measured inside a 10m chamber. The joints of the wire mesh were reinforced with copper tape, and the cage was connected with a ground wire.

For the magnetic field band, a loop antenna was used. For the electric field band, a biconical and an LP antenna were used for transmission, and a ball probe antenna was used for reception. The wire mesh cage was evaluated by configuring the measurement environment with the same setup as the wire mesh inside the 10m chamber. A total of 380 measurement frequencies were selected with reference to the MIL-STD-188-125-1 standard.

III. RESULTS AND DISCUSSION

The measured Shielding Effectiveness (SE) values were analyzed across the 10 kHz to 1 GHz frequency band for the three types of wire mesh. The grid mesh exhibited SE values of 15–20 dB in the 10 kHz to 150 MHz frequency band, and it was observed that the SE value decreases toward higher frequencies.

Similarly, the results for the three types of wire mesh cages—constructed by assembling the wire meshes—demonstrated comparable performance. Consistent with the flat wire mesh, the cages exhibited SE values of 15–20 dB in the 10 kHz to 150 MHz frequency band, confirming that the shielding performance decreases as the frequency increases.

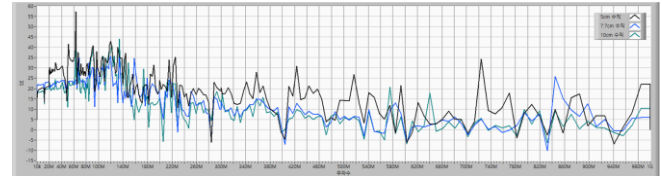


Fig. 1. Wire mesh vertical polarization (10 kHz - 1 GHz)

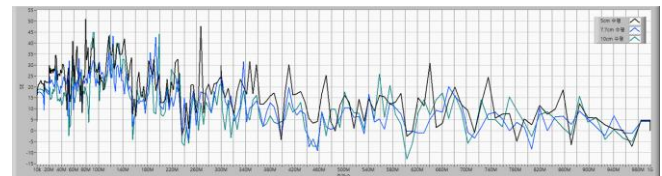


Fig. 2. Wire mesh horizontal polarization (10 kHz - 1 GHz)

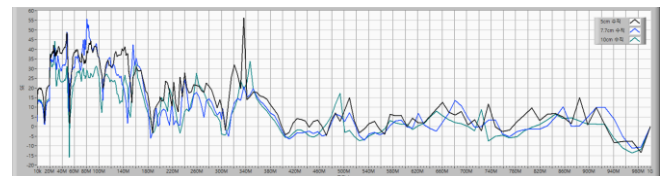


Fig. 3. Wire mesh cage vertical polarization (10 kHz - 1 GHz)

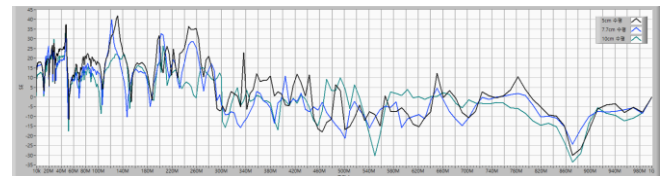


Fig. 4. Wire mesh cage horizontal polarization (10 kHz - 1 GHz)

IV. CONCLUSION

Through this study, the shielding effectiveness of the wire mesh was confirmed, and it was verified that the smaller the grid size, the higher the shielding effectiveness. Furthermore, it was confirmed that a higher shielding effectiveness is produced in the band below 100 MHz, where EMP energy is primarily concentrated.

Considering environmental attenuation factors—such as the building where major facilities are located and the distance from outer walls—it is judged that the wire mesh can be utilized as a practical and low-cost mitigation measure against nuclear EMPs.

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Automated EMS Measurement System for ISO 11452-9 Using a High Precision Mobile Manipulator

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Abstract– Modern vehicles are increasingly exposed to narrowband radiated electromagnetic energy from portable wireless devices used in or near the vehicle. This paper discusses the technical background of the ISO 11452-9 standard and introduces an automated Electromagnetic Susceptibility (EMS) measurement system using a mobile manipulator. By integrating a collaborative robot with an autonomous mobile platform, the system achieves a high positional resolution of 0.05 mm. This approach overcomes the limitations of manual testing, ensuring high repeatability and precision for complex automotive electronic components.

Keywords– ISO11452-9, EMS, Portable Transmitter, Mobile Manipulator, Automotive EMC, Automation.

I. INTRODUCTION

Modern vehicles have evolved into complex assemblies of electronic sensors, ECUs, and communication modules. These systems are susceptible to electromagnetic interference from portable devices like 5G smartphones and Wi-Fi units, which can lead to malfunctions in safety-critical systems such as ADAS. ISO 11452-9 provides a standardized method for evaluating the immunity of automotive components to these portable transmitters. This study introduces an automated testing solution using a mobile manipulator to improve measurement reliability and efficiency over traditional manual methods.

II. ISO 11452-9 STANDARD OVERVIEW

The ISO 11452-9 standard defines test methods for evaluating the immunity of components to narrowband radiated electromagnetic energy in a shielded chamber. The 2021 revision expanded the frequency range to 142 MHz – 7.125 GHz to include 5G and Wi-Fi 6 bands. Testing requires scanning the transmission antenna at a distance of 50 ± 5 mm from all surfaces of the Device Under Test (DUT). Depending on the antenna type, scan cells are defined as 50 mm x 50 mm or 100 mm x 100 mm.

III. AUTOMATED MEASUREMENT SYSTEM

To overcome the poor repeatability and human error inherent in conventional manual jig-based testing, a mobile manipulator-based system (Fig. 1) was developed.

- **System Configuration:** The system combines a collaborative robot with an autonomous mobile platform to cover large and complex test areas.

- **Precision:** It offers a positional control resolution of 0.05 mm, satisfying the strict repeatability requirements of the standard.
- **Software Integration:** The dedicated software automatically generates scan coordinates based on the selected antenna type, eliminating the need for manual recalibration during antenna swaps.



Fig. 1. Developed automated EMS measurement systems

IV. CONCLUSION

As portable wireless devices become more powerful, automated EMS testing is essential for ensuring automotive reliability. The proposed mobile manipulator system significantly enhances testing precision and efficiency. This technology provides a practical solution for the advanced electromagnetic resilience required in the modern automotive industry.

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Frequency Tuning of Relativistic Backward Wave Oscillator

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Abstract– This paper investigates frequency-tuning mechanisms for the Relativistic Backward Wave Oscillator (RBWO), emphasizing factors that limit stable, high-power operation. Building on known constraints related to breakdown, material limits, and mode competition, this work outlines approaches for enhancing tunability through controlled drift-length variation and active signal-based modulation. The study provides a framework for improving frequency agility while maintaining reliable RBWO performance.

Keywords– Relativistic Backward Wave Oscillator, high-power microwaves, frequency tuning, slow-wave structure, beam-wave interaction

I. INTRODUCTION

High-power microwave (HPM) research supports applications across science, industry, medicine, and defense, while gigawatt-level ultrashort-pulse sources reveal important beam-wave interaction physics. Ongoing study of their operating limits and performance-enhancement methods remains essential for advancing both fundamental understanding and practical use.

A literature survey revealed several frequency-tuning approaches proposed in previous studies [1-5], including beam-voltage tuning and magnetic-field tuning, each providing a limited tuning bandwidth of less than 5%. The widest tuning range reported to date was achieved through mechanical adjustments of the drift length before and after the slow-wave structure (SWS), as well as through SWS-period sweeping, with experiments demonstrating that such methods can yield up to approximately 10% tunability. However, these techniques are generally considered impractical due to their reliance on external mechanical actuators and concerns regarding long-term component durability. In our experiments on the SNU RBWO, we observed a tuning bandwidth of about 3% at 350–500 kV for the TM_{01} mode, accompanied by an output-power efficiency of roughly 20%.

II. METHODOLOGY

This study focuses on high-level analysis of tunability mechanisms using non-sensitive theoretical considerations and conceptual modeling. The methodology examines how drift-length adjustments, reflector characteristics, and externally injected reference signals influence the phase relations that establish RBWO oscillation frequency. The oscillator is treated as a resonant feedback system in which controlled perturbations to boundary conditions and beam modulation affect the standing-wave pattern within the slow-wave structure. The goal is to identify general principles for enhancing tunability without providing implementation-specific operational details.

III. RESULTS AND DISCUSSION

By enlarging the interaction volume for the desired mode in coaxial RBWO, it lowers power density and improves stability. Figure 1 illustrates this trend: while most RBWOs operate near breakdown when pushed for maximum output, CRBWOs deliver higher power with lower power density over a larger effective volume, demonstrating their advantage in high-frequency, high-power operation. The superior frequency tunability of coaxial RBWO is shown in Fig 1.

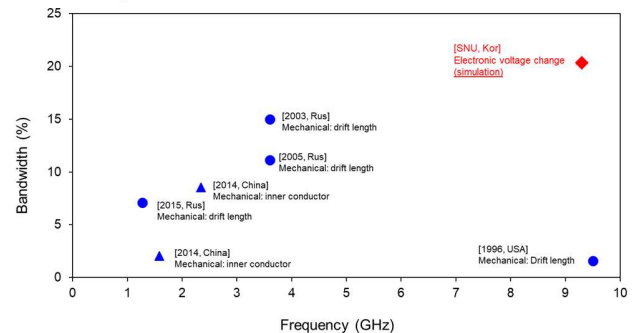


Fig. 1. Frequency tunability in coaxial RBWO

IV. CONCLUSION

Preliminary MAGIC2D PIC simulations of the proposed coaxial RBWO operating in the QTEM mode show a tuning bandwidth exceeding 20% delivering more than 2 GW of output power with over 30% efficiency. Complementary initial studies on a planar RBWO using a sine-shaped ridged SWS further support the feasibility of broad tunability, demonstrating more than 30% frequency-tuning bandwidth with a wide sheet electron beam.

ACKNOWLEDGMENT

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The Effect of Transients Caused by Conductive IEMI Attacks on Electronic Devices

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This study investigates the impact of conductive Intentional Electromagnetic Interference (IEMI) attacks on electronic devices to address the growing threat of portable electromagnetic pulse (EMP) generators. To observe these effects, conductive IEMI pulses were injected into operating personal computers and laptops. The results demonstrate that commonly used electronic devices are highly vulnerable to malfunctions caused by non-standard EMP generators.

Keywords— Conductive IEMI, Electromagnetic Pulse, Transients, Electronic Devices / IT Devices, Vulnerability, Malfunction, Portable EMP Generator

I. INTRODUCTION

Recently, along with the development of industrial technology, electrical and electronic equipment operated in digital control devices for energy, defense, broadcasting/communications, transportation, and meteorological facilities mostly utilize semiconductor components composed of integrated circuits. Such equipment, which pursues miniaturization, lightweighting, and multi-functionality, is highly sensitive to electromagnetic pulses (EMP), and exposure to EMP can cause malfunctions or damage to the devices. In particular, considering the trend where individuals can easily manufacture or purchase EMP generators through the Internet, and their sizes are becoming small enough to be portable, countermeasures against small-scale EMP terrorism are also required. To address these issues, this paper aims to identify and analyze the impact of conductive IEMI (Intentional Electromagnetic Interference) attacks on electronic devices.

II. METHODOLOGY

To observe the changes that occur when conductive IEMI pulses are injected into operating personal computers and laptops, a test environment was configured as shown in Figure 1.

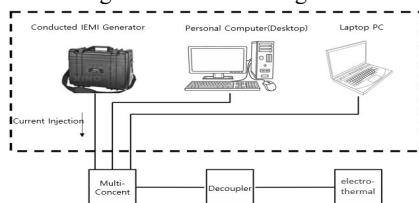


Fig. 1. Test Configuration Diagram

As shown in Figure 2, the conductive IEMI generator used in the test generates an output voltage of 6.1 kV, a pulse width of 1.6 μ s, and a repetition rate of 16 Hz when verified by connecting a 15 Ω resistor. The test was repeated 5 times, and for each test, a pulse was injected for about 2 seconds to observe changes in the Equipment Under Test (EUT). If a malfunction occurred in the EUT after the pulse injection, a stabilization time of approximately 15 minutes was given after a single test to allow sufficient time for normal operation to resume.

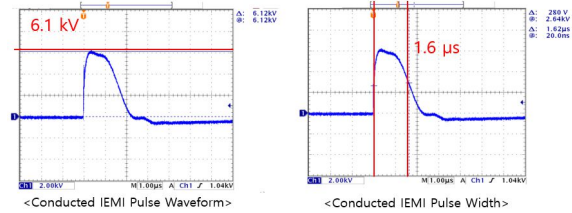


Fig. 2. Conducted IEMI Pulse Waveform & Width

III. RESULTS AND DISCUSSION

The measurement results for the laptop PC showed that out of the five trials, the device experienced three 'Self-resets', required one 'Power reset', and exhibited no malfunction ('NONE') in one instance. In contrast, the personal computer (PC) demonstrated higher vulnerability under this specific test environment; it experienced two 'Self-resets' and required three 'Power resets', resulting in malfunctions in all five trials without a single instance of normal operation. These results are analyzed as malfunctions caused by inductive noise within the internal circuits of the electronic devices, triggered by factors such as high-voltage pulse intensity, extremely rapid rise time, and repetitive pulsing. Furthermore, the findings suggest that while laptop PC can filter transient voltages to some extent using external adapter and battery, personal computer is more susceptible to system interruptions because their internal power supply is directly connected to the mains electricity.

These findings clearly demonstrate that commonly used, general electronic devices are highly vulnerable to transients caused by non-standard, portable EMP generators. The significant rate of critical malfunctions, such as unprompted system resets and complete power failures requiring manual resets, highlights the urgent need for protective countermeasures against small-scale EMP terrorism

IV. CONCLUSION

Through this paper, it was confirmed that the electronic devices we currently use on a general basis are vulnerable to terrorism using non-standard EMP generators. In the future, it is necessary to conduct extensive research on the effects of repetitive high-power electromagnetic pulses on IT devices by varying the number of tests and the test environment (such as cable length and thickness, number of connected EUTs, etc.).

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Ultra-Compact, 4-GW PFN-Marx Generator for HPM Applications

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Abstract– This paper presents the design of a ultra compact 26-stage PFN-Marx generator specifically engineered for high-power microwave (HPM) applications. By integrating a 6-cell pulse forming network (PFN) into a zigzag Marx configuration, the system achieves a peak power of ~ 4 GW and an energy density of ~ 4 kJ/m³ within ~ 0.1 m³. Simulation results demonstrate a peak output of 650 kV under a 106 Ω matched load, providing a 103 ns rectangular pulse. The current design phase confirms the feasibility of megavolt-level outputs in a portable format, with prototype fabrication and experimental validation planned as the next stage.

Keywords– PFN-Marx, High-power microwave (HPM), Ultra-compact, Pulsed power, Energy density, Zigzag Marx structure

I. INTRODUCTION

The demand for HPM sources has shifted towards systems that offer not only high voltage but also extreme compactness and high energy density. This paper presents the design of a 26-stage PFN-Marx generator capable of delivering a peak power of ~ 4 GW. With an energy density of ~ 4 kJ/m³ within an ultra-compact volume of ~ 0.1 m³, that has ~ 0.35 m diameter and ~ 1.2 m length. this design represents an advancement in mobile and space-constrained pulsed power technology for driving HPM devices.

II. SYSTEM DESIGN AND TECHNICAL SPECIFICATIONS

The generator integrates a 26-stage zigzag Marx configuration with optimized Pulse Forming Networks (PFN) to maximize power-to-volume efficiency [1-3].

- **Ultra-Compact Structure:** By utilizing a zigzag assembly, the total system volume is minimized to ~ 0.1 m³, achieving a high energy density of 4 kJ/m³.
- **PFN Unit Parameters:** Each of the 26 stages incorporates a 6-cell PFN. Each cell consists of a 2,100 pF capacitor (50 kV) and a 35 nH inductor, storing 2.625 J. The total PFN impedance (Z_{PFN}) is 4.1 Ω , and the total Marx impedance is 106 Ω .
- **Pulse Characteristics:** The system is engineered to produce a 103 ns rectangular pulse, ensuring maximum energy transfer and frequency stability in HPM loads.

III. DESIGN RESULTS AND DISCUSSION

The PFN-Marx performance was evaluated through circuit simulations. The results confirm that the design reaches performance metrics tailored for HPM applications.

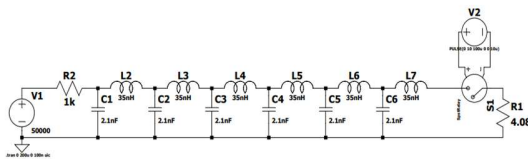


Figure 1 Single Stage PFN-Marx Circuit.

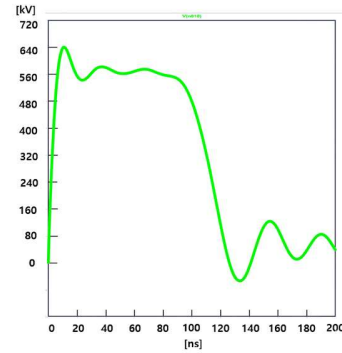


Figure 2 26-Stage PFN-Marx Simulation Output.

Table I. Predicted Design Specifications and Performance

Key Parameter	Unit	Value
Peak Power Output	GW	~ 4
Energy Density	kJ/m ³	~ 4
Total System Volume	m ³	~ 0.1
Peak Output (Open Load)	kV	$\sim 1,300$
Peak Output (106 Ω Load)	kV	~ 650
Pulse Width (FWHM)	ns	~ 103

The simulation results in Fig. 2 show a peak output voltage of 650 kV under a 106 Ω matched load. It should be noted that these results are based on the current design phase and detailed simulations; the physical fabrication of the prototype and subsequent experimental validation are scheduled as the next stage of this research. The 6-cell PFN configuration per stage in Fig. 1 is expected to provide a superior flat-top waveform, critical for HPM source efficiency.

IV. CONCLUSION

This study demonstrates a PFN-Marx generator design that strikes an optimal balance between power, energy density, and physical size. With a predicted 4 GW peak power and an energy density of 4 kJ/m³, the 0.1 m³ ultra-compact design provides a robust solution for advanced HPM applications. Future work will focus on the hardware implementation and experimental testing.

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Analysis of Radiated and Conducted Disturbance Environments for E1 HEMP Based on Protection Concept

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Abstract– IEC TC77 SC77C establishes appropriate equipment immunity testing levels for E1 HEMP threats by considering the electromagnetic environment at the equipment's location. This paper analyzes the radiated and conducted disturbance environments across six principal protection concepts defined in IEC 61000-6-6 and IEC 61000-2-11. While radiated disturbance levels decrease consistently according to the applied protection concept, conducted disturbance levels exhibit certain discrepancies and lack a corresponding reduction in some areas. To ensure clarity for standard implementers, a revision of IEC 61000-6-6 is proposed to explicitly define the basis for establishing conducted environments at each test port.

Keywords– E1 HEMP, Protection Concepts, Radiated Disturbance, Conducted Disturbance, IEC 61000-6-6.

I. INTRODUCTION

IEC TC77 SC77C is responsible for establishing immunity test levels for E1 HEMP, considering the specific electromagnetic environment where equipment is installed. Current standards, such as IEC 61000-6-6, define these environments based on various protection concepts. However, an analysis reveals discrepancies between the minimum attenuation required by these concepts and the actual defined disturbance levels, particularly regarding conducted environments.

II. HEMP PROTECTION CONCEPTS AND ENVIRONMENTS

The HEMP environment is classified into six principal protection concepts based on the level of shielding and protection provided at the equipment location.

- Concept 6: Represents an environment within a room or building with excellent RF shielding (80 dB) and PoE protection (80 dB).
- Concept 1A: Represents a poorly shielded building or residence (0 dB) without lightning protection.
- Minimum Attenuation: IEC 61000-2-11 specifies the minimum required attenuation for electric fields, magnetic fields, and conducted currents for each concept, ranging from 0 dB for Concept 1A to 80 dB for Concept 6.

III. ANALYSIS OF DISTURBANCE ENVIRONMENTS AND DISCREPANCIES

A. Radiated Environment

The radiated disturbance levels for E1 HEMP are reduced in accordance with the protection concept applied at the installation site.

- Consistency: The disturbance level decreases from 50 kV/m for Concept 1 to 0.005 kV/m for Concept 6, aligning with the 0 dB to 80 dB attenuation scale.

B. Conducted Environment

In contrast to radiated levels, conducted disturbance levels do not always show a corresponding reduction relative to the protection level.

- AC Power and Telecom: For Concepts 3 through 6, a damped

sinusoidal signal is applied. However, further justification is needed for the 25 kV value used for underground lines in Concepts 1A and 2A.

- I/O Data Ports: Discrepancies exist where current data show no environmental difference based on conducted protection devices (A/B) in Concepts 1 and 2, despite differences appearing based on radiated shielding. Additionally, the justification for the 75 A current in Concept 1A requires clarification.
- Exterior Antenna Ports: The standard applies a specific frequency band (25 MHz to 450 MHz) only to external antennas, for which the underlying basis needs to be more clearly explained.

TABLE I. CONDUCTED ENVIRONMENT AND PROTECTION CONCEPT

Protection concept	unit	1		2		3	4	5	6
		A	B	A	B				
Conducted environment for E1 HEMP	dB	0	20 ¹⁾	0	20 ¹⁾	40 ²⁾	40 ²⁾	60 ²⁾	80 ²⁾
Conducted disturbance level	AC power and telecom	kV	600 25 ³⁾	60	600 25 ³⁾	60	1.5	1.5	0.15
	I/O data ports	A	75	75	75	75	0.75	0.075	0.0075
	Exterior antennas	kV	1,800f 450>E25	4	1,800f 450>E25	4	0.72	0.72	0.07

1) 1B/2B has 20 dB lightning protection at the exterior antennas and the AC mains.

2) In-door or building interior with nominal PoE protection

3) A 25/500 ns waveform with a 50 Ω source impedance is assumed for underground lines without insulation breakdown; otherwise, the disturbance may be modified.

III. SUMMARY AND PROPOSAL

The analysis confirms that while radiated disturbance levels align with the protection concepts, conducted levels show variations that can lead to confusion.

- Proposal: It is proposed that IEC 61000-6-6 be revised to provide a clearer explanation of the basis for setting conducted environment levels for each test port.
- Conclusion: Explicitly defining how these environments are established will enhance understanding and prevent confusion for readers and implementers of the standard.

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Redefining High-Power Electromagnetic Threats in the LEO/VLEO Satellite Mega- Constellations:

Re-analyze of Space-Weather Events (Solar cycles 21-25: 1977-2025) as HPEM/HEMP Analog Factors

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Current HPEM/HEMP standards (IEC 61000-2-9:2025; DOE benchmark, 2021) define E1, E2, and E3 environments based primarily on Cold War-era nuclear test phenomenology, with limited cross-validation against large-scale empirical data from natural electromagnetic events. This paper presented a comprehensive validation of HEMP waveform standards through extreme space weather events during Solar Cycles 20–25 (1977–2025). By re-analyzing 47 severe solar-electromagnetic events using an OSINT-RAG (Retrieval-Augmented Generation) framework applied to NOAA/SWPC records, IEC/NATO/DOE standards, and infrastructure damage reports, we identified direct physical and operational analogs for HEMP E1, E2, and E3 components. A key finding was the significant “Standardization Gap” in IEC 61000-4-25, where empirical GIC ((Geomagnetically Induced Current) durations persisted up to 12 times longer than current test requirements. We also formalized the “Thermospheric Drag” threat to LEO constellations—demonstrated by the 2022 Starlink and 2024 Gannon events—as a critical vulnerability pathway absent from international HPEM standards. To address these gaps, we proposed the Natural HEMP Equivalence Index (NEEI), a multi-variable metric for global infrastructure risk.

Keywords— HPEM, HEMP, E1/E2/E3 waveform standards, geomagnetic disturbance, GIC, HF blackout, GNSS interference, LEO/VLEO satellites, thermospheric drag, NEEI, IEC 61000-2-9, space weather

I. Research Background & Motivation

HEMP/HPEM standards (e.g., IEC 61000-2-9:2025; MIL-STD-2169C, 2020; DOE Benchmark, 2021) define E1, E2, and E3 waveform environments based on Cold War nuclear test phenomenology. However, a large-scale open empirical validation (e.g., space weather event) across all three components simultaneously is absent from the published literature.

The early-time E1 component (ns– μ s) arises from the Compton-electron mechanism and couples to short conductor systems; the intermediate E2 (μ s–ms) overlaps spectrally with lightning transients; the late-time E3 (s–hundreds of s) results from MHD coupling of geomagnetic field distortion, inducing quasi-DC GIC in long power lines, pipelines, and telecommunications cables [1],[2]. While physically well-motivated, cross-validation against empirical large-scale impact data from natural events remains limited [3]. E3/GIC analogs have been studied in isolation [4],[7], but a simultaneous, statistically rigorous treatment of natural analogs to all three components within a unified framework has not been published.

Concurrently, LEO/VLEO satellite mega-constellations expanded rapidly, with over 7,400 active satellites by 2024. The Feb. 2022 Starlink event (38 de-orbits from a moderate G1–G2 storm inducing 20–50% thermospheric density increase at 210 km) and

the May 2024 G5 Gannon storm (mass Starlink maneuvering; 72-h NORAD tracking degradation) confirm catastrophic infrastructure impacts on LEO/VLEO systems.

This paper addressed three objectives: (i) systematic, statistically validated re-analysis of 47 severe space-weather events in Solar Cycles 21–25 (1977–2025) as natural analogs to HEMP E1, E2, E3; (ii) formalization of the Natural HEMP Equivalence Index (NEEI) as a composite dimensionless scalar commensurable with IEC benchmark waveform levels; and (iii) extension to NEEI+ incorporating LEO/VLEO satellite drag and anomaly terms.

II. METHODOLOGY: OSINT–RAG PIPELINE

This study utilized an Open-Source Intelligence-Retrieval-Augmented Generation (OSINT-RAG) pipeline to ingest 89,000 document chunks covering 47 extreme space weather events in the space weather datasets. The analysis mapped natural parameters (X-ray flux, solar radio flux, geospace metrics) to HEMP waveform characteristics [10].

A) Event Dataset (47 Events, Solar Cycles 20–25)

Beginning with a candidate pool of approximately 1,240 events extracted from NOAA/SWPC solar event catalogs for the period 1977–2025, five successive quality filters are applied:

(1) solar flare intensity $\geq$ X1.0 class (GOES X-ray flux); (2) confirmed associated with a Coronal Mass Ejection (CME) in the CDAW NASA catalog; (3) geomagnetic storm rating $\geq$ G3 ($K_p \geq 7$) in NOAA SWPC records; (4) available GIC measurement or proxy record (USGS, EPRI, NERC); and (5) documented infrastructure impact in at least one independent report. The application of this five-stage filtering process yielded a final dataset of 47 events spanning Solar Cycles 20–25.

The resulting 47 event dataset included: 8 events from 1977–2000 (including the 1989 Quebec storm); 19 events from 2001–2010 (including Halloween 2003 and the December 2006 GNSS event); 12 events from 2011–2020 (including the 2012 Carrington class near miss); and 8 from 2021–2025 (including the 2022 Starlink event and the May 2024 G5 Gannon storm). All selected events were parameterized according to flare class, minimum Dst index, maximum K_p index, GIC peak amplitude and duration, Solar Energetic Particle (SEP) proxy, solar cycle phase, and geographic impact region.

B) 89,000-Chunk Document Corpus (4 Layers) using AI technology

- L1(Layer 1) NOAA/SWPC time-series
- L2(Layer 2) IEC/DOE/EMP Commission standards
- L3(Layer 3) Infrastructure damage reports (NERC, Lloyd's 2024, EPRI)
- L4(Layer 4) 890 peer-reviewed papers (AGU Space Weather, IEEE Trans. EMC, JGR).

III. A NOVEL INDEX FOR THREE HPEM/HEMP: NEEI AND NEEI⁺

We propose the Natural HEMP Equivalence Index (NEEI), a single unified scalar simultaneously mapping natural solar-electromagnetic parameters onto HEMP E1, E2, and E3 domains:

$$NEEI = \alpha_1 \cdot E1_norm + \alpha_2 \cdot E2_norm + \alpha_3 \cdot E3_norm \quad (1)$$

$E1_norm = (SID \text{ duration [min]} \times X\text{-ray flux peak [W/m}^2]) / E1_reference$, where $E1_reference$ corresponds to the DOE 50 kV/m field level. $E2_norm = \text{Type-III burst flux [SFU]} / E2_reference$, set by the Dec 2006 GNSS-disruption event [14]. $E3_norm = |Dst \text{ min [nT]}| \times GIC \text{ peak [A]} / E3_reference$, where $E3_reference$ is the IEC E3 GIC threshold in IEC 61000-4-25.

Component weights ($\alpha_1 = 0.30 \pm 0.04$, $\alpha_2 = 0.15 \pm 0.05$, $\alpha_3 = 0.55 \pm 0.06$; PCA regression, Bootstrap 95% CI, $n=1,000$) are derived by PCA on the 47-event dataset, minimizing residuals against economic damage. The high α_3 reflects dominant E3 coupling to power grids.

Table 1. NEEI classification for representative events.

Event	Dst (nT)	NEEI	E1 Class	E2 Class	E3 Class	S_drag
1989 Mar 13	-589	0.87	Probable	Weak	Confirmed	N/A
2003 Oct 29	-383	0.81	Confirmed	Probable	Confirmed	Low
2006 Dec 5-6	-147	0.44	Probable	Confirmed	Weak	Low
2012 Jul 23	~-1200*	0.97*	Confirmed*	Probable*	Confirmed*	High*
2022 Feb 3	-57	0.19	Weak	None	Weak	High
2024 May 10	-412	0.78	Confirmed	Probable	Confirmed	High

*Estimated for 2012 near-miss. S_drag=High denotes $\delta p/p_{\text{quiet}} > 20\%$.

To address LEO/VLEO constellations, NEEI is extended to NEEI⁺:

$$NEEI^+ = w_1 \cdot S_abs + w_2 \cdot S_RF + w_3 \cdot S_GIC + w_4 \cdot S_drag + w_5 \cdot S_anom \quad (2)$$

The first three sub-scores re-label NEEI components: S_abs (D-region absorption; HF impact), S_RF (GNSS L-band RF interference), S_GIC (long-line induction). $S_drag = (p_storm/p_quiet - 1)$ at mission altitude—a novel pathway not covered in any current IEC/MIL-STD/NATO standard. S_anom indexes Single Event Upset/Latch-up and ESD likelihood [9]. While w_1 – w_5 await full-dataset calibration, this pilot demonstrates simultaneous multi-pathway activation

IV. RESULTS & STATISTICAL ANALYSIS

A) E1 Component: SID–HEMP Early-Time Analog

For E1, 41 of 47 events (6 excluded for NC confidence < 0.40) yielded: Spearman $\rho = 0.78$ ($p < 0.001$), Pearson $r = 0.71$, Kendall $\tau = 0.59$. Classification: 9 Confirmed, 18 Probable, 14 Weak, 6 NC. The causal basis is D-region electron density enhancement absorbing HF radio waves through the same ionospheric layer governing E1 Compton-electron propagation. The E1 analog operates via impact-pathway similarity (HF blackout) rather than waveform equivalence; no solar event produced a 50 kV/m, 2.5 ns rise-time field.

B) E2 Component: Solar Radio Burst–HEMP Intermediate Analog

For E2 (37 classifiable events), Spearman $\rho = 0.63$ ($p = 0.003$). The weaker correlation reflects more complex spectral/directional characteristics of Type-III bursts. Classification: 4 Confirmed, 11 Probable, 22 Weak, 10 NC. The Dec 2006 event [14]—during which solar radio burst flux caused GNSS L1/L2 failures across the sunlit hemisphere—constitutes the strongest E2 Confirmed event. The E2 analog operates as operational impact pathway (GNSS RF interference at 1.2–1.6 GHz).

C) E3 Component: GIC–MHD-EMP Physical Analog

E3 (45 classifiable events) yielded the strongest result: Spearman $\rho = -0.87$ ($p < 0.001$), confirming the expected negative correlation between signed Dst and GIC-based E3 confidence. The physical analog is most direct here: geomagnetic MHD-EMP and HEMP E3 both induce quasi-DC GIC through identical electromagnetic induction physics, differing only in source geometry and duration. Classification: 28 Confirmed, 12 Probable, 5 Weak, 2 NC

Table 2. . HEMP component natural-analog correlations

HEMP Comp.	NaturalProxy	Spearman	Pearsonr	Kendall	pvalue
E1 (Early)	SID $\times$ X-ray	0.78	0.71	0.59	$< 0.001^{***}$
E2 (Interm.)	Type-III burst	0.63	—	—	0.003 **
E3 (Late)	Dst $\times$ GIC_	-0.87	-0.85	-0.71	$< 0.001^{***}$

V. STANDARDS GAPS ANALYSIS & REVISION IEC SC77C

Based on the 47-event analysis, we identified three specific gaps within IEC SC 77C standards gaps. Consequently, we proposed targeted, direct revisions to address for IEC SC77C standards

Gap	Issue	Description	Proposed Revision
Gap 1	E3 Duration Under specification	IEC 61000-4-25 specifies ~300 s. Observed GIC: 600–3,600 s (mean 1,340 s) $\rightarrow$ 4.5 $\times$ systematic underestimation of transformer thermal stress.	$\rightarrow$ Extend mandatory E3 test to 1,800 s; 3,600 s extended class.
Gap 2	No E1/E2 Natural Validation	IEC 61000-2-9:2025 Ed.2 contains no cross-reference to natural event analogs for E1 or E2.	$\rightarrow$ Add Informative Annex in IEC 61000-2-9 Ed.3 or new SC 77C Technical Report.
Gap 3	LEO/VLEO Absent from HEMP Standards	No IEC/MIL/NATO HEMP document covers atmospheric drag as HPEM pathway. 38 Starlink satellites lost in 2022; mass fleet disruption in 2024.	$\rightarrow$ New IEC TS on HPEM threats to space infrastructure;

VI. CONCLUSIONS

This paper presented the first statistically validated, simultaneous re-analysis of 47 severe space-weather events during the Solar Cycles 21–25 (1977–2025) as natural analogs to all three HEMP/HPEM waveform components (E1,E2,E3) using a reproducible OSINT–RAG methodology.

- Statistically validated open-literature demonstration of E1, E2, and E3 natural analogs ($\rho = 0.78 / 0.63 / -0.87$).
- NEEI as a novel, dimensionless composite index mapping natural events to IEC waveform domains.
- NEEI⁺ as a first quantitative HPEM framework incorporating LEO/VLEO atmospheric drag pathway.
- Empirical proof of 4.5 $\times$ IEC E3 test duration gap (600–3,600 s observed vs. ~300 s specified).
- Three actionable IEC SC 77C revision priorities with specific technical justification.

As Solar Cycle 25 activity continues post-maximum (2025–2028) and LEO/VLEO mega-constellations scale toward 100,000+ satellites, integrating natural-analog analysis into IEC standards revision becomes both technically justified and operationally urgent.

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Lightning Protection for Spaceports and Launch Complexes: Historical Lessons, Standards Harmonization, and Resilient Architectures

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Abstract— Lightning is a well-documented, preventable hazard to space-flight operations. Mature high-lightning-density ranges have demonstrated effective launch-complex lightning protection systems (LPS), yet commercial spaceports still show inconsistent requirements and enforcement. This extended abstract consolidates a three-part SIPDA 2025 series into a risk-informed synthesis for launch pads where both direct attachment and launch-vehicle-triggered lightning can drive mission-critical failures (e.g., Apollo 12 and AC-67). We compare key standards and regulatory instruments used for spaceport LPS (e.g., NFPA 780, U.S. range guidance, and IEC 62305) and identify gaps in scope, inspection intervals, and enforceability. We also evaluate common launch pad LPS architectures—catenary systems with insulating masts, metallic-pole configurations, and tower-centric solutions—to motivate a practical resilience direction: harmonized, science-based requirements; competence-based enforcement; explicit exclusion of non-validated “lightning elimination/suppression” claims; and lifecycle-centric inspection/change-control aligned to the lightning electromagnetic environment (LEME).

Keywords— spaceports; launch pads; lightning protection systems (LPS); lightning electromagnetic environment (LEME); EMI; catenary wires; insulating masts; standards harmonization; inspection and maintenance; operational resilience.

I. MOTIVATION: LIGHTNING AS AN OPERATIONAL RESILIENCE PROBLEM

Spaceports concentrate high-consequence assets in exposed environments with long conductor runs and strict schedule constraints. The requirement extends beyond interception: the LPS must control LEME/EMI coupling such that induced transients remain below susceptibility thresholds and remain effective under continual site modifications.

II. HISTORICAL EVIDENCE: RECURRING FAILURE MODES

Program history shows lightning impact is neither hypothetical nor confined to the pad. Early reliance on launcher umbilical towers created shared current paths and elevated coupling exposure on umbilicals and avionics; triggered-lightning events further motivated strict launch commit criteria supported by robust lightning surveillance.

III. STANDARDS AND REGULATORY GAPS

Spaceport stakeholders commonly reference building-centric lightning standards, range manuals, and safety regulations. However, these sources are not consistently launch-pad-specific, and they differ in technical guidance, inspection periodicity, and enforcement mechanisms across agencies and jurisdictions. Key issues include:

- **Scope mismatch:** general-purpose LPS methods may not capture launch-complex coupling and long-conductor transient behavior without additional engineering verification.
- **Inspection and lifecycle ambiguity:** requirements for periodic inspection, post-modification review, and competence of reviewers

vary and are often under-specified for evolving pad configurations.

- **Enforcement capability:** effective compliance requires specialized training in LPS principles, LEME/EMI coupling risk, and evaluation of modifications; absent this, unsafe practices can persist under nominal “compliance.”

Given the strategic and operational importance of spaceports, harmonization through risk-based, science-driven enforcement is positioned as a resilience and mission assurance requirement rather than a discretionary design preference.

IV. ENGINEERING ARCHITECTURES FOR LAUNCH PADS

Launch pad LPS architectures impose different trade spaces in interception probability, EMI coupling, constructability, and maintainability:

- **Catenary wires with insulating masts:** strong interception performance while limiting direct lightning current flow on towers that interface with vehicle umbilicals, thereby reducing coupling risk and supporting defined protected work zones.
- **Metallic poles with catenary wires / stand-alone poles:** potentially viable under constraints but generally increase coupling exposure and may reduce shielding effectiveness relative to insulated-mast catenary solutions.
- **Tower-centric interception:** historically associated with elevated induced environments on umbilicals and shared current paths; may be incompatible with modern sensitive avionics and long interface runs without rigorous mitigation and verification.

Design selection should be explicitly tied to site lightning density, vehicle/payload EMI sensitivity, operational tempo, and the practical ability to sustain low-impedance bonding/grounding and configuration control over time.

V. CONCLUSIONS AND PROPOSED DIRECTION

The consolidated outcome of [1]-[3] is a clear industry need for a unified, enforceable spaceport LPS framework that:

- mandates validated interception strategies and LEME/EMI control appropriate to launch risk;
- requires competent review (training and/or third-party expertise) for design changes and periodic inspections;
- explicitly excludes non-validated “elimination/suppression” devices and claims; and
- treats the LPS as a living system with formal maintenance, inspection, and change-control aligned to operational resilience.

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Summary of Winter Lightning in Japan

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Abstract– Winter lightning in the coastal areas along the Sea of Japan is known to cause significant damage to social infrastructure. Its distinguishing characteristics include upward leader propagation from tall structures, discharge durations on the order of several hundred milliseconds, and transferred charges that can be more than 100 times greater than those of typical summer downward negative lightning. Although winter lightning also occurs in other regions, reports of events causing comparable levels of destruction are rare. This paper presents a comprehensive review of previous studies on the characteristics of winter lightning observed in the coastal areas of the Sea of Japan.

Keywords– Winter lightning, upward lightning, charge transfer, long duration

I. INTRODUCTION

Numerous reports have described cases of catastrophic damage due to lightning flashes along the coast of the Sea of Japan during the winter season. In this study, we referred to lightning flash that occurs along the coast of the Sea of Japan in winter as winter lightning in Japan. Lightning flashes occur in other regions of the world in the winter, but there are few reports of devastating ones, except the coastal areas of the Sea of Japan in winter. Previous studies [1],[2] have demonstrated the characteristics of the winter lightning in Japan were different from those in other regions during winter. We summarized the findings of previous studies on Japanese winter lightning and clarified the differences between that and lightning flashes that occur in winter in other regions.

II. CHARACTERISTICS OF WINTER LIGHTNING IN JAPAN

Winter lightning in Japan only occurs within approximately 30 km of the coast of the Sea of Japan [3]. This finding indicates that water vapor supplied from the Sea of Japan contributes significantly to the occurrence of winter lightning. The most notable features of Japanese winter lightning flashes are that their leaders often propagate upward from tall structures, exhibit prolonged durations, and occasionally transfer charges up to approximately 100 times greater than those of typical downward negative flashes. The duration time of downward negative return stroke is generally on the order of micro-second. However, in some cases, that of initial continuous current pulse can last up to several hundred milliseconds. A previous study [4] exhibited 98% of lightning flashes that hit the wind turbines were initiated from the wind turbines. The median charge transfer of downward negative lightning flashes was 7.5 C [5], whereas lightning flashes with charge transfer exceeding 900 C have been observed in winter lightning in Japan [6]. Approximately 4% of lightning discharges observed by the Rogowski coils at the bottom of the wind turbines exhibited a charge transfer exceeding 300 C in Japanese winter lightning area [7]. A report demonstrated that the percentage of negative, positive and bipolar lightning flashes on wind turbines in winter were 71%, 15%, and 14% respectively [8]. The median value of charge transfer of upward

bipolar flashes that hit them was approximately 107 C [9].

III. DISCUSSION

The heightened energy observed in winter lightning in Japan is attributable to the lower altitude of the -10°C isothermal layer. According to one of previous studies, when the altitude of -10°C isothermal layer was between 1000 m and 2,500 m, the upward lightning flash that propagated from the wind turbines at 500 m above sea level increased [10]. The findings of the study suggested that energetic upward lightning flashes occur in other areas when the distance between the -10°C isotherm altitude and the base attitude of a high structure was similar to that (500 – 2000m) in the coastal area of the Sea of Japan. Recently, upward flashes with transferred charge exceeding 300 C initiated from a tower standing at an altitude of 1300 m have been observed in Austria [11].

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Comparative Study of Lightning Interaction at Sântis Tower and a Wind Turbine in Croatia

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Abstract– Since July 2025, an ongoing systematic investigation of lightning interactions with tall structures and wind turbines is conducted as a collaboration among Switzerland, Croatia and Poland. One objective is to establish a multimodal state-of-the-art wind farm lightning observatory in Croatia within a winter-lightning region, following the same measurement concept as the Sântis Tower facility in Switzerland. The project aims to understand how lightning interactions differ across geographical, site-specific conditions, and wind-turbine specific characteristics, including rotating blades and composite blade materials. To evaluate the impact of the upward lightning current on composite materials and support the development of more resilient materials, a crucial part of the project also involves investigating lightning interaction with wind-turbine composites through destructive testing of manufactured composite samples that will be carried out by the Polish partner. Summarize objective, approach, and key results. Do not include equations or tables.

Keywords– lightning attachment; lightning seasonality; upward flashes; wind turbines; winter lightning

I. INTRODUCTION

The Sântis Tower is 123.5 m tall and is located at an elevation of 2,502 m above sea level on Mount Sântis; measurements have been conducted there since 2010 [1]. These measurements currently constitute one of the largest datasets of upward negative flashes and have provided insights into lightning types, detection network performance, and lightning–ionosphere interactions. In contrast, the Croatian wind turbine is situated in a hilly area at approximately 1,250 m above sea level in a region characterized by intensive winter lightning. Measurements at this site started in July 2022 and by the start of this multilateral collaboration included two Rogowski coils installed at the base of the wind turbine tower and a single high-speed camera [2]. Since July 2025, at the start of the collaboration, an additional high-speed camera and an event camera have been deployed at the nearby 110/20 kV substation, with a view of multiple wind turbines in the wind farm. At the time of writing, a wideband E-field sensor is being tested in the laboratory and is being prepared for field deployment in the first quarter of 2026. The next steps include installing an X-ray sensor and associated acquisition hardware inside the wind turbine tower; installation is

expected by the end of the second quarter of 2026.

II. INITIAL RESULTS SINCE PROJECT LAUNCH (JULY 2025)

Since the start of the project through the end of 2025 (a six-month period), 26 upward lightning flashes (and no downward flashes) were recorded at the instrumented wind turbine. Based on the measured event currents, 22 of the flashes were negative, 3 were positive, and 1 was bipolar. The monthly distribution was as follows: July 2025: 2, August 2025: 3, September 2025: 5, October 2025: 5, November 2025: 9, January 2026: 2. For almost half of the flashes, camera footage is available, with the lightning channel clearly visible between the instrumented wind turbine and the cloud base. The collected measurements are systematically validated against lightning location data for CG strokes and the emitted optical radiance of lightning imager data. The paper will show the onsite setup and present measurements of representative cases.

ACKNOWLEDGMENT

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Impedance-Based Prediction of Common-Mode Current in Non-LISN Conditions

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Abstract– Electromagnetic compatibility (EMC) compliance is mandatory for mass production. Electromagnetic interference (EMI) is typically evaluated through conducted emission (CE) and radiated emission (RE) tests. These tests use a line impedance stabilization network (LISN) to provide standardized impedance for measurement repeatability. However, in practical operation, the LISN is absent, allowing extensive common-mode (CM) return paths to form. This results in discrepancies between standardized test results and actual emission levels. This paper models CM paths with and without a LISN and proposes an impedance-based scaling factor to convert LISN-measured CM current to practical conditions. Experimental validation confirms a deviation within 1.5 dB at the resonance frequency.

Keywords– Electromagnetic compatibility (EMC), Electromagnetic interference (EMI), Line impedance stabilization network (LISN)

I. INTRODUCTION

Electromagnetic compatibility (EMC) is conducted to ensure product compliance. Conducted emission (CE) testing evaluates noise currents that propagate outside the equipment through power or signal lines. CE measurements are typically performed using a LISN (Line Impedance Stabilization Network). The LISN suppresses external noise and enables reproducible measurement of the noise characteristics of the equipment under test (EUT). However, this setup does not represent the power impedance of the actual operating environment. The magnitude and frequency characteristics of the noise current can therefore change depending on the presence of the LISN. As a result, the noise behavior in real environments may differ from the results obtained in CE tests [1].

This paper proposes a method to predict the common-mode (CM) current in an operating condition without a LISN. The method uses the CM current measured with a LISN and a scaling factor (S) derived from the impedance characteristics of the measurement setup.

II. PREDICTION OF COMMON-MODE CURRENT IN NON-LISN CONDITIONS

Fig. 1 shows the CM equivalent circuit of the CE test including the EUT. For scaling factor derivation, the test setup is divided into the power supply, the LISN, and the EUT. With the LISN present, the total CM impedance consists of the LISN and the EUT, whereas without the LISN it consists of the power supply and the EUT.

Since the LISN consists of passive components, its impedance can be calculated analytically. The EUT impedance is obtained by subtracting the LISN impedance from the total CM impedance measured with the LISN. The power supply impedance is then obtained by subtracting the extracted EUT impedance from the total CM impedance without the LISN. Assuming the same noise source,

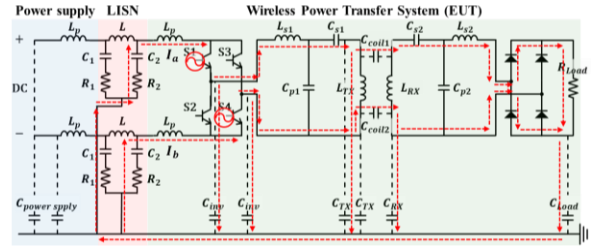


Fig. 1. Common-mode equivalent circuit with LISN

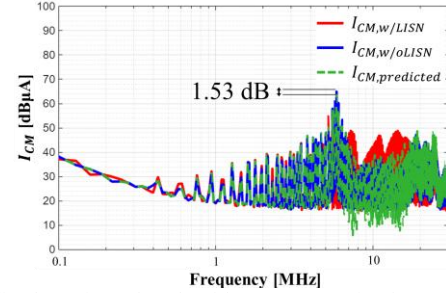


Fig. 2. Predicted experimental results of the CM current using the proposed method the ratio of CM currents equals the ratio of impedances. Based on this relationship, the scaling factor is derived from (1) and (2) to predict the CM current without the LISN.

$$S = \frac{I_{CM\ w/o\ LISN}}{I_{CM\ w\ LISN}} = \frac{Z_{LISN} + Z_{EUT}}{Z_{power\ supply} + Z_{EUT}} \quad (1)$$

$$I_{CM\ w/o\ LISN} = S \times I_{CM\ w\ LISN} \quad (2)$$

As shown in Fig. 2, the proposed prediction method agrees well with the measured results even in the resonance region. The error is 1.53 dB at the major peak and remains within 5.46 dB across the entire frequency band.

III. CONCLUSION

This paper proposes a method to predict the CM current in a non-LISN condition. The proposed method shows an error of 1.53 dB at the resonance point of the common-mode current, confirming its high prediction accuracy.

ACKNOWLEDGMENT

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Time-domain Shielding Effectiveness of a Small Slot on PEC Screen Using Convolution Integral

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Abstract– This study analytically calculates the peak-reduction SE_{TD} for a HEMP-simulating double exponential pulse incident on a slotted conducting plate using convolution integral. The analytical transmitted waveforms and SE_{TD} values were validated against numerical simulations, demonstrating excellent agreement.

Keywords– Slot, convolution integral, time-domain, shielding effectiveness

I. INTRODUCTION

Accurately defining time-domain shielding effectiveness (SE_{TD}) is crucial in EMP research to evaluate transient protection. Since small slots are the primary leakage paths in EMI/EMC shielding, their transient response must be rigorously analyzed. This paper proposes an analytical method to calculate the peak value reduction SE_{TD} ^[1] for a double exponential EMP pulse using convolution integral. The proposed model is validated against numerical simulations, demonstrating excellent agreement and offering a practical tool for rapid shielding evaluation.

II. TIME-DOMAIN SE OF PEC PLATE WITH A SLOT

Based on Bethe's theory, an electrically small slot can be modeled by its magnetic polarizability.^[2] Accordingly, the frequency-domain transfer function of a slot on a PEC plate is expressed as follows:

$$H(\omega) \approx j\omega\tau_a, \quad (1)$$

where $\tau_a = 2\mu_0\alpha_m/Z_0w^2$. μ_0 and Z_0 are permeability and intrinsic impedance of air, respectively, α_m is magnetic polarizability which is dependent on the shape of slot. When the slot is rectangular as depicted in Fig. 1(a), the expression^[3] is

$$\alpha_m = \frac{\pi}{16}lw^2 \cdot \frac{1}{\ln(4l/w)-1}, \quad (2)$$

where w and l are width and length of slot. The time-domain transfer function can be expressed by Fourier transform of (1) as

$$h(t) = \tau_a\delta'(t). \quad (3)$$

For an incident pulse $e_{inc}(t)$, the transmitted time-domain electric field $e_{tran}(t)$ through an apertured plate is given by the convolution integral:

$$e_{tran}(t) = e_{inc}(t) * h(t) = \tau_a e_{inc}'(t). \quad (4)$$

When the incident EMP is modeled as a double exponential pulse, it is expressed as follows:

$$e_{inc}(t) = E_0(e^{-\alpha t} - e^{-\beta t}), \quad (5)$$

where α and β determine the decay and rise rates of the pulse, respectively; for most EMP waveforms, $\beta > \alpha$ is satisfied. The transmitted electric pulse is

$$e_{tran}(t) = E_0\tau_a(\beta e^{-\beta t} - \alpha e^{-\alpha t}). \quad (6)$$

To determine the peak value reduction SE_{TD} , the maximum values of the incident and transmitted pulse are

$$e_{inc}^{max}(t_{max}) = E_0\left(\frac{\beta}{\alpha}\right)^{-\alpha/(\beta-\alpha)}\left(1 - \frac{\alpha}{\beta}\right), \quad (7)$$

$$e_{tran}^{max}(0) = E_0\tau_a(\beta - \alpha), \quad (8)$$

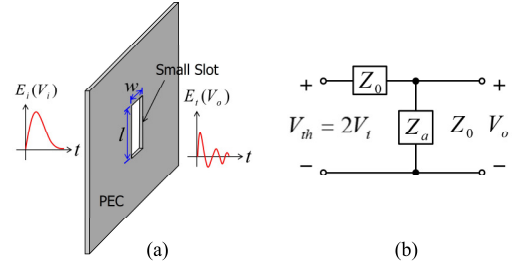


Fig. 1. Incident pulse on a conducting plate with a slot and its equivalent circuit model.

where $t_{max} = \ln(\beta/\alpha)/(\beta - \alpha)$. Finally, the peak value reduction SE in time-domain of a slotted PEC plate is defined when a double exponential pulse is incident.

$$SE_{TD} = 20 \log_{10} \left(\frac{e_{inc}^{max}}{e_{tran}^{max}} \right) = 20 \log_{10} \left(\frac{1}{\tau_a\beta} \left(\frac{\beta}{\alpha} \right)^{-\alpha/(\beta-\alpha)} \right). \quad (9)$$

III. RESULTS AND DISCUSSION

Using (9), three different sizes of slots were calculated SE_{TD} [dB] as shown in Table 1. The accuracy of our analytic model was validated through excellent agreement with CST MWS simulations.^[4] Naturally, the SE in the time domain also decreases as the slot size increases, just as in the frequency domain.

Table 1. Time-domain SE of three different size of slots ($\alpha = 4 \times 10^7 \text{ s}^{-1}$ and $\beta = 6 \times 10^8 \text{ s}^{-1}$).

Slots	Slot #1	Slot #2	Slot #3
$w[\text{mm}] \times l[\text{mm}]$	1×10	1×100	10×100
SE_{TD} [dB]	49.01	34.38	29.01
CST MWS	49.56	35.35	28.92

IV. CONCLUSION

This study analytically evaluates the SE_{TD} of slotted plates under HEMP excitation via convolution integrals. This framework provides a computationally efficient tool for predicting the transient response to high-speed electromagnetic pulses.

ACKNOWLEDGMENT

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Electromagnetic Shielding Analysis of Multi-Layered Panel with Absorbing Materials

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Abstract– This paper analyzes the electromagnetic shielding property of two-dimensional multi-layered panel with slots inserted by absorbing materials by an analytic modal solution for transverse magnetic (TM) wave incidence. The penetrated fields are calculated in terms of the size of slot and the types of absorbing materials.

Keywords– multi-layered panel, absorbing material, shielding

I. INTRODUCTION

Metallic panels are standard for EMP shielding, but gaps at panel joints often become leakage paths that degrade performance and complicate maintenance.^{[1],[2]} To address this, multi-layered structures incorporating electromagnetic (EM) absorbers are proposed to attenuate internal leakage and enhance structural robustness. This paper analytically investigates the shielding characteristics of a multi-layered panel using a two-dimensional (2D) approximation. Under transverse magnetic (TM)-wave incidence, an analytical solution is derived via Fourier transform and mode-matching.^[3] We calculate the penetrated fields to evaluate shielding effectiveness and demonstrate the impact of absorbing layers in suppressing leakage through structural discontinuities.

II. FORMULATION

This paper analyzes the TM-wave incidence case, where the incident magnetic field has no z -component (Fig. 1). This represents the worst-case shielding effectiveness (SE) scenario, as the electric field component aligns with the slot's shorter side (x -axis), maximizing penetration. The 2D multi-panel structure is divided into five regions. In the semi-infinite regions (I) and (V), fields are expressed using Fourier integrals due to continuous modes, while the finite slot regions (II) and (IV) and the internal cavity (III) are represented by Fourier series to capture discrete modes. All field expressions satisfy the boundary conditions on the x -axis conducting surfaces. The remaining unknown coefficients are obtained by enforcing mode-matching at $y = 0$ and $y = -d$. Solving the resulting four-equation system for Regions (II) and (IV) via matrix inversion yields the complete field distribution.

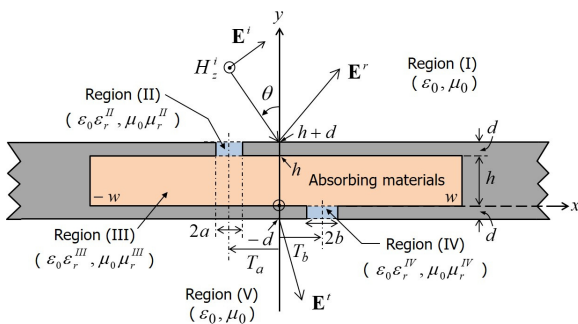


Fig. 1. Two-dimensional approximated geometry of multi-layered conducting plates loaded with absorbing materials.

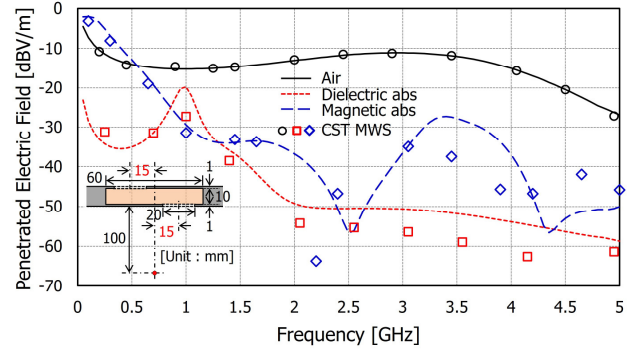


Fig. 2. Analytical and numerical results of penetrated electric field.

III. RESULTS

Using the derived formulation, we calculated the transmitted electric field in Region (V) as shown in Fig. 2. The impact of absorbing materials in the cavity was analyzed when the upper and lower slots are offset. Results indicate a significant increase in transmission loss due to absorbing materials. Notably, incorporating absorbers enhances shielding performance at the frequency above 1.5 GHz by 20–40 dB compared to air. This confirms that the proposed multi-layered panel structure with lossy materials is highly effective. The accuracy of our analytical modal solution was validated through good agreement with CST MWS simulations.^[4]

IV. CONCLUSION

This paper presented an analytical modal solution for a 2D multi-layered panel structure with absorbers using Fourier transform and mode-matching techniques. We investigated transmission characteristics across various geometries and materials, confirming good agreement with numerical simulations. These results provide a practical design reference for optimizing electromagnetic shielding panels.

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This work was supported by the Institute of Information & communications Technology Planning & Evaluation (IITP) grant funded by the Korea government (MSIT) (No. RS-2025-02642985, Development of Identification, Protection, and Detection Technologies for High-Power Electromagnetic Pulse Protection).

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Recent Advances in Equipment for Measuring Soil Properties in the Lightning Current Frequency Range

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Abstract—The work outlines the latest progress in developing measuring equipment for field measurements of frequency-dependent electrical soil properties. Key factors affecting measurement accuracy and ease of use are described. These factors include memory type and size, generator amplitude, and the performance of the microcontroller responsible for fast Fourier transform calculations. The device consists of a main unit (responsible for sampling current and voltage signals), an optically isolated voltage probe, and a smartphone for controlling the main unit. The maximum sampling rate of the analog-to-digital converters tested is 100–140 MHz with a resolution of 12–14 bits (depending on the chip).

Keywords—measurement; SoC; measurement equipment; soil properties; frequency-dependence

I. INTRODUCTION

The frequency-dependent electrical soil properties are important for grounding characteristics [1]. Previously, a device was developed for field measurements of soil properties [2]. Recent improvements have allowed more accurate measurements at lower frequencies and direct current [3]. In addition, the selection of suitable electrodes and the use of higher voltages significantly reduce noise [4]. However, the use of larger memory [3] has significantly increased the measurement time. Thus, a more powerful microcontroller is required for faster measurements. In addition, the static RAM previously used still imposes limitations on the measured signals. Therefore, the use of dynamic memory would be preferable. Furthermore, using an analog-to-digital converter (ADC) with a higher frequency can improve accuracy in the MHz frequency range. An ADC clock source with lower jitter can also improve accuracy. It would also be convenient to incorporate an amplifier into the device [4].

II. MEASUREMENT DEVICE

The previous version of the device is based on the ESP32 microcontroller and first-in-first-out (FIFO) memory for sampled signals [2]. The device is controlled via Wi-Fi (or Bluetooth). Initially, the ESP32 was used for fast Fourier transform (FFT) calculations. As the number of measured points increased [3], it became preferable to transmit the measured signal to a smartphone and calculate the FFT there. However, data transfer still takes a significant amount of time. Therefore, it would be preferable to use a more powerful microcontroller for FFT calculations. These proposed improvements can be achieved with modern system-on-chip (SoC) microchips. For this reason, several printed circuit boards (PCBs) were created to test the capabilities of the SoC-based device. Fig. 1 shows the preliminary boards for signal sampling.

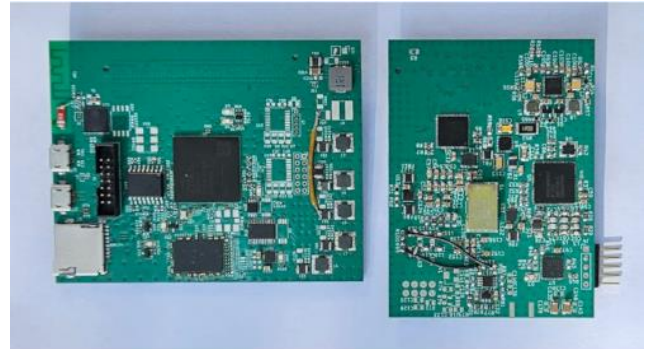


Fig. 1. Development board.

After successful testing of these printed circuit boards, a preliminary version of the device was created. Preliminary tests showed that the generator and sampling circuits are working properly. Future improvements may include improving the fiber optic link between the sensors and the main unit. Since the device is designed to cover the frequency range of lightning currents, other possible uses may include current recording.

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Statistical Analysis and Synthesis of Vehicular Unintended Electromagnetic Emissions Spectra

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Abstract— This paper presents a Gaussian-Process-based framework for synthesizing Unintended Electromagnetic Emission (UEE) spectra from diesel vehicles to mitigate measurement data scarcity in machine-learning classification. Measured spectra are classified into three types by fluctuation intensity, and each spectrum is partitioned at a CUSUM-derived cutoff frequency f_c that separates a high-variance low-frequency region from a low-variance high-frequency region. The Gaussian assumption is verified by Q-Q analysis, with fewer than 5% of empirical quantiles deviating more than 1 dBuV/m from their theoretical quantiles. Fluctuation standard deviations in the low-frequency region exceed those in the high-frequency region by a factor of 3.6–4.6 across all types, with Type 2 and Type 1 exhibiting the strongest and weakest fluctuations in both frequency regions, respectively.

Keywords— Diesel vehicle, Spectral fluctuation, Unintended electromagnetic emissions

I. INTRODUCTION

Unlike electric vehicles, diesel vehicles (DVs) possess fewer electronic components, making their UEE spectra highly dependent on specific vehicle configurations. Predicting these emissions through simulations is often impossible due to a lack of detailed component data. To address this, we treat the emission spectrum as a stochastic process. Because the cumulative emission arises from the superposition of many independent component sources, the Central Limit Theorem motivates a Gaussian-Process model [1]. This allows us to synthesize UEE spectra that retain the spectral characteristics of real vehicles.

II. STATISTICAL ANALYSIS AND SYNTHESIS

To synthesize the signal, the measured spectrum $S(f)$ is decomposed into a deterministic mean $\mu(f)$ and a stochastic fluctuation $\varepsilon(f)$:

$$S(f) = \mu(f) + \varepsilon(f). \quad (1)$$

Vehicles are classified into three types based on their fluctuation intensity. We use the CUSUM (Cumulative Sum) algorithm to identify a cutoff frequency (f_c), which marks the transition between high-variance low-frequency regions and low-variance high-frequency regions of fluctuation intensity [2]. The statistics of f_c reported in [2] show that this transition differs systematically by type.

The fluctuation $\varepsilon(f)$ is assumed to follow a normal distribution. We verify this using Q-Q plots. Analysis shows that deviation from theoretical and empirical quantiles remains negligible: the proportion of points with $D > 1$ dBuV/m, where D is the vertical distance between the empirical points and the reference line of the Q-Q plots, is below 5% [2]. This justifies a Gaussian approximation.

The synthesized spectrum $X(f)$ is sampled from a multivariate normal distribution:

$$X(f) \sim N(\mu, \Sigma), \quad (2)$$

where μ and Σ are the mean and covariance matrix, respectively. μ is estimated from real data. $\Sigma_{ij} = \sigma_i \sigma_j \exp(-|f_i - f_j|/l_c)$, where σ is the standard deviation chosen from Table I and l_c is the correlation length, which is 13.65 MHz, independent of type and frequency region [2]. Examples of realizing the multivariate normal distribution in Eq. (2) are shown in Fig. 1, where differences in standard deviation between types and frequency regions can be observed.

TABLE I. STANDARD DEVIATION STATISTICS OF FLUCTUATION SPECTRUM BY SPECTRUM TYPES AND FREQUENCY REGIONS. UNIT: dBuV/m

		Mean	Median
Type 1	Low	2.097	2.033
	High	0.461	0.452
Type 2	Low	4.048	4.118
	High	1.051	1.059
Type 3	Low	2.510	2.454
	High	0.702	0.675

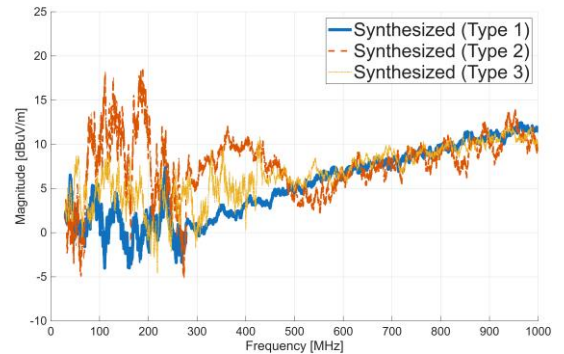


Fig. 1. Examples of synthesized spectra for each type

III. CONCLUSION

The proposed framework reproduces measured DV fluctuation statistics using three empirical parameters — mean $\mu(f)$, frequency-dependent standard deviation σ , and cutoff frequency f_c — enabling scalable synthesis of UEE datasets for classifier training under data-limited conditions.

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Effect Of MOV Ground Impedance On PCI Performance In HEMP Filters

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This paper analyzes the effect of metal-oxide varistor (MOV) ground impedance on Pulse Current Injection (PCI) performance in HEMP power line filters. The ground impedance of the MOV significantly influences current distribution and residual voltage characteristics under high-amplitude pulse conditions. An equivalent circuit model incorporating parasitic inductance and resistance of the grounding path was developed to predict PCI responses. Experimental validation was conducted using PCI test setups in accordance with MIL-STD-188-125 requirements. The results demonstrate that increased ground impedance leads to higher peak residual voltage, thereby degrading PCI performance and highlighting the importance of low-inductance grounding design in HEMP filter implementation.

Keywords– HEMP Power Line Filter, MOV Parasitics, Ground Impedance, Pulse Current Injection (PCI)

I. INTRODUCTION

High-Altitude Electromagnetic Pulse (HEMP) protection systems require robust power line filters to ensure compliance with stringent military standards such as MIL-STD-188-125-1. Under fast transient conditions, such as those represented by PCI single-pulse tests, the grounding path of the MOV plays a critical role in determining the overall surge current distribution and residual voltage behavior. The parasitic impedance introduced by the grounding structure (particularly its inductive component) can significantly affect high-frequency response in the MHz range. In this study, three different grounding configurations for MOV implementation (wire, braided conductor, and busbar structures) were fabricated and evaluated. The impedance characteristics of each grounding type were measured and incorporated into an equivalent circuit model for simulation analysis.

II. METHODOLOGY

Three different grounding conductor configurations were fabricated to evaluate their impact on PCI performance: (1) stranded wire, (2) braided conductor, and (3) busbar, as shown in Fig. 1. All specimens were manufactured with identical length (200 mm) and comparable cross-sectional area (approximately 5 mm²) to ensure consistent electrical comparison.

III. SIMULATION AND DESIGN

Three grounding conductor configurations were prepared, and an equivalent circuit was constructed. The R, L, and C parameters were measured using an impedance analyzer, and the extracted values are summarized in Table I.

TABLE I. EXTRACTED PARAMETERS OF THE GROUNDING PATH

Conductor Type	R_g (m Ω)	L_g (nH)	C_g (pF)
Stranded wire	6.7	130	14.5
Braided Conductor	7.1	92	14
Busbar	20	60	15

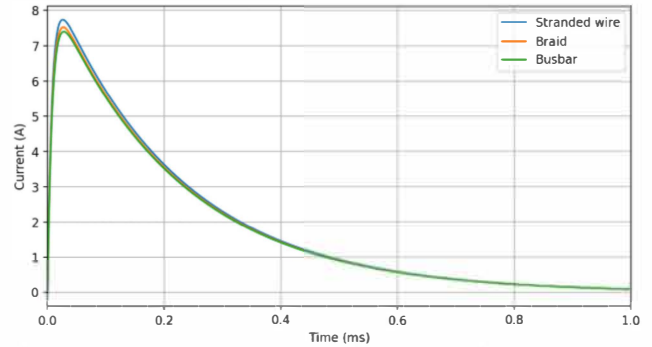


Fig. 1. Simulated transient current waveforms under PCI excitation.

Using the extracted parameters, PCI test simulations were performed. The results are shown in Fig. 1, and the corresponding values are listed in Table II.

TABLE II. EXTRACTED PARAMETERS OF THE GROUNDING PATH

Conductor Type	Peak Current (A)	Risetime (A/s)
Stranded wire	7.73	1.17×10^6
Braided Conductor	7.52	1.09×10^6
Busbar	7.40	1.03×10^6

The busbar configuration, exhibiting the lowest parasitic inductance, resulted in the smallest PCI response. However, all configurations satisfied the criteria specified in MIL-STD-188-125-1.

IV. CONCLUSION

This study investigated the impact of grounding conductor configuration on PCI residual current characteristics in a HEMP protection structure. The results indicate that structural optimization to minimize grounding inductance is more effective for mitigating PCI-induced transient stress than reducing conductor resistance alone, providing important design guidance for HEMP protection systems.

ACKNOWLEDGMENT

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Experimental Analysis of HPEM Susceptibility Thresholds for Electronic Equipment

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Abstract This study experimentally verified the threshold level according to the physical size of the test object, the number of electronic components, and the frequency band of the applied signal. While existing numerical or theoretical analytical methods offer advantages in terms of cost and time, they are predicting threshold levels for arbitrary targets remain challenging. While experimental results cannot provide definitive conclusions, they offer the advantage of representing real-world values and identifying trends. To confirm this, Experimental results were collected and analyzed for various electronic devices.

Keywords– HPEM, Susceptibility, Threshold level

I. INTRODUCTION

High-power electromagnetic environments, such as NEMP or HPM, pose a significant threat to modern electronic infrastructure. Although various numerical analysis methods are used to estimate coupling effects, the vast diversity of target objects and environmental variables (e.g., shielding effectiveness and incident polarization) limits the accuracy of theoretical predictions in practical scenarios. The objective of this study is to analyze experimental data to identify key physical attributes that influence system-level thresholds, thereby improving the predictability of IEMI effect for complex platforms.

II. METHODOLOGY

This study employed an experimental approach to determine the susceptibility thresholds of several representative systems, ranging from small handheld devices to large vehicles. Testing was conducted using two types of HPEM sources: the M30R (mesoband) and T50 (UWB)[1].

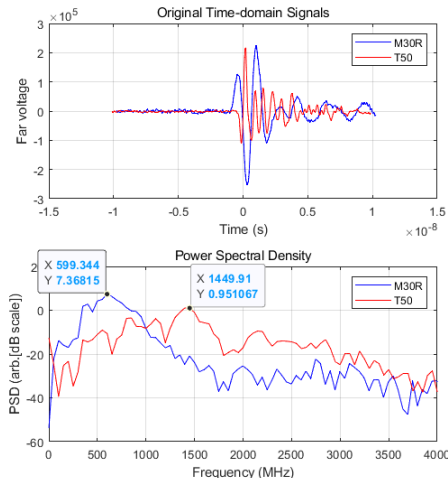


Fig. 1. Waveform and frequency spectrum of M30R and T50

III. RESULTS AND DISCUSSION

The experimental results are summarized in Fig. 2, illustrating the relationship between physical size, components and threshold level.

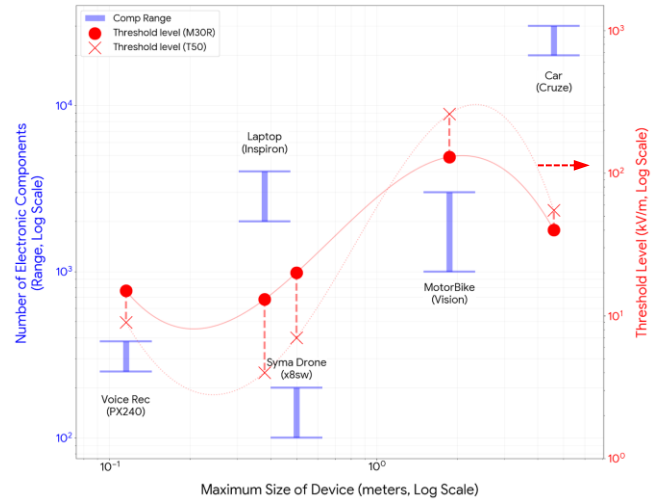


Fig. 2. Threshold level according to physical size, number of electronic components [2],[3] via different HPEM sources.

As shown in Fig. 2, the data reveals the following trends.

Frequency dependency: High-frequency sources (T50) lower susceptibility thresholds in sub-meter devices due to more efficient electromagnetic coupling with compact internal architectures.

Complexity dependency: Among devices of similar size, higher components count significantly heighten vulnerability by providing more potential entry points and coupling paths for interference.

IV. CONCLUSION

This work demonstrates that while theoretical modeling is valuable, empirical trends based on physical attributes provide a more practical overview for susceptibility assessment. Measured results show that system size, component number and frequency band are critical factors in determining HPEM thresholds. It is expected that expanding this dataset to include more diverse subjects will enable the development of predictive method for the HPEM susceptibility of arbitrary target systems.

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Low-Profile Wideband Ka-Band Absorber by Using Dielectric Aperture Tuning

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Abstract– This paper presents a low-profile wideband absorber featuring stable absorption under oblique incidence for electromagnetic interference mitigation in integrated platforms. This proposed absorber employs resistive loops to realize multiple resonances, enabling wideband absorption, and introduces corner holes to improve broadband impedance matching. This invention is expected to offer an effective solution for suppressing EMI in compact form factors.

Keywords– Electromagnetic absorber, Millimeter-wave, Wideband.

I. INTRODUCTION

Recent mmWave (FR2) platforms provide communication capabilities in compact form factors [1]. In order to satisfy these requirements, communication modules such as array antenna and beamformer IC are highly integrated within a compact package. However, the co-located RF modules result in electromagnetic interference (EMI), undesired electromagnetic coupling, leading to degradation of the signal integrity. As an alternative, low-profile electromagnetic absorbers can be a practical solution for dissipating undesired electromagnetic energy at the packaging level [2].

In this paper, a low-profile, single-layer wideband absorber is proposed. This wideband operation can be achieved by combining multi-resonant unit cell geometry, distributed resistive loading, and dielectric aperture tuning, maintaining stable absorption under oblique incidence.

II. LOW-PROFILE WIDEBAND ABSORBER DESIGN

As illustrated in Fig. 1, the proposed absorber unit cell consists of two square loops with resistive sheets, a ground plane, and a low-temperature co-fired ceramic (LTCC) substrate, which has a relative permittivity of 5.9 and loss tangent of 0.002. The unit cell period (p) is 3.7 mm, and the thickness of the substrate (t_{sub}) is 1.0 mm, which are equivalent to $0.29 \lambda_0$ and $0.08 \lambda_0$ at the lowest operating frequency of 23.8 GHz, respectively. The length of the outer loop (l_1) is 2.5 mm, and the inner loop (l_2) is 1.2 mm. The outer loop controls the low-frequency resonance, while the inner loop contributes to the high-frequency resonance. The outer resistive sheet and the inner resistive sheet are set to 90 Ω and 100 Ω , respectively. These different resistance values tune the quality-factor (Q-factor) of each resonance. The multi-resonant geometry generates two distinct resonant modes, resulting in broadband absorption. The corner hole radius (r_{hole}) is 0.8 mm, which lowers the effective permittivity of the dielectric substrate, enhancing broadband impedance matching.

Fig. 2 presents the simulated absorption of the proposed absorber under oblique incidence for transverse electric (TE) polarization. The proposed absorber achieves 90% absorption bandwidth at normal incidence from 23.8 GHz to 41.7 GHz, corresponding to a fractional bandwidth of 54.7%. In addition, the proposed absorber maintains stable absorption up to 45° for TE polarization.

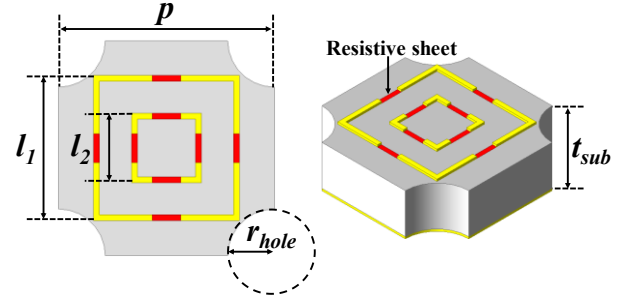


Fig. 1. Schematic of low-profile wideband absorber unit cell.

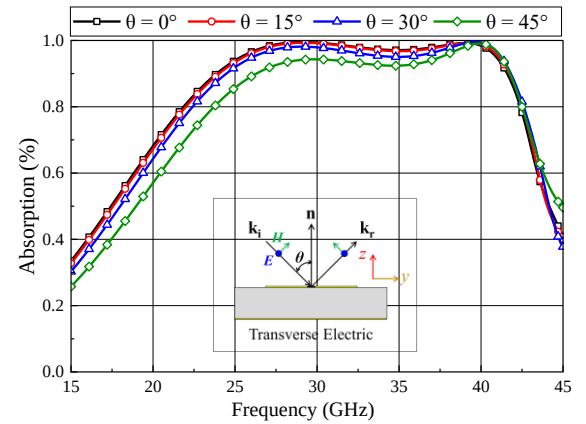


Fig. 2. Simulated absorption under oblique incidence for TE polarization.

III. CONCLUSION

This paper proposes a low-profile wideband absorber using corner hole structure, achieving 23.8 GHz to 41.7 GHz. Side holes in dielectric substrate reduce the effective permittivity for broadband impedance matching. This work features a planar and low-profile structure while maintaining wideband and oblique incidence stable absorption, providing a practical solution for EMI mitigation in compact platforms.

ACKNOWLEDGMENT

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Electromagnetic Noise from a PFN-Marx Generator in High-Power Microwave Systems

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Abstract—Electromagnetic (EM) noise from a gigawatt-level pulse-forming-network (PFN) Marx generator in a high-power microwave (HPM) system is investigated through both simulation and experiment. Frequency spectrum analysis indicates distinct peaks below 1.5 GHz, with field strengths on the order of tens of V/m at a distance of 1 m. The CST simulations exhibited good agreement with experimental measurements, confirming the reliability of the modeling approach. This work provides one of the few experimental validations of CST-based EM noise modeling for PFN-Marx generators and demonstrates that a geometry-based simulation approach can serve as a practical tool for evaluating electromagnetic interference characteristics in HPM system design.

Keywords— EM noise, HPM system, PFN-Marx generator

I. INTRODUCTION

Gigawatt (GW) level pulse generators commonly employed in high-power microwave (HPM) systems are sources of strong electromagnetic (EM) noise due to their rapid switching and high voltage characteristic. [1], [2] The generation of transient EM fields can interfere with surrounding electronic systems; therefore, to ensure system reliability, it is necessary to understand their characteristics. Despite its importance, previous studies have primarily focused on low-voltage systems, and EM noise generation by GW-level pulse generators has not yet been sufficiently characterized. [3] Therefore, we investigated EM noise generated by an HPM system employing a GW-level pulse-forming-network (PFN) Marx generator. The EM noise characteristics were analyzed based on experimental measurements and numerical simulations performed with the time-domain solver of microwave studio in CST Studio Suite. This study provides a comprehensive characterization of EM noise and further presents experimental validation of CST-based EM noise modeling.

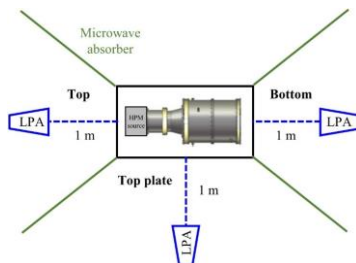


Fig. 1. Schematic of the experimental setup

II. METHODOLOGY

The experiments and CST simulations were performed on the same HPM system. The schematic representation of the experimental setup is shown in Fig. 1. Measurements were conducted at three locations (Top, Top plate, Bottom) positioned 1 m away from the system. The experiments were conducted outdoors, with the device placed on a non-conductive surface. A log-periodic antenna was

employed to measure EM noise in the frequency range of 0.1-1.5 GHz. To minimize interference from microwaves generated by HPM source, measurements were not conducted on the side where microwave radiation occurs. Additionally, microwave absorbers were used to shield the antenna.

The CST model was constructed using the 3D drawings for actual fabrication to accurately reflect the actual physical geometry of the system. The triggering and charging process of the PFN-Marx generator were omitted, and the operation of the high-voltage switches were simplified using voltage ports. [2] The HPM source was simplified as a matched lumped resistor, thereby excluding RF noise generation from the analysis.

III. RESULTS AND DISCUSSION

The spectra of EM noise obtained from the experimental measurements and CST simulations are presented in Fig. 2. The spectra were measured and analyzed for both the vertical and horizontal components. Several distinct peaks with field strengths on the order of several tens of V/m were observed below 1.5 GHz. Spatially, the EM noises were strongest at the bottom, and overall, their horizontal components were stronger than the vertical ones. The CST results showed good agreement with the experimental spectra in terms of magnitude, distribution, and relative noise levels across positions, although some discrepancies were observed.

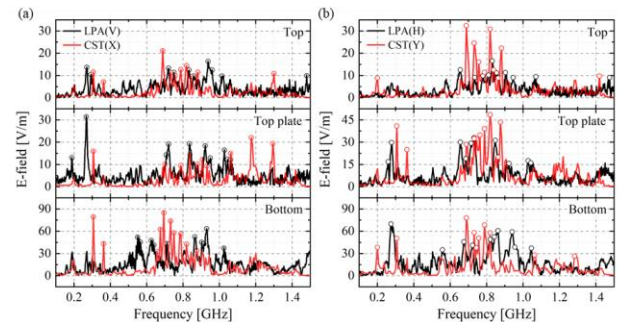


Fig. 2. Frequency spectra of (a) vertical and (b) horizontal components of EM noise. The ten largest peaks in each spectrum are indicated by circles.

IV. CONCLUSION

The results provide valuable input for electromagnetic interference assessment in system construction. Furthermore, by presenting one of the few experimental validations of CST-based EM noise modeling, this work demonstrates that such simulations can serve as a practical tool for HPM system design.

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Rocket Design Evaluation for Rocket Triggered Lightning Experiment in Malaysia

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Abstract– This paper presents a rocket system design for rocket-triggered lightning (RTL) experiment in Malaysia. Test results show a peak thrust of 324 N, burn time of 1.6 s, and total impulse of 369 N·s, indicating stable performance suitable for 200 meters launching conditions with typical cloud base of tropical thunderclouds around 500 meters.

Keywords– rocket-triggered lightning, tropical lightning, thrust measurement.

I. INTRODUCTION

Natural lightning's unpredictability requires artificial triggering for controlled research. Rocket-triggered lightning (RTL), using a rocket to launch conductive wires, is the most reliable method, with 50% success for ground-connected triggering versus 10% for altitude triggering [1]. The first Southeast Asia RTL facility was developed through the SATREPS RTL-3D project between Japan and Malaysia, based on designs from the USA, Japan, China, and Brazil. This paper evaluates RTL rocket design parameters, velocity, thrust, and safety protocols for Malaysia's tropical conditions.

II. METHODOLOGY

The rocket together with its test enclosure was evaluated using a Phantom T2410 high-speed camera and a calibrated HX711-based 200 kg load cell in order to examine its launch characteristics, thrust performance, and operational safety. The high-speed camera was employed to record the ignition process, lift-off, and the initial stage of flight, enabling detailed observation of the rocket motion, stability, and smoke trail development immediately after launch. In parallel, the calibrated load cell was used in the static firing test to obtain the thrust-time response of the KNSB (potassium nitrate-sorbitol) motor for quantitative performance analysis. Fig. 1A shows the rocket equipped with a motor designed for a theoretical altitude of 200 m. Fig. 1B presents the calibrated HX711-based 200 kg load cell used to measure motor thrust. Fig. 1C shows the aluminum-acrylic launcher enclosure integrated with a wireless igniter for safer remote ignition. Fig. 1D shows the smoke trail generated during launch, used to estimate the rocket trajectory and visualize the wire deployment path during ascent.

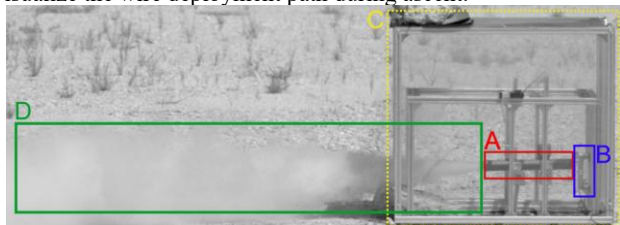


Fig. 1. High-speed camera image from the rocket test at the RTL site in Melaka, Malaysia.

III. RESULTS AND DISCUSSION

Fig. 2 shows the thrust-time history obtained from the static firing test. The load cell was calibrated using known reference masses to ensure measurement accuracy, and the data were recorded directly on a laptop via a USB interface. After ignition, the motor rapidly reached a peak thrust of 324 N, followed by a stable combustion phase and a gradual tail-off over the 1.6 s burn duration. The measured total impulse of 369 N·s places the motor in the Class-I impulse range. The smooth thrust rise and absence of significant oscillations indicate stable combustion behavior. Post-test inspection confirmed that the casing, nozzle, and bulkhead remained intact, demonstrating that the mechanical design withstood the thermal and pressure loads during operation.

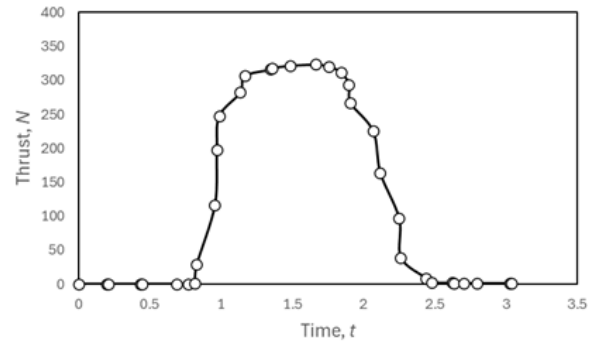


Fig. 2. Thrust-time history of a typical Class-I rocket motor

IV. CONCLUSION

In conclusion, the proposed rocket for rocket-triggered lightning experiments demonstrated stable thrust performance and satisfactory structural integrity during testing. The measured motor characteristics confirm its suitability for operation under Malaysian tropical conditions.

ACKNOWLEDGMENT

This research work is supported by SATREPS project (GERAN ANTARABANGSA-SATREPS/2023/FKEKK/A00052) in collaboration between Japan Science and Technology Agency (JST, JPMJSA2210) and JICA. It is also promoted by Ministry of Higher Education (MOHE) in Malaysia.

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Measurement and Modeling of Connection Process Between the Triggering Wire and Wind Turbine Blade Tip in Altitude-Triggered Lightning

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Abstract—Simultaneous measurements of the connection process between the triggering wire and the tip of the wind turbine blade in altitude-triggered lightning were carried out. The observed wave speed of the connection process was $4.2 \times 10^6 \text{ m/s}$ in two-dimension. Modeling results were compared to the measured electric fields through both the charge transfer modes of return stroke and M-component associated with elevated wind turbine blade structure.

Keywords- altitude-triggered lightning; wind turbine; connection process; electric fields; charge transfer modes.

I. INTRODUCTION

Lightning incidence to modern wind turbines receives great attention due to the large scale structure and the length of the wind turbine blades. Rocket-triggering lightning technique enables lightning to occur at a predictable time and position. Altitude triggering lightning is usually performed through the insulating wire tied to the rocket and the lightning rod on the ground [1]. The insulated gap between the lower extremity of the copper wire and the rod can be used to analyze the leader initiation, development, connection, and related transient effects (e.g., [2]) through simultaneous measurements.

II. RESULTS AND ANALYSIS

In the Summer of 2023, altitude triggering lightning was performed through a height of 20 m long wind turbine blade at triggering lightning site of Conghua, Guangdong Province, China. Simultaneous measurement and comparison of current at the ground level and fast electric fields at 110 m obtained on June 18, 2023, at 13:05:01 were presented in Fig.1. The current was ascribed to the upward positive leader initiated from the wind turbine blade tip and then final bridged successfully in-between the blade tip and lower extremity of the steer wire.

Based on analytical representative of current from measured data, the comparison of the measured data and simulated data were presented in Fig.2. The simulation of return stroke mode was performed by injected the current at the wind turbine blade tip. The transient effects on the electric fields can be found and validated through the charge transfer mode of lightning M-component([3]-[4]).

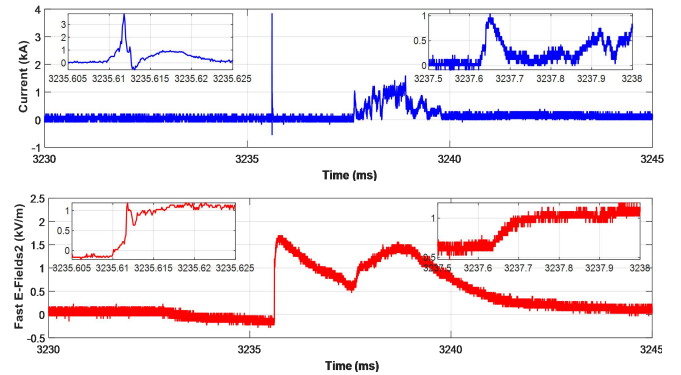


Fig.1 Simultaneous measurement of current and fast electric fields at 110m obtained on June 18, 2023, at 13:05:01(local time).

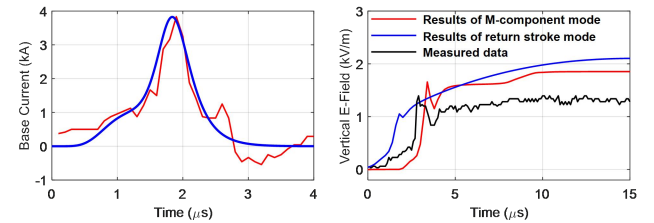


Fig.2 Comparison between the measured data and the analytical representative of current, and the simulated fast electric fields.

III. CONCLUSION

The paper presents simultaneous measurements of the connection process between the triggering wire and the tip of the wind turbine blade in altitude-triggered lightning. The wave speed of the connection process from the wind turbine blade tip was up to $4.2 \times 10^6 \text{ m/s}$ (2D), which is one or two orders of the values than the previous long air gap discharge test. The comparison results demonstrated that the connection process is more suitable to be reproduced through charge transfer mode of M-component.

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Effect of Metasurface Layers on Dual-Elliptical Patch Antenna for Biomedical Applications

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Abstract– This work investigates the effect of single- and double-layer metasurfaces on a dual-elliptical patch antenna for biomedical applications at 1.445 GHz to 4.0 GHz. The proposed double metasurface achieves return losses of -20.159 dB at 2.02 GHz and -36.675 dB at 3.4 GHz. However, the bandwidth decreases from 2.555 GHz to 2.295 GHz with the addition of the metasurface. In contrast, the directivity and gain increase by 1.294 dB and 1.261 dB, respectively, compared to the antenna without a metasurface.

Keywords– dual-ellipse antenna, U-slot, metasurface, biomedical

I. INTRODUCTION

Compact, low-profile wideband antennas are increasingly needed for biomedical applications. FR-4 microstrip patch antennas are widely used for their low profile, light weight, low cost, and RF compatibility [1]. Compared with rectangular patches, elliptical shapes offer smoother current distribution and reduced edge diffraction, improving radiation efficiency and impedance performance [2]. Edge cutting increases electrical length and coupling in dual-elliptical designs, enhancing multiband performance. A metasurface above the antenna improves gain and directivity [3]. This work proposes a dual-elliptical microstrip patch antenna with a U-slot, edge cutting, and metasurface layers.

II. METHODOLOGY

Fig. 1(a) shows the simulated FR-4 dual-elliptical patch antenna ($66\text{ mm} \times 64\text{ mm}$) with a U-slot, edge cuts, and single and double metasurface layers. Fig. 1(b)–1(d) show the fabricated antenna with single and double metasurfaces positioned 15 mm and 30 mm above the antenna.

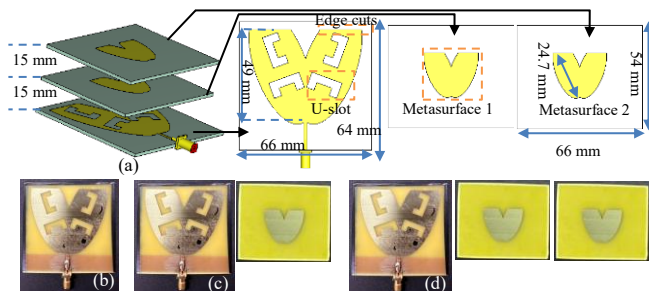


Fig. 1. Antenna with metasurface (simulation), (b) fabricated dual-elliptical microstrip patch, (c) fabricated antenna with single layer of metasurface, (d) fabricated antenna with double layer of metasurface.

III. RESULTS AND DISCUSSION

Fig. 2 compares antenna performance without and with single- and double-layer metasurfaces. The simulated bandwidth spans 1.445–4.0 GHz, while the measured range is 1.315–3.61 GHz. At 2.8 GHz, directivity increases to 5.488 dB and 5.417 dB for single and double metasurfaces, compared to 4.194 dB for the antenna alone, giving a 1.294 dB improvement. Similarly, gain rises to 5.116 dB and 4.848 dB from 3.855 dB, achieving a 1.261 dB enhancement.

This is because the metasurface acts as a superstrate reflector, creating constructive broadside interference while suppressing surface waves and unwanted radiation. At 2 GHz, the antenna without a metasurface shows the highest peak E-field of 127.538 V/m at 15 mm. With single- and double-layer metasurfaces, the field decreases to 29.686 V/m and 28.524 V/m due to reflection, absorption, and phase redistribution of the radiated energy.

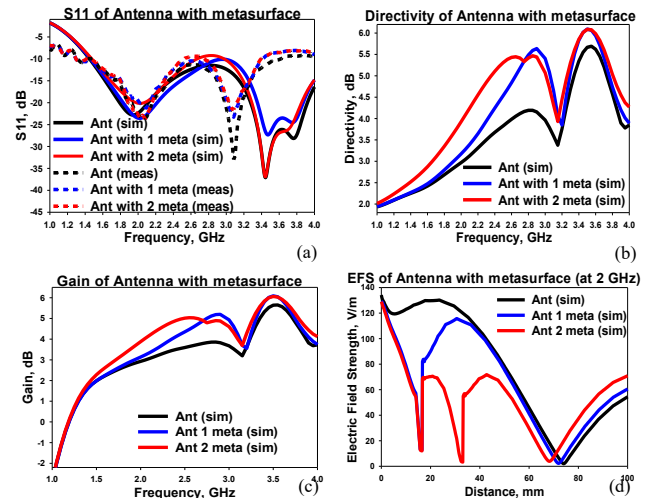


Fig. 2. Performances of the antenna, antenna with single and double layers of metasurface, (a) return loss, (b) directivity, (c) gain, (d) E-field strength.

IV. CONCLUSION

The integration of metasurface structures onto antenna provide the return loss at 2.02 GHz and 3.4 GHz with -20.159 dB and -36.675 dB. However, the bandwidth of antenna reduces from 2.555 GHz to 2.295 dB. In other hands, there are improvement of directivity and gain of 1.294 dB and 1.261 dB, respectively.

ACKNOWLEDGMENT

This research work is supported by SATREPS project (GERAN ANTARABANGSA–SATREPS/2023/FKEKK/A00052) in collaboration between Japan Science and Technology Agency (JST, JPMJSA2210) and JICA. It is also promoted by Ministry of Higher Education (MOHE) in Malaysia.

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Design and Testing of an Air Termination System using a Faraday Cage-Based Lightning-Resistant Drone for Active Lightning Protection

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Abstract– This paper presents the development of an active lightning protection system utilizing a drone-based air termination system. A lightning-resistant drone equipped with a Faraday cage structure is proposed to actively form a Lightning Protection Zone (LPZ). Electromagnetic simulations were conducted to optimize the cage design for both LPZ formation and internal component protection. High-voltage discharge tests confirmed the system's ability to safely capture lightning strikes.

Keywords– Active lightning protection, Drone air termination, Faraday cage, Lightning Protection Zone (LPZ).

I. INTRODUCTION

Current active lightning protection technologies, such as lasers and rocket-triggered lightning, present significant operational challenges. Laser systems require high energy (over 500 MJ) and are highly expensive [1]. To address these issues, this study proposes a new concept: a drone-applied air termination system. By utilizing a drone with hovering capabilities, the system can maintain altitude, move to necessary locations, and be reused, offering superior economic viability. This mobility allows for the rapid formation of a broad Lightning Protection Zone (LPZ) to protect critical national facilities.

II. DESIGN AND FABRICATION OF LIGHTNING-RESISTANT DRONE AIR TERMINATION

To ensure the drone can withstand direct lightning strikes while effectively capturing them, a Faraday cage structure was applied [2,3]. 3D EM simulations (using Maxwell, CST, and EMA3D) were performed to determine the optimal design. The electrostatic field analysis revealed that connecting a down-conductor linearly increases the potential, which is highly efficient for inducing lightning and forming an LPZ. Magnetic field simulations showed that increasing the number of rings in the Faraday cage decreases the internal magnetic field to below 1 mT ($= 795 \text{ A/m}$), protecting the drone's internal electronics. However, since denser cages increase weight and reduce flight performance, an optimized prototype was designed ($\phi 1625 \times h900 \text{ mm}$) keeping the drone weight at 6 kg and the cage under 3 kg.

III. HIGH VOLTAGE DISCHARGE TEST

To empirically validate the lightning resistance and capturing capability of the prototype, high-voltage discharge tests were conducted as shown in Figure 1. The tests were performed by applying a negative polarity 1 MV impulse through a needle electrode. As a result of testing under various conditions, such as mounting the proposed lightning-resistant drone air termination system on grounding rods and insulating rods, it was confirmed that the system successfully captured discharges multiple times.

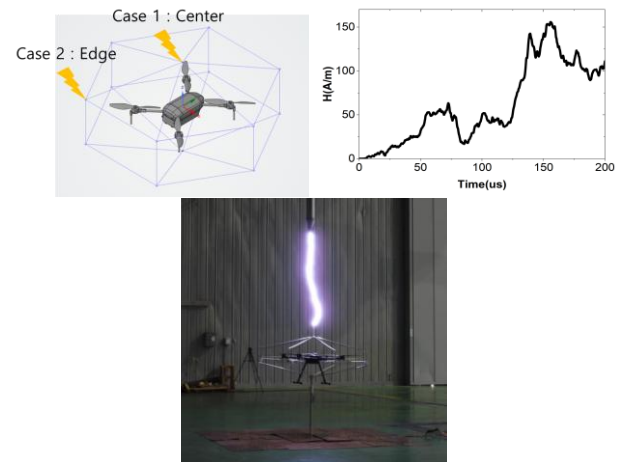


Fig. 1. Electromagnetic simulation of the magnetic field intensity inside the Faraday cage based on the lightning attachment points(top) and high-voltage discharge test of the prototype (bottom).

However, when artificial discharges were applied to a drone without the cage, malfunctions occurred in its internal electronics (e.g., the flight controller board). This indicates that additional insulation and electromagnetic shielding improvements are necessary to protect the drone from direct lightning strikes and lightning electromagnetic pulses (LEMP).

IV. CONCLUSION

The design and preliminary testing of a drone-based air termination system demonstrate significant potential for active lightning protection. The Faraday cage effectively shields internal components while the overall system successfully induces and captures high-voltage discharges. Future work will focus on extensive field flight testing, performance verification of the prototype, and the precise selection of optimal down-conductor materials to ensure complete operational reliability.

ACKNOWLEDGMENT

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A Frequency-Selective Harmonic Blocking Method for Radiated EMI Reduction in WPT Systems

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Abstract– This paper proposes a circuit-level harmonic blocking method to reduce radiated electromagnetic interference (EMI) in wireless power transfer systems employing *LCC–LCC* compensation. By introducing frequency-selective *LC* resonant branches, harmonic currents are diverted away from the transmitter coil at targeted harmonic frequencies. As a result, leakage magnetic fields associated with harmonic components can be effectively reduced while preserving the fundamental power transfer characteristics. The proposed approach provides a compact solution applicable to systems adopting *LCC–LCC* compensation topology.

Keywords– Wireless Power Transfer (WPT), Electromagnetic Interference (EMI), Harmonic Suppression, Resonant Circuits.

I. INTRODUCTION

Wireless power transfer (WPT) systems generate leakage magnetic fields due to magnetic coupling between transmitter and receiver coils, which may interfere with nearby electronic devices. One of the dominant sources of electromagnetic interference (EMI) in WPT systems is harmonic currents generated by switching converters.

Although resonant compensation networks efficiently transfer the fundamental component, harmonic currents can still flow through the coils and become a major source of radiated EMI. Conventional mitigation methods such as shielding plates, auxiliary coils, or filter circuits often increase system size or introduce additional losses [1]–[2].

Therefore, a circuit-level solution that suppresses harmonic currents without additional shielding structures is desirable. This paper proposes a frequency-selective harmonic blocking method for systems employing *LCC–LCC* compensation, which prevents harmonic currents from flowing through the power transfer coils.

II. PROPOSED METHOD FOR REDUCING HARMONIC CURRENT

Fig. 1 shows the proposed *LCC–LCC* system with additional *LC* resonant branches for harmonic current suppression. Fig. 1(a) illustrates the overall WPT system, while Fig. 1(b) shows the transmitter-side circuit extracted from the proposed topology for harmonic current analysis.

$$\omega_0 = \frac{1}{\sqrt{L_f C_f}} \quad (1)$$

When (1) is satisfied, the *LC* branch behaves as a short circuit at the operating frequency and does not affect the fundamental operation of the WPT system.

To suppress harmonic currents, the filter inductance is determined as (2).

$$L_{f(2m+1),TX} = \frac{mL_{S1}}{(2m+1)^2 - 1} \quad (2)$$

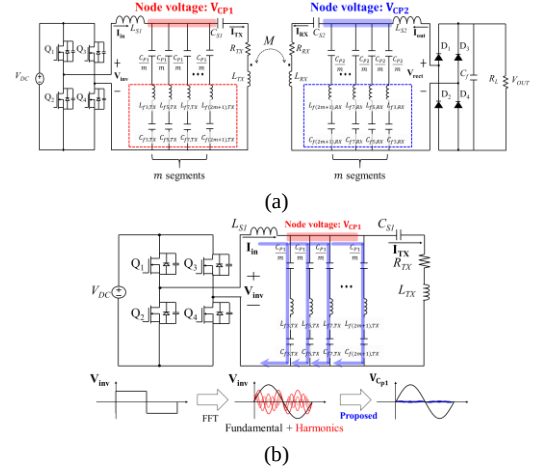


Fig. 1. Proposed WPT system for harmonic current reduction. (a) proposed *LCC–LCC* system with additional *L_fC_f* branches and (b) transmitter-side equivalent circuit extracted from the proposed topology.

where *m* represents the number of targeted harmonic frequencies.

When both (1) and (2) are satisfied, the impedance formed by $\frac{C_p}{m}$, *L_f*, and *C_f* becomes nearly zero at harmonic frequencies, as illustrated in Fig. 1(b). Consequently, harmonic currents circulate through the *LC* branches instead of flowing through the transmitter coil, effectively reducing leakage magnetic fields associated with harmonic components.

III. CONCLUSION

This paper presented a frequency-selective harmonic blocking method for reducing radiated EMI in systems employing *LCC–LCC* compensation. By introducing *LC* resonant branches tuned to harmonic frequencies, harmonic currents are diverted away from the transmitter coil. As a result, leakage magnetic fields associated with harmonic components are effectively reduced while preserving the fundamental power transfer characteristics.

ACKNOWLEDGMENT

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Hardness Surveillance of the HEMP/IEMI Shielded Doors and Filters

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Abstract– A detection system for electromagnetic leakage from shielding doors was designed, implemented, and rigorously validated at 1 GHz through comprehensive simulation and experimental verification, demonstrating a dynamic range in excess of 90 dB. Furthermore, a dedicated detection system for assessing degradation in HEMP/IEMI filters was developed and evaluated at 10 MHz, achieving a dynamic range exceeding 80 dB. The integration of these two systems led to the development of a Hardness Surveillance System capable of simultaneously monitoring potential degradation of the HEMP/IEMI shielding doors and filters.

Keywords– HEMP/IEMI, shielding doors, filters, Hardness Surveillance System, electromagnetic leakage

I. INTRODUCTION

The maintenance of HEMP/IEMI-shielded facilities involves more than routine equipment inspection and represents a critical requirement for sustaining facility survivability and operational continuity. Proper maintenance is essential for preserving shielding effectiveness and the operational continuity of critical equipment, as well as for maintaining reliability and compliance with applicable military standards, including MIL-STD-188-125. This paper presents a monitoring system capable of simultaneously detecting degradation in the most vulnerable and frequently used doors and filters within a HEMP/IEMI shielded room.

II. EM LEAKAGE AND FILTER INSERTION LOSS MONITORING SYSTEM

In this section, EM leakage detection systems are discussed. The proposed method is relatively simple yet effective compared with other methods [1]–[3]. Fig. 1 - Fig.3 illustrate the measurement setup, simulation results, and experimental insertion loss results. Fig. 1 presents the developed antenna and its installation on the inner and outer frames of the shielding door. Fig. 2 presents the simulated shielding effectiveness for different hole sizes and configurations. As the hole size decreases, the insertion loss increases by approximately 15 dB, indicating that the system can effectively monitor changes in shielding performance. Fig. 3 shows the difference between the maximum received signal levels measured with the shielding door open and closed. The results demonstrate an operating dynamic range exceeding 90 dB at 1 GHz. Fig. 4 illustrates a system for monitoring variations in filter insertion loss, defined as the difference in the received signal level between a direct probe connection and a connection through the filter. The dynamic range exceeds 80 dB.

III. INTEGRATED HARDNESS SURVEILLANCE SYSTEM

Fig. 5 presents an integrated system capable of measuring variations in door shielding effectiveness and the level of filter degradation in a HEMP/IEMI shielding facility. This configuration enables simultaneous, real-time monitoring of potential degradation in multiple shielding doors and filters.

IV. CONCLUSION

The method proposed in this paper enables effective, practical, and real-time monitoring of degradation in shielding doors and filters for HEMP/IEMI protection.



Fig. 1. Set up for EM Leakage Measurement

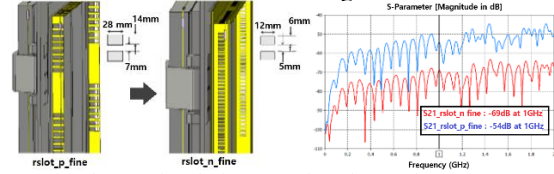


Fig. 2. Simulation Results of Insertion Loss

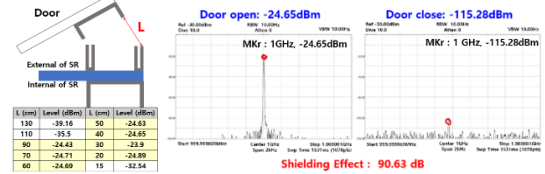


Fig. 3. Measurement Results of Insertion Loss

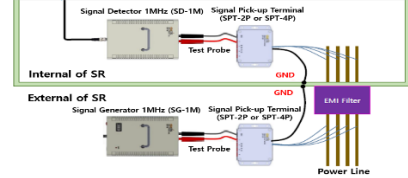


Fig. 4. Filter Characteristics Measurement System

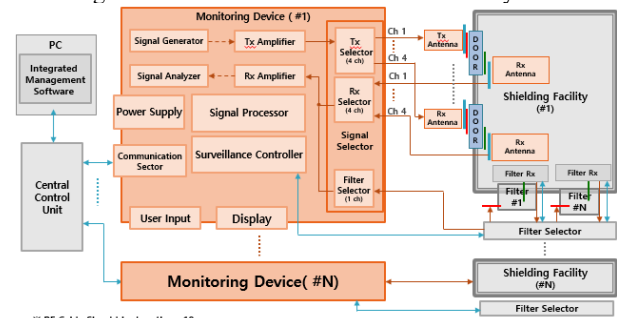


Fig. 5. Hardness Surveillance System of Shielding Facility

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Comparison of Power Supply Conducted Susceptibility Under Quotient-of-Exponential and Damped Sinusoidal Excitations

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Abstract— This paper compares susceptibility thresholds of switched-mode power supplies (SMPS) subjected to quotient-of-exponential (QE) and damped sinusoidal (DS) excitations induced by the coupling of a High-altitude nuclear ElectroMagnetic Pulse (E1 HEMP) on power lines. Destructive tests revealed waveform-dependent failure mechanisms. A statistical comparison of susceptibility levels shows that susceptibility results for QE and DS are not directly comparable, with no consistent trend in the parameters of the associated log-normal distributions.

Keywords— Susceptibility tests, E1 HEMP, log-normal distributions, quotient-of-exponential, damped sinusoidal.

I. INTRODUCTION

Intentional electromagnetic threats, such as High-altitude nuclear ElectroMagnetic Pulse (HEMP), can induce harsh stresses in electrical networks, potentially damaging connected devices [1,2]. Electronic equipment must therefore be tested with different waveforms depending on its network location: QE for devices directly connected to the power network, and DS for devices downstream of HEMP protection devices. Test campaigns on SMPS using both QE and DS injections were conducted to analyze the influence of waveform on susceptibility thresholds in order to reduce the number of required tests if comparison is possible.

II. EXPERIMENTAL SETUP

Power supply functions are critical nodes in low-voltage networks due to their widespread use [2]. SMPS were selected as Equipment Under Test (EUT), with two models covering distinct power categories: 65 W and 15 W. During tests, the SMPS operated under nominal conditions with resistive loads matched to their rated power. A consistent experimental setup (isolated transformer, common-mode choke, capacitive coupler, etc.) was used to minimize variability. Two injection modes were considered: Total Common Mode (TCM – both phase and neutral relative to ground) and Unitary Common Mode (UCM – phase relative to ground). Two sources were used to generate 10/500 ns QE waveforms and DS waveforms at 30 kHz and 100 kHz, with other frequencies possible in future work. Finally, the generators were calibrated on a 30 Ω reference load to allow comparison of susceptibility thresholds across equipment without considering their impedances or nonlinear effects.

III. RESULTS AND DISCUSSION

During tests, both with QE and DS waveforms, the main effects were of different natures: high voltages causing arcing between input traces or dielectric breakdown in differential-mode capacitors; high currents leading to the failure of surface-mount resistors. As shown in Fig. 1, for QE injections, the current increases linearly with voltage (in purple), whereas for DS injections, the current increases sharply beyond a certain voltage, resulting in strong

nonlinearity (in blue). This difference is due to the SMPS's input common-mode impedance, which behaves like a capacitance: lower-frequency disturbances (DS) result in smaller currents, favoring arcing above a voltage threshold.

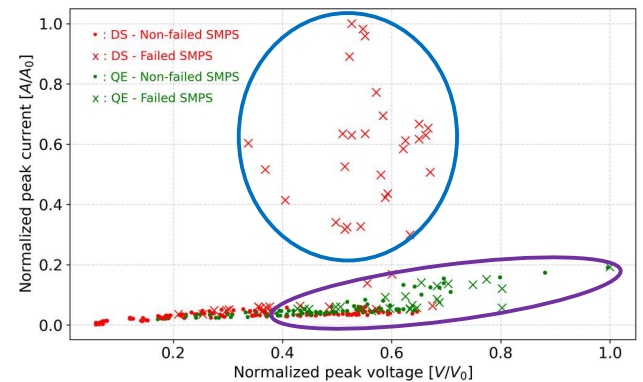


Fig. 1. Peak currents versus peak voltages measured at the power supply inputs for all injections.

Nevertheless, susceptibility distributions were compared for both waveform types, for each SMPS model and injection mode, using injected current levels normalized to 30 Ω . The posterior distributions of the log-normal susceptibility parameters μ (mean) and σ (standard deviation) were obtained via Bayesian inference for QE, DS at 30 kHz, and DS at 100 kHz for each configuration and superimposed for comparison. The results indicate that μ and σ are not directly comparable between waveform types, with no consistent trend across injection modes or SMPS models.

IV. CONCLUSION

The influence of waveform shape on susceptibility thresholds was studied using conducted QE and DS injections. Failure mechanisms differed with the waveform type. Comparison of susceptibility distributions for both SMPS revealed no consistent trend. The estimated means and standard deviations were not directly comparable, regardless of injection mode, waveform type, or SMPS power, indicating that results obtained with QE waveforms cannot be extrapolated to DS injections.

ACKNOWLEDGMENT

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The Effect of RF-DEW Pulse Characteristics on LNA Immunity

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Abstract– Radio Frequency Directed Energy Weapons (RF-DEW) are becoming increasingly relevant in modern battlespace, because of the rise of unmanned aerial systems (UAS). One of the coupling mechanism of RF-DEW is through the RF receive chain. A Low Noise Amplifier (LNA) is essential. This is the first part of this RF-coupling chain and could be damaged by the RF-DEW.

This study focuses on the effect of the duration and recurrence of the incoming RF pulse on the failure point of the LNA. When the LNA fails, the receiver chain will no longer be functional.

Eight modern LNAs have been stress tested under various pulse lengths and repetition rates with increasing envelope power until failure. The experiments show that a significantly higher peak-envelope power is required for 1 μ s pulses than for longer pulses (5 μ s or 10 μ s). A pulse train of 10 pulses resulted in a lower required peak-envelope power compared to a single pulse, although the reduction was modest.

Keywords– Directed energy weapons, low noise amplifier, damage mechanism, pulse duration, pulse repetition

I. INTRODUCTION

Radio Frequency (RF) Directed Energy Weapons (DEW) are a means of disrupting or permanently damaging a target by means of high intensity RF. The rise of drones on the battlefield has increased interest to explore capabilities of RF-DEWs to counter drones.

One essential part of effective deployment of the RF-DEW system is knowledge of the coupling path of the RF energy into the target system, and the power or energy level needed to disrupt the targets. An antenna on the target could be an effective way to coupling RF-energy into the target system. In the RF receive chain the Low Noise Amplifier (LNA) is assumed to be sensitive to RF-energy [1]. The damage level of a LNA as a function of the DEW's pulse shape is instrumental information to predict the effectiveness of an RF-DEW system. An initial estimation on the damage level of LNAs at 1GHz was investigated in [2]. This information should be extended by the effect of various pulse shape parameters on the damage level.

This research focuses on the immunity of LNAs to different RF-DEW pulses and is aimed at giving an insight into the effects of pulse length and pulse repetition on modern LNAs.

II. METHODOLOGY

A. Test procedure

The test procedure consists of directly injecting a pulsed 1 GHz burst into the LNAs. The pulses have varying lengths (1 μ s, 5 μ s, 10 μ s) and consist of either 10 pulses (PRF = 10 kHz) or a single pulse, resulting in 6 unique test sets. Each set is repeated at least 3 times for increased statistical significance. The peak power of the pulses is increased in 1 dB steps until the LNA fails.

B. Components

Each test set consists of 8 different LNAs, which have been selected for a variation of manufacturers, technologies and supply power. In total 195 test samples were used. The selected LNAs can be seen in TABLE I.

TABLE I. PROPERTIES OF THE TESTED LNAs

LNA	Manufacturer	Technology	Supply power
BGA2818	NXP	Si	3.3 V / 20 mA
BGB 707	Infineon	SiGe:C	3.3 V / 10 mA
Gali 52	Minicircuits	InGa	4.4 V / 50 mA
HMC376	Analog Devices	GaAs	5.0 V / 73 mA
HMC8411	Analog Devices	GaAs	5.0 V / 55 mA
MAX2612	Analog Devices	Si	5.0 V / 69 mA
PMA2-33	Minicircuits	Si	3.3 V / 56 mA
QPA0163	Qorvo	SiGe	3.3 V / 14 mA

III. RESULTS AND DISCUSSION

The results of all tests are given in Fig. 1. The figure shows the peak input power required for LNA failure for a certain test pulse train. The values present the median value over the 8 used LNAs and the repetitions for a certain test set. The individual LNAs failure points show similar trend as presented in Fig. 1.

The power required to damage an LNA with a 1 μ s pulse is 2.5–3 dB higher compared to a 5 μ s pulse. The difference between 5 μ s and 10 μ s is less prominent.

The difference in failure power between 1 pulse and 10 pulses is only significant for the 1 μ s pulses as the power difference for the 5 μ s and 10 μ s pulses is within the error margin. This correlates with inspection of individual LNA output waveforms, which frequently shows degraded performance during the first pulse at the failure power.

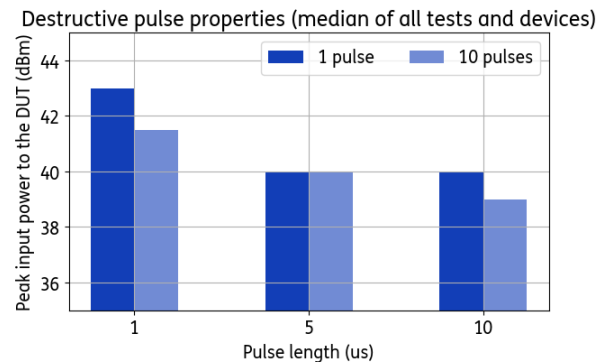


Fig. 1: Peak input power required to damage LNAs for 1 and 10 pulses with 1 μ s, 5 μ s, and 10 μ s pulse lengths.

IV. CONCLUSION

The results show a correlation between the pulse length and the immunity of the LNAs, it shows an initial decline of peak power needed for LNA failure as the pulse length increases from 1 μ s to 5 μ s. However, after 5 μ s, this effect saturates, and 10 μ s pulses require almost the same peak input power. The repetition of pulses lowers the required peak input power, however most LNAs break at the first pulse. From this experiment, it can be concluded that a single 5 μ s pulse is the most effective pulse to damage an LNA with an RF-DEW.

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A Three-Axis Sensor for Direction-Independent Detection of High-Power Electromagnetic Pulses

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Abstract – This paper presents a three-axis Active Electromagnetic Protector (AEP) sensor for detecting high-power electromagnetic pulse (HPEM) threats. The proposed orthogonal structure provides direction-independent sensitivity. Simulation results verify reliable detection capability for arbitrary incident directions and polarizations based on detectable electric field strength, demonstrating robust sensing performance.

Keywords – HPEM, EMP detection, three-axis sensor, electromagnetic pulse sensor, direction-independent sensing

I. INTRODUCTION

High-power electromagnetic pulse (HPEM) threats can induce severe transient disturbances in electronic systems. Conventional EMP sensors often exhibit directional and polarization sensitivity, which limits their effectiveness against arbitrarily incident electromagnetic waves [1]–[3]. Recent approaches, including electro-optic and MEMS-based sensors, have also been studied, but limitations remain in directional coverage and complexity. This paper proposes a compact three-axis Active Electromagnetic Protector (AEP) sensor designed to detect incident EMP fields from all directions [4]. The sensor performance is evaluated in terms of detectable electric field strength (V/m), enabling direct assessment of its capability to sense transient electromagnetic disturbances regardless of propagation direction.

II. METHODOLOGY

The proposed methodology consists of EMP sensing and output signal conditioning. The proposed sensing structure was modeled and simulated to evaluate its response under nuclear and non-nuclear electromagnetic pulse excitation. The induced voltage can be expressed as

$$V_{ind}(t) = h_{eff} E_{inc}(t) \quad (1)$$

Full-wave simulations were conducted using transient plane-wave excitation to evaluate the sensor response under nuclear and non-nuclear EMP conditions. Detection performance was assessed based on whether the induced voltage exceeded the threshold required by the signal processing circuitry, where the threshold was determined from the minimum operating level of the signal processing board.

III. RESULTS AND DISCUSSION

The performance of the proposed AEP sensor was evaluated through full-wave simulations to verify its direction-independent detection capability. As shown in Fig. 1, the proposed sensor is implemented on an FR-4 substrate with a three-axis orthogonal configuration to achieve direction-independent sensitivity. Fig. 2 illustrates representative angular responses for different polarizations in the azimuth and elevation planes. Although the response varies depending on incident direction and polarization, no complete nulls are observed. The minimum induced voltage across all tested conditions is approximately 8.5 V, while higher responses

are obtained in other orientations. This confirms reliable direction-independent detection capability, with the minimum induced voltage remaining above the detection threshold for all tested conditions. In addition, the minimum level exceeds the signal processing board requirement (approximately 5 V), providing stable signal delivery with sufficient margin under all tested directions and polarizations.

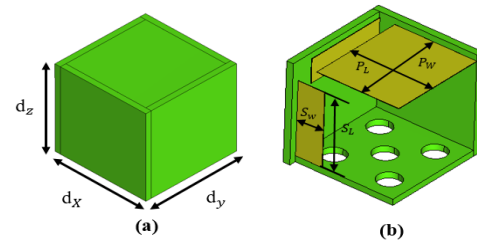


Fig. 1. Geometry of the proposed detection sensor: (a) external view, (b) internal geometry. $d_x=10$, $d_y=10$, $d_z=10$, $P_L=7.4$, $P_W=8.4$, $S_L=7.4$, $S_W=2.4$. (mm)

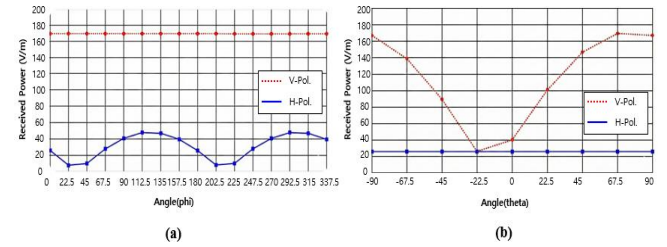


Fig. 2. Simulated induced voltage versus angle for vertical and horizontal polarizations: (a) azimuth directions, (b) elevation directions.

IV. CONCLUSION

This paper presented a three-axis Active Electromagnetic Protector (AEP) sensor for detecting high-power electromagnetic pulse (HPEM) threats. Simulation results verified that the sensor maintains detectable sensitivity over the tested directions and polarizations, demonstrating its suitability for direction-independent EMP detection. Future work will include prototype implementation and experimental validation of the proposed sensor.

ACKNOWLEDGMENT

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Design of Electromagnetic Wave Absorbers Using Sub-skin-depth Conductive Films

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Abstract– This study examines an optimized multilayer structure that enhances electromagnetic absorption by exploiting ohmic losses in sub-skin-depth conductors. While a single resistive sheet is theoretically limited to 50% absorption, this work surpasses that limit using a resonant stack comprising a thin graphite layer, an FR4 dielectric, and a copper ground plane. Simulations at 24 GHz indicate that optimal performance occurs with a graphite sheet resistance of 377 Ω/sq and an FR4 thickness of 1.805 mm. These findings provide essential design parameters for high-performance shielding and absorption in the K-band.

Keywords– Sub-skin, conductor, graphite, ohmic, Salisbury

I. INTRODUCTION

Ohmic loss in sub-skin-depth conductors is pivotal for high-efficiency absorption [1]. While single sheets are limited to 50% absorption, Salisbury Screens achieve total absorption via phase cancellation [2]. This work establishes a design framework for 24 GHz absorbers by incorporating guide wavelength (λ_g) effects to minimize discrepancies between theoretical and waveguide-based models.

II. METHODOLOGY

In this Section, the exploitation of Ohmic losses in sub-skin-depth conductors serves as a fundamental mechanism for high-efficiency electromagnetic wave absorption. Despite its utility, a standalone resistive sheet immersed in free space ($\eta_0 \approx 377 \Omega$) is inherently constrained by the physical trade-off between reflection and transmission Eq. (1) [3].

$$\Gamma = \frac{-\eta_0}{2R_s + \eta_0}, \quad T = \frac{2R_s}{2R_s + \eta_0} \quad (1)$$

Under the critical impedance matching condition of $R_s \approx 188.5 \Omega/\text{sq}$, the power absorption coefficient Eq. (2) reaches a theoretical maximum of 50%, with the remaining power equally dissipated as reflection (25%) and transmission (25%) [1].

$$A_{\max} = 1 - |\Gamma| - |T| = \frac{2\eta_0^2}{4\eta_0^2} = 0.5(50\%) \quad (2)$$

To surpass this inherent 50% absorption limit, a Salisbury Screen configuration—comprising a resistive film, a dielectric spacer, and a conducting ground plane—is employed. The ground plane re-reflects transmitted energy, and the dielectric thickness is adjusted to impose a 180° phase reversal, enabling destructive interference and full absorption at the target frequency [3]. The implemented structure uses a 377 Ω/sq graphite sheet, an FR4 layer, and a metallic backing of sufficient thickness to exceed the skin depth [2]. Under standard free-space conditions, the quarter-wavelength matching thickness is calculated as $d_{\text{free}} = \lambda_0/4\sqrt{\epsilon_r}$ at 24 GHz, this formula yields a thickness of approximately 1.5 mm. However, this calculation is insufficient for practical applications as it neglects the electromagnetic boundary conditions imposed by the waveguide walls. When the TE₁₀ mode propagates within a rectangular

waveguide, the effective wavelength is elongated due to dispersion. The guide wavelength is defined by the following relationship $\lambda_g = \lambda_0/\sqrt{1 - (f_c/f)^2}$ for a WR-42 waveguide, the cutoff frequency (f_c) is approximately 14.1 GHz. This elongation effect shifts the resonance frequency, necessitating a thicker dielectric layer to maintain the phase cancellation condition at the target frequency. The precise optimal thickness is determined by calibrating λ_g with the relative permittivity (ϵ_r) of the substrate in Fig. 1. Numerical analysis confirms that the S₁₁ absorption peak is precisely aligned at 24 GHz when the thickness is adjusted to 1.805 mm. This result demonstrates that waveguide-specific compensation is essential for mitigating the frequency shift typically encountered in high-frequency absorber characterization.

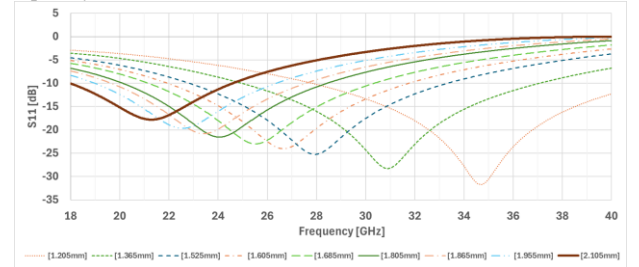


Fig. 1. Simulation results S₁₁ response of the multilayer structure with varying FR4 thickness

III. CONCLUSION

Numerical simulations confirm that the optimized multilayer absorber attains a reflection-loss minimum of −21.5 dB at 24.0 GHz. Employing a waveguide-specific thickness of 1.805 mm and a graphite sheet resistance of 377 Ω/sq effectively eliminates the dispersion-induced frequency shift that would occur with the 1.5 mm free-space design. These results establish a concise yet robust framework for sub-skin-depth absorber design and highlight the importance of waveguide-compensated modeling for accurate high-frequency performance, with future validation to be conducted using a WR-42 measurement setup.

ACKNOWLEDGMENT

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Measurement of Electromagnetic Attenuation of Construction Materials in the HEMP Frequency Band

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Abstract– This study proposes a resonator-based measurement system for evaluating the electromagnetic attenuation characteristics of construction materials in the high-altitude nuclear EMP (HEMP) frequency band. An LC resonant antenna and a compact cavity were employed to establish a stable evanescent-mode measurement environment. Experimental results demonstrated a dynamic range exceeding 59 dB at 1 MHz. In addition, Low-E glass exhibited limited attenuation below 100 MHz, while the attenuation characteristics of reinforced concrete were predominantly governed by the embedded rebar.

Keywords– HEMP, EMP, compact cavity, LC resonant antenna

I. INTRODUCTION

Electromagnetic pulse (EMP) poses significant threats to critical infrastructure systems requiring quantitative evaluation of the electromagnetic attenuation characteristics of building structures for effective protection. However, reliable measurement techniques in the low-frequency band, particularly in the high-altitude nuclear EMP (HEMP) region, have not yet been fully established [1], [2]. This paper proposes a resonator-based measurement method for assessing the electromagnetic attenuation characteristics of construction materials in the HEMP band and experimentally validates the approach using reinforced concrete and glass specimens.

II. METHODOLOGY

According to IEEE Std. 299, low-frequency shielding measurements are conducted under near-field conditions, where the magnetic field dominates below 20 MHz [3]. Conventional 30 cm loop antennas exhibit limited sensitivity at low frequencies, and general antenna structures become impractically large due to long wavelengths. To address these limitations, an LC resonant antenna with 24 discrete resonance points was designed to operate over the 10 kHz – 160 MHz frequency range.

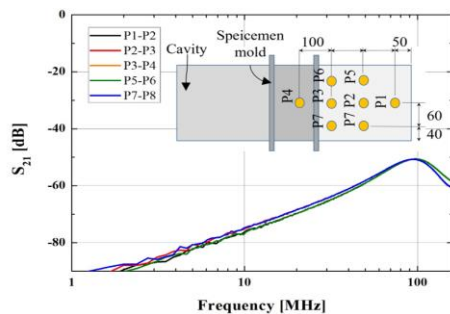


Fig. 1. Variation of S_{21} with cavity position (> 1 MHz)

A compact square waveguide resonator with a 200 mm x 200 mm cross-section was designed, considering typical rebar spacing for specimen installation. The cutoff frequency was calculated to be approximately 750 MHz. Since the maximum measurement frequency (160 MHz) is well below this value, the structure operates in an evanescent mode. The LC resonant antenna was placed inside an electrically small cavity, and as shown in Fig. 1, measurements

confirmed that no internal resonant modes are formed. As a result, spatial variation in field magnitude is minimized, enabling stable and repeatable attenuation measurements in the low-frequency band.

III. RESULTS AND DISCUSSION

Based on the proposed structure, a measurement system for the HEMP frequency band was implemented. Experimental results confirmed that an operating dynamic range exceeding 59 dB at 1 MHz can be achieved. The developed system was applied to characterize the electromagnetic attenuation properties of various construction materials, including glass, reinforced concrete, plain concrete, and rebar specimens as shown Fig. 2. For the glass samples, Low-E glass with thicknesses of 18 mm and 24 mm was fabricated and evaluated. The results indicate that negligible attenuation was observed below 100 MHz, while attenuation of approximately 10 dB was measured near 100 MHz. In the case of reinforced concrete, the measurements demonstrated that, within the HEMP frequency band, the shielding performance is predominantly governed by the embedded rebar rather than by the plane concrete.

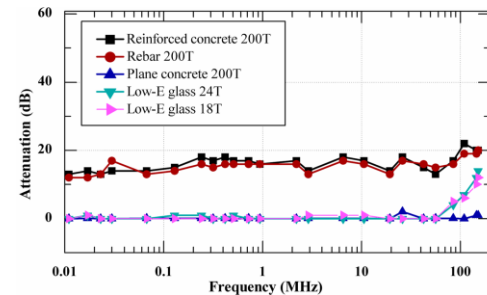


Fig. 2. Comparison of attenuation measurements by specimen type

IV. CONCLUSION

This study proposed a resonator-based measurement system applicable to the HEMP low-frequency band. By employing evanescent-mode operation and an electrically small cavity structure, the influence of internal resonant modes was eliminated, enabling a stable measurement environment. Experimental results demonstrated sufficient sensitivity in the low-frequency region and confirmed that rebar predominantly governs the electromagnetic attenuation characteristics of reinforced concrete. Future work will include comparison with other measurement methods, evaluation of repeatability and measurement uncertainty, and electromagnetic analysis of the test configuration to further improve the reliability of the proposed measurement system.

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Influence of Pantograph Cutting Magnetic Induction Lines on Traction Power Supply System

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Abstract—To investigate the impact of dynamic electromotive force generated by the pantograph cutting magnetic induction lines around the 27.5 kV contact wire on the traction power supply system, this paper proposes a train-network coupling model based on electromagnetic induction theory. This model determines the induced electromotive force at different operating speeds and its propagation characteristics within the traction system. Simulation results reveal the relationship between train speed and the amplitude and distribution of currents in each traction network line. This research provides a reference for electromagnetic interference assessment and traction system design.

Keywords—electrified railway, traction power supply system, pantograph, electromagnetic induction, magnetic induction lines

I. INTRODUCTION

During electrified train operation, the pantograph inevitably cuts the magnetic induction lines around the contact wire, generating a dynamic electromotive force (EMF). This force can couple into the traction power network, degrading power quality and interfering with signaling and communication systems^{[1],[2]}. To quantify the EMF and evaluate its impact on various lines of the traction power supply system, this paper proposes a train-network coupling simulation model. Its key contribution is revealing the relationship between train speed and the amplitude of currents in traction network lines. This study fills a research gap and provides reference for electromagnetic interference assessment in electrified railways.

II. METHODOLOGY

According to Faraday's law of electromagnetic induction^[3], when a conductor cuts magnetic induction lines while moving through a magnetic field, the induced dynamic electromotive force is proportional to the magnetic flux density, the effective conductor length, and the velocity. The distance from the contact line varies along the pantograph, leading to corresponding variations in magnetic induction intensity. In this study, the pantograph is discretized into infinitesimal segments of length dL . The total dynamic electromotive force generated by the entire pantograph cutting the magnetic induction lines is obtained by integrating the contributions from each segment, as in (1).

$$E = \int_r^n \frac{\mu_0 I v}{2\pi L} dL, \quad (1)$$

where n is the vertical distance between the train roof and the contact line, r is the radius of the contact line, μ_0 is the vacuum permeability, I is the current in the contact line, v is the train operating speed, and L is the length of the pantograph.

The various lines in the traction power supply system are modeled using a chain circuit model based on multiconductor transmission lines. The dynamic EMF is equivalently represented by a voltage source, which is placed between the contact line and the train. This yields the coupled train-network model shown in Fig.1.

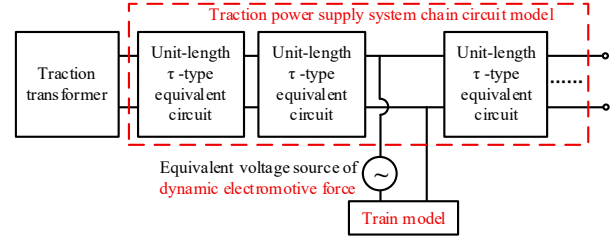


Fig. 1. Simulation model for pantograph cutting magnetic induction lines.

III. RESULTS AND DISCUSSION

The values of the dynamic electromotive force at different train speeds are shown in Fig. 2, and the current amplitudes in each line of the traction power supply system are presented in Table I.

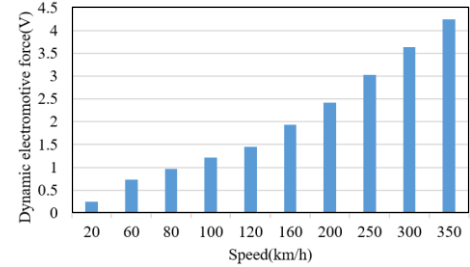


Fig. 2. Relationship diagram between dynamic EMF and train speed.

TABLE I. CURRENT AMPLITUDE AT DIFFERENT SPEEDS

Speed (km/h)	Current amplitude (A)				
	Contact line	Feeder	protective wire	Ground wire	Rail
20	67.6304	67.5073	$3.03172553e^{-2}$	0.500429	20.9227
100	67.6325	67.5094	$3.03172595e^{-2}$	0.500444	20.9233
200	67.6352	67.5119	$3.03172649e^{-2}$	0.500462	20.9241
300	67.6378	67.5145	$3.03172702e^{-2}$	0.500480	20.9249
350	67.6391	67.5158	$3.03172728e^{-2}$	0.500489	20.9253

IV. CONCLUSION

Results show that as train speed increases, the dynamic EMF rises, leading to increased current amplitudes in all traction network lines. Among these, the currents in the contact wire and feeder increase significantly, while the variation in the protective wire current is negligible. These findings support design guidelines for improved traction system electromagnetic resilience.

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Test Methodology for HPM Protection Limiters

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Abstract – As modern electronic infrastructures increasingly depend on wireless communications the threat of High-Power Microwave (HPM) and Intentional Electromagnetic Interference (IEMI) is getting a critical concern for public safety. This paper proposes a comprehensive testing framework that focuses on the non-linear dynamic characteristics of HPM protection limiters, including spike leakage, flat leakage, and microsecond-level recovery times. The proposed methodology provides a technical basis for standardization of test method of HPM protection limiters against sophisticated electromagnetic threats.

Keywords– HPM, protection, limiter, test method

I. INTRODUCTION

In the context of wireless telecommunications, RF front-end components such as low-noise amplifiers (LNAs) are highly susceptible to HPM/EMI environments, which can result in permanent physical damage or temporary degradation. To mitigate these risks, HPM protection limiters can be employed as essential protection mechanisms. However, the test method for HPM protection limiter performance is not yet standardized. Historically, the international testing standard IEC 61000-4-24 has focused on high-altitude electromagnetic pulse (HEMP) protection. However, as was recognized at the 2025 IEC SC 77C plenary meeting, the threat landscape for IEC 61000-4-24 will expand to include HPM and IEMI. This paper addresses the technical requirements for the performance testing of HPM protection limiters for the revision of IEC 61000-4-24.

II. KEY PERFORMANCE INDICATOR

A standardized evaluation of an HPM limiter must be based on the following technical parameters:

- **Insertion Loss (S_{21}):** Defines the impact of the limiter on the system's signal-to-noise ratio during normal operation.
- **Clamping Threshold (Input-Output Linearity):** The input power level at which the limiter transitions from a linear state to a protective (high-impedance) state.
- **Flat Leakage Power:** The steady-state power level that passes through the limiter during sustained HPM exposure
- **Pulse Response and Spike Leakage:** The peak power that leaks through before the limiter fully activates.
- **Recovery Time (t_{rec}):** The time required for the device to return to a low-loss state..

III. PROPOSED TEST METHODOLOGY

A. Test Signal Specifications

The methodology categorizes HPM threats into two types of

signals to evaluate the limiter's response across the time and frequency domains.

- **Narrowband HPM:** Utilizes high-power continuous wave (CW) or pulsed-RF at the system's operating frequency (f_0). This evaluates the "Flat Leakage" level and the thermal endurance of the PIN diode or solid-state elements.
- **Broadband IEMI:** Utilizes ultra-fast transients with sub-nanosecond rise times. This is critical for measuring "Spike Leakage" - the transient energy that passes through before the limiter reaches its full isolation state.

B. Key Measurement Procedures

- **Linear State Characterization:** Prior to high-power injection, the insertion loss and return loss (S_{11}) must be verified using a VNA to ensure the limiter does not degrade normal system performance.
- **Spike and Flat Leakage Analysis:** Using a directional coupler and high-speed real-time oscilloscope, the time-domain pulse response is captured. The peak power of the initial spike and the steady-state flat leakage must remain below the damage threshold of the protected LNA.
- **Recovery Time Measurement:** After the HPM pulse ceases, the time required for the limiter to return to its low-loss linear state (typically <1.0 dB insertion loss) is measured. This ensures the communication link can be re-established within microseconds.

IV. CONCLUSION

The methodology proposed in this paper provides a standardized approach to testing HPM protection limiters, emphasizing the critical metrics of spike leakage and recovery time. This test methodology can form the basis for verifying a consistent, high level of protection for critical RF infrastructures against evolving HPM/IEMI threats.

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A Compact Hardware Module using an Oscilloscope IC for HPEM Detection and Identification

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Abstract– This paper proposes a method for measuring radiated high-power electromagnetic (HPEM) fields using a compact HPEM detection and identification module based on an oscilloscope integrated circuit (OSC IC). The proposed module was implemented using a custom-designed OSC IC, a small microcontroller unit (MCU), and an aluminum jig. Experimental results demonstrate that the proposed module equipped with an antenna measures radiated damped sinusoidal (DS) fields in a gigahertz transverse electromagnetic (GTEM) cell with performance comparable to that of a commercial oscilloscope.

Keywords– Integrated circuit (IC), High-power electromagnetic (HPEM), Damped sinusoidal (DS)

I. INTRODUCTION

With the increasing adoption of artificial intelligence (AI) and internet of things (IoT), the risk of damage from HPEM threats, such as EMP, is increasing due to the growing dependence of critical systems on interconnected electronic devices and automated decision-making. For effective HPEM protection, information on HPEM event occurrence and the corresponding waveform is required.

This paper proposes a compact HPEM detection and identification module implemented with a custom-designed OSC IC.

II. HPEM DETECTION AND IDENTIFICATION MODULE

The core component enabling the operation of the HPEM detection and identification module is the custom-designed OSC IC, and its operating principle is similar to that of [1]. The major circuit blocks of the OSC IC are high-speed analog-to-digital converter (ADC), shift registers, and an event detector. The OSC IC samples the input signal into 5-bit binary data using the eight ADCs, and the resulting 5-bit data containing the waveform information from the detected signal are stored in the shift registers. The stored data are read out and stored by an external equipment.

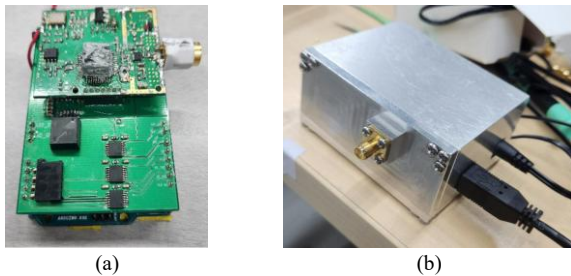


Fig. 1. Photographs of the proposed HPEM detection and identification module: (a) circuitry (b) module jig.

As shown in Fig. 1, the proposed module comprises a circuitry and an aluminum jig. The circuitry includes a compact OSC IC module and an MCU, which interfaces with an external control program to control OSC IC operation. The circuitry is shielded from external interference by a 5 mm-thick aluminum jig with dimensions of $65.1 \times 84 \times 40.6 \text{ mm}^3$. An SMA connector at the OSC IC input enables interfacing with external measurement devices, such as an antenna. An experiment was conducted to verify HPEM detection and identification using the proposed module. As shown

in Fig. 2, a DS waveform was radiated inside a GTEM cell and measured with the same antenna connected in sequence to a commercial oscilloscope and the proposed module. The DS source is structurally similar to that in [2], and the results are shown in Fig. 3.

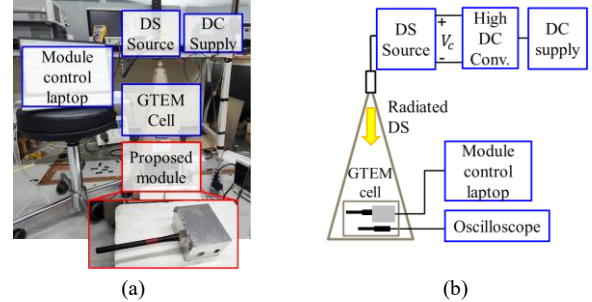


Fig. 2. Radiated DS measurement experiment: (a) photograph; (b) block diagram.

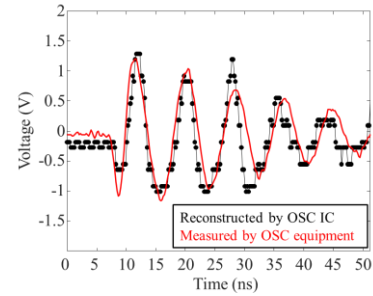


Fig. 3. Comparison of radiated DS measurement results.

III. RESULTS AND DISCUSSION

As shown in the commercial oscilloscope measurement in Fig. 3, the radiated DS measured by the helical antenna exhibits an amplitude of up to approximately $2.5 V_{pp}$. The waveform measured by the proposed module shows good agreement with that measured by the commercial oscilloscope.

IV. CONCLUSION

This paper proposes a method for measuring an HPEM event using the proposed module implemented with a custom-designed OSC IC. Experimental results indicate that the proposed sensing and detection module can functionally replace a commercial oscilloscope.

ACKNOWLEDGMENT

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System-Level Far-Field Attenuation and Noise Scan Evaluation of Electroless-Plated Ni–Cu Alloy Shield

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A coupled evaluation of far-field shielding and near-field (H-field-dominated) magnetic noise suppression is presented for an electroless-plated $\text{Ni}_{0.2}\text{Cu}_{0.8}$ (N2C8) alloy film (330 nm) on PET. Far-field shielding effectiveness is extracted from X-band S-parameters, while near-field behavior is mapped by noise-scan imaging on a Raspberry Pi 4. The film achieves ~ 30 dB SE and lowers peak radiated noise by ~ 8 dBuV ($\sim 2.5\times$ lower voltage, $\sim 6\times$ lower power). The results suggest roles of low surface impedance, eddy-current counter-fields, and Ni-related damping.

Keywords – electroless plating, Ni–Cu alloy, EMI shielding, far-field SE, noise scan

I. INTRODUCTION

Electromagnetic interference (EMI) poses reliability challenges in compact and flexible electronic systems, where both far-field radiation and near-field coupling generate system-level noise. As platforms become thinner and more densely integrated, lightweight and compliant shielding materials are increasingly required. Copper thin films provide high electrical conductivity but are susceptible to oxidation and long-term degradation, which can compromise shielding performance. Alloying copper with nickel enhances corrosion resistance and environmental stability while maintaining sufficient conductivity for electromagnetic attenuation.^[1] Thus, Ni–Cu alloys offer a balanced combination of conductivity and durability for thin-film shielding applications. In this work, a $\text{Ni}_{0.2}\text{Cu}_{0.8}$ alloy film (N2C8) was fabricated on a PET substrate via electroless plating. The Cu-rich composition preserves high conductivity, while Ni incorporation improves stability. The far-field shielding effectiveness and near-field noise suppression were evaluated to validate its system-level electromagnetic mitigation capability.

II. METHODOLOGY

The N2C8 film was fabricated on a PET substrate via electroless plating after surface activation to ensure uniform nucleation and adhesion. The alloy composition was controlled at $\text{Ni}_{0.2}\text{Cu}_{0.8}$ with micrometer-scale thickness. Far-field shielding effectiveness was measured using a vector network analyzer (VNA) in the X-band. The total SE was calculated as

$$SE = -20 \log_{10} |S_{21}| \quad (1)$$

where S_{21} the forward transmission coefficient.

Near-field electromagnetic emission was evaluated using a noise scanning system with a near-field probe. The probe was scanned at a fixed height above the device, and emission amplitudes were recorded over the target frequency range. Radiated noise from a Raspberry Pi 4 board was measured before and after applying the N2C8 shielding film to quantify suppression performance.

III. RESULTS AND DISCUSSION

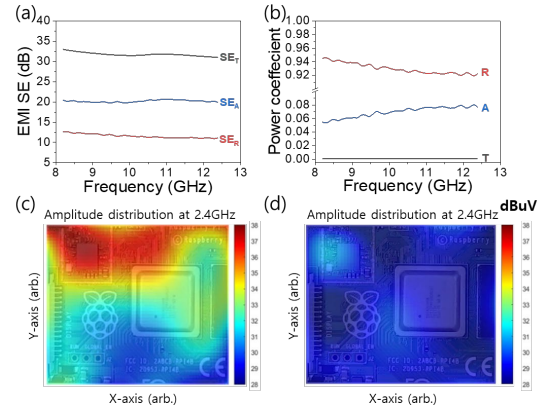


Fig. 1. Electromagnetic shielding performance of the electroless-plated N2C8 thin film. (a) Far-field shielding effectiveness, (b) Power coefficients in the X-band. (c) Near-field noise scan amplitude distribution (dBuV) of the reference device. (d) Noise scan amplitude distribution after applying the N2C8 film.

Fig. 1(a) shows the far-field shielding effectiveness of the N2C8 film, maintaining ~ 30 dB across X-band. SE decomposition indicates reflection-dominant behavior. Fig. 1(b) presents the corresponding power coefficients, where reflection remains dominant and transmission is strongly suppressed, suggesting impedance mismatch at the air–metal interface governs attenuation.

Near-field emission maps of the Raspberry Pi 4 (Fig. 1(c,d)) show localized hot spots >38 dBuV in the reference, whereas the N2C8 film suppresses the peak to ~ 30 dBuV, an ~ 8 dBuV reduction ($\sim 2.5\times$ lower voltage, $\sim 6\times$ lower power). This trend indicates reduced magnetically coupled interference, plausibly aided by eddy-current counter-fields in the conductive layer. The Cu-rich matrix maintains low sheet resistance, while Ni incorporation improves environmental robustness and may add magnetic damping.

IV. CONCLUSION

The electroless-plated $\text{Ni}_{0.2}\text{Cu}_{0.8}$ film achieved ~ 30 dB far-field attenuation and ~ 8 dBuV radiated noise reduction. The Cu-rich alloy preserves high conductivity, while Ni incorporation enhances environmental stability, enabling reliable EMI mitigation for flexible electronic platforms.

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A study on Measurement Method and Application of Reflection-based Absorptances for Planar Materials

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Abstract—Electronic metal housings suffer from internal electromagnetic compatibility (EMC) issues due to signal reflections. Traditional 2-port measurements are insufficient for planar absorbers applied directly to metallic surfaces. This paper proposes a 1-port reflection-based waveguide measurement with a metal backing to accurately evaluate absorptance, analyze dimensional tolerance errors, and demonstrate practical performance within a metallic housing

Keywords— EMI shielding, absorptance, rectangular waveguide, reflection measurement, internal EMC.

I. INTRODUCTION

While metallic housings protect electronic systems from external threats, they create reflective cavities that lead to internal EMC problems. Planar absorbing materials are used to mitigate this, but general 2-port measurements cannot reflect their real-world application against a metal backing. Therefore, this paper introduces and evaluates a reflection-based rectangular waveguide method to accurately quantify the performance of metal-backed absorbers in practical settings.

II. METHODOLOGY

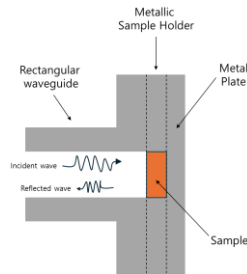


Fig. 1. 1-port reflection measurement

To accurately represent the physical conditions of planar absorbing materials within a metallic housing, a 1-port reflection measurement was developed using rectangular waveguides. As illustrated in Fig. 1, the sample is placed inside a metallic sample holder and terminated with a solid metal plate, mimicking the exact environment where the material is applied to the housing wall. We also investigated dimensional tolerances, as tiny air gaps between the material and the sample holder can introduce significant errors. Furthermore, the full-wave electromagnetic simulation (CST) is conducted to model the internal EMC environment. To simply depict an internal electronic circuit, a 50-ohm microstrip line was modeled inside the mock metallic housing.

III. RESULTS AND DISCUSSION

The reflection-based measurement approach effectively quantifies the true absorptance of the metal-backed material. To validate its practical impact, simulations were performed, as shown

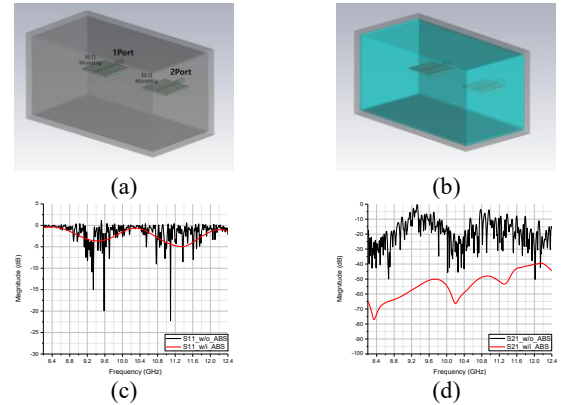


Fig. 2. CST simulation models and results: (a) bare metallic housing, (b) housing with planar absorber, (c) simulated reflection coefficient (S11) and (d) transmission coefficient (S21)

on Fig. 2. Fig. 2(a) and (b) depict the 3D models of metallic housing and housing with the planar absorber, respectively, both containing the 50-ohm microstrip line circuitry. The simulated S-parameters clearly demonstrate the absorber's performance. As shown in the reflection coefficient (S11) plot in Fig. 2(c), the metallic housing suffers from severe internal cavity resonances. However, these resonant peaks are significantly suppressed when the planar absorber is applied. Furthermore, the transmission coefficient (S21) in Fig. 2(d) demonstrates a substantial reduction in electromagnetic cross-coupling between the 50-ohm microstrip lines. Consequently, these simulation results verify that the proposed reflection-based absorptance data reliably predicts practical internal EMC improvements.

IV. CONCLUSION

This study proposed a 1-port reflection-based waveguide measurement with a metal backing to evaluate planar absorbers for metallic housings. To verify the practical EMC improvements, CST electromagnetic simulations incorporating a 50-ohm microstrip line were conducted. The results demonstrated that the applied absorbers significantly suppressed internal cavity resonances (S11) and reduced electromagnetic cross-coupling (S21). Therefore, the proposed 1-port measurement provides a highly reliable and direct evaluation of absorptance for mitigating internal EMI.

ACKNOWLEDGMENT

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Development of an EMP Protection Filter with Multi-Stage PIN Diode Limiters for Communication Systems

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Abstract– This paper presents an EMP protection filter for communication systems using a compact interdigital band-pass filter (BPF) and cascaded PIN diode limiters. The target specification was normal operation up to 1 kW at 1.3 GHz with insertion loss below 1 dB and leakage power below 100 mW. A miniaturized interdigital BPF fabricated on Rogers 6010 showed measured insertion losses of 0.59 dB at 1.25 GHz and 0.83 dB at 1.35 GHz. Limiter-only measurements showed insertion losses of 0.20 dB, 0.36 dB, and 0.61 dB for the 1-stage, 2-stage, and 3-stage designs, respectively. Leakage power decreased as the number of limiter stages increased, and the integrated 3-stage prototype achieved leakage below 10 mW with approximately 0.7 dB insertion loss. These results confirm the expected trade-off between low-loss transmission and high-level EMP protection, and indicate that the 3-stage configuration is the most robust solution for protecting communication front-end circuits.

Keywords– EMP protection, PIN diode limiter, insertion loss, leakage power

I. INTRODUCTION

Communication front-end circuits are highly vulnerable to externally coupled high-power electromagnetic interference and EMP threats. Therefore, RF protection devices must satisfy two conflicting requirements: low insertion loss during normal operation and strong suppression of high-power incident signals. A practical approach is to combine a PIN diode limiter with a band-pass filter. In this work, an EMP protection module was developed to compare 1-stage, 2-stage, and 3-stage limiter configurations and to identify an appropriate trade-off between transmission performance and protection robustness.

II. DEVICE DESIGN

The proposed module was designed for 1.25–1.35 GHz operation with a maximum input power of 1 kW at 1.3 GHz as shown in Fig. 1. The target insertion loss was less than 1 dB, and the target leakage power was below 100 mW. An N-type coaxial interface was used. For miniaturization, an interdigital BPF was designed on Rogers 6010 substrate, resulting in a compact size of $35 \times 10.6 \times 0.625$ mm. The overall concept was a limiter-first and filter-assisted structure, where the PIN diode stages initially suppress the high-power signal and the BPF then provides frequency selectivity.

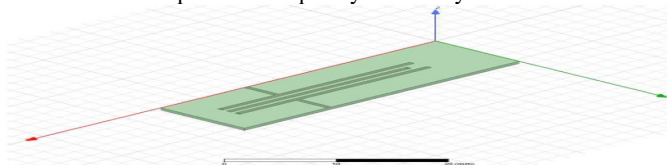


Fig. 1. Schematic view of PIN diode limiter

III. RESULTS AND DISCUSSION

The measured BPF insertion loss was 0.59 dB at 1.25 GHz and 0.83 dB at 1.35 GHz as given in Fig. 2. For limiter-only structures, insertion loss increased with stage number, from 0.20 dB for the 1-stage design to 0.36 dB for the 2-stage design and 0.61 dB for the 3-stage design. In contrast, leakage performance improved as additional stages were introduced. The integrated 3-stage prototype showed approximately 0.7 dB insertion loss with leakage below 10 mW, confirming that higher stage count provides stronger protection at the cost of slightly increased passband loss. The 1-stage design is suitable when minimum loss is the highest priority, the 2-stage design offers a balanced compromise, and the 3-stage design provides the strongest protection margin for severe EMP environments.

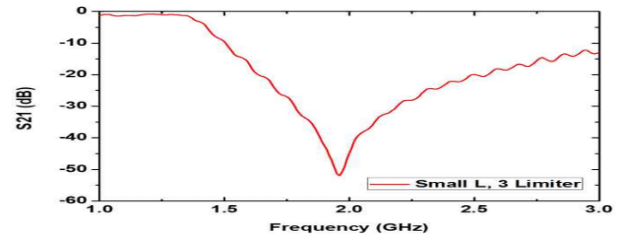


Fig. 2. Transmission performance of demonstrated PIN diode limiter

IV. CONCLUSION

An EMP protection filter for communication systems was developed using a compact interdigital BPF and multi-stage PIN diode limiters. The results showed that insertion loss increased with limiter stage number, whereas leakage power decreased. Among the tested structures, the 3-stage integrated design provided the best protection capability while maintaining acceptable communication-band performance. Therefore, the proposed multi-stage limiter architecture is a practical and robust candidate for EMP-hardened communication front-end protection.

ACKNOWLEDGMENT

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Optimization of air-based laser-driven X-ray generation with a 25-TW femtosecond laser

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Abstract— Laser–solid interactions driven by high-intensity femtosecond lasers generate hot electrons that produce strong electromagnetic pulses (EMP) and bremsstrahlung X-rays. In this work, X-ray emission generated in ambient air using a 25-TW femtosecond laser interacting with solid targets is investigated as a diagnostic of hot-electron dynamics. The X-ray dose and spectrum were measured using an ionization chamber and a multi-filter imaging plate method. Assuming a synchrotron-like spectrum, the reconstructed photon distribution indicates a characteristic energy of ~ 10 keV. These results provide insight into hot-electron generation and related electromagnetic phenomena such as EMP emission.

Keywords— laser-matter interaction, hot electrons, laser-driven X-rays, electromagnetic pulse, atmospheric laser plasma

I. INTRODUCTION

High-intensity laser–solid interactions generate energetic electrons at the target front surface through mechanisms such as resonance absorption, vacuum heating, and $J \times B$ (ponderomotive) heating. These hot electrons can produce strong electromagnetic effects, including electromagnetic pulses (EMP) as well as bremsstrahlung and $K\alpha$ X-ray emission. Because bremsstrahlung radiation originates from energetic electrons interacting with the target material, X-ray measurements provide useful information on the energy distribution and dynamics of hot electrons.

In this work, we investigate X-ray emission generated by a 25-TW femtosecond laser interacting with solid targets in ambient air. By measuring the X-ray dose and reconstructing the photon spectrum using a multi-filter transmission method, we estimate the characteristic electron energy and discuss its relevance to electromagnetic processes such as EMP generation.

II. METHODOLOGY

Experiments were performed using a Ti:sapphire laser system delivering 1.6 J pulses with a duration of 60 fs at a repetition rate of 10 Hz, corresponding to a peak power of ~ 25 TW [1]. The laser beam was focused onto solid targets in ambient air without a vacuum chamber. Key parameters affecting X-ray generation, including the effective focal position influenced by nonlinear propagation in air, target position relative to the focal plane, and laser incidence angle, were systematically investigated [2]. Different target materials and thicknesses (3M tape, aluminum, and tungsten) were tested, and the emitted X-ray dose was measured using an ionization chamber. The photon energy distribution was estimated using a multi-filter method with metal foils (Fe, Cu, Ti, Ni, Mo, Zr, Pb, etc.) [3]. Assuming a synchrotron-like spectrum, the measured transmission through the filters was fitted to reconstruct the X-ray spectrum.

III. RESULTS AND DISCUSSION

Fig. 1(a) shows the normalized X-ray transmission through multiple metal filters together with the best-fit curve assuming a

synchrotron-like spectrum. The fitted model reproduces the measured transmission across the filters within the experimental uncertainty. To enhance the X-ray output, several experimental parameters were optimized, including the effective focal position influenced by self-focusing in air, target positioning and translation, and the laser incidence angle. The X-ray dose was measured using an ionization chamber, and additional enhancement was obtained by varying the target material and thickness, including 3M tape, aluminum, and tungsten. Based on the measured transmission data, the X-ray photon spectrum was reconstructed as shown in Fig. 1(b). The reconstructed spectrum indicates a characteristic photon energy of approximately 10 keV. The shaded region represents the uncertainty resulting from $\pm 10\%$ variations in the measured signals.

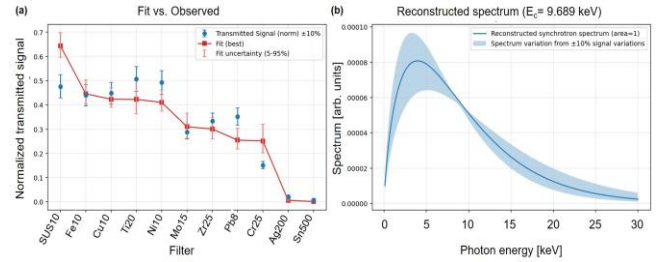


Fig. 1. (a) Measured X-ray transmission through metal filters and fitted curve as assuming a synchrotron-like spectrum. (b) Reconstructed X-ray spectrum with $\pm 10\%$ uncertainty.

IV. CONCLUSION

X-ray emission from laser–solid interaction in ambient air was investigated using a 25-TW femtosecond laser. The X-ray spectrum was reconstructed using a multi-filter transmission method assuming a synchrotron-like distribution, yielding a characteristic photon energy of ~ 10 keV. Since the emission originates from bremsstrahlung radiation produced by energetic electrons, the reconstructed spectrum provides information on the hot-electron population at the target surface. These results provide insight into electron dynamics and related electromagnetic phenomena such as EMP generation in high-intensity laser experiments.

ACKNOWLEDGMENT

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Numerical Simulation of Plasma Density and Thermal Channel Evolution in 1 kHz Laser Lightning Control Ignoring Turbulence Effects

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[#]The authors share some contribution

Abstract– This paper investigates the simulation of femtosecond laser filament in air. The (2D+1) nonlinear Schrödinger equation (NLSE) was adopted in this study. Comparison to the results from Ma et al. demonstrate the reasonableness of the simulation methodology in this study. The analysis of plasma density and thermal channels are then performed by adopting the parameters from laser guided lightning test by Houard et al. The simulated filament length of around 50m is comparable to the overlapped lightning channel in laser guided lightning test by Houard et al.

Keywords– Femtosecond laser filament, Kerr effects, plasma effect, laser plasma density, thermal channels

I. INTRODUCTION

Lightning pose threats to humans from various aspects. Based on the Benjamin Franklin's famous kite experiment in the 18th century, the lightning rod has been known and utilized for around three hundred years[1]-[2]. The laser lightning control was first tested in Japan through carbon dioxide laser in 1995. Recently, the laser lightning control test with initial input laser parameters of 720mJ, 860fs, 1030nm and 1kHz has been successfully carried out in Santis Tower in Europe in 2021[3]. The novelty of this study lies in the evolution of plasma density and thermal channels in air.

II. METHODOLOGY

The (2D+1) nonlinear Schrödinger equation (NLSE) was adopted to simulate femtosecond laser filament in air. The expression is,

$$\frac{\partial A(r, \tau, z)}{\partial z} = \left(\frac{i}{2k_0} \nabla_{\perp}^2 - i \frac{k_0'}{k_0} \frac{\partial^2}{\partial z \partial \tau} - \frac{k_0'}{k_0} \frac{\partial^2}{\partial z \tau^2} \right) A(r, \tau, z) +$$

$$ik_0 n_K A(r, \tau, z) - \frac{ik_0}{2} \beta^{(K)} |A|^{2K-2} \left(1 - \frac{n_e}{n_{at}} \right) A(r, \tau, z) \quad (1)$$

where $A(r, t, z)$ is the electric field, diffraction term and group velocity dispersion item is in the first row, Kerr effect is the first item in the second row, multi-photon ionization, and plasma effect are the last item. The turbulence effect is neglected in simulation.

III. RESULTS AND DISCUSSION

Simulations of axis intensity and axis electron density from Ma et al. [2] were presented in Fig.1. The well agreement with [2] demonstrate the suitability of the simulation methodology. Fig.2 shows the evolution of plasma density and laser-induced thermal channels toward a steady state during the cumulative process with laser parameters of 500mJ, 1030nm, 7ps and 1kHz (e.g., [3]).

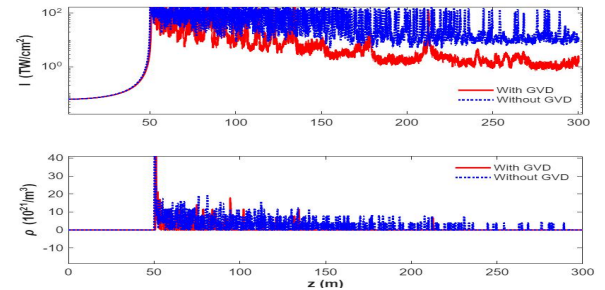


Fig. 1 Simulation of axis intensity and axis plasma density to be compared with Fig.3a2 and Fig.3b2 of the reference [2].

IV. CONCLUSION

The simulation methodology is reasonable by comparing to previous simulation results. The obtained filament length of around 50m are characterized by the evolution of plasma density and thermal channels, which is comparable to the length of the overlapped lightning channel from twice experiments of the laser guided lightning test by Houard et al.

ACKNOWLEDGMENT

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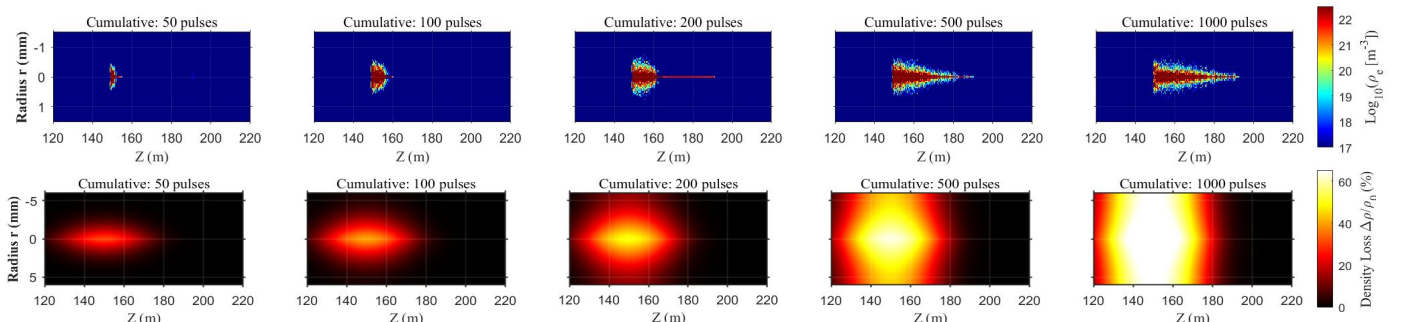


Fig. 2. Simulation of plasma density and thermal channels during laser propagation. (input laser parameters: 500mJ, 1030nm, 7ps and 1kHz)

Simulation Analysis of Incident-Angle Effects on Electromagnetic Fields in a UAV

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Abstract– This study investigates the effect of incident electromagnetic wave angle on internal field penetration in a UAV platform using electromagnetic simulations. A three-dimensional UAV model including major printed circuit boards (PCBs) was developed in CST Studio Suite. The incident angle was varied from 0° to 90°, and the maximum electric fields induced on key components were analyzed. The results show that the main board and ESC board experience higher electric fields compared with other components. The field levels vary depending on component location and incident angle, with noticeable increases observed around certain angles. These results provide insight into incident-angle-dependent electromagnetic coupling in UAV structures.

Keywords– High-Power Microwave (HPM), UAV

I. INTRODUCTION

The increasing use of unmanned aerial vehicles (UAVs) has raised interest in effective countermeasure technologies. High-Power Microwave (HPM) systems provide a non-kinetic and cost-effective approach for disabling low-cost drones [1]. However, experimental evaluation of HPM effects is constrained by the need for large interference-free test areas, difficulty in measuring incident fields on flying drones, and limited repeatability due to crash-induced damage. In addition, most tests assume frontal irradiation, which does not represent realistic operational environments where waves may impinge from various angles. This study employs electromagnetic simulations to analyze incident-angle-dependent field penetration, enabling prediction of electromagnetic effects under realistic conditions.

II. METHODOLOGY

A three-dimensional electromagnetic model of the UAV was constructed using CST Studio Suite, incorporating the airframe geometry and major printed circuit boards (PCBs). The internal electric field distributions under external electromagnetic wave excitation were analyzed. The electromagnetic wave used in the simulation were at specific frequency within the L/S bands, with a pulse width of 100 ns.

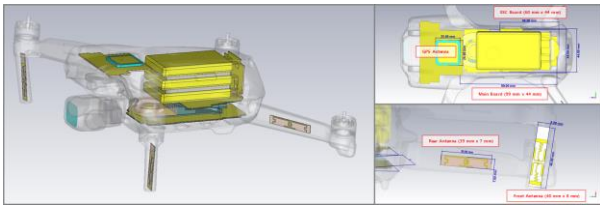


Fig. 1. 3D model of the UAV with PCBs

The incident angle of the electromagnetic wave was varied from 0° (frontal incidence) to 90°. For each incident angle, the maximum electric field induced on the PCB was calculated and compared. In addition, the internal field distribution within the UAV was examined to evaluate the variation in electromagnetic coupling

characteristics with respect to incident angle.

III. RESULTS AND DISCUSSION

The maximum electric field induced on the main board, ESC board, GPS board, front antenna, and rear antenna was obtained through electromagnetic simulations for incident angles ranging from 0° to 90°.

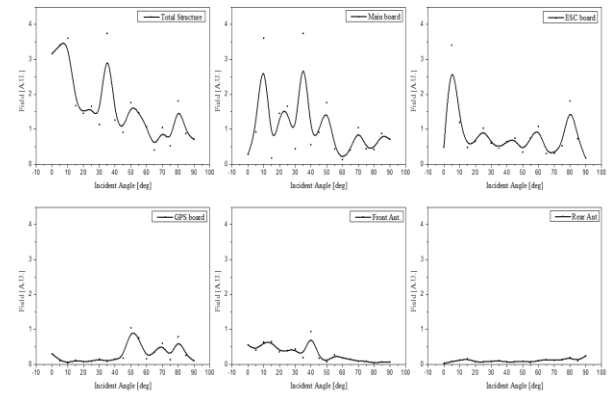


Fig. 2 Maximum electric field induced on UAV components as a function of incident angle

The results show that the main board and ESC board experience relatively higher electric fields compared with the GPS board and antennas, likely due to their central location and larger conductive structures within the airframe. The field variation with incident angle differs depending on component location. The main board and ESC board exhibit fluctuating field levels, while the front antenna shows a gradual decrease as the incident angle approaches 90°. In contrast, the rear antenna shows a slight increasing tendency, and the GPS board exhibits higher field levels in the 50°–90° range compared with the 0°–40° range. A noticeable increase in the electric field was also observed around incident angles of approximately 30°–40° in some components, which may be related to structural resonance or coupling effects within the UAV structure.

IV. CONCLUSION

Electromagnetic simulations were performed to evaluate the influence of incident angle on internal field penetration in a UAV platform. The results show that electromagnetic coupling characteristics vary significantly depending on component location and incident angle.

ACKNOWLEDGMENT

This work was funded by the government and conducted by the Agency for Defense Development.

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Analysis of EM Coupling in a UAV with Various Cable Configurations

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Abstract– The increasing use of unmanned aerial vehicles (UAVs) in civilian and defense contexts has driven interest in high-power microwave (HPM) technology as an effective countermeasure. This paper investigates HPM-induced electromagnetic coupling in a commercial UAV by varying internal cable configurations to examine the resonance-related coupling behavior. The results show that the frequencies of strong induced E-field response depend on cable configuration, indicating that the dominant HPM coupling characteristics can shift with internal cable configuration.

Keywords– Cable, coupling, high-power microwave (HPM), unmanned aerial vehicles (UAVs)

I. INTRODUCTION

In recent years, commercial unmanned aerial vehicles (UAVs) have been adopted in both civilian and military sectors due to their low cost and operational convenience. However, their wide availability and ease of deployment have also increased the risk of misuse, surveillance, and attacks in sensitive areas. Consequently, the need for practical countermeasures has intensified, and high-power microwave (HPM) sources have attracted attention because of their rapid response capability and cost-effectiveness in disrupting onboard electronic systems [1], [2].

The effects of HPM on UAVs depend strongly on platform configuration and the associated electromagnetic coupling paths. Various structural and electrical features, including airframe apertures, internal cable length and routing, and shielding conditions, significantly influence induced responses, and resonance may result in enhanced electromagnetic coupling at specific frequency bands. For effective HPM-based neutralization of UAVs, prior identification of dominant coupling mechanisms and vulnerable frequency bands is important. In this study, electromagnetic simulations are performed to investigate HPM induced coupling to internal cable in a commercial UAV. In particular, the analysis examines the impact of cable variations on coupling behavior and high E-field response across frequency.

II. MODELING & SIMULATION

To investigate HPM-induced electromagnetic coupling in the UAV, a simulation-based analysis is performed. Fig. 1 illustrates the simplified UAV model and the simulation configuration employed in this study. As shown in Fig. 1(a), the components most relevant to electromagnetic coupling, including internal motor cables, shielding structures, and a simplified onboard chip module, are included. A plane wave is incident from the frontal direction of the UAV, as depicted in Fig. 1(b). The induced E-field distribution on the metal block and in its nearby region is analyzed to assess how variations in cable length and cable type influence coupling behavior under HPM exposure. Fig. 2 presents the E-field distributions on the UAV surface induced along the motor cables under HPM exposure, highlighting the effects of cable length and cable type. In Fig. 2(a), the cable length is varied to examine its influence on HPM-induced coupling. The results show that the

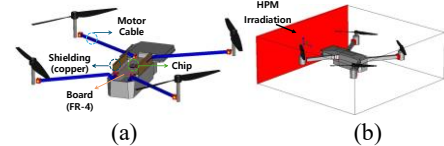


Fig. 1. Simulation configuration: (a) simplified UAV model with internal components and (b) frontal plane-wave HPM exposure.

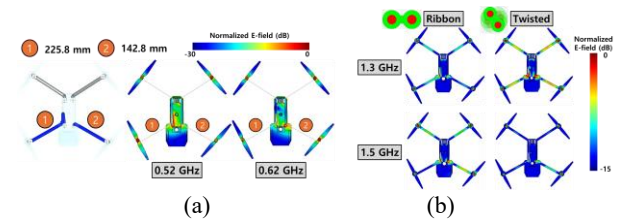


Fig. 2. Simulation results induced E-field distribution for (a) different cable lengths and (b) cable types.

frequency at which strong coupling occurs shifts with cable length, indicating a length-dependent change in the dominant coupling behavior. The influence of cable type on the induced E-field distribution is illustrated in Fig. 2(b). The frequencies shown in Figs. 2(a) and (b) are selected to highlight the most representative coupling responses for each case, rather than for direct comparison. Different cable types exhibit peak responses in different frequency regions, showing that the coupling characteristics are also affected by cable configuration. These results suggest that the motor cable can form a dominant internal coupling path, and that the induced E-field trends can be used to identify frequency bands susceptible to strong HPM coupling [1].

III. CONCLUSION

We investigated the frequency-dependent electromagnetic coupling of internal UAV cables under HPM exposure by varying cable configurations. The results showed that strong coupling bands shift with cable configuration, indicating configuration-dependent HPM coupling into the UAV interior. These findings provide an initial indication of susceptible frequency bands based on induced E-field magnitude, while detailed system-level responses are not addressed. This insight supports HPM countermeasures by guiding configuration-aware susceptibility assessment and test planning.

ACKNOWLEDGEMENT

This research is performed based on the cooperation with Soongsil University – LIG Defense&Aerospace Cooperation.

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The Development of a drone Detection, Tracking, and Steering System applicable to the HPM anti-drone system

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Abstract – This paper presents the development of a drone detection, tracking, and steering system which is applicable to the High Power Microwave (HPM) anti-drone system. The approach integrates a detection and tracking framework with a steering mechanism, enabling precise alignment of the HPM system with its targets. Experimental results demonstrate that the method effectively detects and tracks drones while accurately directing the HPM device toward the target.

Keywords– ADS, RADAR, EO/IR CAMERA, LRF, HPM

I. INTRODUCTION

Recent advancements in drone technology have accelerated alongside the expansion of their applications in civilian and military domains. In recent conflicts, including those involving the United States and Iran, highlight the growing role of reconnaissance and strike drones in modern warfare [1]. As a result, the demand for effective counter-drone technologies has grown substantially.

This paper presents an integrated system for drone detection and tracking using radar and EO/IR cameras. The system generates precise steering commands for a HPM actuator, enabling accurate targeting of electromagnetic energy toward the detected drone.

II. METHODOLOGY

A counter-drone system is typically composed of four sequential stages: detection, tracking, identification, and response, with the first three forming the detection and identification system and the fourth serving as the response stage [2]. The proposed system adds a precision steering function to direct the HPM device toward the detected drone.

The system employs four Active Electronically Scanned Array (AESA) radars for 360° coverage to determine the position (latitude, longitude, and altitude) of detected drones. This information is displayed on the operator's interface and converted into azimuth and elevation angles for the EO/IR camera to track and identify the selected target. While the EO/IR camera tracks the drone, the Laser Range Finder (LRF) mounted on it measures the precise distance when the drone enters a predefined range. The system refines the drone's position from this measurement. The software converts these coordinates into azimuth and elevation angles for the HPM-mounted actuator, which then steers to emit the microwave signal.

In the proposed system, The HPM-mounted actuator is large-scale and located at a distance from the control station, where the surveillance radar and EO/IR camera are installed. Accurate knowledge of the positions of both the EO/IR camera and the actuator is therefore critical for precise target engagement.

Fig. 1 illustrates the configurations of the AESA radar, the EO/IR camera, and the actuator system.

III. RESULTS AND DISCUSSION



Fig. 1 The configurations of the AESA radar, the EO/IR camera, and the actuator system

To obtain precise positional information, a drone equipped with Real-Time Kinematic (RTK) positioning capability was utilized to measure the actual latitude, longitude, and altitude of the system components. These measurements enabled the enhancement of positional accuracy for steering and targeting, ensuring reliable guidance of the HPM system toward the detected drone.

Fig. 2 shows the EO/IR camera's drone tracking display and the view from the actuator-mounted observation camera, respectively.

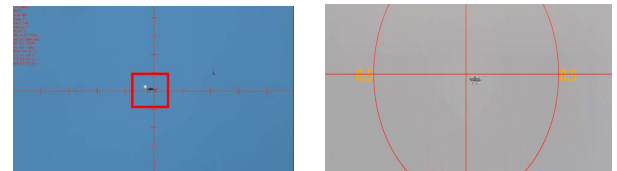


Fig. 2 The EO/IR camera's drone tracking display and the view from the actuator-mounted observation camera

IV. CONCLUSION

This paper presents an integrated system for detecting and tracking drones using AESA radars and EO/IR cameras. A LRF mounted on the EO/IR camera measures the distance to the drone, enabling calculation of its latitude, longitude, and altitude. These coordinates are then converted into azimuth and elevation angles and transmitted to the HPM-mounted actuator, which is positioned separately from the control station. This process allows accurate steering of the HPM toward the target drone.

ACKNOWLEDGMENT

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Analysis of an Oscillating Electric Field Intensity on a Target Illuminated by an HPM Source

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Abstract—The oscillating electric field amplitude on a target E_T [V/m] at a distance r [m] in the far-field zone of an HPM source irradiating the output power P_S [W] through an HPM antenna might be estimated and two scenarios of high-power microwave electromagnetic radiation use to destroy hostile electronics are discussed.

Keywords—oscillating electric field, high-power microwave antenna, electrical breakdown of hostile electronics.

The oscillating electric field amplitude on a target E_T [V/m] at a distance r [m] in the far-field zone of an HPM source irradiating the output power P_S [W] through an HPM antenna might be estimated as follows [1],

$$E_T = \sqrt{\frac{P_S \cdot K \cdot Z_0}{4\pi} \cdot 10^{G_A/10} \cdot \frac{1}{r}}, \quad (1)$$

where P_S [W] is usually measured, calculated, and analyzed as the “root-mean-square” value of the oscillating amplitude of the output power, $Z_0=376.37 \Omega$ is the impedance of free space, 4π [sr] is the number of steradians in a sphere surrounding the HPM antenna, G_A [dBi] is the HPM antenna gain showing a relation of (i) power density radiated by the HPM antenna in the direction of its maximum gain to (ii) power density of an isotropic oscillator having the same total radiated power P_S , and K is the parameter which equals either $K=1$ when E_T defines the “root-mean-square” amplitude of the oscillating electric field on a target (**Scenario #1**), or $K=2$ when E_T defines the maximum or peak amplitude of the oscillating electric field on a target (**Scenario #2**).

The engagement **Scenario #1** assumes that the oscillating electric field E_T induces an electrical breakdown inside an illuminated target after a repeated number of the CW electric field oscillations. A typical example of such a scenario is an illumination of a semiconductor material within the target by the oscillating electric field E_T . This may result in a gradual in time accumulation of an electrical charge inside a semiconductor material with its following electrical breakdown due to an induced electrical discharge between (i) the accumulated electrical charge and (ii) either some random electrical potential or a closest grounded electrode inside an illuminated target. The engagement **Scenario #2** assumes that the oscillating electric field E_T induces an electrical breakdown/discharge inside an illuminated target during one or few short pulses of the electric field oscillations. One of a typical examples of such a scenario is an electrical surface or air breakdown occurring due to a potential difference formed between (i) some random electrode of the illuminated target, where the instantaneous value of the oscillating electric field E_T achieves its maximum amplitude value, and (ii) some other random electric potential inside an illuminated target or a closest grounded electrode of an illuminated target. The distance between the two electrodes on an illuminated target is to be around the half of the wavelength, $\lambda/2$, from the electrode where the maximum amplitude value E_T is achieved. Another example of Scenario #2 is an electrical discharge through the existing lumped elements of the target circuitry (resistors, inductors, diodes, etc.), when an extremely high pulsed electric current, caused by the potential difference between the above-mentioned two random electrodes of the illuminated target, results in abrupt melting of those lumped elements, non-linear junctions, etc. down the circuit path.

Estimations of the oscillating electric field amplitude E_T [V/m] on a target at a distances r [m] up to 10 km has been done using (1) assuming that the radiated average HPM power P_S is either 500 MW,

which is typical for the S-band part of the HPM radiation spectra, or 1 GW, which may be achieved in the L-band part of the HPM radiation spectra. In both estimations, the HPM antenna gains $G_A=20$, 30 and 40 dBi are used. Parameter $K=2$, which assumes Scenario #2 of operation, when the HPM effect is caused by the maximum or peak amplitude of the oscillating electric field on a target, is assumed. Results of simulations are shown in Fig. 1 and in Fig. 2, respectively, against two levels of electric field amplitude providing (i) lethal level of $E_T=3$ kV/m, and (ii) disruptive level of $E_T=1$ kV/m on a target, in accordance with [2], [3].

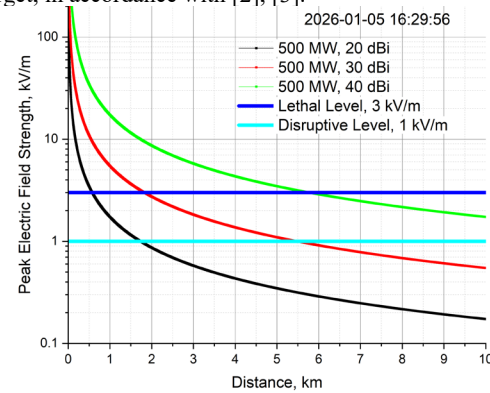


Fig. 1. Estimations of peak electric field amplitude E_T [kV/m] on a target at range of distances r [km] at average $P_S=500$ MW.

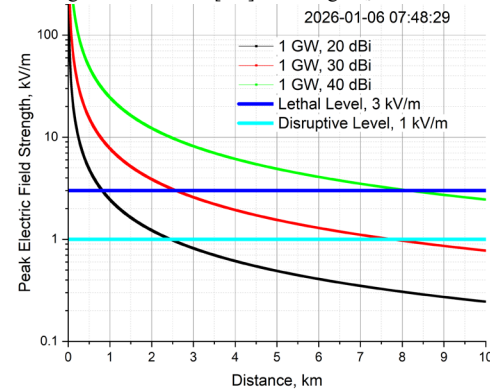


Fig. 2. Estimations of peak electric field amplitude E_T [kV/m] on a target at range of distances r [km] at average $P_S=1$ GW.

Analysis of the estimations (Fig. 1, Fig. 2) shows that the lethal level, $E_T=3$ kV/m, of peak electric field oscillations on a target may be achieved with the high gain HPM antenna, $G_A=40$ dBi, at (i) distance $r \sim 5.5$ km from an HPM source proving $P_S=500$ MW of output HPM power (Fig. 1.), and (ii) at distance $r \sim 8$ km from an HPM source proving $P_S=1$ GW of output HPM power (Fig. 2).

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Design and Experimental Investigation of a 450 kV PFN Marx Generator

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Abstract– This paper presents the experimental results of a 20-stage pulse forming network (PFN)–Marx generator with a pulse width of 100 ns. Each stage consists of four 5 nF, 50 kV capacitors connected in parallel to form a single PFN unit. At the maximum charging voltage, the generator produces an output voltage of 450 kV across a 100 Ω high-power microwave (HPM) load. The characteristic impedance of the PFN–Marx generator is estimated to be 54 Ω . The generator performance is experimentally verified based on results obtained with a 100 Ω resistive load.

Keywords– Marx generator, pulse forming network, high power Microwave System, Spark gaps.

I. INTRODUCTION

This work extends the study in [1] on a 300 kV PFN–Marx generator, with the objective of increasing the output voltage to 450 kV while maintaining efficient operation with a 100 Ω HPM load. The number of stages is increased from 12 to 20 to meet this requirement. The system is designed to achieve a pulse FWHM of ~ 100 ns, a minimum rise time of 20 ns, and an output voltage up to 450 kV. The generator is capable of operation at a repetition rate of 2 pulses per second.

II. MARX GENERATOR DESIGN

Based on the specified performance requirements, the initial estimation of the FWHM and the characteristic impedance of the PFN–Marx generator is carried out using following analytical expressions.

$$T_{FWHM} = 2.S.\sqrt{LC} \quad (1)$$

$$Z_{PFN} = N.\sqrt{\frac{L}{C}} \quad (2)$$

where S represents the number of cells in each PFN unit, L is the inductance associated with the stripline geometry, C is the capacitance used in each cell, and N denotes the number of stages in the PFN–Marx generator.

Each PFN unit consists of four 5 nF capacitors connected in parallel using aluminum strips, with an estimated interconnection inductance of 35 nH. Based on (1) and (2), the expected output pulse FWHM is ~ 100 ns, and the characteristic impedance of the PFN–Marx generator is 54 Ω .

The capacitors are arranged in a zigzag configuration [2] to enhance output performance through effective utilization of stray capacitance. The stray capacitance of the first four stages is further increased using wider holder strips to improve the erection process. Stainless steel electrodes with a 2.8 mm spark gap are employed for the first stage, which is triggered via a field-distortion technique using a thyatron and coaxial cable, as described in [3].

Each stage is supported by charging and discharging inductors of approximately 80 μ H. The assembly is housed in a pressurized vessel filled with SF₆ gas at pressures up to 3.5 bar for full-power operation. A high-density polyethylene (HDPE) feedthrough connects the PFN–Marx generator to the load.

The system was tested under multi-shot operation, with the output connected to a diagnostic unit comprising parallel resistive

TABLE I. MARX GENERATOR DESIGN PARAMETERS

Param	Charge voltage	Marx stage	PFN cells	Total Impedance	Open load vol.	Pulse width
Value	40 kV	20	4	54 Ω	800 kV	~ 100 ns

loads. Current and voltage measurements were obtained using Rogowski coil and V-dot sensors, respectively. A summary of the design parameters of the generator is shown in Table 1.

III. EXPERIMENTAL RESULTS

The characteristic impedance of the PFN–Marx generator was evaluated through short-circuit and matched-load testing, yielding an estimated value of approximately 54 Ω . The performance of the proposed PFN–Marx generator is experimentally validated and compared with both circuit-level and full-wave electromagnetic simulations carried out using CST Microwave Studio Suite. Figure 1 shows the measured current waveforms, recorded with a Rogowski coil during discharge into a 100 Ω resistive load at different gas pressures. The experimental results demonstrate that the output rise time of less than 20 ns and pulse width of ~ 110 ns are within the acceptable range of the initially defined design requirements.

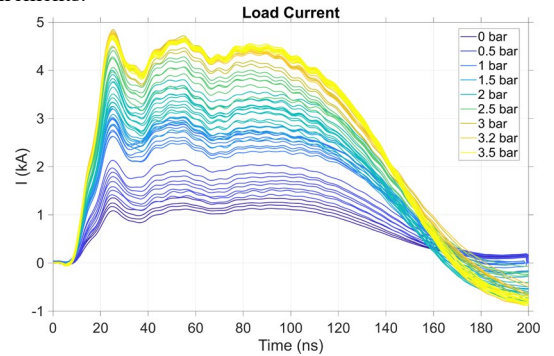


Fig. 1. Marx-PFN load current measured through Rogowski coil.

IV. CONCLUSION

The design and experimental validation of a 20-stage PFN–Marx generator have been presented for operation with a 100 Ω resistive load at an SF₆ pressure of 3.5 bar. The triggering system, based on a thyatron and coaxial cable configuration, enabled reliable triggering of the generator. During full-load testing, output voltages of 450 kV were successfully achieved.

For future work, efforts will focus on extending the operation to higher repetition rates by implementing multi-stage triggering schemes. Additionally, challenges related to spark gap alignment in the horizontal configuration were identified, and improvements in mechanical alignment and stability are planned.

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Numerical Verification of the Stochastic Lightning Progression Model

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Abstract– The stochastic lightning propagation model is an effective tool for engineering assessment of lightning risk. To address the issue of insufficient model credibility, numerical verification is carried out in this study. The root-mean-square (RMS) of the residual is employed as the convergence criterion, and a safety factor is introduced to regulate numerical errors. Numerical experiments indicate that accurate lightning paths can be obtained only when the safety factor is sufficiently small.

Keywords– lightning propagation model, numerical verification, convergence criterion, safety factor

I. INTRODUCTION

Lightning poses a significant threat to the reliable operation of power systems. To improve upon the traditional electro-geometric model (EGM), the stochastic lightning propagation model (SLPM) was first proposed in 2000^[1] as an alternative approach for lightning risk assessment. In this framework, the spatial electrostatic field is computed using the finite difference method, and the direction of leader growth is determined probabilistically through a dielectric breakdown model, enabling the reproduction of the stochastic and branching characteristics of lightning channels.

Over the past two decades, extensive studies have been conducted on SLPM, including parameter calibration and validation as well as applications in various engineering scenarios. Traditionally, the fractal dimension of the lightning channel was used as the primary indicator for evaluating model reliability. In 2018^[2], Xie et al. adjusted model parameters so that the simulated fractal dimension, charge distribution, and near-ground electric field agreed with observational data from a specific event. In 2022^[3], Guo et al. further assessed the model by comparing lightning attachment probabilities between laboratory long-gap discharge experiments and stochastic lightning simulations.

According to the ASME V&V20-2009^[4] standard, verification refers to the process of ensuring that the computational model correctly implements the underlying mathematical model, whereas validation evaluates the extent to which the model accurately represents the real physical system. Importantly, validation must be built upon prior verification. However, previous studies on SLPM have focused exclusively on validation, while formal verification of the computational implementation has never been conducted.

To address this issue, the present study performs solution verification. Numerical error estimation is carried out, and a convergence criterion based on the residual RMS is established. A safety factor is introduced to ensure that the iterative error remains within a prescribed multiple of the discretization error. Finally, numerical experiments are conducted to demonstrate the practical procedure and significance of the proposed verification framework.

II. METHODOLOGY

In the two-dimensional SLPM formulated on a uniformly spaced Cartesian grid, the discretized electric potential satisfies the following discrete Laplace equation at all interior nodes

$$L\phi = L_h\phi_h = 0, L_h e_d = L_h(\phi - \phi_h) = \tau_h = \frac{h^4}{12} \left(\frac{\partial^4 \phi}{\partial x^4} + \frac{\partial^4 \phi}{\partial y^4} \right) + O(h^6), \quad (1)$$

where L and L_h denote the continuous and discrete Laplace operators, e_d and τ_h represent the discretization error and the truncation error. Let $e_i^k = \phi_h - \phi_h^k$ be the iterative error and $r^k = -L_h\phi_h^k$ be the residual, we have the relation $L_h e_i^k = r^k$. Taking the RMS of the residual as the convergence criterion, it satisfies

$$r_{\text{rms}}^k = \sqrt{\frac{1}{N} \sum_{i=1}^N (L_h \phi_h^k)_i^2} \rightarrow \sqrt{\frac{1}{N} \sum_{i=1}^N (L_h e_d)_i^2} \approx \text{tol} = f |\phi_t| N^{-1}, \quad (2)$$

where N denotes the total number of grid points, f is the safety factor, ϕ_t represents the electric potential at the upper boundary.

III. RESULTS AND DISCUSSION

As shown in Figure 1, the simulated lightning paths vary depending on the value of the safety factor.

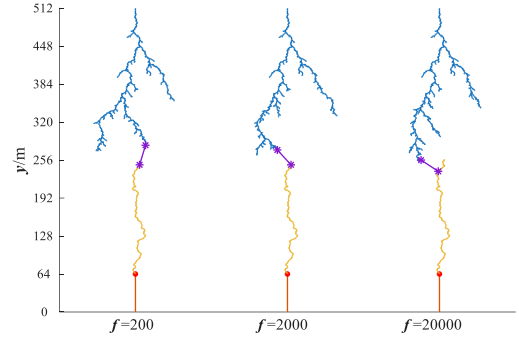


Fig. 1. Lightning paths obtained for different safety factors with $N=512^2$

IV. CONCLUSION

Accurate lightning paths are achievable only after the iterative error has been reduced to a sufficiently small magnitude.

ACKNOWLEDGMENT

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A PEXP-Based Modified Description of the Full-Time HEMP Waveform (E1, E2, E3)

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Abstract– The 2025 IEC 61000-2-9 standard provides a foundational description for HEMP waveforms. However, it employs double exponentials (DEXP) for E1/E2 with derivative discontinuity at the initial time and E3 is described via an approximate formulation derived from the convolution of the magnetic field. This paper extends the p -power of double exponentials (PEXP) and its superposed variants to model the full-time HEMP waveform. Optimized PEXP expressions are obtained via Nondominated Sorting Genetic Algorithm II (NSGA-II), thereby establishing a unified description of the full-time HEMP waveform. The proposed method eliminates derivative discontinuity and achieves high accuracy, providing a rigorous waveform model for HEMP assessment and EMC simulations.

Keywords– full-time HEMP waveform, PEXP, waveform fitting

I. INTRODUCTION

The 2025 IEC 61000-2-9 standard describes the E1 and E2 components of HEMP using DEXP [1], which suffers from a derivative discontinuity at $t = 0$. The E3 component is defined by an approximate convolution of the magnetic field. As an alternative description, the quotient of double exponentials (QEXP) is inherently continuous in its derivative at $t = 0$, and with a time shift factor, it brings the waveform near zero at $t = 0$ [2]. In contrast, the PEXP not only resolves the derivative discontinuity but also eliminates the need for a time shift. This paper extends the p -power of double exponentials (PEXP) and its superposed form to model the full-time HEMP waveform covering the E1, E2, and E3 components. The results demonstrate that the proposed method maintains high fitting accuracy while enabling a unified formulation. It provides a more rigorous waveform input model for HEMP effect assessment, protection design, and EMC simulations.

II. METHODOLOGY

The proposed modified description of the full-time HEMP waveform is based on the p -power of double exponentials (PEXP). The basic form of PEXP, as proposed in [2], is given by

$$E(t) = kE_0(1 - e^{-\alpha t})^p e^{-\beta t}, \quad (1)$$

where E_0 is the peak amplitude, α and β are the rising/decaying rate parameters, and p controls the smoothness of the wavefront. The factor k ensures the peak value equals E_0 . While a single PEXP term suffices for the E1 and E2 components, which exhibit a rapid rise followed by a decay, the E3 component exhibits a more complex shape involving multiple time scales. Therefore, a superposed form of PEXP is introduced:

$$E(t) = \sum_{i=1}^n k_i E_{0i} (1 - e^{-\alpha_i t})^{p_i} e^{-\beta_i t}. \quad (2)$$

This representation provides the flexibility to model the slowly decaying long-duration E3 waveform, and the E3 considered in this study corresponds to a soil conductivity of 10^{-3} S/m.

All fitting procedures were implemented in MATLAB with Nondominated Sorting Genetic Algorithm II (NSGA-II). For E1/E2, the PEXP parameters were optimized to match the DEXP waveforms defined in the 2025 IEC 61000-2-9 standard. The optimization targets included the rise time t_r , the full width at half

maximum t_{FWHM} , and the total energy flux. For E3 component, due to its long duration and slow decay, the optimization focused on the following metrics: peak amplitude, time to peak, zero-crossing time, and the energy fluxes of both positive and negative polarities. These parameters were selected based on the coupling effects of the E3 waveform on long-line infrastructure (e.g., power grids), as they effectively capture key characteristics such as amplitude, rate of change, polarity reversal, and energy accumulation. It should be noted that the reference waveform used for fitting was obtained by convolving the analytic magnetic field expression given in the standard to derive a more accurate electric field.

III. RESULTS AND DISCUSSION

The optimized PEXP parameters are summarized in Table I. A single-term PEXP as defined in (1) was employed for the E1 and E2, while a three-term superposed PEXP as defined in (2) was used for the E3, with the coefficients k_i taking both positive and negative values to accurately reproduce its bipolar waveform. The corresponding waveforms are shown in Fig. 1.

TABLE I. OPTIMIZED PEXP PARAMETERS

HEMP	Optimized PEXP parameters				
	k	E_0 (V/m)	α (s ⁻¹)	β (s ⁻¹)	p
E1	1.3061	50000	9.82×10^8	3.68×10^7	17
E2	1	100	8.36×10^8	1000	45
E3	126.24	8.425×10^{-2}	2.44×10^{-2}	6.21×10^{-2}	5
	-21.95	8.425×10^{-2}	1.58×10^{-2}	2.45×10^{-2}	9
	-6.89	8.425×10^{-2}	4.3×10^{-3}	1×10^{-2}	7

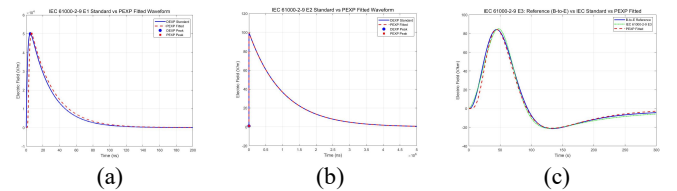


Fig. 1. Comparison of baseline optimized waveforms. (Red dashed: PEXP; blue solid: IEC) (a) E1 waveform (b) E2 waveform (c) E3 waveform

As shown in Fig. 1, the PEXP-fitted waveforms exhibit agreement with the IEC standard waveforms, demonstrating the high accuracy of the proposed method.

IV. CONCLUSION

This paper presents a unified PEXP-based description of the full-time HEMP waveform. Optimized via NSGA-II, the proposed method accurately fits the E1, E2, and E3 components. It resolves the derivative discontinuity inherent in the DEXP formulation for E1/E2 and provides a consistent representation for the E3 component while maintaining high fitting accuracy. The resulting unified description offers a rigorous waveform input for HEMP effect assessment, protection design, and EMC simulations.

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Equivalent Circuit Model of BCI Probes Accounting for Internal and External Parasitic Capacitances

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Abstract

In the conventional equivalent circuit model of a bulk current injection (BCI) probe, it was not clear whether it represented the parasitic capacitance of the inner or outer sides of the probe. This ambiguity may cause errors with a large-diameter cable. Thus, this study proposes a BCI probe model accounting for both capacitances. Using full-wave simulations, we confirmed that our model accurately represents insertion loss and probe capacitance, outperforming traditional methods for thick conductors. **Keywords**– Bulk current injection (BCI) probe, parasitic capacitance, equivalent circuit model.

I. INTRODUCTION

The bulk current injection (BCI) test is widely used to evaluate the immunity of an in-vehicle device. In this test, the BCI probe clamps on the cable and injects currents into the device under test (DUT). Since the setup affects the applied current, the equivalent circuit model enables estimation of the current, resulting in an efficient design. Especially, the equivalent circuit model of the BCI probe is most important component as the injection source.

In the previous study [1], the model for the BCI probe was developed under the assumption of a thin conductor (ideally, a conductor with negligible thickness). However, this assumption makes it unclear whether the equivalent parasitic capacitance represents the inner or outer side. This ambiguity can cause model degradation when the internal capacitance is comparable to the external capacitance, for example, in a large-diameter cable such as a high-voltage cable.

This study proposes an equivalent circuit model that accurately represents the internal and external capacitances. The model is verified focusing on the insertion loss and using full-wave simulation.

II. METHODOLOGY

This study assumes the thick conductor of the large-diameter cable is clamped by the BCI probe, as shown in Fig. 1. For this scenario, we propose an equivalent circuit model in Fig. 2. The proposed model accounts for two distinct parasitic capacitances: C_{int} , which represents the internal capacitance between the ideal thin cable and the probe shield, and C_{ext} , which represents the external capacitance between the shield and the system ground.

In the conventional equivalent circuit model [1], the equivalent parasitic capacitance C_p represents the shunt admittance of the lumped- π equivalent circuit. Based on our proposed model in Fig. 2, C_p can be derived from C_{int} and C_{ext} as follows:

$$C_p = \frac{0.5C_{int}C_{ext}}{C_{int}+C_{ext}} \approx 0.5C_{int} \quad (C_{int} \ll C_{ext}). \quad (1)$$

When the cable is thin as shown in Fig. 3(a), $C_{int} \ll C_{ext}$ holds, allowing C_p to be approximated by $0.5C_{int}$. However, as shown in Fig. 3(b), the internal capacitance increases for a large-diameter cable and is denoted as C'_{int} . To represent this increase, we introduce a factor κ defined as $\kappa \equiv C'_{int}/C_{int}$. As κ increases, C_{ext} becomes

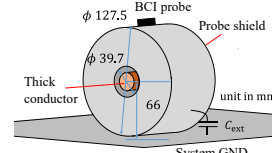


Fig. 1 Assumed scenario.

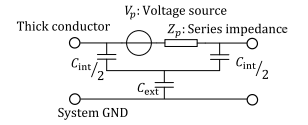
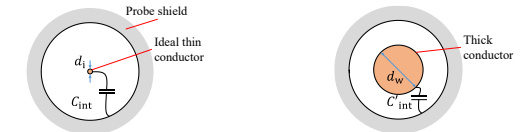
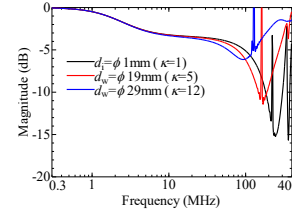


Fig. 2 Equivalent circuit model.

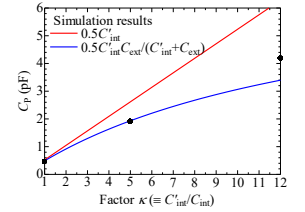


(a) Ideal thin conductor. (b) Thick conductor.

Fig. 3 Definition of internal capacitances C_{int} and C'_{int} .



(a) Insertion loss.



(b) Equivalent parasitic capacitance.

Fig. 4. Impact of internal capacitance increase.

non-negligible, and the approximation in (1) no longer holds.

To validate the proposed model, full-wave simulations were conducted based on the measurement setup described in the previous study [1]. It consists of an F-140 BCI probe clamped to a cable whose diameter is $\phi 1$ mm and a height of 66 mm above system GND. Fixtures were installed at both ends of the wire to correspond with the experiment. Subsequently, the effects of the fixtures were de-embedded based on [1]. The factor κ was set as 1 ($d_i=1$ mm), 5 ($d_w=19$ mm), and 12 ($d_w=29$ mm), and the changes in the insertion loss were evaluated. To investigate the cause, C_p was also derived from the lumped-PI circuit, which is identified from the Y-parameters.

III. RESULTS AND DISCUSSION

Fig. 4(a) indicates that the insertion loss increases with frequency, with the influence of the conductor diameter becoming more pronounced beyond 30 MHz. The sharp peaks observed above 100 MHz seem to be due to the de-embedding process. Fig. 4(b) shows the results of identifying C_p for each κ . These results indicate that the proposed method is suitable and effective for cables with thick conductors. Future work will evaluate the effects of cable diameter on applied current in a practical BCI test environment.

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Equivalent Circuit Modeling and Validation of a Laboratory Noise Reproduction Setup for Railway Vehicle EMC Evaluation

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Abstract– This study validates a laboratory noise reproduction setup that employs a single-wire system for simplicity to replicate actual railway vehicle noise environments. Equivalent circuit models for the noise generator and the injection clamp, including their associated wiring and termination loads, were developed. By identifying circuit constants through S-parameter fitting, the developed model demonstrated strong agreement with measured results in the frequency range of 0.3–30 MHz.

Keywords– railway, transient noise, equivalent circuit modeling, injection clamp, noise reproduction, EMC

I. INTRODUCTION

Dead section noise contains broadband frequency components [1]. Previous studies have presented techniques to reproduce transient noise currents that penetrate equipment during dead section passage in laboratory environments, focusing on single-wire systems [1] and shielded cable systems [2]. Rectangular pulse EMC test instruments are widely used for product reliability evaluation, yet their equivalent circuit models have not been reported. Therefore, while actual railway environments utilize shielded cables (Fig. 1(a)), this study initially employs a single wire to simulate these cables in the reproduction setup (Fig. 1(b)). This paper focuses on establishing equivalent circuit models for the noise generator and the injection clamp as a primary step toward a theoretical validation of the entire system. Integration of a detailed shielded cable model is reserved for future work to complete the comprehensive reproduction setup model.

II. EQUIVALENT CIRCUIT MODELING

Equivalent circuit models of the noise generator and injection clamp were derived using the test setup shown in Fig. 2. A noise generator (NoiseKen, INS-4020) was connected to the injection clamp (NoiseKen, CA-806) via a coaxial cable with a 50 Ω termination. Voltage waveforms at the termination were measured and FFT-analyzed for frequency characteristics. For the noise injection setup, a pulse width of 1 μ s was selected to capture the low-frequency peak observed around 1 MHz in the actual vehicle measurement. In the equivalent circuit models, the noise generator was represented as a voltage source with internal resistance, the coaxial cable as a transmission line, and the termination resistor as an RLC element. Circuit constants were identified by fitting S-parameters measured with a vector network analyzer (Keysight E5061B) using a circuit simulator (Keysight ADS).

III. COMPARISON RESULTS

Comparisons between measured and simulated frequency-domain

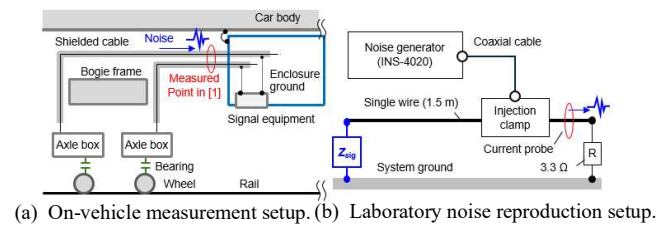


Fig. 1. Measurement configurations.

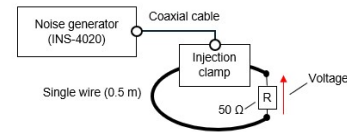


Fig. 2. Test setup for equivalent circuit modeling of noise generator and noise injection clamp.

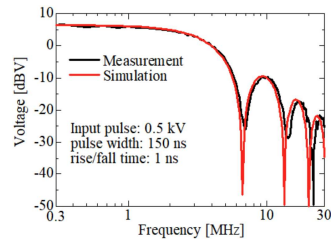


Fig. 3. Terminal voltage spectrum. (Fig. 2. setup).

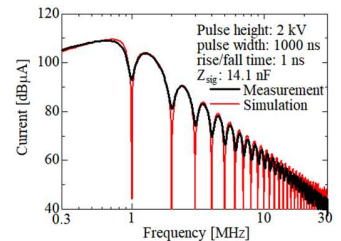


Fig. 4. Injected noise current spectrum. (Fig.1(b). setup).

characteristics are shown in Figs. 3 and 4. Considering that the main noise components are below 10 MHz [1, 2], the verification was conducted in the frequency range of 0.3–30 MHz. The simulated results demonstrated good agreement with the measured results within this range.

IV. CONCLUSION

The derived models for the noise generator and injection clamp were successfully validated. Future work will focus on incorporating a detailed shielded cable model into the equivalent circuit, as outlined in the Introduction. This will lead to a comprehensive model of the entire laboratory reproduction setup, further enhancing the accuracy of railway vehicle noise evaluations.

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A Novel Modeling Method adapted to Arbitrary 3D Airborne Transient Electromagnetic Antennas

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Abstract— We present a modeling method for airborne transient electromagnetic (TEM) transmitting antennas with arbitrary-shaped structures. By discretizing the transmitting antenna into a combination of edges or nodes within the finite element mesh, the mesh discretization preserves the geometric fidelity of complex antenna configurations while maintaining computational efficiency, and enables flexible integration of transmitting antennas into complex 3D geoelectric models (incorporating heterogeneous structures such as topography and bathymetry model). Unlike traditional approximation methods based on infinitesimal horizontal electric dipoles (HED) or vertical magnetic dipoles, this approach accurately characterizes the geometric properties and current distribution of practical transmitting sources by directly imposing current density on the degrees of freedom (DoF) corresponding to the Nédélec function space.

I. INTRODUCTION

This paper investigates the reliability of transient electromagnetic (TEM) forward modeling under arbitrary transmitter antenna configurations. A novel finite-element method (FEM) framework is presented to accurately characterize the antenna's geometric configuration, spatial position, and current distribution, and evaluate the resulting electromagnetic field propagation in subsurface media. Theoretical formulations demonstrate that explicitly discretizing the transmitter into edge or node combinations eliminates near-field errors inherent in point-source approximations, providing guidance for high-precision EM exploration and reliable multi-component data interpretation [1].

Idealized dipole approximations in conventional TEM modeling, such as those applied to finite-sized land loops, marine towed antennas, and dynamic airborne systems, can cause severe field strength errors exceeding 30% and systematic inversion biases of 10%–15% in high-resolution surveys [2]. The objective of this study is to overcome the limitations of existing total-field (TF) and secondary-field (SF) methods by developing a physically compatible modeling scheme in Nédélec spaces. The key contribution is a high-fidelity source representation that strictly enforces the divergence-free condition, which improves near-field accuracy and inherently avoids curl-field spurious solutions compared with prior idealized source models.

II. METHODOLOGY

In airborne TEM methods, the operation usually employs a lower frequency, thus displacement currents are neglected. This approximation enables us to focus on the electromagnetic induction process dominated by conduction currents. The governing equation for the electromagnetic fields can be expressed as

$$\nabla \times \mu^{-1} \nabla \times \mathbf{E} + i\omega \sigma \mathbf{E} = -i\omega \mathbf{j}_e \quad (1)$$

In full-wave finite element analysis, the contribution of source terms requires discrete treatment on the right-hand side of the system. All transmitter sources are defined by a series of coordinate points. In the total electric field calculation method, the Nédélec degrees of freedom associated with the source current are located on the edges of the mesh. When an edge coincides with the coordinates defining a source segment, the corresponding basis function can be regarded as a horizontal electric

dipole (HED) carrying current density $\mathbf{j}_e$. According to reference [3], this current density is defined as

$$\mathbf{j}_e = I \cdot d\mathbf{l} \cdot \delta(\mathbf{r} - \mathbf{r}_0) \quad (2)$$

Substitute expression (1) into the source term of the weak form equation, and obtain the integral form of the right-hand vector.

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$$f^i = \int_{\Omega} \omega \mathbf{j}_e \cdot \mathbf{N}_i d\Omega = \int_{\Omega} \omega (I \delta(\mathbf{r} - \mathbf{r}_0) d\mathbf{l}) \cdot \mathbf{N}_i d\Omega \quad (3)$$

We apply the current to the imaginary part. We constructed a semi-airborne ground-electric model of a circular transmitting antenna with terrain structure, as shown in Fig. 1.

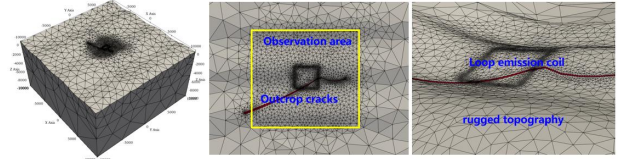


Fig. 1. A semi-airborne transient electromagnetic geoelectric model of circular sources with topography structure

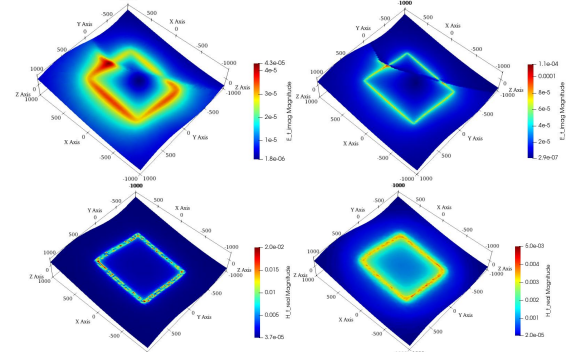


Fig. 2. Fig. 1. The imaginary parts of the electric and magnetic fields for the two planes, $z = 0$ and $z = 50$.

III. RESULTS AND DISCUSSION

The simulated electromagnetic field distributions across different elevation planes (as shown in Fig. 2) demonstrate that the proposed edge-based source integration technique accurately captures the near-field characteristics of the circular antenna over complex topography. Furthermore, by directly mapping the current density onto the Nédélec degrees of freedom, the method inherently preserves the divergence-free nature of the source, effectively mitigating the spurious solutions typically encountered in conventional dipole approximations.

IV. CONCLUSION

The proposed method overcomes traditional dipole-approximation limitations by directly imposing current density on Nédélec degrees of freedom, enabling accurate and efficient modeling of arbitrary airborne TEM antennas in complex 3D geoelectric media for high-precision electromagnetic exploration.

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Analysis of Screening Effects in Patterned VACNT Arrays Using Model Based Analysis

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Abstract – Cold cathode emitters offer fast response, low turn-on voltage, and high efficiency, with vertically aligned carbon nanotube (VACNT) arrays as representative sources. However, dense VACNT arrays suffer from electrostatic screening that degrades emission performance. This study analyzes screening effects in patterned VACNT arrays using model-based analytical approaches. A full 17×17 array is compared with representative reduced patterns formed by selective tip removal. Electrostatic and particle-tracking-based models confirmed by experiments are employed, and experimental data are further analyzed using the Murphy–Good (MG) framework. Reduced patterns show improved emission performance due to effective screening mitigation.

Keywords – Cold cathode, VACNT, Field emissions, Fowler Nordheim, Murphy good

I. INTRODUCTION

Cold cathode emitters have attracted attention due to their fast response, low turn on voltage and high efficiency compared to thermionic cathodes. Among them, vertically aligned carbon nanotube (VACNT) arrays are promising candidates owing to their high aspect ratio and strong field enhancement. However, dense VACNT arrays suffer from screening effect that degrades emission performance. Although patterning strategies based on selective tip removal have been proposed to mitigate this effect, a systematic physical interpretation of pattern-dependent emission behavior stays necessary [1].

II. METHODOLOGY

Figure 1 illustrates the schematic configuration of the VACNT array models and representative patterned structures derived from a 17×17 baseline. All patterns share an identical electrical environment, enabling direct comparison between experiments and analytical evaluations. Electrostatic analysis and Fowler–Nordheim–based particle tracking analysis, along with corresponding experimental measurements, were conducted following the method reported in [2].

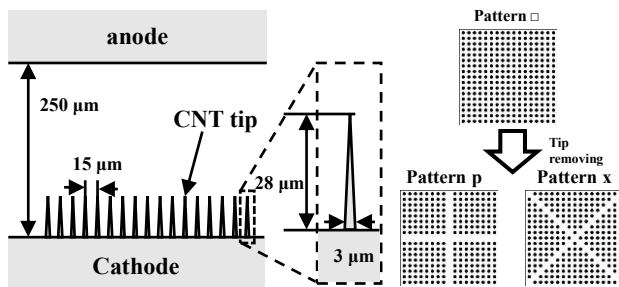


Fig. 1. Schematic illustration of VACNT arrays.

III. RESULTS AND DISCUSSION

Patterned VACNT arrays show enhanced field emission performance compared to the fully populated array. The observed improvement is consistently associated with reduced electrostatic screening due to selective tip removal. Detailed experimental results and quantitative comparisons are available in prior publications.

The MG-based analysis incorporates image-charge–changed tunneling barriers, providing a more physically consistent description of high-field emission behavior than conventional Fowler–Nordheim models [3]. The implications of pattern-dependent screening effects and model applicability are discussed, along with ongoing assessment of integrating MG-based formulations into model-based analysis.

IV. CONCLUSION

This study investigated field emission behavior in patterned VACNT arrays with a focus on inter-tip electrostatic screening effects. Selectively reduced patterns showed improved emission performance compared to a fully populated 17×17 array, emphasizing the importance of tip distribution in array design. The observed trends are consistently explained using electrostatic and particle-tracking–based analytical models. More analysis using the Murphy–Good framework offers extended physical insight into high-field emission behavior.

ACKNOWLEDGMENT

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Long-Term Testing Of Overstressed GaN-SiC Transistors

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Abstract– We report on long-term testing of overstressed Gallium Nitride on Silicon Carbide (GaN-SiC) transistors. To increase the peak output power from the transistors, they are driven at a higher drain-source voltage than the specifications recommend. If the duty cycle is kept well below the specified duty cycle, thermal heating is not a problem, but increasing the drain-source voltage by 50-100 % may introduce new failure modes. To investigate this, a test setup for long-term testing has been designed and built. It is shown that even though the peak output power from the transistors is increased by almost 100 %, the failure rate is low.

Keywords– gallium nitride, silicon carbide, failure mode, long-term testing, overstressing

I. INTRODUCTION

With the advent of semiconductor-driven phased-array High Power Microwave systems, the need for high peak power transistors becomes a crucial factor. Semiconductor-based systems may never reach the power density of electron tube systems, but the possibility of combining the output power from hundreds to thousands of transistor amplifiers in air using Active Electronically Scanned Array (AESA) systems offers many other advantages. The possibility of generating tailored waveforms, where pulse length, pulse repetition rate and frequency can be adjusted over a wide parameter space, represents a huge advantage compared to more traditional HPM-systems. Digital beam steering and lobe shaping offer fast aiming without the need for mechanical motion. Even with the benefits listed above, there will always be a need for high peak power from the transistors driving the array. To find the reduction in lifetime, or total number of pulses, that can be expected from a high peak power GaN-SiC transistor that is driven far outside the specifications, an automated test setup has been designed. It consists of signal and pulse generators together with gate and drain-source voltage supplies. The output signal is monitored by a spectrum analyzer where output power and phase noise are continuously analyzed. An infrared camera monitors the temperature of the entire amplifier and enables quick shutdown in the event of extreme temperature rise. The entire test setup is computer-controlled and automatically shuts down all power supplies in the event of transistor breakdown. The number of pulses and operational time are stored together with information on phase noise and output pulse power for the duration of the test. A schematic overview of the setup is depicted in Fig. 1.

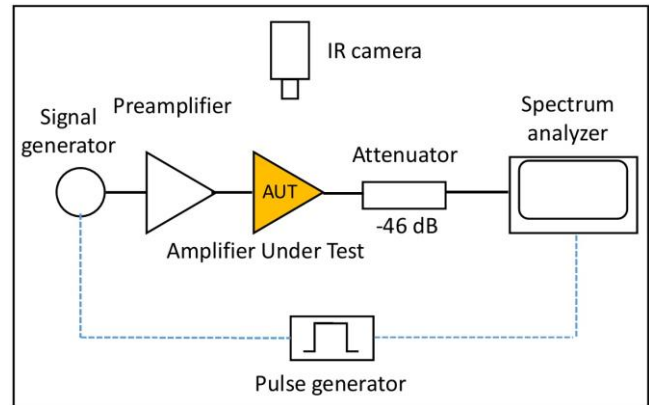


Fig. 1. Schematic overview of the test setup.

II. METHODOLOGY

The experiments performed use a Qorvo QPD1025 L-band transistor with a peak pulse power of 1.5 kW at 100 μ s long pulses and a 10 % duty cycle [1]. The transistor is designed to be driven at a drain-source voltage of 65 V. In these experiments, it is driven at 80 V with a pulse length of 10 μ s and with a repetition rate of 1 kHz, resulting in a duty cycle of 1 %.

III. CONCLUSION

The transistors proved to be more resilient than anticipated. At the time of writing, the long-term testing is still ongoing. We will report on both continuous and intermittent operation and compare the results from both experiments.

ACKNOWLEDGMENT

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A Conceptual Study Of A Ten-Cavity Relativistic Magnetron With TE11 Output Mode

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Abstract– This work investigates a ten-cavity relativistic magnetron suitable for generating high power microwaves. The objective is to modify the diffraction output to generate a TE11 mode. Using RF finite element simulations, we found a shape that converts the $2\pi/4$ and π modes to a single TE11 waveguide mode. The beam-to-microwave conversion efficiency is $\sim 23\%$ for the $2\pi/4$ mode (~ 750 MW) and $\sim 7\%$ for the π mode (~ 1.4 GW), which was verified using particle-in-cell simulations.

Keywords– relativistic magnetron, diffraction output, mode conversion

I. INTRODUCTION

Relativistic magnetrons are high efficiency microwave sources suitable for generating high power microwaves (HPM) [1,2]. The magnetron extracts the radiation axially using a diffraction output structure. The role of the diffraction output is to extract the radiation with high transmission efficiency and to convert the excited mode into a suitable TE or TM waveguide mode.

This paper investigates a ten-cavity relativistic magnetron, which has recently received attention for microwave generation e.g. in X-band [3]. For ten-fold symmetric diffraction outputs, the π mode is typically converted into a TE51 mode. The purpose of this work is to break the ten-fold symmetry of the diffraction output and instead create new symmetries that promotes conversion to a TE11 mode. The approach is similar as in reference [4], which shows how to convert the π mode to a TE11 mode in a six-cavity magnetron.

II. SIMULATION RESULTS

The relativistic magnetron geometry is shown in Fig 1. The interaction region has ten equally shaped cavities. Six of the cavities are extruded into the diffraction output, while four of the cavities are tapered (i.e. $\beta=90^\circ$).

The interaction region has six modes that can potentially be excited. Using eigenmode calculations with finite element software, the cutoff frequencies of each mode were calculated. The cutoff frequencies of the 2π mode is 4.55 GHz, the π mode 2.96 GHz, the $2\pi/3$ mode 2.93 GHz, the $2\pi/4$ mode 2.84 GHz, the $2\pi/5$ mode 2.59 GHz and $2\pi/6$ mode 1.80 GHz. For reasonable applied voltages, we expect to excite the π , $2\pi/3$ and $2\pi/4$ modes.

To study mode conversion when the waves pass through the diffraction output, 3D finite element simulations in the frequency domain was performed. For the proposed magnetron, the π and $2\pi/4$ modes convert to a TE11 mode (see Fig. 2), while the $2\pi/3$ and $2\pi/5$ modes convert to a TE21 mode.

To verify the performance and if electron-wave interaction can excite the magnetron modes, we performed particle-in-cell calculations. We studied a case with an applied magnetic field strength of 0.34 T in the interaction region, with varying applied voltages. The π , $2\pi/3$ and $2\pi/4$ modes were observed and oscillating near their respective cutoff-frequencies. However, only the $2\pi/4$ has a high efficiency of $\sim 23\%$ with a peak power of ~ 750 MW (see Fig. 3). The π mode has an efficiency of 7%, with a peak power of

~ 1.4 GW.

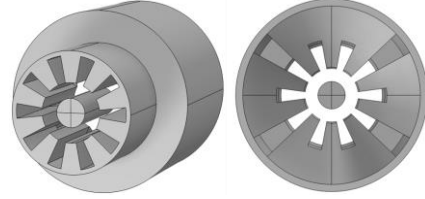


Fig. 1. Left: The interaction region, with $r_c=1$ cm, $r_a=2.05$ cm, $r_v=3.9$ cm, $\psi_v=15^\circ$. Right: The diffraction output, with $\alpha=21^\circ$, $\beta=90^\circ$ and $r_{cyl}=6$ cm, where α is the cone angle of the diffraction output, and β is the angle of the tapered cavities.

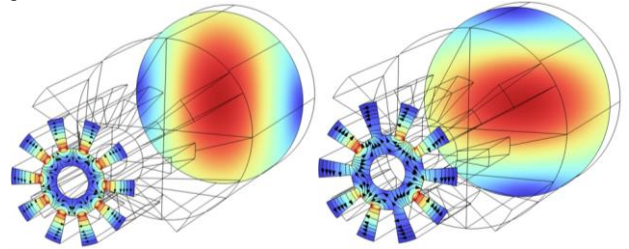


Fig. 2. Finite element simulation of the diffraction output. Left: π mode converts to a TE11 (vertical polarization) mode within 2.96-3.05 GHz. Right: $2\pi/4$ mode converts to a TE11 mode (horizontal polarization) between 2.84-2.95 GHz

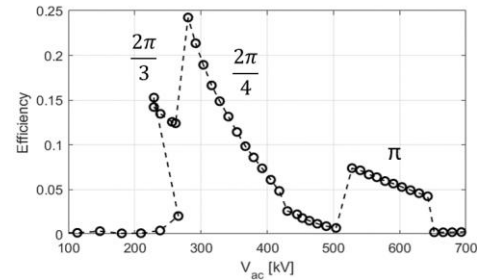


Fig. 3. Beam-to-microwave efficiency calculated with particle-in-cell simulations for $B=0.34$ T and an interaction region length of 13.5 cm.

III. DISCUSSION AND CONCLUSIONS

The proposed magnetron has the potential to convert both the π and $2\pi/4$ mode to a TE11 mode. The advantage is that a TE11 mode can be generated at two frequencies and polarization. The main problem is the efficiency of the magnetron, in particular when using the π mode. Future work therefore includes improving the efficiency by e.g. modifying the interaction region further.

ACKNOWLEDGMENT

This work was sponsored by the Swedish Armed Forces.

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EMP Generation by Ultra-High-Power Laser in Air

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Abstract— The enormous peak power of ultra-high-power (>10 TW) laser pulses is advantageous for generating strong electromagnetic pulses (EMP). In this work, we generated laser pulses with peak powers exceeding 20 TW in air and focused them onto a solid target to produce EMP. The emitted EMP from a single laser shot were simultaneously measured in three different directions by using D-dot sensor placed 30 cm from the target, yielding maximum voltages up to ~ 4 V in the GHz range. Our results demonstrate that EMP can be efficiently generated using ultra-high-power lasers in air, opening new opportunities for generating strong EMP in the GHz range.

Keywords— Electromagnetic pulses, High-power femtosecond laser in air, Laser-plasma interaction

I. INTRODUCTION

Recent advances in high-power femtosecond laser technology have enabled the generation of laser pulses with multi-petawatt peak power.^[1] The interaction of such ultraintense lasers with matter serves not only as a key tool for investigating strong-field physics but also as a platform for various applications, including particle acceleration and the generation of secondary radiation sources. Powerful electromagnetic pulses (EMP), which are inevitably produced during these interactions, can damage electronic equipment; therefore, most studies have focused on methods to avoid or mitigate their effects.^[2]

Attempts to utilize advanced high-power femtosecond laser technology in atmospheric environments have continued for many years; however, the strong nonlinearity of air has severely limited the use of ultraintense lasers in open air. Recently, we developed a novel approach based on a helium-filled compressor, which enables the stable generation of ultra-high-power laser pulses exceeding 20 TW directly in the atmosphere.^[3] This approach opens opportunities to extend ultraintense laser-matter interaction studies, previously conducted mainly in vacuum environments, to open-air conditions. As a result, EMP, which have traditionally been regarded as effects to be avoided in this field, can instead be explored as useful phenomena for various applications.

In this study, we measured EMP generated by tightly focusing an ultra-high-power laser pulse onto a thin metallic foil in air. To maximize the emitted EMP, we investigated its dependence on target parameters. This work demonstrates the development potential of atmospheric ultra-high-power-laser-driven EMP generation and lays the foundation for producing strong GHz EMP in air.

II. METHODOLOGY

The experiment was performed using a 20 TW Ti:sapphire laser capable of producing 60 fs, 1.3 J pulses after compression in air. The laser pulse is focused onto a copper foil target by an off-axis parabolic mirror with a f-number of 1.7. Targets with thicknesses ranging from 0.5 to 100 μm and sizes from 1 to 50 mm were tested using a wide range of aspect ratios. The generated EMP was detected using D-dot sensors placed 30 cm from the target at three different positions (along the beam propagation direction, 50° from

the propagation direction, and along the beam reflection direction), and the signals were recorded with a sampling rate of 40 GS/s.

III. RESULTS AND DISCUSSION

Figure 1 shows the EMP signal produced by the interaction of a laser pulse with a copper foil target. The strongest signal was obtained when a target with a thickness of 50 μm and dimensions of 50×2.5 mm was oriented along the laser polarization direction. The difference in signal intensity among the measurement positions was not larger than the shot-to-shot fluctuations of the laser. Despite variations in the target configuration, the FFT spectra of the measured EMP signals were consistently observed in the 1–10 GHz range, with pronounced peaks around 5 GHz. The present results remain at a very preliminary level, making it difficult to discuss the EMP generation mechanism in detail. For this purpose, a precise spatiotemporal characterization of the focused laser beam in air is essential, along with more extensive experimental and theoretical investigations.

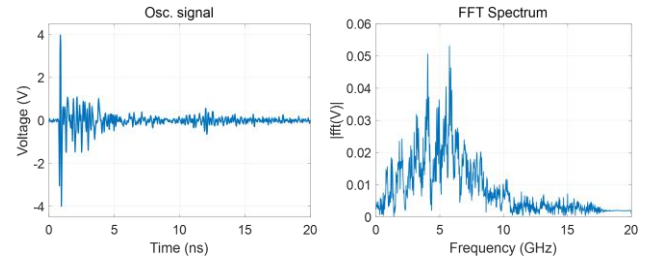


Fig. 1. Raw EMP signal driven by an ultra-high-power laser and its FFT spectrum. The results were not calibrated using the transfer function of the measurement system.

IV. CONCLUSION

The interaction between ultra-high-power lasers and matter is a well-established and highly effective method for generating strong EMP in the GHz range. In this work, we extended this approach to atmospheric conditions and demonstrated the basic feasibility of EMP generation in air. With further studies, it is expected that much more efficient and powerful EMP generation can be achieved.

ACKNOWLEDGMENT

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Diagnosis of 20-TW Femtosecond Laser Focusing in Air for Laser-Plasma EMP Generation

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Abstract— A femtosecond 20-TW laser was operated at atmospheric pressure by filling the compressor chamber with helium, and the compressed beam was extracted through a window into ambient air and focused. Several diagnostics were employed to determine whether the beam reached the relativistic-intensity regime. The focal spot size was estimated from the crater size produced on a solid target, while the pulse duration was measured using a SPIDER system and the pulse energy was measured with an energy meter. Based on these measured parameters, the laser intensity at focus in air was calculated. The focused laser was then irradiated onto a metal target, and the generation of X-rays and GHz-band electromagnetic pulse (EMP) was confirmed.

Keywords— laser-plasma interaction, laser-plasma EMP generation, Focal intensity diagnosis

I. INTRODUCTION

The focusing of femtosecond lasers in ambient air has attracted considerable interest because it can be applied to a variety of radiation sources. [1] In such experiments, a central question is whether the laser remains sufficiently well focused after propagation in air to reach an intensity high enough to initiate plasma formation at the target surface. The objective of this study is to assess the air-focusing performance of a 20-TW laser by examining the target response under laser irradiation. The key contribution of this work is an experimental diagnosis of near-relativistic interaction conditions based on focal-spot characterization. Although the present measurements were conducted as part of a broader project aimed at X-ray and EMP generation, this presentation focuses on verifying the relativistic plasma formation.

II. METHODOLOGY

A 20-TW femtosecond laser system was operated at atmospheric pressure by filling the compressor chamber with helium. [2] The compressed beam was then extracted through an output window into ambient air and focused to evaluate whether the on-target interaction reached a regime relevant to relativistic plasma formation as shown in Fig. 1. For a high-power femtosecond laser such as that used in this experiment, the electric field at focus is sufficiently strong to ionize matter immediately, making a direct measurement of the focal intensity impossible with conventional diagnostics. Therefore, the focal intensity was estimated indirectly by measuring several parameters and calculating the intensity using the following equation:

$$I = \frac{1.88U}{\pi w_0^2 \tau}, \quad (1)$$

where I is the peak laser intensity at focus, U is the laser pulse energy delivered to the focus, w_0 is the $1/e^2$ radius of the focal spot, and τ is the full width at half maximum (FWHM) pulse duration. In

this approach, three parameters are required to determine the focal intensity, namely the focal spot size, the pulse duration, and the laser energy delivered to the focus. In this experiment, the focal spot size was determined by focusing the laser onto a solid target and measuring the size of the resulting crater using confocal microscopy and SEM. The laser energy delivered to focus was measured using the knife-edge method, and the pulse duration after passing through the focus was measured using a SPIDER. In addition, the intensity was also estimated by analyzing the spectrum of the plasma emission generated when the laser was focused in air. As a complementary method, the laser intensity was calculated from analysis of the X-ray spectrum produced when the focused laser irradiated a solid target.

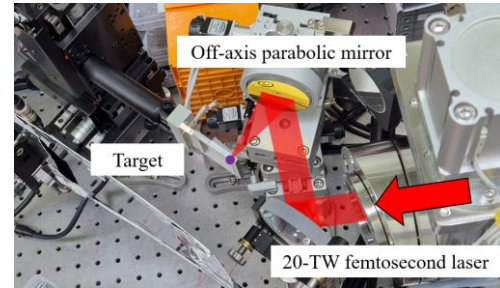


Fig. 1. Photograph of the experimental setup, showing the target, the off-axis parabolic mirror, and the propagation direction of the 20-TW femtosecond laser (red arrow).

IV. CONCLUSION

A 20-TW femtosecond laser was successfully operated at atmospheric pressure and focused in ambient air. The focal intensity was evaluated indirectly from the focal spot size, pulse duration, and laser energy, supported by plasma-emission and X-ray spectral analysis. The results indicate that the focused beam reached a regime relevant to relativistic plasma formation on a solid target and was also used for laser-plasma EMP generation and measurement. [3]

ACKNOWLEDGMENT

This research was supported by the Challengeable Future Defense Technology Research and Development Program through the Agency for Defense Development (ADD) funded by the Defense Acquisition Program Administration (DAPA) in 2025 (No.915107201).

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Using Complex Time Delay to Control Pulse Propagation in Electrically-Large Enclosures

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Abstract— We introduce the concept of complex time delay (CTD) and show that it governs the time and frequency shifts of Gaussian pulses propagating through complicated multi-modal enclosures. We show that chirped pulses offer the opportunity for extraordinary control of pulse propagation in generic scattering systems. The theory is tested experimentally for pulses propagating in 1-, 2-, and 3-dimensional complex systems with overlapping modes, and their time and frequency shifts are in good agreement with predictions.

Keywords— delay effects, electromagnetic scattering, pulse propagation, scattering matrix, time delays

I. INTRODUCTION

Our goal is to explore new and simple ways to extract useful information about arbitrary and generic scattering systems through measurements and analysis of the scattering (S)-matrix. We assume that these systems are reverberant in the sense that waves propagate across the length and breadth of the system multiple times before significantly decaying in amplitude. We introduce the concept of *complex time delay* and demonstrate its utility in describing the time and frequency shifts that both chirped and un-chirped pulses suffer as they propagate through complex multi-port scattering systems. Chirped pulses are utilized in many applications, as well as for fundamental studies, so understanding and controlling their propagation in complex settings will have wide impact. Predicting how chirped pulses propagate through electrically-large enclosures is also important for addressing electromagnetic compatibility issues.

II. THEORY

Complex time delay (CTD) examines the dependence of scattering matrix elements on frequency, and includes the phase variation (real part) and the amplitude variation (imaginary part) of the scattering matrix element. Here we focus on just one type of CTD, namely the transmission complex time-delay [1, 2],

$$\tau_T^{(ij)} = \frac{-i}{2\pi} \frac{d}{df} \ln(S_{ij}), \quad (1)$$

where $i \neq j$ are two distinct ports, and S_{ij} is the complex and frequency dependent scattering matrix element between those two ports, and f is frequency.

We experimentally tested the predicted equivalence of CTD measured in the frequency domain ($\tau_T^{(ij)}$) to the time (D_t) and frequency (D_f) shifts of pulses measured in the time-domain, for the same scattering system [3]. This relationship is expressed as [4],

$$D_t = \text{Re} \left[\tau_T^{(ij)} \right], \quad (2)$$

$$D_f = -\tilde{\Delta}^2 \text{Im} \left[\tau_T^{(ij)} \right], \quad (3)$$

where $\tilde{\Delta}$ is the bandwidth of the pulse.

We have discovered that chirping the pulse allows one to explore the CTD over a range of frequencies, thus modifying the time shift of a time-domain pulse as,

$$D_t^{\text{chirp}} = \text{Re} \left[\tau_T^{(ij)} \right] - \Omega' T^2 \text{Im} \left[\tau_T^{(ij)} \right], \quad (4)$$

where Ω' is the linear chirp rate and T is the duration of the pulse. This expression shows that the time shift suffered by a chirped pulse is a unique linear combination of the real and imaginary parts of the system CTD, modified by properties of the chirped pulse that we have under our control, namely the center frequency, chirp rate (positive or negative), and the pulse duration. Equation (4) offers a remarkable degree of control over pulse propagation between two ports, through a combination of pulse shaping (choosing center frequency, Ω' , and T), and cavity shaping (to modify the frequency complexation of $\tau_T^{(ij)}$).

III. EXPERIMENTAL RESULTS AND DISCUSSION

We have verified the accuracy of Eq. (4) using time-domain pulses sent through microwave graphs, over a range of chirped-pulse time-bandwidth products ($\Omega' T^2$) ranging from -10^3 to $+10^3$. We also demonstrated the principle of pulse shaping to achieve a specific scattering goal, utilizing chirped Gaussian pulses engineered to have $D_t^{\text{chirp}} \approx 0$ in a given scattering system, over a broad range of center frequencies [5].

IV. CONCLUSION

We have introduced the concept of complex time delay and shown that it correctly predicts the time shift and frequency shift of chirped and un-chirped pulses that propagate through complex multi-port scattering systems. This quantity offers new opportunities to control pulsed signal propagation through generic scattering systems.

ACKNOWLEDGMENT

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Impedance Recurrence Statistics of Complex Scattering Systems

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Abstract— We investigate the question of how frequently the impedance presented by a complex multi-modal system repeats the same desired value. Specifically, we investigate the expected difference in frequency between which impedance values of a given system are approximately repeated. The answer is statistical in nature and applies to any generic scattering system that is many wavelengths in size. The predictions are tested experimentally in microwave graphs, billiards, and three-dimensional cavities, and are found to be in reasonable agreement with theory.

Keywords - correlation functions, electromagnetic impedance, electromagnetic scattering, random matrix theory, recurrence

I. INTRODUCTION

We consider the propagation of waves through complicated scattering systems with a finite number of asymptotic scattering channels coupled to the outside world. When the dimensions of the systems are much larger than the wavelength of the waves propagating within, the scattering properties become extremely sensitive to cavity details, such as the shape of the boundaries and the presence of internal objects. This sensitivity suggests that a statistical approach to understanding the system is appropriate.

II. METHODOLOGY

We have developed the random coupling model (RCM) to incorporate non-universal physical features that practical systems invariably have, such as coupling details of ports or antennas and short orbits (also known as direct processes), with a random matrix theory (RMT) description of the wave propagation within the enclosure [1]. One of the applications of the RCM is identifying the underlying universal fluctuations of impedance from experimental ensemble data, which is done by normalizing the measured cavity impedance matrix by the radiation impedance specific to the ports or antennas used in the measurement. The radiation impedances of the ports are then used to normalize the actual impedance matrix, yielding a dimensionless impedance matrix Z that is independent of many details of the system, allowing for comparison with universal RMT predictions.

We address this question: If a port of the system presents a particular value of the complex impedance Z^t at a frequency f , what is the average frequency interval Λ before the same value of impedance Z^t , within some specified tolerance, is seen again? Empirically, we find that the expected frequency interval Λ between repetitions of a target impedance depends on the mean mode spacing Δ of the system, the degree of absorption α of the system, the value of the desired target impedance Z^t , and the tolerance ϵ on the target value. Our predicted scaling law for Λ provides a predictive tool for how often a desired impedance condition occurs naturally in frequency for a given cavity or resonator, which has practical relevance for electromagnetic compatibility.

The predicted scaling law for the frequency interval Λ between

repetitions of a target impedance Z^t with tolerance ϵ is,

$$\Lambda(Z^t) = 2\sqrt{\pi\alpha} \frac{\Delta}{\epsilon} \frac{\mathcal{P}(Z^{peak})}{\mathcal{P}(Z^t)},$$

which applies in the high-loss limit ($\alpha \gg 1$), and the expression involves the ratio of the peak of the probability density function (PDF) of impedance $\mathcal{P}(Z^{peak})$ to the value of the PDF at the target impedance value $\mathcal{P}(Z^t)$. These in turn have relatively simple analytical forms in the high-loss limit for the diagonal and off-diagonal components of impedance [2].

III. EXPERIMENTAL RESULTS AND DISCUSSION

Experiments are performed in one-dimensional microwave graphs, two-dimensional billiards, and three-dimensional complex enclosures [2, 3]. We compile impedance statistics by varying phase shifters embedded in the microwave graph, and tunable reflective metasurfaces in the billiards and 3D cavity, and measuring the scattering matrix as a function of frequency. The two-frequency correlation functions of impedance are first examined, and found to be in very good agreement with the predictions of RMT and the RCM. We also examine the PDF of $\mathcal{P}(Z)$ and follow their evolution with system loss parameter α . Finally, we examine the scaling of $\Lambda(Z^t)$ with tolerance ϵ , system loss parameter α , and the distance of the target impedance Z^t from the target impedance Z^{peak} , for both diagonal and off-diagonal components of the impedance matrix. In general, we find excellent agreement between theory and experiment in the high loss regime.

IV. CONCLUSION

We have derived a unique prediction for the recurrence interval between two similar values for the system impedance for generic complex scattering systems. The predictions have been verified by measurements on a wide variety of experimental scattering systems.

ACKNOWLEDGMENT

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A Stochastic Framework for Target Effect Probability Estimation in Complex Enclosures

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Abstract— Assessing the vulnerability of electronic systems operating inside electrically large enclosures is challenging under high-power electromagnetic (HPEM) exposure because wave-chaotic behavior produces highly variable internal field coupling. Even small changes in boundary conditions or frequency can cause substantial variations in the electromagnetic fields within the enclosure, rendering deterministic analysis unreliable. The random coupling model (RCM) statistically characterizes the coupling behavior of electromagnetic fields to target ports under such conditions. Here, we develop a stochastic framework that combines this statistical description with the susceptibility of the target to predict the probability of target effects in complex enclosures.

Keywords— High-power electromagnetic (HPEM), random coupling model (RCM), stochastic framework, target effect

I. INTRODUCTION

High-power electromagnetic (HPEM) exposure poses a significant threat to electronic systems, making prior vulnerability assessment essential. However, when such systems are located inside electrically large and irregular enclosures, small variations in boundary conditions or frequency can cause unpredictable variations in the power induced at internal electronic components. This sensitivity prevents deterministic prediction and therefore requires a statistical framework for vulnerability assessment.

The random coupling model (RCM) provides a statistical description of the power coupled to internal ports in wave-chaotic enclosures [1]. By characterizing enclosure variability through a loss parameter, it predicts the distribution of power coupled to a target port regardless of its specific position within the enclosure. The susceptibility threshold of a target device is not deterministic but varies across samples, and when represented as a distribution, it can be combined with the coupled power statistics to evaluate the probability of target effects in wave-chaotic enclosures [2].

In this paper, we propose a stochastic framework that integrates the RCM-based coupled power distribution with the susceptibility threshold distribution of the target. This formulation enables prediction of target effect probability in wave-chaotic enclosures.

II. TARGET EFFECT PREDICTION FRAMEWORK

As shown in Fig. 1, wave-chaotic enclosures exhibit sensitive and irregular internal field behavior, where small changes in excitation conditions or target position produce large variations in coupled responses. This sensitivity makes deterministic prediction impractical, requiring a statistical framework for target coupling. Fig. 2 illustrates the proposed statistical framework for target effect prediction. Within this framework, the RCM predicts the statistical distribution of power coupled to a target port inside a wave-chaotic enclosure at a given frequency and incident power level, once the loss parameter α is extracted from a finite ensemble of measured configurations [2]. In this way, enclosure-induced variability can be quantified without explicit modeling of the internal field

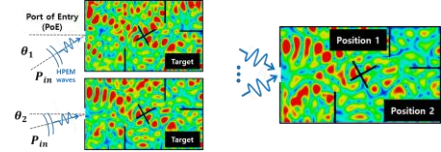


Fig. 1. Variability of field coupling in a wave-chaotic enclosure.

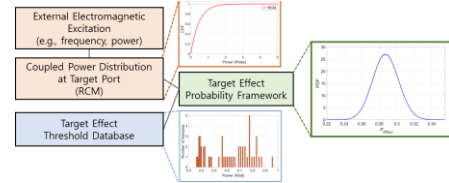


Fig. 2. Framework integrating RCM-based coupling statistics and target threshold characterization.

configuration. A target effect is defined as the transition of a device into a malfunction or non-operational state when the power coupled to the target exceeds its susceptibility threshold. Therefore, prediction of target effect probability requires not only the statistical description of coupled power but also characterization of the device threshold. The threshold may be obtained from direct injection tests, prior experimental data, or other available information depending on the test conditions. The threshold probability density function (PDF) is denoted $f_x(x)$, and the cumulative distribution function (CDF) of the coupled power predicted by the RCM is denoted by $F_{P_{\text{Target}}^{\text{RCM}}}(x)$. The probability of target effect is expressed as

$$P_{\text{Effect}} = \int_0^{\infty} [1 - F_{P_{\text{Target}}^{\text{RCM}}}(x)] f_x(x) dx \quad (1)$$

This formulation combines the statistical description of enclosure-induced coupling with the statistical representation of the target threshold, enabling quantitative prediction of target effect probability in wave-chaotic enclosures.

III. CONCLUSION

We present a stochastic framework for predicting target effect probability in wave-chaotic enclosures under high-power electromagnetic exposure. By combining the statistical power coupling described by the RCM with the susceptibility characteristics of the target, the formulation enables quantitative vulnerability assessment. This approach provides a basis for evaluating electromagnetic effects in wave-chaotic environments.

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Electric Field Anisotropy Induced by Dimensional Ratios in Rectangular Electromagnetic Chambers

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Abstract – This study analytically predicts field anisotropy in unstirred enclosures using modal wall-loss. Driven theory shows that internal field distribution is frequency-independent and determined by cavity geometry. The proposed model achieves a four-time less RMSE, enhancing statistical EM analysis for practical EMC and EMP testing.

Keywords – unstirred enclosure, field anisotropy, EMC, EMP

I. INTRODUCTION

As operational frequencies increase, deterministic electromagnetic analysis becomes computationally challenging, necessitating statistical approaches. Probabilistic methods to replace deterministic methods rely on the "well-stirred" hypothesis, which assumes field isotropy. However, practical unstirred enclosures show significant anisotropy influenced by their specific geometry and source characteristics. This study proposes an analytical method to predict these anisotropic behaviors by adopting directional weighting factors based on the enclosure's dimensional ratios.

II. METHODOLOGY

Conventional methods assume isotropic electric field intensity through proper stirring techniques [1]. Without stirring, however, the field distribution depends on source properties and enclosure geometry, modeled as a weighting function in a local coordinate system. Defining l , t_1 , and t_2 as the longitudinal and orthogonal transverse axes—which also correspond to the enclosure's edge lengths. The energy weighting factor w_l for the l -axis is formulated based on [2], neglecting source effects:

$$w_l = \frac{E_l^2}{E_{total}^2} = \frac{E_{l,TM}^2}{E_{TM}^2 + E_{TE}^2} = \frac{E_{l,TM}^2/E_{TM}^2}{1 + E_{TE}^2/E_{TM}^2} = \frac{E_{l,TM}^2/E_{TM}^2}{1 + Q_{TE}/Q_{TM}} \quad (1)$$

The weighting factor is obtained by calculating the ratio between the electric field components in numerator and the quality factor in denominator, as defined in (1). In over-moded regimes, where both parameters depend on the wavenumber, a statistical interpretation is employed by applying spherical averaging on the k -space.

$$\langle |E_{l,TM}|^2 \rangle = \frac{2}{\pi} \int_0^{\pi/2} \int_0^{\pi/2} \sin^3 \theta \, d\theta d\phi = \frac{2}{3} \quad (2)$$

$$\frac{\langle Q_{TE} \rangle}{\langle Q_{TM} \rangle} = \frac{\int_{\Omega} 1/Q_{TE} \, d\Omega}{\int_{\Omega} 1/Q_{TM} \, d\Omega} = \frac{3l + 3h(t_1, t_2)}{5l + h(t_1, t_2)} \quad (3)$$

In (3), the function $h(t_1, t_2) = 2t_1t_2/(t_1 + t_2)$ is the harmonic mean. Since the harmonic mean satisfies the commutative property, the weighting function is ultimately determined by the ratio between the longitudinal and transverse components in the local coordinate system. By substituting Eqs. (2) and (3) into (1), the final expression for the weighting function is

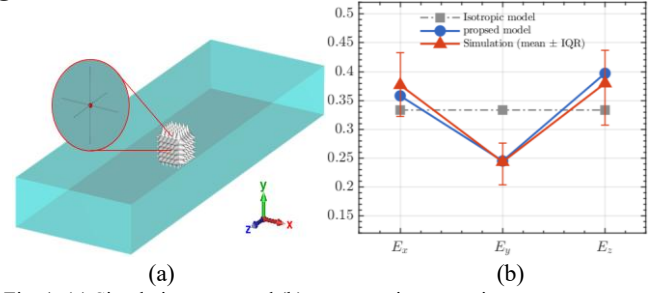


Fig. 1. (a) Simulation setup and (b) energy ratio comparison

$$w_l = \frac{1}{3} \times \frac{5l + h(t_1, t_2)}{4l + 2h(t_1, t_2)} \quad (4)$$

When the axis of interest aligns with the z -axis of the global coordinate system, the local axis l corresponds to the edge length along the z -direction, while t_1 and t_2 correspond to the lengths along the x and y axes, respectively. For the enclosure shown in Fig. 1(a), the predicted energy ratios by (4) are 0.358 for the x -direction, 0.243 for the y -direction, and 0.397 for the z -direction. These values deviate by up to 27% compared to the conventional isotropic assumption.

III. RESULTS AND DISCUSSION

For validation, a $30 \times 10 \times 90$ cm enclosure (conductivity: 5000 S/m) was modeled. To minimize source effects, results from a centered tri-pole antenna rotated in 90° steps were ensemble-averaged. Band-averaged electric field analysis from 5 to 10 GHz showed excellent agreement with theory. Notably, relative errors for the x , y , and z axes were reduced by 2.5, 45, and 3.2 tiems, respectively, proving the proposed model's superior accuracy over the isotropic assumption.

IV. CONCLUSION

This paper presents a weighting function model that accounts for enclosure geometry in predicting directional electric field energy. The results demonstrate that incorporating structural dimensions significantly improves accuracy over the standard isotropic assumption in over-moded, non-stirred environments.

ACKNOWLEDGMENT

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Control and Diagnostics System for High-Power Microwaves (HPM)

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Abstract– This paper presents the design of a control and diagnostic system for the stable operation of a 10 GHz, gigawatt-class pulsed high-power microwave (HPM) source. Because such systems operate in extremely strong electromagnetic environments, protection against self-induced electromagnetic interference is essential. The proposed architecture employs an electromagnetic shielding shelter, physical separation of the control console, and fiber-optic communication isolation. The system consists of eleven electrically isolated modules designed to minimize interference between analog and digital subsystems.

Keywords– Relativistic Backward Wave Oscillator, high-power microwaves, control and diagnostic system, EMI

I. INTRODUCTION

High-power microwave (HPM) systems capable of generating gigawatt-level pulsed electromagnetic radiation have attracted significant attention due to their applications in defense systems, advanced electromagnetic testing, and high-power RF research. The reliable operation of such systems requires sophisticated control and diagnostic infrastructures because the devices operate in extremely harsh electromagnetic environments where strong self-generated electromagnetic fields can disrupt electronics and instrumentation. Therefore, robust monitoring, precise timing control, and effective electromagnetic protection are essential for stable and repeatable operation. Developing integrated control and diagnostic systems that can withstand intense electromagnetic interference while providing accurate system monitoring has become a critical component in modern HPM experimental platforms.

II. METHODOLOGY

The control and diagnostics system was designed according to several engineering principles aimed at minimizing electromagnetic interference in an extreme HPM environment. First, physical isolation and electromagnetic shielding were implemented by installing all control and diagnostic units in individually enclosed metallic housings to block external electromagnetic radiation. Second, signal integrity was ensured by placing measurement and control modules close to the target system to minimize cable length, while employing low-impedance circuit design and shielded signal lines to reduce noise coupling. Third, communication and power isolation were achieved using fiber-optic communication links for all external data transmission and electrically isolated internal power supplies to prevent electromagnetic noise from entering the system. Finally, a modular architecture was adopted in which the overall system is divided into eleven functional modules, consisting of three control modules, five diagnostic modules, two communication modules, and an integrated control console, enabling flexible operation and improved robustness against electromagnetic disturbances.

III. RESULTS AND DISCUSSION

Based on the proposed design, the system is expected to enable stable control and reliable data acquisition under high-power HPM operation. Separation of analog and digital grounds within each module effectively reduces mutual interference. The integrated control console is physically separated from the shielded shelter, allowing safe remote operation while maintaining accurate system monitoring. Fiber-optic communication and electrically isolated diagnostic lines further enhance immunity to electromagnetic interference. The overall block diagram of the control and diagnostic system is shown in the figure.

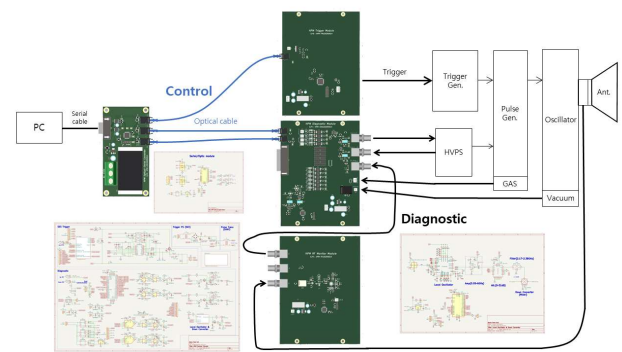


Fig. 1. Control and diagnostic system of HPM

IV. CONCLUSION

This study presents a protection-oriented design of a control and diagnostic system for the safe operation of an ultra-high-power pulsed electromagnetic source exceeding 1 GW. The proposed approach employs individual metallic shielding, fiber-optic-based signal transmission, and electrical isolation of power and subsystem lines to mitigate malfunction and potential damage in extreme electromagnetic environments. The presented design framework is expected to serve as a key enabling technology for the future development and practical deployment of high power electromagnetic source systems.

ACKNOWLEDGMENT

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Gauge Invariance in Classical Electrodynamics

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Abstract: This paper deals with the classical subject of Gauge Invariance. Jackson [1] who has stated, “It seems necessary from time to time to show that the electric and magnetic fields are independent of the choice of gauge for the potentials.” We have demonstrated the veracity of this statement in classical electrodynamics with an example.

I. INTRODUCTION

Vector and scalar potentials and an arbitrary gauge relationship are used for convenience in solving boundary value problems involving electromagnetic (EM) fields. Two of the most common gauges are the Lorenz and Coulomb gauges. It is less commonly known that the Lorenz and Coulomb gauges are two special cases of a generalized gauge called the velocity gauge (or v -gauge). This generalized gauge is usually not presented to the students of electrodynamics, and it is characterized by a non-physical parameter v , that has the units of a velocity and can vary between c (the speed of light) and infinity. The use of gauges prompts several truly relevant questions. If one uses the different gauges provided by v , are the electromagnetic fields the same? Is gauge invariance a sacred principle in classical electrodynamics? Is the set of gauges that can be used to compute an electromagnetic field discrete, or is it continuous, as the generalized velocity gauge suggests? Within this generalized gauge framework, we have

$$\nabla \cdot \vec{A}_v(\vec{r}, t) + \frac{1}{v^2} \frac{\partial \phi_v(\vec{r}, t)}{\partial t} = 0. \quad (1)$$

There has been some controversy regarding gauge invariance and uniqueness. Engelhardt [2] has claimed that Maxwell’s equations have non-unique solutions. It is also interesting to note that Wu and Wu [3] have shown that gauge invariance is violable in Quantum Physics by considering the decay of Higgs Boson into two photons! Given these considerations from published works, reviewing gauge theory and establishing its fundamental precepts is worthwhile.

II. EXAMPLE USING A HERTZIAN DIPLOE.

For example, consider the ideal Hertz dipole shown in Figure 1. We assume a time-harmonic variation of a current element $I(t) = I_0 \cos \omega t$ flowing in an infinitesimally thin conductor along the z -axis over a

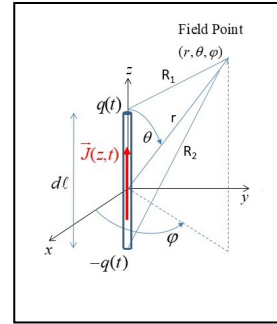
differential length $-d\ell/2 \leq z' \leq d\ell/2$. The current density may be expressed using Dirac delta functions and Heaviside unit step functions as

$$\vec{J}(\vec{r}', t) = \hat{z} I(t) \delta(x') \delta(y') [u(z' + d\ell/2) - u(z' - d\ell/2)].$$

Due to charge continuity $I(t) = dq/dt$, there exist equal and opposite time-varying charges $\pm q(t)$ at the ends of the differential filament, where $q(t) = (I_0 / \omega) \sin \omega t$.

The resultant charge density is given by

$$\rho(\vec{r}', t) = q(t) \delta(x') \delta(y') [\delta(z' - d\ell/2) - \delta(z' + d\ell/2)]$$



For this example, we have computed the scalar and vector potentials for various values of v ranging from c to infinity. We have conclusively shown that while the potentials are non-unique for various values of v , the resulting EM fields are always unique.

Fig. 1. Illustration of a Hertzian dipole.

III. RESULTS AND DISCUSSION

This paper shows that gauge invariance is not violated in classical electrodynamics for the continuous set of potentials encompassed in the v -gauge. Furthermore, we will show that the parameter v need not be limited to the range of c to infinity, but it can take on any value in the range $-\infty < v < \infty$ except $v = 0$.

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Gauge Invariance in Classical Electrodynamics

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Abstract: This paper deals with the classical subject of Gauge Invariance. Jackson [1] who has stated, “It seems necessary from time to time to show that the electric and magnetic fields are independent of the choice of gauge for the potentials.” We have demonstrated the veracity of this statement in classical electrodynamics with an example.

I. INTRODUCTION

Vector and scalar potentials and an arbitrary gauge relationship are used for convenience in solving boundary value problems involving electromagnetic (EM) fields. Two of the most common gauges are the Lorenz and Coulomb gauges. It is less commonly known that the Lorenz and Coulomb gauges are two special cases of a generalized gauge called the velocity gauge (or v -gauge). This generalized gauge is usually not presented to the students of electrodynamics, and it is characterized by a non-physical parameter v , that has the units of a velocity and can vary between c (the speed of light) and infinity. The use of gauges prompts several truly relevant questions. If one uses the different gauges provided by v , are the electromagnetic fields the same? Is gauge invariance a sacred principle in classical electrodynamics? Is the set of gauges that can be used to compute an electromagnetic field discrete, or is it continuous, as the generalized velocity gauge suggests? Within this generalized gauge framework, we have

$$\nabla \cdot \vec{A}_v(\vec{r}, t) + \frac{1}{v^2} \frac{\partial \phi_v(\vec{r}, t)}{\partial t} = 0. \quad (1)$$

There has been some controversy regarding gauge invariance and uniqueness. Engelhardt [2] has claimed that Maxwell's equations have non-unique solutions. It is also interesting to note that Wu and Wu [3] have shown that gauge invariance is violable in Quantum Physics by considering the decay of Higgs Boson into two photons! Given these considerations from published works, reviewing gauge theory and establishing its fundamental precepts is worthwhile.

II. EXAMPLE USING A HERTZIAN DIPOLE.

For example, consider the ideal Hertz dipole shown in Figure 1. We assume a time-harmonic variation of a current element $I(t) = I_0 \cos \omega t$ flowing in an infinitesimally thin conductor along the z -axis over a

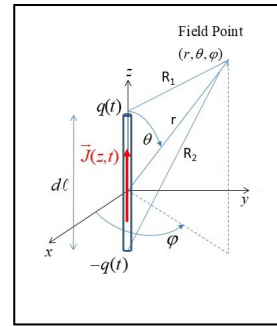
differential length $-d\ell/2 \leq z' \leq d\ell/2$. The current density may be expressed using Dirac delta functions and Heaviside unit step functions as

$$\vec{J}(\vec{r}', t) = \hat{z} I(t) \delta(x') \delta(y') [u(z' + d\ell/2) - u(z' - d\ell/2)].$$

Due to charge continuity $I(t) = dq/dt$, there exist equal and opposite time-varying charges $\pm q(t)$ at the ends of the differential filament, where $q(t) = (I_0 / \omega) \sin \omega t$.

The resultant charge density is given by

$$\rho(\vec{r}', t) = q(t) \delta(x') \delta(y') [\delta(z' - d\ell/2) - \delta(z' + d\ell/2)]$$



For this example, we have computed the scalar and vector potentials for various values of v ranging from c to infinity. We have conclusively shown that while the potentials are non-unique for various values of v , the resulting EM fields are always unique.

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III. RESULTS AND DISCUSSION

This paper shows that gauge invariance is not violated in classical electrodynamics for the continuous set of potentials encompassed in the v -gauge. Furthermore, we will show that the parameter v need not be limited to the range of c to infinity, but it can take on any value in the range $-\infty < v < \infty$ except $v = 0$.

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Overview of Lightning Current Observation at Tokyo Skytree (2012-2025)

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Abstract– Direct observation of lightning current is a fundamental approach to evaluating lightning parameters, which are among the most basic factors for the lightning protection design of various grounded objects. However, for a tall grounded structures, upward lightning flashes initiated from the top of the structure typically occur, and direct observation results of downward flashes remain limited worldwide. Tokyo Skytree is a free-standing independent broadcasting tower with a height of 634 m, located on flat terrain at 2 m ASL about 8 km inland in the Tokyo metropolitan area, Japan (35.71°N, 139.81°E). The authors have been conducting lightning current observation using Rogowski coils installed on the tower since 2012. Over a 14-year period ending in 2025, we have confirmed 45 downward-initiated and 92 upward-initiated lightning flashes. This paper provides an overview of the lightning strikes observed at Tokyo Skytree.

Keywords– downward flash, upward flash, lightning current, tall structure, lightning observation

I. INTRODUCTION

Tokyo Skytree is a 634-m-high freestanding broadcasting tower constructed on flat terrain at 2 m ASL about 8 km inland in Tokyo, Japan. Because of its height, lightning flashes were expected to strike the tower many times during and after the construction. As of the end of 2025, 137 lightning flashes striking the tower, 45 downward-initiated lightning flashes, and 92 upward-initiated lightning flashes to the tower, have been observed since the end of February 2012.

II. METHODOLOGY

Since 2012, the authors have conducted lightning current observations using Rogowski coils installed on the tower. Two types of coils are employed: the L-coil for low-frequency components (0.5 Hz to 250 kHz) and the H-coils for high-frequency components (1.5 kHz to 5 MHz)^[1]. These coils are installed at a height of 497 m and are enclosed in a hexagonal protective tube surrounding the six main tower columns, as shown in Fig. 1.

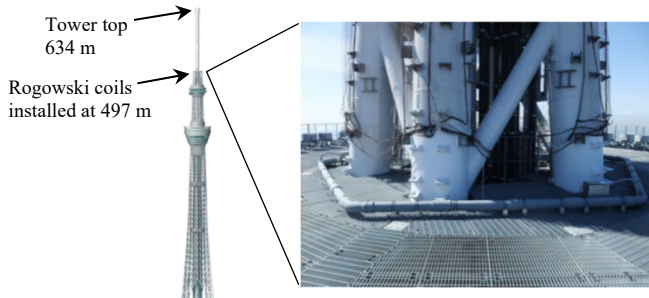


Fig. 1. Photograph of Rogowski coils (497-m level) enclosed in a hexagonal protective tube surrounding the six main tower columns.

III. RESULTS AND DISCUSSION

Table I lists the lightning strike records to date. Fig. 2 shows the distribution of the altitude of the -10°C isotherm level during lightning events. While it is difficult to identify a clear threshold due to the accumulation and variance of the 14-year data, both downward and upward flashes generally occur when the -10°C altitude is between 5000 and 6000 m. Above this range, downward flashes become dominant, whereas upward flashes are more prevalent below this range.

TABLE I. CONFIRMED FLASHES AT TOKYO SKYTREE (2012-2025)

Year	Downward /Upward		Positive /Negative /Bipolar			Return Stroke [times]
2012-2025 Total ^d	D: 45 ^{a,b} (32.8%)	U: 92 ^c (67.2%)	P: 8 (6.0%)	N: 116 (87.2%)	B: 9 ^e (6.8%)	229
Averaged annual flashes (14 years)	D: 3.21	U: 6.57	P: 0.57	N: 8.29	B: 0.64	16.4

^a The authors confirmed 4 downward flashes to lower and middle level of tower side occurred in May 2012, May 2014, August 2016 and August 2023.

^b includes subsequent return strokes branched from a downward flash which struck elsewhere.

^c 2 upward flashes were confirmed by optical and EM observations in April and May 2012 (no recorded current).

^d 3 flashes were confirmed by optical observations, so the authors couldn't obtain the polarity of these flashes (Downward flash in May 2012 and upward flashes in April and May 2012).

^e All the bipolar flashes started with positive upward leaders.

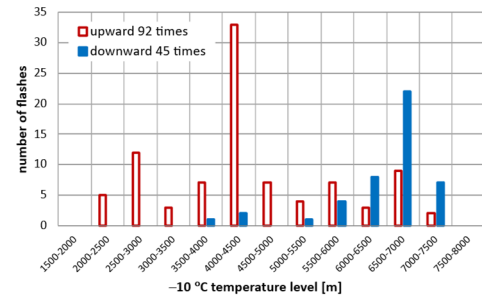


Fig. 2. Histogram of altitude of -10°C isotherm when lightning strikes Tokyo Skytree (2012-2025)

IV. CONCLUSION

This paper outlined 14 years of lightning strike data at Tokyo Skytree. As this observation project enters its 15th year in 2026, the authors intend to steadily continue these observations.

ACKNOWLEDGMENT

The authors would like to thank Tobu Tower Skytree Co., Ltd., Tobu Railway Co., Ltd., and all other contributors for their extensive support and cooperation in this observation project.

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Twelve-Year Tower-Based Validation of ALDIS Performance Using Direct Lightning Current Measurements at the Säntis Tower

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This paper presents a preliminary performance analysis of the Austrian Lightning Detection and Information System (ALDIS) [1] using direct lightning current measurements obtained at the instrumented Säntis Tower (see a description of the facility in [2]) in northeastern Switzerland over a 12-year period from June 2010 to June 2022. The study builds upon previous investigations performed for the European Cooperation for Lightning Detection (EUCLID) network for the period 2010–2013 [3] and significantly extends the database both temporally and statistically.

The Säntis Tower (124 m, located at 2502 m above sea level) is equipped with calibrated Rogowski coils and B-dot sensors enabling direct measurement of lightning currents associated predominantly with upward flashes. Such tower-based measurements provide exact strike location and ground-truth current parameters, making them particularly suitable for evaluating lightning location system (LLS) performance in terms of flash detection efficiency (DE), pulse detection efficiency, location accuracy (LA), and peak current estimation.

During the 12-year observation period, several hundred upward negative flashes were recorded, comprising thousands of individual pulses, including return strokes and Initial Continuous Current (ICC) pulses. Using synchronized GPS timing and strict spatial–temporal correlation criteria, ALDIS-reported events within a predefined radius (5 km) around the tower were matched to directly measured current waveforms. Flashes containing only initial continuous current (ICC-only events), which are inherently undetectable by LLSs, are excluded from detection efficiency analysis.

The extended dataset enables a statistically robust evaluation of pulse detection efficiency as a function of peak current amplitude. The long-term dataset also allows quantification of year-to-year variability and assessment of the impact of network upgrades and algorithmic refinements implemented during the study period.

Peak current estimates provided by ALDIS are also compared with directly measured current peaks at the tower. Location accuracy is evaluated using the absolute distance error between ALDIS-reported stroke locations and the known tower position. The median location error is analyzed as a function of current amplitude and pulse characteristics.

The 12-year dataset further allows investigation of long-term trends in location accuracy and the effectiveness of propagation and terrain-related corrections in mountainous regions.

The presented study constitutes one of the longest tower-based validation campaigns of a national lightning detection network in Europe. The results provide valuable insight into the performance of ALDIS for upward lightning in alpine terrain and contribute to improved understanding of LLS behavior for tall structures such as broadcast towers and wind turbines. The findings are relevant for lightning protection design, risk assessment, and future algorithm development for lightning location systems operating in complex topography.

Acknowledgments – This project has been partially supported by the Swiss National Science Foundation (Grants 200020-204235 and IZ11Z0_230951 (Second Swiss Contribution MAPS)).

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Simulation-Based Analysis of Conducted Disturbance Propagation Induced by Electromagnetic Pulses in Building Power Networks

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Abstract– This paper presents a simulation-based approach for evaluating conducted current disturbances within building power grids induced by the coupling of a high-altitude nuclear electromagnetic pulse (HEMP E1) onto overhead electrical distribution lines. Experimental results are used as a reference to parameterize and validate numerical models describing conducted propagation in low-voltage electrical networks. An electromagnetic simulation of an electrical cable is first developed using CST, then translated into an equivalent transmission-line model implemented in LTspice. To quantify parasitic currents at the input interfaces of sensitive equipment, the simulation approach is applied to a low-voltage network prototype and validated against experimental measurements.

Keywords– HEMP-induced disturbances, circuit simulation, building power networks, parasitic current.

I. INTRODUCTION

Intentional electromagnetic threats, such as High-Altitude Nuclear Electromagnetic Pulse (HEMP), can induce high-amplitude currents and voltages in electrical networks, potentially disrupting or damaging connected devices [1,2]. Assessing the influence of a building's electrical network is essential to predict conducted parasitic currents level at the input interfaces of sensitive equipment's. Previous experimental work has highlighted the influence of network components and the propagation mechanisms in transmission-line-type conductors [3]. Building on this, an iterative approach was implemented to verify the agreement between simulations and measurements, starting with a simple case of a single power line before extending to a more complex low-voltage (LV) electrical network.

II. SIMULATION RESULTS

A 50-m power line composed of phase, neutral, and protective earth conductors is reproduced using a three-dimensional simulation (CST) during high current injection [4]. Fig.1 compares the measured current at the load terminals with the current obtained from the simulation. Based on previous work [3], a circuit-based approach using transmission-line modeling is proposed to define a numerical model of the power line. The correlation of results, presented in Fig. 1, confirms the transmission-line behavior of the conductors and validates the interpretation derived from the experimental measurements. The circuit-based approach is then extended to a complete prototype representative of a LV residential distribution network. For the five power lines composing the prototype, the characteristic impedance and propagation delay are theoretically estimated from the conductor properties. For all simulated power lines, good agreement between measurements and simulations is observed for both the injected current and the current measured at the load.

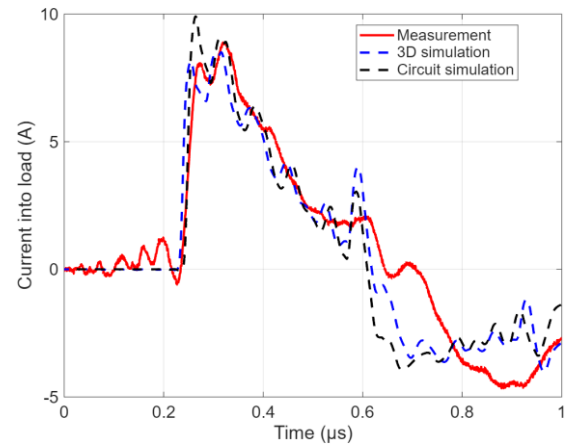


Fig. 1. Measurement /3D simulation / circuit simulation CM current comparison of 50 m electrical line.

These results are currently valid for a TT earthing system and the applicability of the transmission-line-based circuit approach to other neutral configurations, such as TN - S has also been demonstrated.

IV. CONCLUSION

The interest of this approach lies in the use of a circuit-based model, which is more easily exploited for determining the constraints applied to the terminal equipment of an installation. As an alternative to 3D simulation, this approach enables the statistical use of the proposed methodology by constructing distributions around the parameters that most strongly influence the network.

ACKNOWLEDGMENT

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Investigation of Polarization-Dependent Back-door Coupling and Shielding Effectiveness in Quadrotor UAVs under Electromagnetic Pulse Exposure

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Abstract– This paper presents a CST-based study of polarization-dependent EMP coupling in a small quadrotor UAV at 200 MHz. Simulations at 0 deg, 45 deg, and 90 deg with local shielding show that arm wiring is the dominant back-door coupling path. At 0 deg, arm copper shielding reduces the peak field from 1.21E-09 V/m to 1.19E-11 V/m, while general copper replacement has little effect. At 45 deg, coupling becomes multi-path and localized shielding loses effectiveness. At 90 deg, the overall induced field is lower, but motor-side vulnerability remains. The results provide practical guidance for UAV EMP hardening and low-power anti-drone EMP concepts.

Keywords– EMP, UAV, CST simulation, shielding effectiveness, loop antenna, back-door coupling

I. INTRODUCTION

Small UAVs are increasingly used in civil and military fields, but their lightweight structures and compact electronics make them vulnerable to strong electromagnetic environments. Earlier studies mainly addressed hard-kill effects such as component damage, whereas recent work showed that soft-kill failure can also occur when transient coupling disturbs control signals without permanent destruction. In particular, interference on motor-control paths can degrade flight stability even if communication links remain active.

This study examines angle-dependent shielding effectiveness in a quadrotor UAV under 200 MHz loop-antenna excitation. The main novelty is a compact comparison of local shielding performance at 0 deg, 45 deg, and 90 deg. The results show that the dominant coupling path and shielding effectiveness strongly depend on incident angle and local geometry, providing practical insight for UAV hardening.

II. METHODOLOGY

Simulations were performed in CST Studio Suite with open boundary conditions. The source was a 1×1 m square loop antenna excited by a double-exponential pulse with 600 V peak amplitude, 20 ns rise time, and 120 ns full width at half maximum. The quadrotor diagonal length was 270 mm. As in earlier UAV EMP studies, the arm wiring can be approximated as a dipole-like conductor with resonant frequency

$$f_{\text{resonant}} = c/2l \quad (1)$$

where l is the arm length and c is the propagation speed. For the present platform, f_{resonant} is about 556 MHz, so 200 MHz is a below-resonance condition dominated mainly by magnetic-flux linkage. The induced electromotive force follows Faraday's law,

$$\text{emf} = -d\Phi_B / dt \quad (2)$$

making the effective loop area of the arm wiring and return path a key factor. Three loop angles, 0 deg, 45 deg, and 90 deg, were analyzed. For each case, a baseline polyimide model, a general

copper model, and an arm-copper-shielded model were examined. A motor-case-only shielding configuration was also reviewed qualitatively.

III. RESULTS AND DISCUSSION

At 0 deg, the strongest coupling occurs along the drone arm near the motor hotspot. The peak field decreases from 1.21E-09 V/m in the baseline model to 1.07E-09 V/m in the whole-copper case, but drops sharply to 1.19E-11 V/m with arm-only copper shielding. This indicates that, under aligned polarization, the arm wiring is the main back-door coupling path, and motor-case-only shielding is insufficient. At 45 deg, coupling becomes more distributed. The baseline, general copper, and arm-shielded cases show peak fields of 4.07E-10, 6.82E-10, and 5.19E-10 V/m, respectively, indicating that localized arm shielding is less effective due to multi-path coupling through the arms, body, and internal wiring asymmetry. At 90 deg, the overall induction is lower than at 0 deg, but vulnerability remains. The baseline and general copper models show similar peaks of 6.62E-10 and 6.55E-10 V/m, while the arm-shielded case still reaches 9.33E-10 V/m at the motor side. Thus, lower field levels do not imply immunity, as induced energy can still concentrate locally and disturb PWM-based motor control. Overall, the CST results show that arm shielding is highly effective only at 0 deg, whereas 45 deg and 90 deg require broader loop-area suppression or dedicated protection at the motor-control interface.

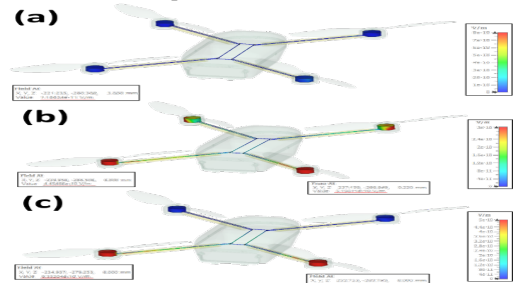


Fig. 1. Electric field distribution of drones with arm shielding: (a) 0 deg aligned (b) 45 deg aligned (c) 90 deg aligned.

IV. CONCLUSION

A compact CST analysis at 200 MHz showed that arm shielding is highly effective only at 0 deg, where the drone arm is the dominant coupling path. At 45 deg and 90 deg, distributed coupling and motor-side vulnerability remain. Thus, practical UAV EMP hardening requires system-level motor-control path protection beyond localized shielding.

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Forced Unlocking of Smart Locks via IEMI-Induced Command Injection

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Abstract— This paper highlights a hardware security threat where intentional electromagnetic interference (IEMI), even at power levels insufficient to cause physical damage, can compromise system integrity by inducing unintended control signals. As a proof of concept, we demonstrate that a smart lock equipped with secure wireless authentication can be forcibly unlocked if its internal motor control relies on simple, unencrypted wiring.

Keywords— IEMI, Hardware security, Command injection

I. INTRODUCTION

Traditionally, intentional electromagnetic interference (IEMI) has been studied in the electromagnetic compatibility (EMC) community as a high-power threat causing physical damage or severe disruption. However, even low-power IEMI can threaten system security by compromising signal integrity.

Prior research has shown that malicious voice commands can be remotely injected into smartphones via electromagnetic coupling on analog cables [1]. Similar vulnerabilities exist in Internet of Things (IoT) devices. Although external wireless communications are typically protected by strong encryption, internal circuit-level communications—such as those between microcontrollers and actuators—often remain unencrypted. If external electromagnetic (EM) waves couple into these paths, attackers may spoof legitimate signals and execute malicious commands. In this study, we investigate whether a smart lock can be forcibly unlocked, despite secure wireless authentication, by injecting continuous-wave EM signals into its unencrypted motor control lines.

II. VULNERABILITY OF SMART LOCKS TO EM COUPLING

Fig. 1 shows the block diagram of a typical smart lock system. Authentication between the smart key and the lock is securely performed via wireless communication. After successful authentication, a control signal is sent to the actuator to rotate the cylinder and unlock the door. However, because this internal motor control signal is unencrypted, spoofing it through external EM coupling through metal parts of door could enable forced unlocking.

Motor control signals use various transmission schemes, including serial protocols such as I²C, simple three-wire configurations, and two-wire schemes that independently control rotation direction and stopping. The two-wire scheme is the simplest, using separate lines for forward and reverse rotation, which makes signal emulation relatively easy for an attacker. When a continuous-wave (CW) EM signal couples into the device, it can be rectified by non-linear components, such as protection diodes at the motor driver IC input. If the rectified voltage exceeds the circuit’s logic threshold, the system may incorrectly get an unlocking command.

III. FORCED UNLOCKING EXPERIMENT ON A SMART LOCK WITH A TWO-WIRE MOTOR

Fig. 2 shows the experimental setup used to demonstrate forced unlocking via CW signal injection into the motor control circuitry. To emulate EM coupling onto the wiring, a bulk current injection (BCI) probe (FCC F-140) was used. A sinusoidal signal from a

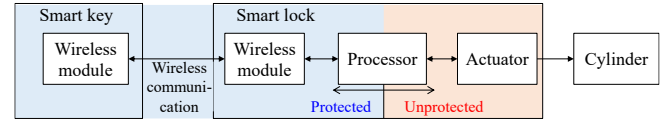


Figure 1 system block diagram of smart lock

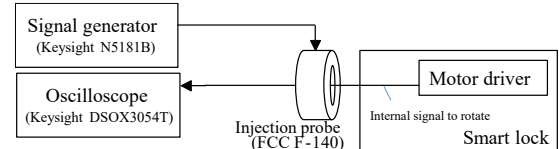


Figure 2 Block diagram of forced opening smart lock by IEMI

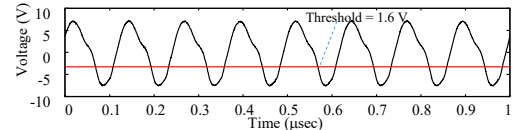


Figure 3 Time domain waveform on wire during EM irradiation

Table I Injected frequency and minimum power to open, amplitude on control signal wire.

Injected	Frequency (MHz)	Injected power (dBm)	Amplitude on wire (V)
	8	23.6	14.7
	9	24.2	16.3
	10	24.8	17.5
	11	30	18.2

signal generator was injected into the control line corresponding to the motor’s unlocking direction, which was temporarily routed outside the board for testing. Injection frequency was swept from 0.1 MHz to 1 GHz, covering the probe’s operating bandwidth. Voltage fluctuations on the control line were monitored with an oscilloscope while the physical behavior of the smart lock was observed. Experiments on a device using the two-wire scheme showed that forced unlocking occurred only within a specific frequency band from 8 MHz to 11 MHz, as summarized in Table I. Because it uses direct injection, this frequency response depend on the characteristics of rectification within the internal circuitry.

Fig. 3 shows the time-domain voltage waveform on the control line during 8 MHz injection, which required the lowest injected power to trigger unlocking. Although the induced sinusoidal waveform is distorted and exceeds the logic-high threshold (1.6 V) only briefly in each cycle, the attack still succeeded. These results indicate that even transient periodic signals can trigger forced unlocking. We hypothesize that repeated high-frequency pulses may accumulate sufficient energy to drive the motor, or that internal circuitry integrates these transient voltages, enabling the system to respond despite the short duration above the threshold.

ACKNOWLEDGMENT

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System-Level Modeling of EFT-Induced Oscillatory Responses in DCDC Converters

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Abstract– Electrical fast transient (EFT) disturbances can excite high-frequency oscillations in power converters and cause functional failures in downstream circuits. This work investigates the oscillation mechanism at the output of a Buck-type DCDC converter under EFT injection and proposes a black-box modeling workflow based on online measurement and transfer-function identification. The validated equivalent transfer-function model shows good agreement with the measured results at each oscillation point, indicating that the established model can accurately describe the input–output relationship of the power supply. The proposed method is applicable to immunity assessment and engineering-oriented optimization of power modules under conditions where the circuit or chip internal structure is unknown.

Keywords– Electrical Fast Transient (EFT), DCDC Converter, Black-Box Modeling, Transfer-Function Identification

I. INTRODUCTION

High-power electromagnetic environments such as EFT, HPM, and IEMI can induce fast transient disturbances that couple into power distribution networks and trigger high-frequency oscillations, leading to functional upset or damage in modern electronic systems^[1]. This paper investigates the oscillatory response of a Buck-type DCDC converter under EFT excitation using a combined measurement-and-modeling approach. A black-box transfer-function-based model is established to reproduce EFT-induced ringing from limited online measurements. The key contribution is a low-order rational modeling workflow validated through time-domain replay, enabling analysis of resonant behavior and evaluation of mitigation strategies such as filtering and damping optimization. The proposed method provides a practical framework for immunity assessment and engineering-oriented hardening of power modules.

II. METHODOLOGY

Due to the lack of internal structural information, the DCDC converter is treated as a black-box two-port system, as illustrated in Fig. 1. The 12-V supply and digital ground are defined as the input port, while the regulated 5-V output and digital ground are defined as the output port. During EFT injection, the input and output voltages are measured simultaneously using two voltage probes and a high-bandwidth oscilloscope. Based on the measured input-output waveforms, a low-order rational transfer function is identified to characterize the DCDC converter dynamics. The model is expressed as

$$H(s) = \frac{V_{out}(s)}{V_{in}(s)} = \frac{b_0 + b_1s + \dots + b_ms^m}{1 + a_1s + \dots + a_ns^n}, \quad (1)$$

where the coefficients a_i and b_i are identified from the measured input–output waveforms using system identification.

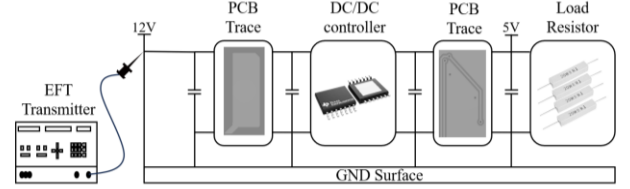


Fig. 1. EFT injection and measurement configuration for black-box modeling of the DCDC converter.

III. RESULTS AND DISCUSSION

To validate the identified model, the measured EFT-induced input waveform $v_{in}(t)$ is applied to $H(s)$, and the corresponding time-domain output response is calculated as

$$v_{out}(t) = \mathcal{L}^{-1}\{H(s)V_{in}(s)\}. \quad (2)$$

The calculated output waveform is then compared with the measured output waveform, as shown in Fig. 2. Good agreement is observed in the dominant ringing frequency, peak amplitude, and decay trend, indicating that the identified transfer-function model captures the transient input-output behavior of the DCDC converter under EFT excitation.

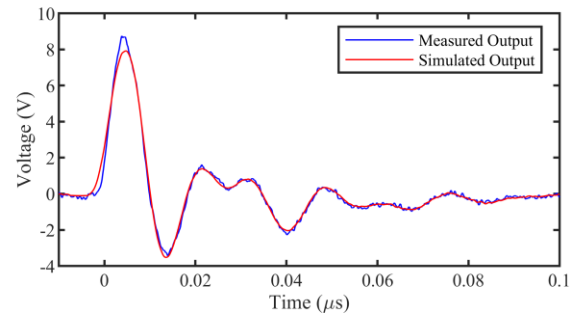


Fig. 2. Time-domain comparison of measured and simulated output voltage of the DCDC converter.

IV. CONCLUSION

A time-domain black-box transfer-function model is established to characterize the output-voltage ringing of a DCDC converter under EFT excitation. The identified model agrees well with measurements in transient peak and damping characteristics, demonstrating its ability to reproduce EFT-induced oscillations. This approach provides a practical framework for EMC analysis and mitigation design in switching power converters.

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SSPP-Based Low-Power Pass Filter for Back-End Protection of Energy Selective Antenna

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This paper presents a spoof surface plasmon polariton (SSPP)-based low-power pass filter for secondary protection of receivers at the back end of an energy selective antenna (ESA). The proposed filter operates in a fully passive device without any external DC power supply, employing a PIN diode whose switching state is determined by the incident signal power level. Under low-power conditions, the filter provides a passband response with an insertion loss of approximately -2.5 dB over the 2.0–2.6 GHz range. Under high-power excitation, strong blocking performance is achieved with more than -56.4 dB suppression and a 300 MHz bandwidth. Fabrication and measurement results confirm effective power-dependent transmission and suppression.

Keywords— Energy Selective Antenna, Spoof Surface Plasmon Polariton, Power-Dependent Filter, PIN Diode

I. INTRODUCTION

This paper investigates the system-level susceptibility of RF receiving systems under high-power electromagnetic (HPM) and electromagnetic interference (EMI) environments, focusing on remnant high-power energy at the back end of an ESA. ESA is a promising front-end protection technology that integrates nonlinear elements such as PIN diodes into the antenna structure, allowing normal reception under low-power conditions while suppressing high-power incident signals [1]. However, residual high-power energy may still reach the back-end circuitry. To address this issue, this paper proposes an SSPP-based low-power pass filter as a passive secondary protection device for ESA systems. SSPP employ periodically corrugated metallic surfaces to mimic optical surface plasmon polaritons at microwave frequencies. Their geometry-dependent dispersion characteristics enable compact transmission structures with enhanced filtering capability, making them suited for passive and bias-free back-end protection filters [2]–[4].

II. METHODOLOGY

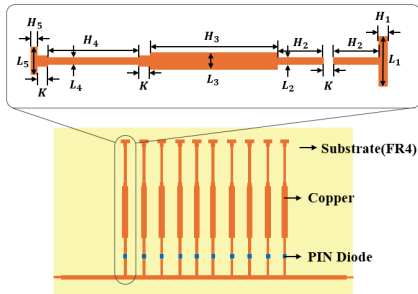


Fig. 1. Proposed SSPP unit cell and overall structure

TABLE I. SSPP UNIT CELL GEOMETRY PARAMETERS

Unit Cell Parameters (mm)									
K	L ₁	L ₂	L ₃	L ₄ /H ₅	L ₅	H ₁	H ₂	H ₃	H ₄
1	8	0.65	1.5	0.5	2.5	1.95	8.3	16.2	15.2

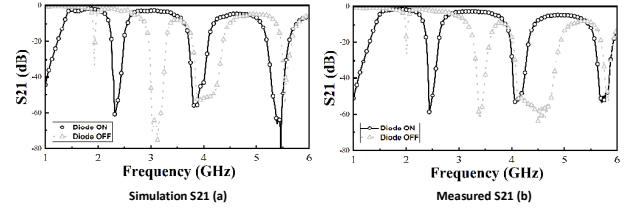


Fig. 2. characteristics of the proposed SSPP-based filter
The proposed filter employs SSPP transmission line with periodically corrugated unit cells. The structure and geometry are shown in Fig. 1 and Table I. A PIN diode enables passive power-dependent operation, providing low-loss transmission in the OFF state and strong attenuation in the ON state, as illustrated in Fig. 2.

III. RESULTS AND DISCUSSION

Figure 2(a) shows a passband from 2.0 GHz to 2.6 GHz with approximately -2.5 dB insertion loss in the diode OFF state. When the diode is ON, 2.45GHz as -56.4 dB suppression with a 300 MHz bandwidth is achieved. The measured results in Fig. 2(b) is response exhibits an approximate 0.15 GHz frequency shift compared with the simulation, which can be attributed to fabrication tolerances and diode modeling inaccuracies. Nevertheless, the overall transmission behavior shows good agreement.

IV. CONCLUSION

A passive SSPP-based low-power pass filter for ESA back-end protection has been presented. The filter achieves low insertion loss under normal conditions and strong power-dependent suppression without external bias

ACKNOWLEDGMENT

This work was supported in part by the Priority Research Center Program through the National Research Foundation under Grant RS-2019-NR040074, and in part by Korea Basic Science Institute (National research Facilities and Equipment Center) grant funded by the Ministry of Education (RS-2022-NF000835), and in part Institute of Information & communications Technology Planning & Evaluation (IITP) grant funded by the Korea government (MSIT) (No. RS-2025-02642985, Development of Identification, Protection, and Detection Technologies for High-Power Electromagnetic Pulse Protection)

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Modelling of E1 HEMP Coupling to a Cable on the Ground and in a Trench

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Abstract– This paper provides a summary of some modeling results of some E1 HEMP coupling to shielded control cables in a typical high voltage electrical substation yard. Several different scenarios were modelled to understand if different treatments could be used to reduce E1 HEMP coupling to the cable shield.

Keywords– HEMP, EM Coupling, Computational EM Modelling

I. INTRODUCTION

Since 2018 QinetiQ and Statnett have been working together on a research and development program to identify new approaches to provide High Power Electromagnetic (HPEM) resilience. In August 2019 QinetiQ were employed to assess the EMP protection performance of a new built High Voltage (HV) substation [1]. Our work identified some areas of incremental improvement to the protection scheme design. One area, which will be discussed in this paper is evaluation of the E1 HEMP, as defined in [2], coupling to shielded cables which are typically 100m to 150m long and are placed within a concrete trench or culvert. The Transmission Line Matrix (TLM) method, based on the CST/Simula suite of tools, was used for the analysis and ten basic configurations and three variants, 12 total, were modelled with the objective of:

- Estimating the E1 HEMP coupled voltage (peak) and current (peak) on a shielded cable – for each of the given configurations
- Comparing the peak voltage and peak current for the given configurations
- Identifying the configuration that has most merit for reducing coupling
- Suggesting other configurations or modifications to the given configurations that may have increased benefit.

II. METHODOLOGY

A Figure showing the baseline configuration (Configuration 1) is shown in Figure 1.

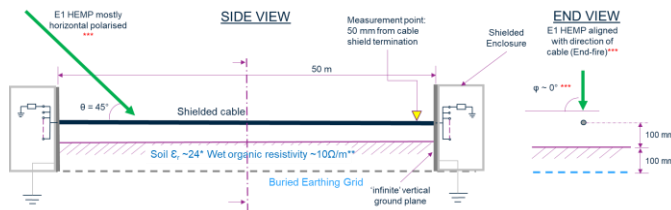


Fig. 1. Baseline configuration

The nine basic and three variant physical configurations that were evaluated are summarized in Table I.

TABLE I. CONFIGURATIONS EVALUATED

Config. No.	Description
1	Shielded cable suspended 100 mm above the ground.
2	Shielded cable lying on the ground.
3	Shielded cable within an open plain concrete trench.
4	Shielded cable within a covered plain concrete trench.
5	Shielded cable within a covered plain concrete trench with a partial metal ground plane (1 m).
6	Shielded cable within a covered plain concrete trench with a partial metal ground plane (0.6 m).
7	Shielded cable within an open plain concrete trench with a 1 m block of basalt aggregate lossy filler.
8i	Shielded cable within an open plain concrete trench wholly filled with basalt aggregate.
8ii	Shielded cable within an open plain concrete trench wholly filled with granite aggregate.
8iii	Shielded cable within an open plain concrete trench wholly filled with asphalt aggregate.
9	Shielded cable within an open Steel Fibre Reinforced Concrete (SFRC) trench.
12	Shielded cable within a plain concrete trench but with an SFRC lid

III. RESULTS AND DISCUSSION

The results suggested that Config. 1 and 2 show high peak currents and multiple resonances (three main length-related resonances), shown in the frequency domain data. Filling the trench with various lossy materials, Configs 6, 8i, 8ii and 8iii reduces the higher frequency resonances (leaving one main resonance) and slightly reduces the peak current in this fundamental resonance. This produces a more pronounced damping effect. Config. 9, which uses SFRC [3] as the trench construction material, provided the lowest induced peak current on the shield of the shielded cable. Config. 12 also provides promising results.

ACKNOWLEDGMENT

This work was supported by Statnett SF, Trondheim, Norway

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Design Improvement of Connector Shield Plates for Enhanced Shielding Performance

Experimental Analysis Using Transfer Impedance and Near-Field EMI measurement

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Abstract– Weak shield continuity at the connector interface significantly degrades the overall shielding performance of the cable system. This paper proposes a method to improve shielding performance through an optimized shield-plate structure. The improvement is verified experimentally through transfer impedance (Z_t) measurements and near-field measurements on a production vehicle. In addition, Z_t modeling is used to identify the factors contributing to the improvement, indicating the importance of connector-interface design.

Keywords– connector shield plate, transfer impedance, shield effectiveness (SE), EMI noise reduction

I. INTRODUCTION

Non-uniform shielding interfaces between the cable and connector can weaken shield continuity and significantly degrade the overall shielding performance of the harness. Previous studies [1] have shown that, even when the cable provides high-performance shielded braid, gaps and increased contact resistance at the cable–connector interface can still degrade the overall shield effectiveness of the cable system. In this paper, we present a design approach to improve connector shielding performance. We verify the shield effectiveness of the proposed design experimentally, by reducing contact resistance and gaps at the connector interface to maintain shield continuity.

II. DESIGN MODIFICATION AND TEST

We perform structural design modifications on the connector shield plate used in commercial products, as shown in Fig. 1. We expand the contact area between the connector and the device housing to reduce contact resistance and gaps at the shield interface.

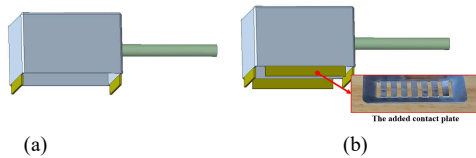


Fig. 1. Design modifications: (a) commercial products; (b) design modification

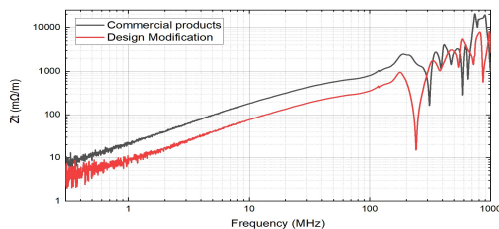


Fig. 2. Measurement of Transfer Impedance(Z_t)

We compare the transfer impedance (Z_t) before and after the improvement in accordance with IEC 62153-4-15 [2]. As shown in Fig. 2, the shielding performance improves by 106 mΩ/m at 10 MHz and by 446 mΩ/m at 100 MHz following the shield-plate design modification.

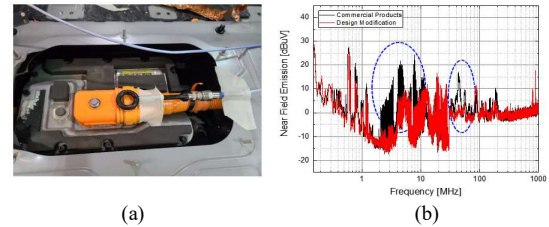


Fig. 3. Near-Field antenna test: (a) test setup; (b) test results

As shown in Fig. 3, the improved shield plate is applied to the vehicle, and the noise characteristics are measured using a near-field antenna after operating the connected components. The results indicate an improvement in shielding performance.

III. RESULTS AND DISCUSSION

We analyze the improvement mechanism based on the Vance model [3], which is widely used to analyze the transfer impedance of shielded cables.

$$Z_T \approx R_0 \frac{y d}{\sinh y d} + j \omega M_{12} \quad (1)$$

The design improvement applied in this study increases the contact area between the connector and the device housing, thereby reducing the contact resistance R_0 , and minimizes gaps at the shield interface, which in turn reduces the leakage component M_{12} .

IV. CONCLUSION

Measured results show that increasing the contact area improves the shielding performance. By reducing contact resistance and minimizing gaps, the proposed design achieves enhanced EMI characteristics. These findings demonstrate that connector interface design is equally important to PCB design improvement for EMI noise reduction.

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Practical Design of an EMC Filter for Power Line Noise Reduction

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Abstract— One of the key challenges in designing power line EMC filters is the uncertainty of source and load impedances, which can lead to significant deviations from the expected filter performance in actual installations. By accurately characterizing the impedance of the noise source and incorporating it into the design process, a more realistic and application-oriented power line filter can be developed. The filter component values are determined based on the noise source impedance to achieve the required attenuation for compliance with electromagnetic compatibility (EMC) standards, and the filter is designed accordingly. The performance of the proposed practical EMI filter design is verified through actual EMC compliance testing. Furthermore, this method contributes to reductions in both design time and overall development costs.

Keywords— power line EMI filter, noise source impedance, EMC standard, EMC compliance test.

I. INTRODUCTION

Switching mode power supplies are widely used in various applications due to their high efficiency and power density. However, the high dv/dt and di/dt characteristics of these systems generate significant electromagnetic interference (EMI).

Although EMI filters are commonly used to reduce such noise, conventional EMC filter design methods rely on numerous assumptions and trial-and-error procedures[1]. As a result, they require substantial design time and often lead to uncertainty in component selection, increasing both the size and cost of the filter.

This paper proposes a practical and systematic filter design method based on the measurement of noise source impedance. By utilizing the measured noise source impedance, appropriate filter component values can be accurately determined to meet compliance requirements.

II. DESIGN OF EMC FILTER

The overall filter design procedure is illustrated in the flowchart in Fig. 1. In Fig. 2, The noise voltages measured by the LISN on the power lines L1 and L2 are expressed by (1) and (2), respectively, and the V_{L1} is presented in Fig. 3.

$$\begin{aligned} V_{L1} &= V_{CM} + V_{DM} & (1) \\ V_{L2} &= V_{CM} - V_{DM} & (2) \end{aligned}$$

To satisfy the compliance requirements specified in the relevant standard, the noise measured after filtering must remain below the limit values shown in Fig. 3.

The common-mode and differential-mode noise voltages, V_{CM} and V_{DM} , are separated using a CM/DM noise separator based on the hybrid circuit shown in Fig. 2. The separated results are also presented in Fig. 4 (a) and (b).

Based on Fig. 4 (a) and (b), the results of the common-mode (CM) and differential-mode (DM) noise analysis are summarized as follows:

Noise	Max. Level	Compliance Limit	Target Level
CM	95 dBμV at 317 kHz	60 dBμV	55 dBμV
DM	92 dBμV at 156 kHz	65 dBμV	60 dBμV

The target level was established with a 5 dB margin, indicating that the CM and DM noise components must be attenuated by approximately 40 dB and 32 dB, respectively

The CM and DM noise source impedances were measured over the frequency range of 10 kHz to 30 MHz using the noise analyzer developed by our company based on the insertion loss method [2]. The measured source impedances were 1.41 kΩ at 317 kHz for CM and 6.28 Ω at 156 kHz for DM. These frequencies are those with the highest noise level in the noise spectrum of each mode.

The filter component values were determined so that the impedance ratio between the CM/DM noise source and the filter corresponds to the required attenuation level. Through an optimization process, a power line filter with a low-pass filter characteristics was designed, as shown in Fig. 5. Fig. 6 shows the effectiveness of the designed filter.

III. RESULTS AND DISCUSSION

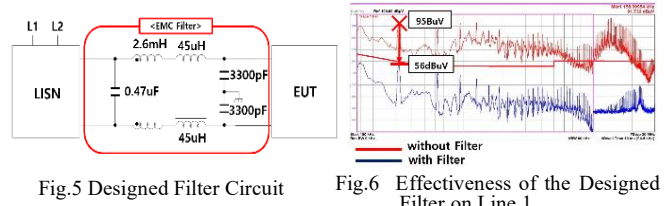
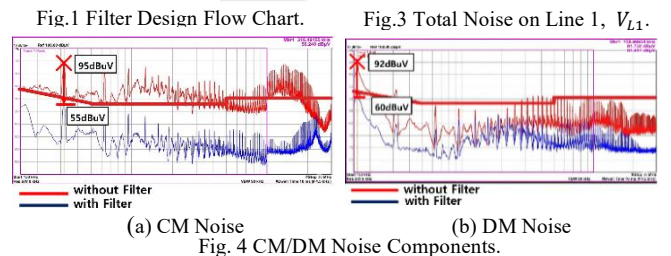
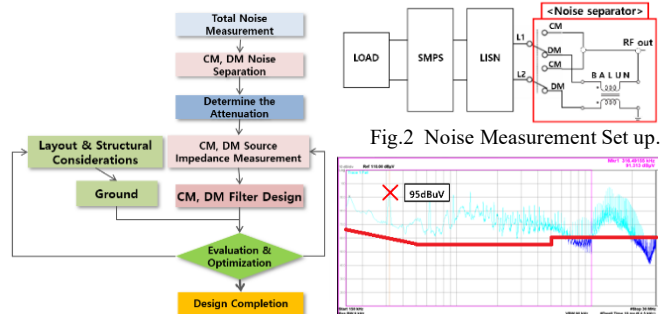
From Fig. 4, it can be seen that the difference between the measured value and the target level for CM/DM noise is less than 2 dB. This deviation is considered to result from the non-ideal characteristics of the inductors and the capacitors.

As shown in Fig. 6, the designed filter satisfies the specified compliance requirements. This result indicates that the method proposed in this paper can be effectively applied to filter design.

This study demonstrates that the proposed noise filter design method based on noise source impedance is more practical and systematic than conventional approaches. Furthermore, since the method is relatively simple, it can reduce design time and cost.

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De-embedding-Based CISPR-25 Coupling Attenuation VNA method Considering DUT Operating Conditions and Cable Environment

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Abstract– This paper proposes a de-embedding-based CISPR-25 coupling attenuation (CA) measurement method using a vector network analyzer (VNA) considering DUT operating conditions and HV-LV cable environments. Artificial networks (ANs) are used as standardized EMC interfaces, enabling CA measurement under DUT operating conditions while reducing uncertainty caused by adapters. To extract the intrinsic CA characteristics of the DUT, the effects of ANs and HV-LV cables are removed using an ABCD-parameter-based de-embedding technique. Measurement results show good agreement with the CISPR-25 VNA method, confirming the applicability of the proposed approach for CA evaluation.

Keywords– coupling attenuation, de-embedding, VNA

I. INTRODUCTION

With increasing vehicle electrification, electromagnetic noise coupling between HV and LV systems has become a critical EMC issue [1], requiring quantitative evaluation of CA. The CISPR-25 VNA CA method [2] provides high measurement efficiency but cannot represent DUT operating conditions and suffers from measurement uncertainty caused by unstandardized adapters. This paper proposes a VNA-based CA measurement method considering DUT operating conditions and HV-LV cable environments. By applying a de-embedding technique, the intrinsic CA characteristics of the DUT can be extracted while maintaining physical equivalence to the CA definition in CISPR-25.

II. METHODOLOGY

CA measurements were performed using the conventional CISPR-25 VNA setup and the proposed AN-based setup including HV-LV cables, as shown in Fig. 1. The CA is calculated as (1)

$$CA_{VNA}(dB) = -|S_{21}|(dB) \quad (1)$$

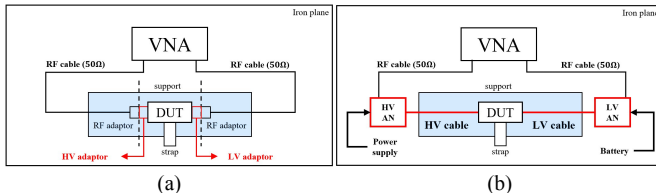


Fig. 1. Measurement setups. (a) CISPR-25 VNA method. (b) Proposed VNA method.

To eliminate measurement system effects, the S-parameters of the HV-LV cables were measured individually (Fig. 2). The S-parameters were converted into ABCD parameters, referred to as the T-matrix, following the procedure described in (2).

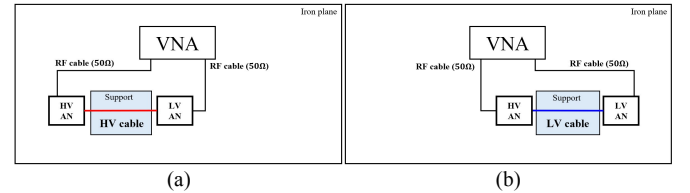


Fig. 2. Measurement setups. (a) HV cable. (b) LV cable.

$$T_{DUT} = T_{HV\ cable}^{-1} \cdot T_{HV\ AN}^{-1} \cdot T_{system} \cdot T_{LV\ AN}^{-1} \cdot T_{LV\ cable}^{-1} \quad (2)$$

As shown in Fig. 3, the results of the proposed de-embedded VNA method closely match those of the standard CISPR-25 VNA method, demonstrating that CA measurement under DUT operating conditions can be mapped to CISPR-25-equivalent CA.

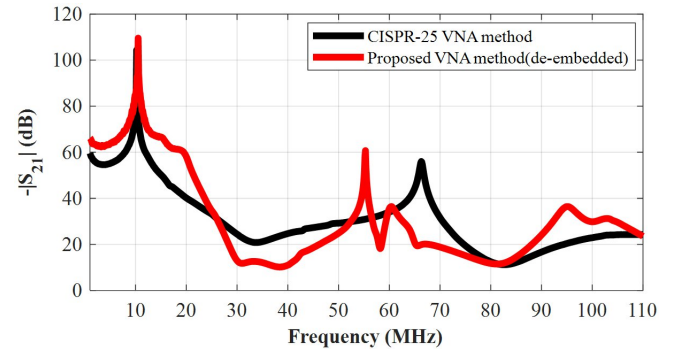


Fig. 3. Comparison of CA measurement results between the proposed de-embedded VNA method and the CISPR-25 VNA method.

III. RESULTS AND DISCUSSION

This paper proposes a VNA-based CA measurement method that represents DUT operating conditions as well as HV and LV cable and AN configurations, and verifies that the CA characteristics derived through the de-embedding process are physically equivalent to the CA quantities defined in the CISPR-25 VNA method.

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Deep Neural Network–Based Analysis of Electric Field Strength in a Cavity

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Abstract – This study aims to rapidly predict the electric field strength inside a rectangular cavity excited by a plane wave incident through an aperture using a deep neural network (DNN). The proposed model achieves reliable prediction performance while significantly reducing the overall prediction time.

Keywords – aperture coupling, cavity, deep neural network (DNN), electric field strength, electromagnetic interference (EMI)

I. INTRODUCTION

As electronic systems become more densely integrated and operate in complex electromagnetic environments, predicting the electric field strength inside devices under external electromagnetic interference (EMI) is becoming increasingly important. Although conventional full-wave simulation tools can provide accurate results, they typically require substantial computation time and resources, making them less suitable for iterative design, rapid validation, and real-time applications. To overcome this limitation, this study employs a deep neural network (DNN)-based model to learn the relationship between operating conditions and internal electric-field distribution. The trained model enables much faster prediction than full-wave simulation, making it a practical tool for rapid analysis.

II. METHODOLOGY

This study employs a simulation-driven learning framework in which training data are generated using CST Microwave Studio (CST MWS) and then used to train a deep neural network (DNN) for rapid prediction of the electric field strength inside a cavity excited by an external plane wave incident through an aperture. To reflect the complexity of realistic internal environments, a dielectric object is additionally introduced inside the enclosure. The input features include frequency, aperture dimensions, cavity dimensions, dielectric volume, relative permittivity, polarization, and angle of incidence. Time-domain simulations are carried out in CST MWS while systematically varying the enclosure and aperture geometries to construct the baseline dataset. Additional samples are generated by modifying the size and position of the dielectric object inside the enclosure. For each configuration, electric field values are sampled using field probes placed in the region of interest, and the spatially averaged field is used as the target value for training. The resulting simulation data are arranged in a form suitable for supervised learning. In total, 64,800 simulation cases are prepared for DNN training, excluding the frequency sweep points within each case. The proposed DNN consists of six fully connected layers designed to model nonlinear electromagnetic behavior. ReLU activation functions are used in the hidden layers to learn nonlinear relationships. The Adam optimizer is used for training, and a regression output layer is employed to predict the electric field strength.

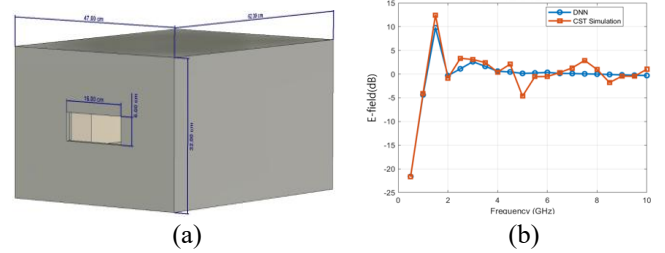


Fig. 1. (a) Cavity Geometry and (b) Prediction Results (volume of 0.081 m³, polarization of 90°, and an incident angle of 0°)

III. RESULTS AND DISCUSSION

For a representative case with an enclosure volume of 0.081 m³, a polarization angle of 90°, and an incident angle of 0°, the DNN prediction shows frequency-dependent behavior that agrees well with the CST MWS simulation results. The predicted and simulated responses follow a similar overall trend across the frequency band, suggesting that the proposed model effectively captures the dominant influence of the input parameters on the internal electric-field response. In the low-frequency region, the resonance peak exhibits an amplitude error of about 3 dB compared with the CST result. A quantitative summary of the prediction performance is provided in Table I. Notably, this level of accuracy is obtained with a significant reduction in computation time, as the evaluation time decreases from more than 10 minutes for a single CST simulation to about 10 seconds with the trained DNN model.

TABLE I. EVALUATION METRICS	
	VALUE
COEFFICIENT OF DETERMINATION	0.96
MSE	3.59
MAE	1.28

IV. CONCLUSION

The DNN-based model enabled rapid and reliable prediction of the internal electric-field strength in a cavity excited by an external plane wave incident through an aperture. These findings indicate that the proposed approach can be applied to a wide range of practical aperture-coupling problems.

ACKNOWLEDGMENT

This work was supported by Korea Research Institute for defense Technology planning and advancement(KRIT) - Grant funded by Defense Acquisition Program Administration (DAPA) (KRIT-CT-23-005).

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Shielding Effectiveness Measurements for Three Types of Objects Based on Electro-optic Sensors

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Abstract– We present a measurement method of a shielding effectiveness (SE) using electro-optic sensors and a vector network analyzer. The SE is obtained by taking the difference between the logarithmic values of the ratios of transmitted power to the input power for conditions with and without objects. We experimentally performed SE measurements for three types of objects.

Keywords– Shielding Effectiveness, Electro-Optic Sensors, Vector Network Analyzer

I. INTRODUCTION

As electronic devices are now widely used in our society, the intentional electromagnetic interference can become a threat due to high-power electromagnetic sources. Considering electromagnetic compatibility issues, the shielding effectiveness (SE) is an essential factor to investigate the immunity and interference of the electronic devices. Electro-optic (EO) sensors are useful in electromagnetic environment because most of parts of these sensors consist of dielectric materials. Also, a size of sensor is small (approximately 10 mm), it can be installed in a small space.

In this paper, we used our technique to measure the SE of three types of objects. First and second objects are a personal computer (PC) case [1] and a small unmanned aerial vehicle (UAV), respectively. Due to the small size of sensors, the sensors can be placed inside objects and also can overcome the interference to the electromagnetic field. Third object is a large structure such as a building. The SE measurement of objects inside the building is very difficult because of complicated conditions of incident wave by the building. Using high sensitivity of the sensors and a vector network analyzer (VNA), we measured the SE of the building.

II. EXPERIMENTAL METHOD

The SE is usually determined by insertion loss measurements, as described in IEEE Standard 299 [2] and is defined by:

$$SE(\text{dB}) = 10 \log \left(\frac{P_{\text{ref}}}{P_{\text{object}}} \right), \quad (1)$$

where P_{ref} and P_{object} are the powers without and with a object, respectively. For SE measurement in a wide frequency range, a combination of a spectrum analyzer and a tracking signal source can be used for time-efficient measurement. Here, we introduce VNA that is inherently equipped with signal source tracking to a built-in receive module. The SE can be obtained by measuring the scattering parameter S , the transmission coefficient of two-port network. The difference of transmission coefficient S_{21} 's in dB between with and without the object is defined as the SE of the object:

$$SE(\text{dB}) = S_{21,\text{ref}}(\text{dB}) - S_{21,\text{object}}(\text{dB}), \quad (2)$$

where $S_{21,\text{ref}}$ is the transmission coefficient without a object, and $S_{21,\text{object}}$ is the one with the object.

III. EXPERIMENTAL RESULTS

Unlike conventional SE measurements based on commercial electrical-field sensors, we used a self-developed EO receiver with three EO sensors. The dimensions for a single axis sensor are 6 mm x 7 mm x 30 mm. Fig.1(a) shows linearity of the three-axis EO sensors measured using TEM cell. The signal emerges from the approximately 2 mV/m level and increases up to 7 V/m at 1 GHz. Fig. 1(b)-(d) present the SE values for the PC case, the small UAV, and the building, respectively. In the case of the PC case, transmitted wave inside the case has totally different polarization due to the multiple reflection in such as case. Using our technique, we can measure the SE values for three axes. Also, the SE values of the building at several tens of meters can be measured by using EO sensors with high sensitivity.

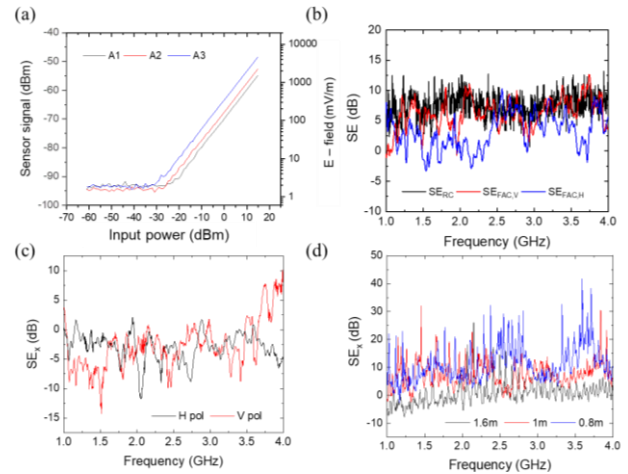


Fig. 1. (a) Measured sensor signals (left axis) and absolute electric field strengths (right axis) versus power at 1 GHz for the three sensors. (b) SE values of the PC case considering three axes simultaneously in the case of fully anechoic chamber (FAC) and reverberation chamber (RC) systems. (c) SE_x values at incident angle of 0° for horizontal and vertical polarization in the FAC. (d) SE_x values at three different heights of sensors inside the building.

IV. CONCLUSION

In this paper, we present the SE measurements of the PC case, the UAV, and the building using EO sensors. In our technique, SE values ranging from 1 GHz to 4 GHz for a single measurement are obtained by utilizing a VNA and a 10-W amplifier based on EO sensors. We hope that our technique will contribute to efficient and reliable SE measurements for various objects.

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Plasma Behavior In High-Power Electron Beam Diodes Under Varying Electrode Surface Structures

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Plasma behavior in high-power electron beam diodes was investigated using a high-speed intensified sCMOS camera by comparing planar and groove-type cathodes. Groove edges, with a 2.4 times greater field enhancement factor, yield a 2.8-fold lower EEE threshold electric field and promote uniform plasma distribution, resulting in slower plasma expansion velocity of 4.27 cm/ μ s compared to 6.45 cm/ μ s for the planar type.

Keywords– High power microwave, High-power electron beam, Explosive electron emission, Cathode flare

I. INTRODUCTION

Explosive electron emission (EEE), in which high-energy electron beams are emitted from dense plasma on cold cathodes under intense electric fields, is essential in high-power microwave (HPM) technologies. Previous studies mainly relied on voltage-current waveforms, as fast imaging lacked nanosecond-scale temporal resolution [1-3]. Using an ultrafast intensified sCMOS camera, we captured plasma images and compared them with voltage-current waveforms for detailed analysis of plasma formation, uniformity, expansion, and gap closure.

II. METHODOLOGY

A 16-stage PFN-Marx generator (25 kV/stage) powered a diode with a 7.7 mm gap and 70 mm diameter. Voltage and current were measured using a capacitive voltage divider and a Rogowski groove, respectively. Plasma images were captured with an Andor iStar sCMOS 18U-73 camera (minimum gating time < 2 ns). Two copper cathode types were tested: flat and groove-type (3 mm depth, width, and interval), with the groove edge exhibiting a field enhancement factor approximately 2.4 times that of the flat surface.

III. RESULTS AND DISCUSSION

The threshold electric fields for EEE initiation were 217 kV/cm and 78 kV/cm for the flat-type and groove-type cathodes, respectively, with breakdown voltages of 340 kV and 248 kV, as shown in Fig. 1. The higher field enhancement factor lowers the threshold, thereby increasing the diode impedance drop rate. The average terrain roughness index (TRI) [4] values were 1618.2 and 981.6, respectively, indicating more uniform plasma distribution in the groove-type, where regularly spaced field enhancing structures promote uniform plasma formation unlike the random formation in the flat-type.

A non-uniform plasma distribution inherently exacerbates the non-uniformity itself according to the Child-Langmuir law. As shown in Fig. 2, the electron density of the flat-type increases more rapidly than that of the groove-type, and the effective gap distance, the distance between the cathode plasma front and the anode, decreases correspondingly faster. The resulting plasma expansion velocities were 6.45 cm/ μ s and 4.27 cm/ μ s for the flat-type and groove-type, respectively.

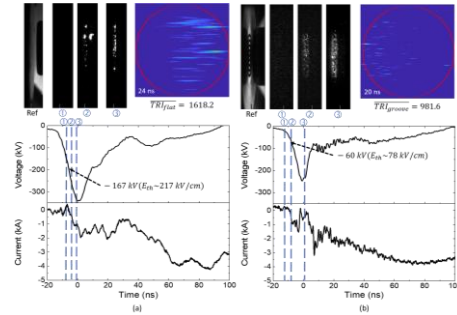


Fig. 1. Current and voltage curve of flat-type(a) and groove-type(b) with plasma images

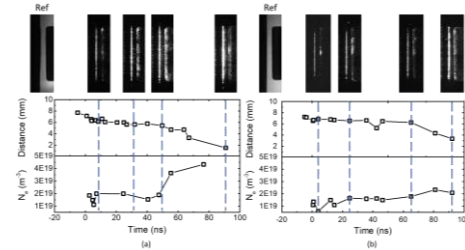


Fig. 2. Effective gap distance and electron density curve of flat-type(a) and groove-type(b) with plasma images

IV. CONCLUSION

In this study, plasma behavior in high-power electron beam diodes was investigated by comparing flat-type and groove-type cathodes. The groove edges exhibited a 2.8 times lower threshold electric field for EEE and a more uniform plasma distribution, resulting in a slower plasma expansion velocity of 4.27 cm/ μ s compared to 6.45 cm/ μ s for the flat-type. These results suggest that arranging high-field-enhancement structures at regular intervals on the cathode surface can serve as an effective strategy to enhance EEE cathode performance.

ACKNOWLEDGMENT

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Thermal Cooling Paint Effect on a Vacuumized HPM Source Chamber

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Abstract– This study investigates the effect of applying a high-emissivity Thermal Cooling Paint (TCP) as a passive method to reduce heat absorption and improve vacuum retention in cylindrical metal chambers used in high-power microwaves (HPM) sources. The goal is to stabilize the pressure level performance inside the chamber when exposed to outdoors conditions. The approach combines thermal analysis and a controlled experimental evaluation of a coated chamber surface.

Keywords– high-emissivity paint, high-power microwaves, thermal management

I. INTRODUCTION

High-Power Microwave (HPM) systems rely on vacuum chambers to ensure stable, controlled operational conditions and optimized beam propagation [1]. In outdoor environments, external heat loads can degrade vacuum stability by increasing chamber temperature and internal pressure [2]. Maintaining the vacuum over time is critical for performance, yet external heat from ambient outdoor conditions can compromise the vacuum seal and stability. This paper investigates the use of Thermal Cooling Paint (TCP) to reduce external heat transfer, thereby helping the chamber to vacuumize faster, and hold vacuum longer to enhance overall HPM system performance. We assess the impact of the TCP on thermal stability, vacuum retention, which consequently will have influence on electromagnetic performance.

II. METHODOLOGY & PRELIMINARY RESULTS

An increase in the temperature of a vacuum chamber can lead to a significant increase in pressure inside it. Gas desorption from the surface, diffusion of gases from the volume, partial pressure of the polymer's residuals and stabilizers are part of the phenomena that can be observed when the environmental temperature rises around a vacuumized enclosure. The increase in pressure, in turn, leads to a reduction in the RF output power extracted from the diode region, as well as, pulse shortening, frequency instability, among other undesired effects [1]. TCP applied to a metal surface of a vacuum chamber exposed to direct sunlight primarily modifies the surface's radiative heat balance, due to TCP's high solar reflectance and high thermal emissivity [3]. The dependence of temperature of a stainless steel, approximately 5 mm thick, vacuum chamber on the ambient air temperature was determined using a thermal balance model that includes absorption of incoming solar radiation, as well as heat losses due to thermal radiation and convection under light wind conditions typical for outdoor experimental testing.

Let us consider an evaluation of the TCP's influence on the operating regime of a vacuum chamber using, as an example, the diffusion through the polymer. The well-known Arrhenius equation (1) describes the temperature dependence of the diffusion permeability:

$$P(T) = P_0 e^{\frac{E_a}{RT}}, \quad (1)$$

where P_0 is the pre-exponential Arrhenius factor, E_a is the activation energy, and R is the universal gas constant. Gas diffusion flow Q through output window of vacuum chamber can be estimated as [4]:

$$Q = \frac{P \cdot A \cdot \Delta p}{h}, \quad (2)$$

where A is the output window's area, h is the output window's thickness, and Δp is the pressure difference at both end of the window. Fig. 1 shows expected gas diffusion flow Q through a Polymethyl methacrylate (PMMA) output window for two types of vacuum chamber: with & without TCP extra coating.

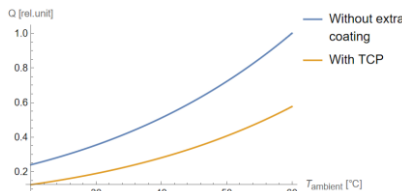


Fig. 1. (Theoretical) Expected gas permeability of PMMA output window.

As demonstrated in Fig. 1 above, the gas diffusion flow Q is significantly higher (roughly double) without the coating. This implies that the coating would significantly hold pressure inside the chamber and consequently hold the vacuum for a longer time; hence, optimizing the performance of the HPEM system.

To experimentally validate the pressure change of the vacuum chamber, a vacuum-based HPM system chamber was vacuumized using a turbopump outdoors under ambient conditions exposed to peak sunlight hours. Pressure, ambient and surface temperatures, and exposure time were recorded using a pressure sensor with a built-in

time log, an infrared-based thermal sensor, and a thermocouple. Several tests were conducted for the uncoated HPM vacuum chamber. Tests for the TCP-coated configurations are planned to be completed before the conference. The primary evaluation parameter was the pressure change rate, enabling quantitative assessment of heat transfer reduction and system vacuum stability improvement due to the coating.

While in lab, HPM testing follows usual procedures, however conducting outdoors tests presents challenges. For example, it is challenging to achieve minimum threshold (~ 15 mPa) vacuum for HPM operation at relatively higher ambient temperatures.

Experiments to record pressure change with time while outdoors were conducted at two different ambient temperatures, values were plotted below in Fig. 2 and Fig. 3

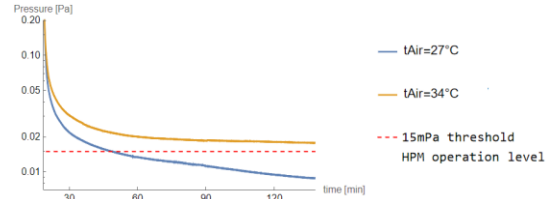


Fig. 2. Plot of Pressure vs Time of outdoors HPM (without coating) vacuumizing at different ambient temperatures.

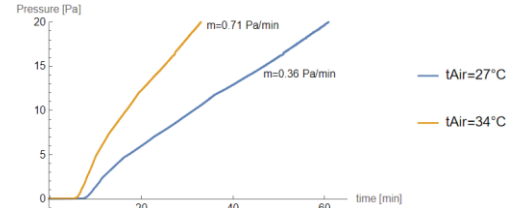


Fig. 3. Plot of Pressure vs Time of natural release of HPM (without coating) vacuum (turbopump is switched off) outdoors at different ambient temperatures.

The recorded surface temperature of the HPM chamber was 33 degrees Celsius and 43 degrees Celsius at ambient temperatures of 27 degrees Celsius and 34 degrees Celsius respectively. It was recorded using an infrared-based thermal sensor, and a thermocouple.

The plots in Fig.2 indicate that at a lower ambient temperature (and consequently a lower surface temperature), it is possible to reach the threshold HPM vacuum operation level of 15 mPa. The plots in Fig.3 indicate that once switching off the turbopump, at a lower ambient temperature (and consequently a lower surface temperature), the pressure increase rate is reduced from 0.71 Pa/min to 0.36 Pa/min at ambient temperatures of 34 degrees Celsius and 27 degrees Celsius respectively. It is anticipated that the TCP will drop the surface temperature by about 7 – 10 degrees Celsius according to previous experimentation; therefore, the performance (and plots) at 27 degrees Celsius ambient temperature almost mimics the expected performance (and plots) of TCP-coated HPM at relatively higher ambient temperatures. Experimental results with the TCP will give the final confirmation of this assumption.

IV. CONCLUSION

The TCP can reduce the surface temperature of a vacuumized HPM chamber when operating in outdoor conditions. By enhancing the thermal stability, it is anticipated that the TCP would maintain the threshold vacuum level more efficiently and consequently optimize the performance and output power of the HPEM system. Results in Fig. 1 demonstrated that the TCP would almost half the gas diffusion flow Q , which is promising for experimenting with the TCP on a vacuumized chamber. Experimental results in Fig. 2 and Fig. 3 show that a reduced ambient temperature and consequently a reduced surface temperature of the HPM vacuum chamber, would result in successfully acquiring the minimum threshold vacuum pressure for HPM operation (15 mPa). They also demonstrate that the pressure increase rate is almost halved from approximately 0.71 Pa/min to 0.36 Pa/min at a 10 degrees Celsius reduction of surface temperature. Experiments are currently ongoing and will be shared during the conference.

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Revised Methodologies for the Assessment of Novel Pulsed Sources

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Abstract– This paper proposes revised methodologies for more accurate representation of the Electromagnetic environment (EME) produced by pulsed Radio Frequency (RF) sources. The proposed methods can aid assessment of RF sources within civilian policy.

Keywords– Propagation Model, Policy, Pulsed Source, Wavelet.

I. INTRODUCTION

In the UK, the Office of Communications manages electromagnetic (EM) spectrum, as defined by The Communications Act 2003 [1] and the Wireless Telegraphy Act 2006 [2]; this includes managing the risk to third party equipment from radiated RF fields, which is achieved through licensing of EM spectrum. It is apparent, however, that the associated policies and processes are not suitable for novel pulsed RF sources. In this paper, revised methodologies are described to enable more accurate representation of the radiated EME produced by novel pulsed sources, and the risk to equipment.

II. CONSIDERATIONS FOR REPRESENTING RADIATED FIELDS FROM NOVEL PULSED SOURCES

Whilst traditional, Continuous Wave (CW) transmitters place almost all their radiated power into a narrow frequency range, novel pulsed sources can have wider bandwidths. To accurately model the EME produced by a pulsed source, the spectral content of the radiated waveform must be used. In this work, the *fft* function in MATLAB® has been used to compute the Discrete Fourier Transform (DFT) coefficients of the radiated time-domain waveform, using a Fast Fourier Transform (FFT) algorithm [3]. From the DFT coefficients, the power spectrum was found [4]. The *coverage* function in MATLAB had been used to display the radiated EME of an isotropic transmitter over terrain at the National Spectrum Centre, using the Irregular Terrain Model (ITM) [5]. The RF carrier frequency plays a dominant role in RF propagation over terrain. MATLAB's implementation of the ITM requires a single frequency and power; however, since pulsed sources can have a large bandwidth, it would be misleading to model these at a single frequency and power. Instead, the power spectrum is used to model coverage at several frequencies, and their outputs are summed. Fig. 1 shows a comparison between the single-frequency and a multi-frequency model.

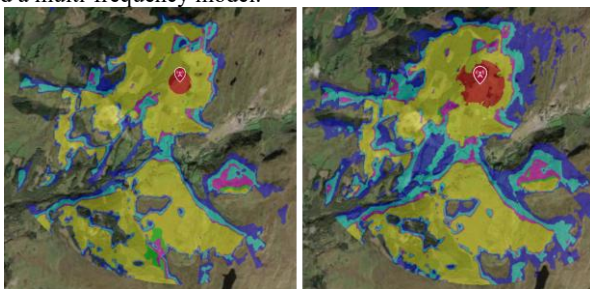


Fig. 1. Comparison of single (left) and multi (right) frequency models. Within the ITM, there also exist independent variables for 'time variability' and 'situation variability'. These are used to represent the fraction of time during which the actual path loss is less than or

equal to the model's prediction due to long term variability or random variables respectively. The situation variability is the shorter-term effect and will have the larger impact within the model; the impact of different values can be seen in Fig. 2.

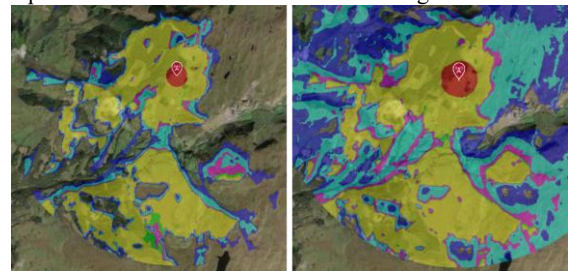


Fig. 2. Comparison of 0.5 (left) and 0.999 (right) situation variability. Time variability and situation variability cannot be easily predicted; but can be determined experimentally. The method proposed here is to model a range of different values and adopt those which most closely match the experimental measurements of the radiated EME. A final consideration is that pulsed sources may operate in bursts rather than be continuously radiating. It is important to recognise the average power output separately from the peak power output (which may exist for only a few nanoseconds). Wavelet analysis can be an effective tool for visualising the temporal patterns of different spectral components of pulsed waveforms, as shown in Fig. 3.

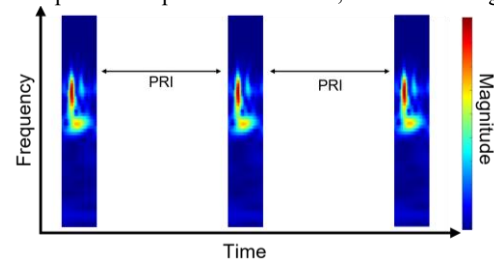


Fig. 3. Magnitude scalogram of a repetitively pulsed transient RF source

IV. CONCLUSION AND FUTURE PROSPECTS

A number of revised methodologies for the assessment of novel pulsed sources have been presented. Further work will experimentally validate the methodologies and devise visualisation methods to communicate the results to the regulators of civilian spectrum policy. Once representative EME are known, third party equipment must also be tested in order to accurately evaluate the risk.

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Optimal Design of Broadband and Polarization Insensitive Multilayer-Dielectric Metamaterial

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Abstract—In this paper, an optimal design method for a broadband and polarization insensitive multilayer-dielectric metamaterial is proposed that consists of three-acrylic layers including air-filled holes. To achieve polarization insensitive refractive indices, the sizes of the holes are determined optimally by using the genetic algorithm by maintaining the mirror symmetry of the holes of the top and bottom layers for the center layer. Based on simulation results from 2.775 to 3.225 GHz of which the fractional bandwidth is 15%, broad refractive indices are found for the three independent polarizations of which the errors are less than 0.03% for the target refractive index 1.5.

Keywords— Metamaterial, Multilayer dielectric, Broadband, Polarization insensitive, Genetic algorithm.

I. INTRODUCTION

Broadband dielectric metamaterials composed of a fiberglass with a polytetrafluoroethylene (F4B) [1] or a woven fiberglass with an epoxy resin (FR4) [2] substrates are essential to realize meta-devices such as electromagnetic cloaks [1,2] in the three-dimensional (3D) space. However, there exist a limitation that the refractive indices of them change sensitively for polarization states [1,2]. To overcome the limitation, an optimal design method for a broadband and polarization insensitive multilayer-dielectric metamaterial is proposed that consists of three-acrylic layers including air-filled holes.

II. METHODOLOGY AND RESULTS

To design a broadband metamaterial, three acrylic substrates are utilized of which the relative permittivity and the thickness are 2.6-i0.01 and 2 mm, respectively. To achieve a target refractive index 1.5, an air-filled-cylindrical hole is included in each substrate, and the three substrates are combined each other. For the polarization insensitive characteristic, the sizes of the holes are determined optimally by using the genetic algorithm (GA) [3,4] by maintaining the mirror symmetry of the holes of the top and bottom layers for the center layer. The figure of merit (FOM) of GA is set to an average of the refractive indices for three different polarizations shown in Fig. 1. GA is applied four times until FOM of the first chromosome converges as shown in Fig. 2(a). The radii of the holes included in the outer and the center layers are determined as 1.3 and 3.5 mm, respectively. From simulation results from 2.775 to 3.225 GHz shown in Fig. 2(b), broad refractive indices are found for the three independent polarizations of which the errors are less than 0.03% for the target refractive index 1.5.

III. CONCLUSION

An optimal design method for a broadband and polarization insensitive multilayer-dielectric metamaterial is proposed based on GA. From full-wave simulations with the center frequency 3 GHz,

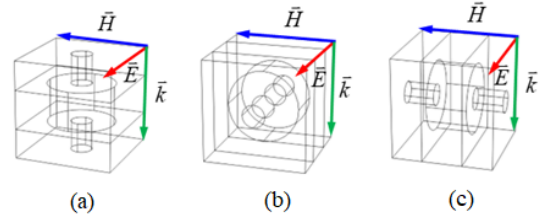


Fig. 1. Polarizations for full-wave simulations. (a) Polarization of which the wave vector is parallel with the axes of the holes. Polarizations of which the wave vector is orthogonal to the axes of the holes are parallel with (b) Electric or (c) Magnetic fields.

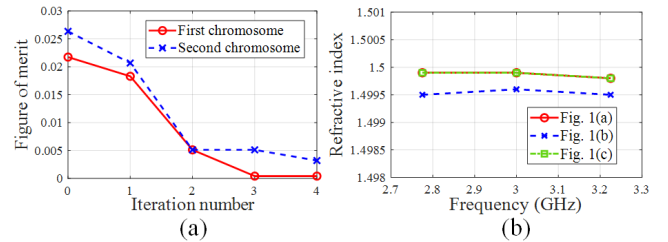


Fig. 2. Simulation results. (a) Figure of merit of the genetic algorithm. (b) Refractive indices for three different polarizations from 2.775 to 3.225 GHz.

the polarization insensitive refractive indices are confirmed for three different polarizations within a fractional bandwidth of 15%.

ACKNOWLEDGMENT

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Leakage Current Measurement System for Surge Protection Devices in Electromagnetic- Pulse Protective Filters

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Abstract– Leakage-current growth in metal-oxide varistor surge protective devices is a well-known precursor to deterioration, thermal runaway, and line-to-ground failure. In electromagnetic-pulse protective power-entry filters, surge protective devices are exposed not only to repetitive surge stress but also to fast conducted transients, making continuous condition monitoring important for mission-critical infrastructure. This paper presents a leakage-current measurement and monitoring system integrated with an electromagnetic-pulse protective filter for a three-phase, four-wire, 380 V, 60 Hz input. The system measures per-phase leakage current, line voltage and current, temperature, and surge-event count, and supports both local visualization and remote supervision through optical-network communication. The implemented platform provided microampere-level leakage-current readings together with concurrent temperature and surge-count information. The work also addresses measurement issues specific to electromagnetic-pulse protective filters, including wide dynamic range, transient immunity, and interpretation of leakage-current components. The results demonstrate the feasibility of continuous condition monitoring for in-service surge protective devices and provide a basis for future calibration, uncertainty analysis, and standards-oriented validation.

Keywords– electromagnetic pulse, high-altitude electromagnetic pulse, surge protective device, metal-oxide varistor, leakage current, condition monitoring

I. INTRODUCTION

High-altitude electromagnetic-pulse protection for critical facilities commonly combines shielding, grounding, and power-entry filtering. Within such filters, surge protective devices absorb transient energy but may gradually deteriorate during service. An increase in leakage current is therefore a useful early indicator of degradation and possible failure. This work develops a leakage-current monitoring system integrated with an electromagnetic-pulse protective filter and focuses on interference-robust sensing for practical in-service diagnosis.

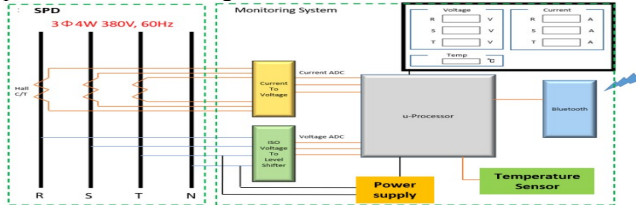


Fig. 1. System-level block diagram of SPD leakage-current monitoring integrated with EMP filter (sensors → isolation/ADC → embedded CPU → optical/LAN + local GUI).

II. SYSTEM DESIGN

The proposed system targets a three-phase, four-wire, 380 V, 60

Hz power-entry filter. The monitoring platform acquires per-phase leakage current, line voltage and current, temperature, and cumulative surge count. For reliable operation under severe conducted disturbances, the system uses phase-separated cable routing, shielded cables and connectors, and a metal enclosure. Remote supervision is supported through optical-network conversion, while local operating status is displayed through an embedded user interface.

III. RESULTS AND DISCUSSION

The developed platform successfully demonstrated simultaneous reporting of microampere-level leakage current, temperature, and surge-event history. The user interface provided phase-by-phase status information and enabled rapid identification of abnormal operating conditions. These results confirm that the proposed system is more than a simple data-acquisition unit; it functions as an integrated monitoring platform for surge protective device health supervision in electromagnetic-pulse protective filters. Future work should include explicit calibration, uncertainty analysis, and immunity validation under standards-based conducted-transient test conditions.

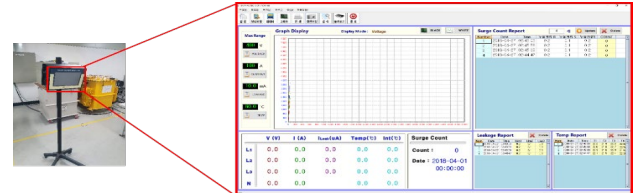


Fig. 2. Example GUI outputs

IV. CONCLUSION

A leakage-current monitoring system for surge protective devices in electromagnetic-pulse protective filters has been developed and demonstrated. The system integrates electrical, thermal, and event-history sensing with local visualization and remote optical-network supervision. The results verify the feasibility of continuous condition monitoring for early detection of surge protective device deterioration and support future standards-aligned validation for electromagnetic-pulse-resilient power-entry protection.

ACKNOWLEDGMENT

This research was supported by the Institute for Information & Communications Technology Planning & Evaluation (IITP) grant funded by the Korea Government (Ministry of Science and ICT (MSIT)) under Grant No. RS-2025-02642985 (Development of Identification, Protection, and Monitoring Technologies for High-Power Electromagnetic Defense).

Port Protection Against HPEM in Complex Enclosures via Coherent Perfect Absorption

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Abstract—We demonstrate coherent perfect absorption (CPA)-enabled port protection in a complex enclosure with aperture by coherently controlling multi-port excitations to protect devices from high power electromagnetics (HPEM). Here, an external field enters through aperture, and the cavity port is actively tuned so that the incident power is prevented from being absorbed internally and is instead driven to exit the enclosure through the aperture via reflection/reradiation, without requiring an additional dedicated sink port. Full-wave simulations show that, even under external excitation through the aperture, tuning the cavity port to satisfy the CPA minimizes the input-to-output power ratio while simultaneously reducing the power absorbed by the protected load port to a negligible level.

Keywords— Coherent perfect absorption, HPEM, energy-sink.

I. INTRODUCTION

High-power electromagnetics (HPEM) entering through an aperture can couple into an indoor space and expose internal electronics to hazardous energy levels. Ensuring port (device) protection therefore requires an electromagnetic protection strategy that prevents the incoming power from being absorbed by internal ports (devices). Coherent perfect absorption (CPA) offers a physics-based way to steer power by coherently controlling the relative amplitudes and phases of two or more excitation channels, enabling selective routing of energy among available loss channels^[1].

In this paper, we demonstrate a CPA-enabled port-protection operation that suppresses power absorption at an internal load under external illumination with complex enclosure with aperture. A key advantage is that the aperture itself can be treated as the dominant loss/leakage channel of the overall system (including the cavity and external port), allowing the incident energy to exit through the aperture without introducing an additional dedicated sink port.

II. METHODOLOGY AND RESULTS

The simulation testbed consists of a 3D metallic enclosure with a rectangular aperture on the front wall. A mechanical stirrer is placed at the enclosure center to promote mode mixing and emulate a strongly reverberant environment. External illumination is provided by a waveguide-fed source positioned in front of the aperture; the waveguide is separated from the aperture by 85mm, satisfying the far-field condition to approximate an incident plane-wave-like excitation at the aperture. In addition to the external excitation path, an internal drive port is located on the side wall of the cavity to allow controlled injection of signals from inside. A 50Ω load is terminated on the top of the cavity, which can represent a protected element inside the enclosure, and the objective is to suppress the power coupled into these loads.

At the CPA frequency of 5.5945 GHz (Fig. 2), the input-to-output power ratio drops to 1.244E-4, indicating near-complete suppression of the targeted output under the CPA condition.

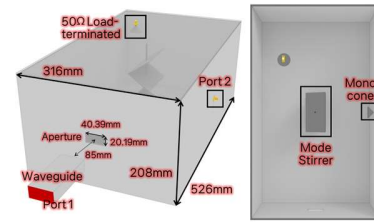


Fig. 1. 3D metallic cavity used to emulate reverberant environment.

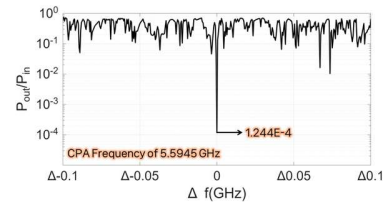


Fig. 2. Input to output power ratio as a function of frequency offset.

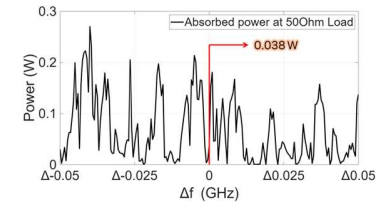


Fig. 3. Absorbed power at 50Ω load when CPA condition is satisfied.

Consistently, when the input power is 0.5W, the absorbed power at the 50Ω load is only 0.038W (Fig. 3), confirming that power absorption by the protected load is negligible when CPA is satisfied. These results imply that an effective port protection can be realized through the aperture itself—without adding a dedicated sink port—thereby preventing energy from being delivered to internal loads.

IV. CONCLUSION

We demonstrated CPA-enabled port security in a complex enclosure with aperture, where coherent amplitude/phase control suppresses absorption at internal loads and redirects energy through the aperture without a dedicated sink port. At 5.5945 GHz, simulations show low input-to-output power ratio and absorbed power of 0.038W at 50Ω load, confirming effective load protection.

ACKNOWLEDGMENT

This work was supported by the National Research Foundation (NRF) of Korea [Grant NRF-2023R1A2C1007983].

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Online Test of 10-kV Arresters Excited by Nanosecond-Level Transient Electromagnetic Disturbances

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Abstract– This paper investigates the response behaviors of 10-kV arresters under the nanosecond-level transient electromagnetic disturbances excitation online. A 10-kV equivalent distribution power network test platform was built, the non- and externally-gapped arresters were tested and the discharge current were compared and analyzed. Test results demonstrate that the continuous operating voltage of power system has little effect on the transient response of the arrester. The arc-extinguishing effect of externally-gapped arrester's air gap affects the pulse width of the discharge current. These conclusions provide guidelines for the protection in system and support that the results obtained in the laboratory are of significant value for the arresters installed in the power network.
Keywords– Arrester, Nanosecond-Level Transient Electromagnetic Disturbances, Online test, Response behavior.

I. INTRODUCTION

Nanosecond-level transient electromagnetic disturbances (TED), such as high-altitude electromagnetic pulse (HEMP), can cause fault or damage in the distribution power systems. The response characteristics and protection effects of the arresters are worthy of study. However, most of the arresters' tests in the existing relevant literature were carried in an offline state [1]–[3]. The objective of this study is to compare the effects of online and offline tests on the response characteristics of the arresters. The key contribution is to build a 10-kV equivalent distribution power network test platform and inject conducted HEMP into non- and externally-gapped arresters online, which enables a clear understanding of the protective effectiveness of arresters in real power systems.

II. METHODOLOGY

The test platform consists of the conducted HEMP injection system, the arrester samples and the 10-kV equivalent distribution power network, shown in Fig.1. The conducted HEMP injection system includes the conducted HEMP generator and the coupling network. The waveform of the conducted HEMP is 20-ns rise time and 550-ns pulse width. And the coupling network plays two important roles during the test. One is that when the distribution network is operating normally, the coupling network presents a high-resistance state, isolating the HEMP generator from the distribution network. The other is that when the HEMP generator outputs nanosecond-level TED, it can change to a low-resistance state quickly, coupling the pulse current into the distribution network.

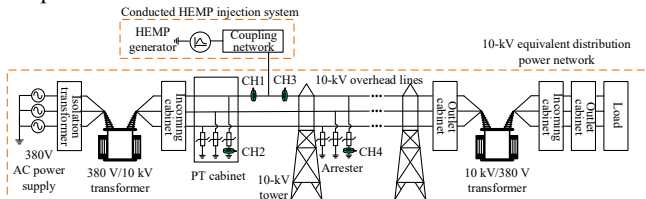


Fig. 1. 10-kV equivalent distribution power network test platform. This paper mainly focuses on the response characteristics of the tested arresters installed on the 10-kV tower. Three types of 10-kV

arresters were selected, namely NG-1, EG-1, and EG-2, respectively.

III. RESULTS AND DISCUSSION

Taking the test waveform of NG-1 as an example, as shown in Fig. 2 (a), whether the system is online has little effect on the waveforms of current flowing through the arrester. Since the normal operating current of the system is very small compared to the injected pulse current, and the normal operating current of the system almost overwhelms in the injected pulse current. The tests were carried out at different phases of the system operating voltage, the same conclusion was obtained. The current I_0 injected into the 10-kV distribution network through the coupling network are basically the same, the discharge currents of the three different arresters are compared and analyzed, as shown in Fig. 2 (b). It can be seen that the current flowing through the non-gapped arrester NG-1 is wider than that flowing through the externally-gapped arresters EG-1 and EG-2. The resistance and inductance of the overhead line will cause a loss in the current amplitude and make the rise time of the current slow to hundreds of nanoseconds. Under a hundreds of nanosecond-level pulse current impact, the response speeds of NG-1, EG-1 and EG-2 are approximate, while the pulse width of the discharge current is related to the arc-extinguishing gap of EG-1 and EG-2.

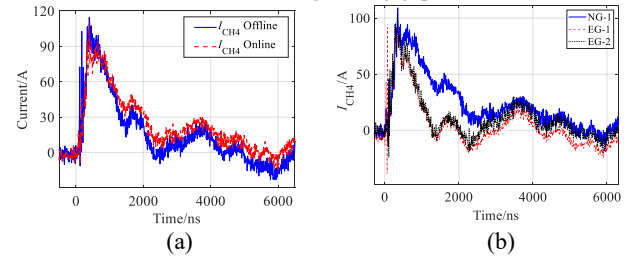


Fig. 2. The discharge current waveforms of the arresters under tests.

IV. CONCLUSION

Measured results show that whether the system has a continuous operating voltage and the phase of the continuous operating voltage have little effect on the transient response of the arrester. The waveforms of non-gapped arrester's discharge current differ from that of externally-gapped arrester's, due to the arc-extinguishing effect of the air gap. These findings support that the conclusion tested in the laboratory are of value of reference for the arresters in the distribution power network.

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HPEM in Controlled Fusion Facility

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Abstract— High-power electromagnetic pulse (HPEM) environments pose significant challenges to controlled fusion facilities. This paper overviews HPEM effects in magnetic confinement fusion (MCF) and laser inertial confinement fusion (ICF) facilities. MCF devices generate stray magnetic fields from plasma and coil currents. The primary focus is ICF facilities, where laser-target interactions generate Electron-EMP from suprathermal electrons, and X-rays induce Cable-SGEMP, External-SGEMP, Cavity-SGEMP, and SREMP. Understanding these HPEM mechanisms is essential for developing protection strategies for next-generation fusion reactors.

Keywords— controlled fusion, HPEM, Electron-EMP, SGEMP, SREMP

I. INTRODUCTION

Controlled nuclear fusion is widely regarded as the ultimate solution to global energy challenges, offering the advantages of environmental friendliness, abundant fuel, and high sustainability compared to traditional fission energy. Two primary approaches have been developed to achieve fusion conditions: MCF and ICF. Both types of fusion facilities generate complex and intense electromagnetic environments that pose significant challenges to the reliable operation of electronic systems. This paper provides an overview of the electromagnetic pulse environments in both MCF and ICF facilities, with a primary focus on the complex HPEM sources and effects in laser inertial confinement fusion facilities.

II. HPEM IN MCF FACILITIES

Magnetic confinement fusion facilities, particularly tokamaks, generate complex electromagnetic environments during operation. The main electromagnetic interference sources in MCF facilities fall into two categories: stray magnetic fields from coil systems and pulsed electromagnetic radiation from high-voltage switching operations. Auxiliary systems such as Marx generators also contribute pulsed emissions. Measurements by Meng's team on the SUNIST device at Tsinghua University show that the Marx generator produces electric field peaks of approximately 850 V/m, 400 V/m, and 200 V/m in the X, Y, and Z directions, respectively (Fig. 1). Although these values remain within the limits specified by IEEE Std C95.1-2009, the pulsed nature of these emissions adds to the overall complexity of the electromagnetic environment.

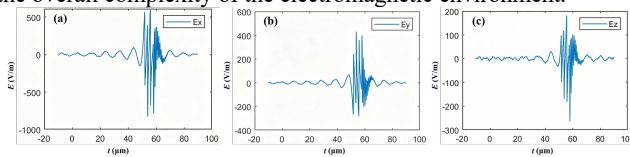


Fig. 1. Time-domain waveform of the electric field of the Marx generator: (a) E_x ; (b) E_y ; (c) E_z .

III. HPEM IN ICF FACILITIES

Laser inertial confinement fusion facilities generate far more complex and intense high-power electromagnetic pulse environments compared to MCF devices. The interaction of high-

power lasers with targets produces multiple HPEM mechanisms across different regions of the facility, including the target chamber interior, target chamber room, and energy storage area, as shown in Fig. 2(a).

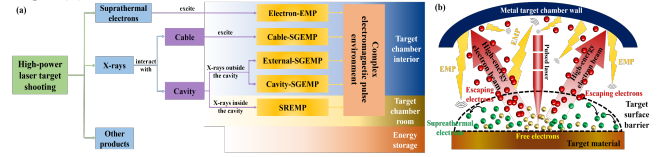


Fig. 2. (a) Classification of electromagnetic environment in ICF facilities and (b) schematic diagram of EMP excited by suprathermal electrons.

During high-power laser-target interactions, intense laser pulses generate suprathermal electrons through various plasma instabilities. A fraction of these suprathermal electrons escape the target and propagate into the target chamber interior. As these escaped electrons move through vacuum and impact the metal chamber wall, they form transient currents that excite broadband electromagnetic pulses, referred to as Electron-EMP, as shown in Fig. 2(b).

Figure 3 illustrates that X-ray interactions with the target chamber cavity produce three distinct effects. When X-rays interact with the outer surface of the cavity, they generate backscattered electrons that excite External-SGEMP in the system space. When X-rays penetrate to the inner cavity surface, forward-scattered electrons excite Cavity-SGEMP within the cavity interior. Additionally, high-energy X-rays can penetrate through the chamber wall and generate Source Region EMP (SREMP) in the target chamber room, extending the electromagnetic disturbance beyond the confinement volume.

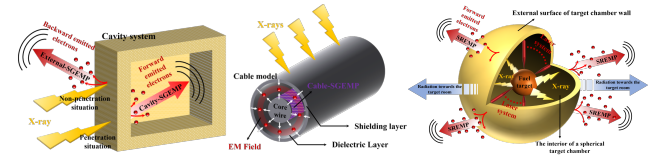


Fig. 3. Schematic diagram of SGEMP and SREMP excitation.

IV. CONCLUSION

In MCF facilities, HPEM includes stray magnetic fields from plasma and coil currents and emissions from high-voltage switching. In ICF facilities, HPEM is more complex, comprising Electron-EMP from suprathermal electrons, Cable-SGEMP from X-ray irradiated cables, and cavity-related effects: External-SGEMP, Cavity-SGEMP, and SREMP. Energy storage switching also contributes. Understanding these mechanisms is essential for developing shielding and protection strategies for next-generation fusion reactors.

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Studies and Challenges on High-Power Microwaves at CEA-Gramat

Innovative Methods for Analyzing Component Susceptibility and Examples of Technological Developments of Various Waveforms

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Abstract – HPEM systems are characterized by the multiplicity of possible waveforms (ultra-wideband, narrowband, etc.) [1]. It is therefore necessary to have increasingly optimized tools for understanding phenomena and to illustrate the feasibility of concepts through technological demonstrators. This contribution shows how CEA-Gramat addresses these challenges. For this purpose, two innovative techniques for analyzing component susceptibility are presented, as well as two examples of technological developments.

Keywords– HPEM sources, vulnerability studies, HPEM.

I. INTRODUCTION

The first studies associated with HPEM focused on achieving maximized microwave power. The objective was to have the highest possible power, as it would allow neutralizing any type of electronic component. This search for a generic HPEM effector ultimately faced little credible concepts, leading to cumbersome solutions that might not produce any effect on targets. Studies were reoriented during the 2000s-2010s towards the search for concepts dedicated to targets (or families of targets), resulting in the demonstration of more credible capabilities, leading to more compact emitters [2 - 3]. However, these achievements reveal a plurality of radiated waveforms (ultra-wideband, narrowband, etc.). To analyze and understand the susceptibility of components, it has therefore become crucial to have both innovative techniques for analyzing and understanding phenomena, and to study the technological feasibility of effector concepts allowing the generation of very diverse waveforms. This presentation details two examples of techniques developed for analyzing susceptibility and two examples of technological concepts realized to meet these challenges.

II. ANALYSIS OF SUSCEPTIBILITY

The two approaches aim to reconcile the advantages of the three techniques commonly used for analyzing the susceptibility of components and to avoid their pitfalls. These three methods are summarized in Tab 1.

TABLE I. ADVANTAGES AND DISADVANTAGES OF COMMONLY USED METHODS FOR ANALYZING SUSCEPTIBILITY

Method	Advantage	Disadvantage
Far-field illumination	Non-intrusive	Understanding of phenomena
Near-Field scanning	Slightly intrusive Localized illumination	Does not allow knowing the voltage deposited at the input of a component
Direct power injection	Precise control of the injected signal	Very intrusive

The first of the two innovative methods consists of merging the principle of direct injection with near-field scanning, through the development of an analytical injection model. This model reconciles the advantage of localized illumination allowed by near-field scanning with the determinism of direct power injection. The second method consists of performing localized non-intrusive plane wave illumination and targeting the illuminated component area. Examples of results will be shown.

III. STUDIES ON SOURCE TECHNOLOGY

Work on high-power microwave source technology focuses on the ability to have evolving waveforms but always seeks to improve the performance of electronic vacuum tubes. Two examples of work carried out by CEA-Gramat are shown, with, on the one hand, the development of a model and a prototype of a nonlinear transmission line. This type of technology has the advantage of increased waveform agility with moderate radiated power. The second example, on the other hand, shows the design of a relativistic-type electronic vacuum tube combining electromagnetic power of several hundred MW with an high operating frequency. Combining high power and high frequency was made possible through the use of higher-order propagation modes of the TM₀₂ type. The electromagnetic performances of these systems will be detailed.

IV. CONCLUSION

The work presented in this communication highlights the challenges related to HPEM addressed by CEA-Gramat in connection with the plurality of waveforms characteristic of HPEM. Two innovative techniques for analyzing component susceptibility are first shown, as well as two examples of technological developments resulting in various waveforms.

ACKNOWLEDGMENT

The authors would like to express their gratitude to the laboratories of LAAS, XLIM, and SIAME for their valuable contributions and insightful discussions. Additionally, they extend their thanks to the French Direction Générale de l'Armement (DGA) for their financial support.

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A Multi-Megawatt Microwave Power Divider

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Abstract– This work presents a high-power 2-way WR-340 waveguide power divider optimized for field uniformity and robust operation at 3 GHz. The refined septum-based topology reduces current discontinuities and electric-field hotspots, supporting reliable performance under multi-megawatt conditions. EM-thermal analysis indicates stable operation up to 5 MW peak-power in pulsed regime. The design achieves $S_{11} < -25$ dB. Fabricated from marine-grade Aluminum with anodizing and silver plating, the prototype demonstrates close agreement between simulation and measurement, validating its suitability for next-generation high-power microwave systems.

Keywords– breakdown, divider, high-power, manufacturability, S-band, septum, thermal, and waveguide.

I. INTRODUCTION

Power dividers – PD are fundamental components in high-power antennas and allied systems enabling the controlled distribution of RF energy across multiple paths while maintaining low loss, high isolation, and stable impedance characteristics [1]. Waveguide-based dividers are particularly suitable for demanding applications due to their superior power-handling capability, mechanical robustness, and low insertion loss compared to planar counterparts [2]–[3]. However, traditional T-junction configurations often suffer from non-uniform field distribution, current crowding, and localized hotspots that limit their performance under multi-megawatt excitation. To address these challenges, this work investigates an improved 2-way power-divider with a novel topology that reduces the electrical stress, enhances the thermal performance and supports modular operation (i.e., septum replacement) for easier tuning and maintenance. The presented design targets reliable operation in harsh, high-duty-cycle environments, combining electromagnetic optimization with thermal and mechanical considerations to achieve stable high-power performance in S-band.

II. DESIGN AND SIMULATION

The standard WR-340 waveguide ($a = 86.36$ mm, $b = 43.18$ mm) is used due to superior power-handling capability at the targeted frequency band compared to WR-284. The proposed power divider provides a balanced power division and match i.e., < -25 dB over 2.9–3.1 GHz band by means of a septum as illustrated in Fig. 1. The surface current distribution shown in Fig. 1(b) illustrates the current flow pattern for dominant propagating mode. The surface currents are flowing parallel to the assembly interface resulting in minimal

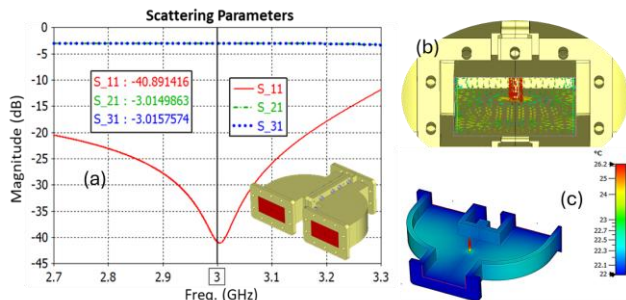


Fig. 1. Simulated (a) scattering parameter of the proposed power divider and, (b) surface current distribution along the physical junction.

mode distortion and undesired leakage through assembly joints. Fig. 1(c) shows the thermal stress on the PD when it is excited with HPM signal from its input port.

III. RESULTS AND DISCUSSION

The PD was manufactured by using CNC process utilizing standard marine grade Aluminum (6081). After anodizing coating, a thin layer of silver plating was applied to enhance the RF performance. The measured S -parameters are in good agreement with the simulation. Fig. 2 shows the measured transmission parameters, S_{21} and S_{31} , in magnitude and phase verifying the stability of the power at the output ports across the working band.

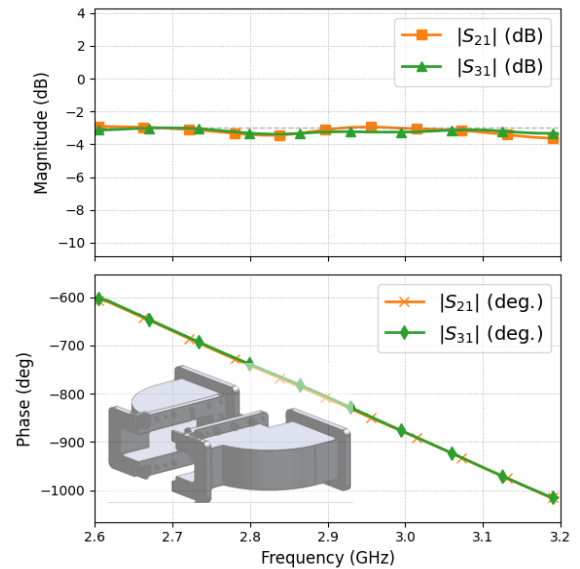


Fig. 2. Measured transmission characteristics of the proposed power divider: magnitude and phase response, 3D mechanical model in inset.

IV. CONCLUSION

The proposed WR-340 power divider demonstrates stable, well-matched performance at frequency of interest, with measurements aligning closely to simulations. Its optimized septum structure improves field uniformity for multi-megawatt pulsed operation. Experimental results at high-power regime will be shared during the conference.

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Z-transform-based Estimation of Lightning Current Waveforms from Far-field Waveforms Generated by Lightning Strikes to a Tall Structure

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Abstract—The relationship between the FDTD-computed far vertical electric-field waveform over infinitely conducting ground and that over lossy ground, generated by a lightning strike to flat ground, is represented by a transfer function in the z -domain. In the FDTD simulations, a current propagates upward along a vertical lightning channel. Using the obtained transfer function, far vertical electric-field waveforms over finitely conducting ground for a lightning strike to flat ground and to a tall structure, assuming a typical return-stroke current, are converted into the corresponding waveforms over infinitely conducting ground.

Keywords—lightning; ground conductivity; FDTD method; tall structure; z -transform

I. INTRODUCTION

A method has been proposed to estimate lightning current waveforms from far-field electric or magnetic field waveforms generated by lightning strikes to a tall structure over infinitely conducting ground [1]. If the far-field waveforms generated by lightning strikes to a tall structure over finitely conducting ground can be converted to those corresponding to infinitely conducting ground, the lightning current waveforms can subsequently be estimated using the method [1]. In this paper, the far-field electric fields generated by a lightning strike to flat ground are computed using the FDTD method [2] in a 2D cylindrical-coordinate system for cases of infinitely and finitely conducting ground, and the relationship between the two results is represented by a transfer function in the z -domain [3]. This transfer function is then applied to the far-field electric-field waveforms generated by a lightning strike to a tall structure over finitely conducting ground, and the characteristics of the waveforms converted to those corresponding to infinitely conducting ground are studied.

II. RESULTS AND CONCLUSION

Fig. 1 (a) shows FDTD-computed waveforms of vertical electric field E_z on the ground at distance $d = 200$ km from a lightning channel attached to a 180-m-tall structure on the ground of $\sigma = \infty$, 5 S/m, 10 mS/m, 1 mS/m, and 0.5 mS/m for the case that the short-circuit lightning current I_{sc} has a magnitude of 11 kA and a risetime of 1 μ s. Fig. 1 (b) shows waveforms of E_z on the infinitely conducting ground at $d = 200$ km from a lightning channel attached to a 180-m-tall structure on the ground, rebuilt with the z -transform transfer function from the FDTD-computed

waveforms for $\sigma = 5$ S/m, 10 mS/m, 1 mS/m, and 0.5 mS/m shown in Fig. 1 (a). Note that the parameter values of the z -transform transfer functions are determined from the relationship between the FDTD-computed E_z waveform for $\sigma = 5$ S/m, 10 mS/m, 1 mS/m or 0.5 mS/m and that for $\sigma = \infty$. It follows from Fig. 1 that the E_z waveform for $\sigma = \infty$ is rebuilt properly from that for $\sigma \geq 10$ mS/m. When $\sigma = 1$ mS/m or 0.5 mS/m, the rebuilt E_z peak is 5 or 10% lower than the E_z peak for $\sigma = \infty$. Note that the original E_z peak for $\sigma = 1$ mS/m or 0.5 mS/m is 58 or 61% lower than that for $\sigma = \infty$.

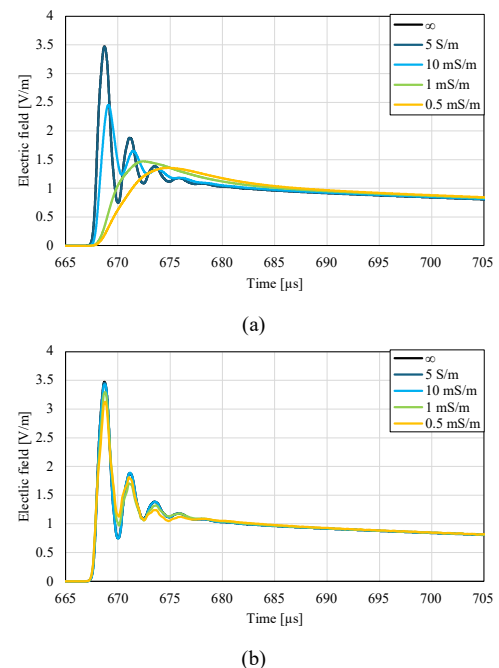


Fig. 1. (a) FDTD-computed waveforms of E_z on the ground of $\sigma = 5$ S/m, 10 mS/m, 1 mS/m, and 0.5 mS/m at $d = 200$ km from a lightning channel attached to a 180-m-tall structure, and (b) waveforms of E_z on the ground of $\sigma = \infty$ rebuilt with the z -transform transfer function from the FDTD-computed waveforms for $\sigma = 5$ S/m, 10 mS/m, 1 mS/m, and 0.5 mS/m.

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Thundercloud Location Estimation for Lightning Warning in Coastal Wind Farms Using Two Electric Field Mills and Wind Direction

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Abstract– This paper proposes a thundercloud location estimation algorithm to provide early lightning warnings for coastal wind farms. Using two electric field mills, the distance is estimated based on the image charge method, where the cloud is modeled as a finite vertical line charge. Geometric ambiguity is resolved by integrating wind direction data. Measurement results demonstrate a strong correlation between the estimated trajectories and actual lightning strikes. The proposed method enables proactive protection measures, such as securing turbine blades, prior to the initial discharge.

Keywords– Thundercloud Location Estimation, Lightning, Electric Field Mills, Wind Direction, Coastal wind farm

I. INTRODUCTION

High-power electromagnetic environments such as lightning can cause severe damage to wind turbine blades on the coast, where lightning strikes have recently led to frequent blade fractures. The objective of this study is to estimate the location of thunderclouds to prevent such lightning-related disasters. We propose a novel location estimation method using only two electric field sensors combined with wind direction data. This approach improves the accuracy of cloud tracking compared with conventional triangulation methods that typically require three or more sensors. This study is crucial for mitigating lightning hazards and enhancing the resilience of offshore wind infrastructure.

II. METHODOLOGY

The proposed algorithm estimates the real-time location of thunderclouds using two electric field mills (EFMs) and wind direction data. The method consists of distance estimation based on the image charge method and ambiguity resolution using meteorological information.

A. Distance Estimation using Image Charge Method

Based on the image charge method, the vertical electric field E measured at the ground level ($z=h$) is expressed as follows:

$$E(R) = \frac{1}{4\pi\epsilon_0} \int_{z_1}^{z_2} \left[\frac{\rho(z-h)}{(R^2 + (z-h)^2)^{3/2}} - \frac{\rho(z+h)}{(R^2 + (z+h)^2)^{3/2}} \right] dz \quad (1)$$

where ϵ_0 is the vacuum permittivity, ρ represents the linear charge density, R is the horizontal distance, and z_1 and z_2 denote the altitude range of the thundercloud. Based on this equation, the horizontal distances R_1 and R_2 from each sensor are calculated.

B. Intersection Calculation and Ambiguity Resolution

In Fig. 1, two potential cloud locations, P_1 and P_2 are determined as the intersection points of two circles centered at sensors S_1 and S_2 with radii R_1 and R_2 , respectively. To resolve this geometric

ambiguity, wind direction and the time rate of change of the electric field are used. The final position is selected by determining whether the thundercloud is approaching or receding from the sensors according to the wind direction and electric field variation.

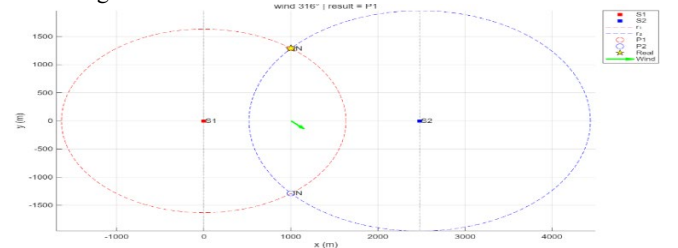


Fig. 1. Selection of the thundercloud location between intersection points P_1 and P_2 based on wind direction.

III. RESULTS AND DISCUSSION

The proposed method successfully predicted the temporal movement of thunderclouds. The estimated thundercloud positions showed strong spatial correlation with the actual lightning strike locations. The errors from the actual lightning strike points were 2.97 km, 0.94 km, and 2.65 km, respectively. These results validate the effectiveness of the proposed algorithm for real-time thundercloud localization using only two sensors with wind direction information.

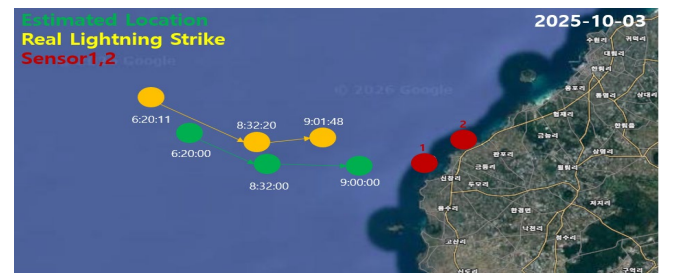


Fig. 2. Comparison between estimated thundercloud movement and actual lightning strike locations.

IV. CONCLUSION

Locating thunderclouds before the first lightning strike provides crucial benefits. Specifically, it allows for predicting lightning threats to coastal wind farms. By detecting thunderclouds near the sensors and locking the wind turbine blades in advance, lightning damage can be minimized.

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Classification of Lightning Electromagnetic Pulse Waveforms with a Machine Learning Method

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Abstract—A simple machine learning algorithm, the random forest (RF), is applied to the classification of lightning electromagnetic pulse (LEMP) waveforms into those due to positive cloud-to-ground (CG) discharges, those due to -CG, those due to positive intracloud (IC) discharges, and those due to -IC. A total of 1,128 LEMP waveforms are used, with 80% assigned to the training dataset and the remaining 20% to the test dataset. In both datasets, the proportions of +CG, -CG, +IC, and -IC are 26%, 57%, 9%, and 8%, respectively, indicating that +IC and -IC events are relatively scarce. The overall accuracy is $100\% \times (50 + 122 + 10 + 4) / 226 = 82\%$, but the least frequent -IC events tend to be misclassified as -CG.

Keywords—lightning, lightning electromagnetic pulse (LEMP), machine learning, cloud-to-ground discharge, intracloud discharge

I. INTRODUCTION

Accurate classification of cloud-to-ground (CG) and intracloud (IC) lightning discharges from their electromagnetic field waveforms is of considerable importance, as it can significantly enhance the performance of lightning locating systems (LLSs). Recent studies have demonstrated that machine learning techniques are highly effective for classifying lightning types based on lightning electromagnetic pulse (LEMP) waveforms [1]–[3]. In this study, the random forest (RF), is applied to the classification of LEMP waveforms into +CG, -CG, +IC, and -IC types, and its classification performance is evaluated. RF is an ensemble machine learning method that builds multiple decision trees and combines their predictions by majority voting for classification. A single decision tree tends to overfit and is sensitive to small data variations, but RF reduces these weaknesses by training each tree on a random subset of samples and features, which increases robustness, improves generalization performance, and reduces variance.

II. RESULTS AND CONCLUSION

Table I gives the numbers of +CG, -CG, +IC, and -IC, extracted from the data measured by the University of Tokyo [4], for training and testing. A total of 1,128 LEMP waveforms are used, with 80% assigned to the training dataset and the remaining 20% to the test dataset. In both datasets, the proportions of +CG, -CG, +IC, and -IC are 26%, 57%, 9%, and 8%, respectively, indicating that +IC and -IC events are relatively scarce. Fig. 1 shows the confusion matrix for classifying +CG, -CG, +IC, and -IC based on RF. Table II shows the precision, recall, and F1 score obtained using RF. Note that these metrics are calculated from Fig. 1 as follows: For example, the

precision of +CG is $100\% \times 50 / (50 + 2 + 2 + 2) = 89\%$, the recall of +CG is $100\% \times 50 / (50 + 6 + 3 + 0) = 85\%$, and the F1 score of +CG is $100\% \times 2 \times 50 / (2 \times 50 + 2 + 2 + 6 + 3 + 0) = 87\%$. It appears from Table II that the precision, recall, and F1 score for discriminating +CG and -CG using RF are about 90%. On the other hand, for the discrimination of +IC and -IC, these metrics are low, which are approximately 20 to 60%. Particularly, the least frequent -IC events tend to be misclassified as -CG, which needs to be improved as future work. The overall accuracy is $100\% \times (50 + 122 + 10 + 4) / 226 = 82\%$.

TABLE I. NUMBERS OF +CG, -CG, +IC, AND -IC EXTRACTED FROM THE MEASURED DATA [4] FOR TRAINING AND TESTING.

	+CG	-CG	+IC	-IC	Total
Training data	234	516	78	74	902
Testing data	59	129	20	18	226

True class	+CG	50	6	3	0
	-CG	2	122	3	2
	+IC	2	6	10	2
	-IC	2	11	1	4
		+CG	-CG	+IC	-IC
Predicted class					

Fig. 1. Confusion matrix for classifying +CG, -CG, +IC, and -IC based on the RF method.

TABLE II. CLASSIFICATION RESULTS BASED ON THE RF METHOD.

	Precision, %	Recall, %	F1 score, %	Total number
+CG	89	85	87	59
-CG	84	95	89	129
+IC	59	50	54	20
-IC	50	22	31	18

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The 3D VHF Mapping Analysis of Tropical Intracloud Flashes Initiation Process

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Abstract– This paper presents preliminary analysis utilizing 3D VHF interferometry technique of initiation process for three intracloud (IC) flashes initiated by a narrow bipolar event (NBE) and compared to ordinary IC flashes. The initiation altitudes and propagation speeds of the NBE-initiated IC flashes were higher than the ordinary IC flashes. This finding is important for the lightning initiation study which is still unclear.

Keywords– intracloud flash; narrow bipolar event; very high frequency lightning map.

I. INTRODUCTION

Some lightning flashes have been observed to be initiated by a narrow bipolar event (NBE) accompanied by strong VHF bursts with unique characteristics of fast propagation speed (10^7 to 10^8) [1]. On the other hand, ordinary flashes were initiated by an ordinary small bipolar pulse with less intense VHF burst [2]. In this paper, six intracloud (IC) flashes with three flashes were initiated by positive NBE are analyzed with the aim of analyzing the initiation process in 3D VHF interferometry mapping perspective. The main contribution of this study lies in the utilization of the method of 3D VHF interferometer technique to understand the IC initiation process and how it relates to the NBE flash. The role of NBE in the initiation process of IC flashes is still unclear whether it is responsible for the initiation or there is no relationship at all.

II. METHODOLOGY

A two-station VHF ITF system was designed to investigate the 3D propagation of lightning flashes. The stations are separated by 18.25 km. For 3D mapping, mean azimuth angles at the zero-crossing time from both stations were intersected to locate the source horizontally, while the horizontal distance was calculated using the Haversine formula and the source height was derived from station geometry following established two-station VHF interferometer method [3].

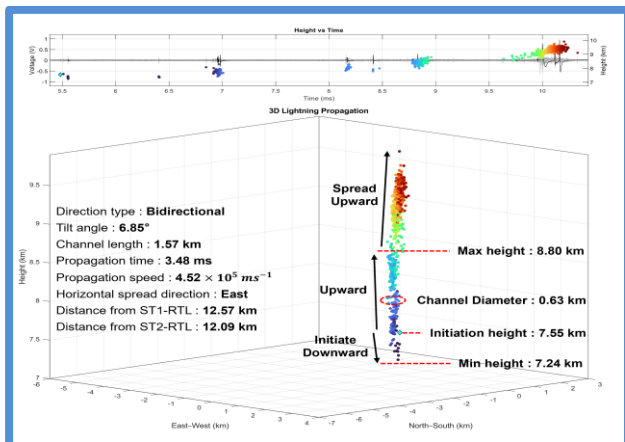


Fig. 1. The 3D map of initiation of an ordinary IC flash.

III. RESULTS AND DISCUSSION

The significant findings of the initiation process analysis between ordinary IC type and NBE-initiated IC flashes are:

- Both types show predominantly vertical bidirectional propagation during initiation with initial propagation was downward as shown in Fig. 1 and Fig. 2.
- The initiation altitudes of NBE-initiated IC flashes were higher (~9-10km) compared to ordinary IC initiation altitudes (~7-8km).
- NBE-initiated IC flashes VHF sources propagation speeds were higher (order of 10^7ms^{-1}) than ordinary IC (order of 10^5ms^{-1}).

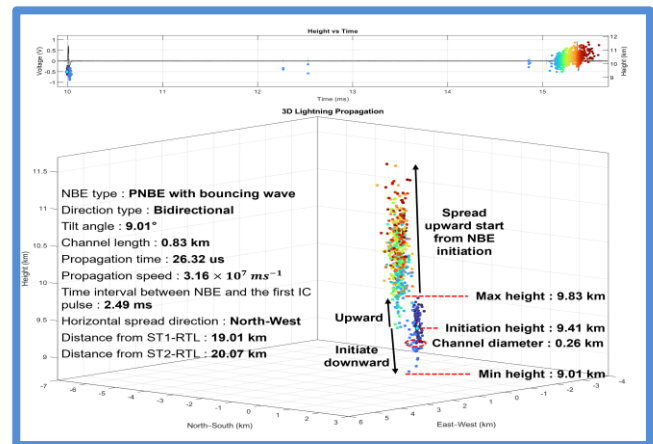


Fig. 2. The 3D map of an IC flash initiated by positive NBE.

IV. CONCLUSION

In conclusion, the initiation process of NBE-initiated IC flashes was different than ordinary IC flashes with relatively higher initiation altitudes and faster propagation speeds.

ACKNOWLEDGMENT

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First Recordings of Lightning Currents at 220 m Skyscraper in Poland

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Abstract– In March 2025, a lightning current measurement system has been launched on a 220 m high skyscraper in Rzeszow, Poland. The first recording together with the corresponding electric field signature obtained from 3 MHz flat-plate antenna is presented and analyzed. The main goal of the future research carried out on the tower will be a comprehensive analysis of the development of different types of lightning discharges, modeling and simulation of the lightning electromagnetic field pulses, as well as, contribute to global databases on lightning interactions with tall structures.

Keywords– lightning; direct lightning current measurement; lightning electric field, tall structures.

I. INTRODUCTION

The Lightning Current Monitoring System (LCMS) installed on the highest residential building in Poland and operated by the Lighting Research Group from Rzeszow University of Technology is based on a DEHN system originally dedicated for wind turbines. The measurement system is equipped with two coils intended to recording of both, fast lightning return stroke current and long continuing/continuous current. Moreover, lightning electric and magnetic fields and the development of the lightning channel will be simultaneously recorded using different sensors and an optical system belonging to the Lightning Observatory in Rzeszow (LOR) and operating at the 1.5 km distance from the skyscraper (Fig. 1).

II. LIGHTNING CURRENT MONITORING SYSTEM

During the last thunderstorm season one direct lightning event was recorded by LCMS (Fig. 2). This event, which has not been registered by LLS, occurred simultaneously with the negative downward cloud-to-ground return stroke of three-stroke lightning flash localized by the lightning location network PERUN about 226 m away from the skyscraper. The corresponding peak currents of the first and subsequent return strokes was 45.18 kA, 14.14 kA, and 5.46 kA, respectively. Results given in Fig. 2(a-b) reveal the oscillatory character observed at the tail of the current waveform which agrees with similar data found in literature [2].

IV. CONCLUSION

First lightning event recorded by a new system LCMS have been presented together with the E-field recording. The obtained lightning current measurements showed that both coils of LCMS were operating correctly. In the following seasons, both upward and downward lightning are expected to be recorded due to the location (197 m above sea level) of the skyscraper.

ACKNOWLEDGMENT

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Fig. 1. The skyscraper equipped with the Lightning Current Monitoring System and the Lightning Observatory in Rzeszow which are 1.5 km apart

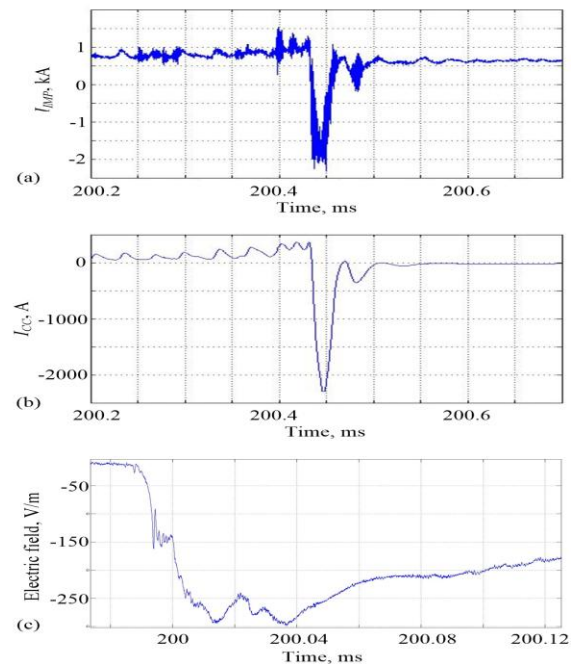


Fig. 2. First lightning event recorded by Lightning Current Monitoring System in Rzeszow: (a) lightning current from high range channel; (b) lightning current from low range channel; (c) corresponding electric field. Zero timescales in plot corresponding to the reported timestamps: E-field recording timestamp: 2025-09-04 at 04:24:56.457041(UTC) Electric current timestamp: 2025-09-04 at 04:24:56 (UTC)

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Simulation-Based Modeling of Conducted E1 HEMP Coupling Effects on a Simplified Power Substation

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Abstract– In this paper, a hybrid simulation model is proposed to analyze conducted E1 HEMP coupling in a simplified power substation. The proposed model integrates BLT equation-based calculations with transmission line and CCVT circuit models. Simulations in MATLAB/Simulink demonstrate that E1 HEMP-induced disturbances propagate along the conduction path, generating transient overvoltages at relay terminals. This study provides an analytical basis for assessing conducted HEMP threats in power systems.

Keywords– baum–liu–tesche (BLT) equation, conducted E1 HEMP coupling, power substation, transmission line

I. INTRODUCTION

High-Altitude Electromagnetic Pulse (HEMP) is an intense electromagnetic pulse generated by a nuclear detonation at altitudes above 30 km and can severely affect critical infrastructure such as power grids and control systems [1]. To analyze the propagation and terminal voltage responses of E1 HEMP-induced disturbances in power systems, simulation-based modeling incorporating line-coupling mechanisms is required [2]. This paper proposes a hybrid simulation model integrating BLT (Baum–Liu–Tesché) equation-based induced voltage calculations with transmission line and CCVT (Coupling Capacitor Voltage Transformer) circuit models. The conducted E1 HEMP coupling effects and the resulting terminal voltage transients are analyzed using MATLAB/Simulink.

II. SIMULATION OF CONDUCTED E1 HEMP COUPLING ON A SIMPLIFIED POWER SUBSTATION

The conducted E1 HEMP modeling in this study begins with the definition of the E1 HEMP electric field waveform, followed by the calculation of the induced voltage on the transmission line using the BLT (Baum–Liu–Tesché) equation. The calculated voltage is applied to a transmission line model to analyze conducted coupling effects within a simplified power system. The proposed hybrid model integrates BLT-based analytical calculations, a transmission

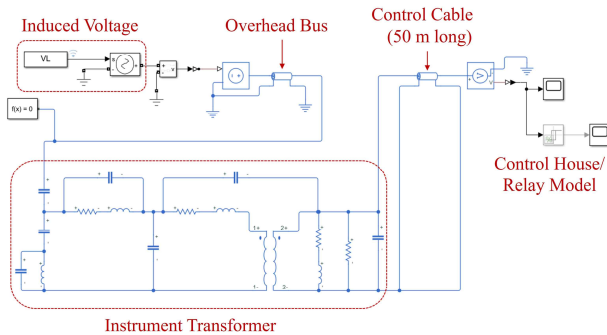


Fig. 1. Simulation model of conducted E1 HEMP coupling on a simplified power substation with applied induced voltage source.

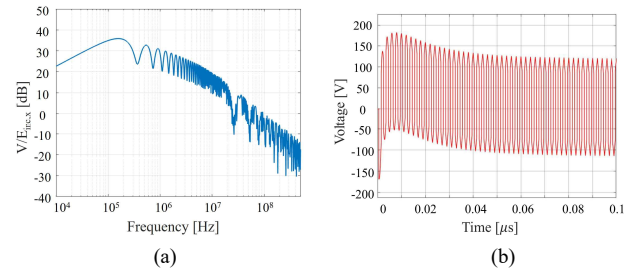


Fig. 2. (a) Frequency response of the terminal voltage under the incident electric field (b) Time-domain voltage waveform at the relay input.

line model, and a circuit-based CCVT model. The simulation is implemented in MATLAB/Simulink, capturing the conduction path from the induced voltage injection point through the instrument transformer and a 50 m control cable to the relay circuit. The instrument transformer is modeled based on the lumped-parameter representation of a 245 kV CCVT specified in the EPRI report [3].

Simulation results are shown in Fig. 2. Fig. 2(a) presents the frequency response of the terminal voltage obtained using the BLT equation, while Fig. 2(b) illustrates the time-domain voltage waveform at the relay input. These results confirm that E1 HEMP-induced disturbances propagate along the conducted path and generate transient overvoltages within the substation system.

III. CONCLUSION

This study proposes a hybrid simulation model for analyzing conducted E1 HEMP coupling characteristics in a simplified substation model. The simulation results confirm that the induced voltage calculated on the transmission line propagates through the conduction path, resulting in transient overvoltages at the relay terminals. These findings provide a critical analytical foundation for evaluating the electromagnetic vulnerability of power system control units and can be further utilized to design optimal mitigation strategies against HEMP-induced threats.

ACKNOWLEDGMENT

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Various Layers and Levels of System-Generated Electromagnetic Pulse Environment Experimental Simulation Based on Electromagnetic Topology

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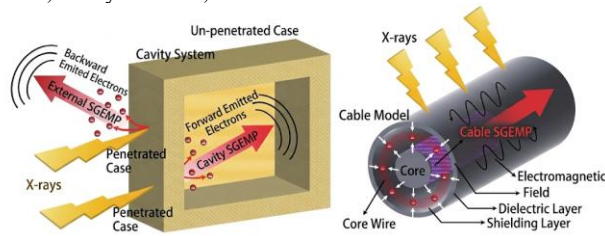
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Abstract—This article explores the various layers and levels of system generated electromagnetic pulse (SGEMP) environment experimental simulation based on electromagnetic topology. Based on the physical process of X-ray excitation of SGEMP and electromagnetic topology theory, the SGEMP environmental experimental simulation is divided into different electromagnetic topology layers from system to device. Various layers and levels of SGEMP environmental experimental simulation is proposed for the physical parameters generated and coupled by SGEMP at different electromagnetic topology layers. The physical quantities of SGEMP environment experimental simulation are divided into photons, source electrons, source current density, electromagnetic field, surface current coupled into the system, and charge density, etc., simulating the effects of SGEMP from the system to the pins of electronic system devices step by step.

Keywords— SGEMP, electromagnetic topology, experimental simulation, various layers and levels, vulnerability assessment

I. INTRODUCTION

When an electronic system is exposed to intense X-ray environments, photoemission and Compton scattering occur on metallic materials. This process generates electron emission and deposition, which in turn produce currents that excite system-generated electromagnetic pulses (SGEMP). Based on the interaction region, SGEMP is primarily classified into external SGEMP, cavity SGEMP, and cable SGEMP.



(a) External SGEMP and Cavity SGEMP (b) Cable SGEMP

Fig. 1. Schematic diagram of X-ray excitation SGEMP types.

Traditional SGEMP experiments face significant bottlenecks. The limited spot size and dose rates of X-ray sources make it extremely difficult to conduct full-system irradiation experiments. Furthermore, extrapolating system-level vulnerability from component-level test results remains technically challenging. To realistically construct experimental environments and quantitatively evaluate electronic system vulnerability, this paper introduces a comprehensive methodology that integrates electromagnetic topology to classify SGEMP coupling layers and excitation levels.

II. METHODOLOGY

Because EM environmental effects are closely related to factors like incidence frequency and object size, scalable models cannot be used

for SGEMP testing. Based on C. E. Baum's electromagnetic topology theory, the spatial geometry of a complex system is divided into bounded sub-regions representing different energy coupling stages. The experimental objects are categorized into five layers: System, equipment, circuit board, integrated circuit, and component; the simulation environments are defined across six levels based on physical quantities: photons, source electrons, source current density, electromagnetic fields, coupled surface currents/charges, and coupled internal currents/voltages.

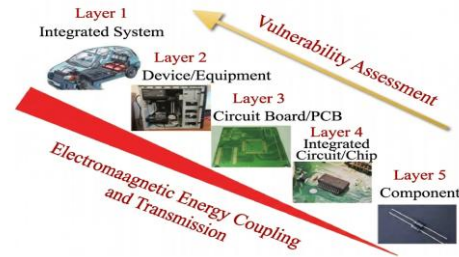


Fig. 2. Layering of systems based on electromagnetic topology.

III. RESULTS AND DISCUSSION

The proposed framework clearly defines the correspondence between various simulation techniques. Level 1 simulation uses pulsed photon sources. While this is the most direct method, it is highly restricted by the facility's source size. Level 2 utilizes electron beams to mimic source electrons, providing larger spot sizes but still facing facility constraints. Level 3 employs large-area current injection to mimic surface current densities, offering a cost-effective alternative for system-level testing. Level 4 utilizes EM field illumination. Levels 5 and 6 rely on direct current or voltage injection onto external cables and internal component pins. This significantly reduces experimental costs while maintaining topological equivalence.

IV. CONCLUSION

The proposed various layers and levels experimental simulation method systematizes the assessment of SGEMP effects. By dividing the SGEMP experimental simulation into different layers and levels, the correspondence between SGEMP experiments can be made clearer, and the equivalence research from the internal field to the external field can be based on evidence. This provides a more scientific description of the SGEMP effect of complex electronic systems under X-ray irradiation, and provides technical support for the quantitative assessment and test evaluation of SGEMP vulnerability in electronic systems.

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On Field-to-TL Coupling Equations

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Abstract – Field-to-transmission line coupling is commonly analyzed using transmission-line-based formulations such as those proposed by Taylor et al., Agrawal et al., and Rachidi, which differ in the representation of the electromagnetic source terms. These formulations are theoretically equivalent and can be expressed using different combinations of electromagnetic field components [1]. This paper addresses a fourth coupling formulation that retains electric-field-related source terms as in the Agrawal et al. model, but is expressed in terms of total voltage, a readily measurable quantity, unlike scattered voltages. The formulation, previously introduced only at a theoretical level, is here compared with the other established models and implemented within advanced time-domain numerical schemes. In addition, boundary conditions are analyzed for which the proposed formulation does not require additional voltage generators (“risers”), thereby avoiding ambiguities in physical interpretation.

Keywords– Field-to-transmission line coupling, FDTD second order, lightning-induced voltages, EMC, numerical methods.

I. AIM OF THE PAPER

Field-to-transmission line coupling is commonly analyzed using transmission-line-based formulations such as those proposed by Taylor et al., Agrawal et al., and Rachidi, which mainly differ in the representation of the electromagnetic source terms [1]. Most of these formulations have been extensively implemented and validated through time-domain numerical schemes for the analysis of lightning-induced effects on overhead lines. Formulation of the coupling equations in terms of scalar and vector potentials proposed by Cooray et al [2] are also available.

Within this context, a formulation expressed directly in terms of the total line voltage would offer a relevant practical advantage, since the total voltage is a quantity that can be readily measured. By contrast, scattered and incident voltages cannot be directly accessed for natural electromagnetic events and are typically available only through controlled experimental setups [3] or numerical simulators. The purpose of this paper is to present a new formulation of the transmission line coupling equations in which the source terms involve only electric field components, as in the Agrawal et al. formulation, but is expressed in terms of the total voltage rather than the scattered voltage used in the Agrawal et al. model. Such a formulation – introduced in [1] – has not yet been compared with the other three mentioned above, nor has it been implemented within advanced time-domain numerical schemes.

A further contribution is the clarification of termination boundary conditions, achieved without introducing additional voltage generators (“raisers” in the Agrawal et al. model), thereby avoiding potential ambiguities in physical interpretation.

II. A NEW FORMULATION

For a lossless single-conductor overhead line, the coupling equations discussed in this work can be written as [1]

$$\frac{\partial V(x,t)}{\partial x} + L' \frac{\partial I(x,t)}{\partial t} = E_x^e(x,h,t) - \frac{\partial}{\partial x} \int_0^h E_z^e(x,z,t) dz \quad (1)$$

$$\frac{\partial I(x,t)}{\partial x} + C' \frac{\partial V(x,t)}{\partial t} = -C' \frac{\partial}{\partial t} \int_0^h E_z^e(x,z,t) dz \quad (2)$$

where $V(x,t)$ and $I(x,t)$ are the line voltage and currents, respectively,

and $E_x^e(x,z,t)$ and $E_z^e(x,z,t)$ are the horizontal and vertical components of the external electric field due to lightning, L' and C' are the per-unit-length inductance and the capacitance of the line. The conductor runs along the x -axis and is located at a height h above the ground on the z -axis. Eqs. (1) and (2) are discretized using a second-order FDTD scheme formally analogous to that previously developed for the Agrawal formulation [4], [5].

III. RESULTS AND DISCUSSION

Induced overvoltages by an indirect lightning return stroke are computed at the termination of a 10-m high, 1-km long single-conductor overhead line on a lossless soil are computed using both the proposed formulation and the classical Agrawal et al. model. The stroke is 50 m distant from the line center. The channel-base current is assumed to exhibit a linear front with a flat top, characterized by a peak amplitude of 11.2 kA and a risetime of 2 μ s. The spatial-temporal distribution of the current along the channel is obtained adopting the TL model by Uman et al. [5]. Fig. 1 compares the overvoltage waveforms obtained with the two approaches. The results are identical, except for unavoidable numerical approximations, confirming that the alternative coupling formulation yields numerically equivalent induced voltages when a consistent FDTD implementation is adopted. This provides a numerical validation of the theoretical equivalence between the two coupling models. Further discussion on advantages of the model and boundary conditions will be provided. Fig. 1 also shows the contribution due to the horizontal electric field, and the sum of the incident voltage and of the riser.

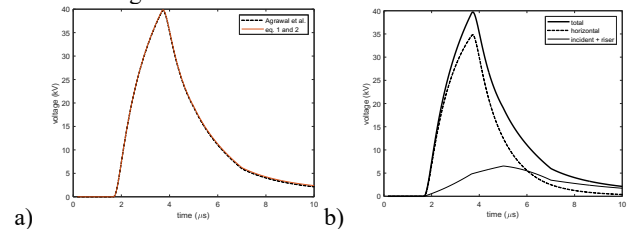


Fig. 1. a) Lightning-induced voltage at the line terminations of a single-conductor overhead line computed using the classical Agrawal formulation and the new coupling formulation (the two results are identical within numerical tolerance). b) Contribution of the horizontal electric field and of the sum of the incident voltage and the “risers” in the Agrawal formulation, corresponding to the contribution of the vertical electric field in the new formulation.

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Design of an Adjustable Multi-Stage Common-Mode Filter for Ultra-Wideband Noise Suppression

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Abstract– In this paper, an adjustable multi-stage common-mode filter (AMS-CMF) is proposed for ultra-wideband common-mode (CM) noise suppression. The inserted mid-stage strengthens the Bloch bandgap and induces interstage coupling, achieving attenuation exceeding 20 dB from 10.5 to 30.4 GHz (19.9-GHz bandwidth). In addition, a width-based unit-cell tuning technique reduces the overall filter length, overcoming the cascading requirement of conventional electromagnetic bandgap (EBG) filters and improving compactness for high-density packaging. The performance of the proposed filter is validated in both the frequency and time domains through 3-D electromagnetic (EM) simulations.

Keywords– Common-mode filter, electromagnetic bandgap (EBG), differential signaling, signal integrity, wideband noise suppression.

I. INTRODUCTION

The increasing data rates of high-speed interconnects demand robust signal integrity solutions. Differential signaling is widely adopted to mitigate noise. However, manufacturing-induced skew causes mode conversion, generating CM currents that lead to electromagnetic interference (EMI) [1]. In addition, SMD-type common-mode chokes suffer from parasitic elements that degrade high-frequency performance and require additional components. PCB-embedded EBG-based filters offer a promising alternative; however, conventional designs require lengthy cascading to achieve sufficient stopband depth and bandwidth. In this paper, an AMS-CMF is proposed to overcome these limitations through the introduction of a mid-stage and a width-based unit-cell tuning technique, thereby achieving broadband CM noise suppression with a compact footprint.

II. METHODOLOGY

The unit cell of the proposed filter is shown in Fig. 1. The conventional two-stage EBG filter consists of low- and high-stage, where the odd-mode impedance is maintained constant while the even-mode impedance alternates between the low- and high-stage. The low-stage length is set to a quarter wavelength at the target frequency, and filtering is achieved through a half-wavelength phase difference caused by reflection [2].

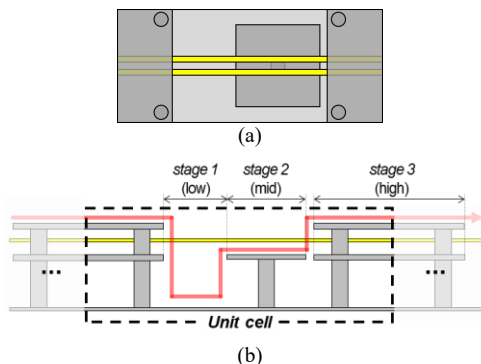


Fig. 1. Proposed AMS-CMF structure.
(a) Top view of the unit cell; (b) Cross-sectional view of the unit cell.

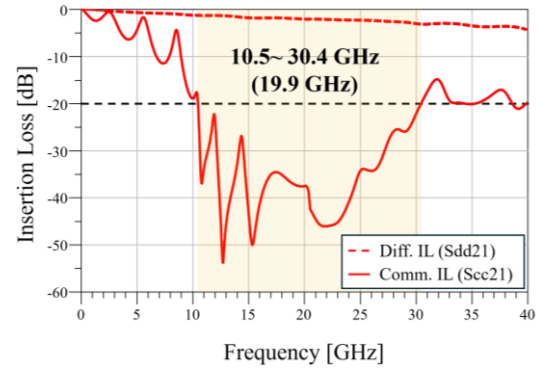


Fig. 2. Filter characteristics of the proposed filter structure in the frequency domain.

The proposed AMS-CMF introduces a mid-stage between the low- and high-stages. This mid-stage induces a parallel LC resonance, enabling deep attenuation of CM currents while introducing an additional admittance term, Y_{mid} . From the perspective of Bloch theory, this additional term increases the magnitude of the Bloch equation, resulting in stronger stopband attenuation than in the conventional two-stage structure. In addition, to mitigate the length penalty inherent in cascaded EBG filters, the unit-cell length is reduced and compensated by width expansion, thereby achieving miniaturization without performance degradation.

III. CONCLUSION

The proposed AMS-CMF was evaluated through 3-D electromagnetic simulations. By incorporating the mid-stage, the proposed filter enables a compact PCB implementation without requiring discrete components. Consequently, the proposed filter achieves broadband CM noise suppression from 10.5 to 30.4 GHz (19.9-GHz bandwidth) while maintaining a differential insertion loss of -4.4 dB at 40 GHz.

ACKNOWLEDGMENT

This work was supported by ANSYS Korea and IDEC (EDA tool provision), and by the Korean government through IITP/MSIT (Grant No. 2020-0-00839).

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An Octave Bandwidth 400Watts Planar Gysel Combiner for C-Band Applications

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Abstract—In this paper, a 2:1 planar Gysel combiner operating from 3-6 GHz frequency range is presented. This combiner is capable of handling high power combination up to 400 Watts while offering a wider fractional bandwidth, improved return loss performance and a compact form factor compared to commercially available designs.

It can be noted that the insertion loss is less than 3.8 dB from 2.6 GHz to 6 GHz reaching more than one octave bandwidth operation.

I. INTRODUCTION

The Gysel power combiner is widely used in high power applications because of its enhanced power handling capability and improved thermal performance compared to Wilkinson power combiners [1]. In the Wilkinson design, the isolation resistor is embedded within the network, which can limit heat dissipation. In contrast, the Gysel configuration employs two isolation resistors connected to ground, allowing more effective thermal management. These resistors can be implemented as high-power terminations, making the Gysel combiner particularly well suited for high-power operation.

II. ARCHITECTURE

Gysel combiner typically employs quarter-wave transmission lines with higher impedance values based on standard design equations. This will result in narrow bandwidth and lower power handling. In the proposed design, the impedances of the conventional Gysel combiner are optimized to enhance power handling capability and wideband performance. This approach differentiates it from the traditional configuration by prioritizing thermal reliability, compactness, power handling capacity and wider bandwidth without compromising impedance matching. The input/output return loss achieved are better compared to other wideband Gysel combiners in the literature [2] while maintaining the insertion loss to the minimum.

III. RESULTS

The combiner was designed using Cadence AWR software for the schematic and electromagnetic simulation. The design was realized on a 60 mils Rogers 6035HTC substrate with 70um copper thickness.

The assembled and tested Gysel combiner is shown in Figure 1. The S-parameters were measured using a 4-port Rhode & Schwarz ZNA43 vector network analyser. The measured results are shown in Figure 2.

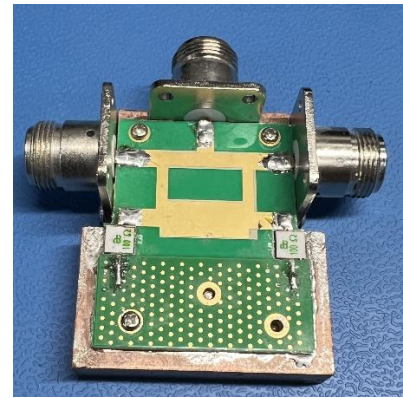


Figure 1: Assembled 2:1 Gysel combiner

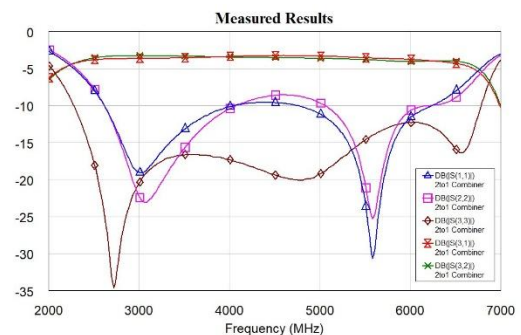


Figure 2: Measured Return loss and Insertion loss

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Analyzing Radiation Characteristics of an HPEM Source based on the IDM

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Abstract—The infinitesimal dipole modeling (IDM) is utilized to analyze radiation characteristics of a high-power electromagnetics (HPEM) source. Infinitesimal dipoles are extracted from a set of measured E-field and E-field at arbitrary points can be calculated accurately, which shows that the IDM is superior than fast Fourier transform-based method. The IDM is helpful in analyzing HPEM source's radiation characteristics and effective range.

Keywords— Infinitesimal dipole modeling (IDM), high power electromagnetics, near-field analysis

I. INTRODUCTION

In order to evaluate the performance of a high-power electromagnetics (HPEM) source, analyzing its radiation characteristics at its operating bandwidth is necessary. An HPEM source often includes a large-scaled antenna. Therefore, the radiation pattern is often achieved by measuring the near-field and applying the fast Fourier transform (FFT).

Analyzing the near-field radiation characteristics of an HPEM source is important, because an HPEM effect testing is often done in the near-field region as the Fraunhofer distance of an HPEM source's antenna can be substantially large. However, it is demanding to evaluate an HPEM source's performance using the FFT-based method as the spatial resolution and the calculation window size are fixed to the initial sampling parameters [1].

This paper demonstrates the analysis of radiation characteristics of a HPEM source based on the infinitesimal dipole modeling (IDM). Based on the IDM, radiated E-field at an arbitrary position can be calculated with greater resolution and spatial range than the FFT-based method. Moreover, an HPEM source's effective range can be predicted more accurately in the near-field region.

II. METHODOLOGY

Through the IDM, the E-field radiated from an AUT is described as

$$\mathbf{E}(\mathbf{r}) = \frac{i}{4\pi\omega\epsilon_0} \sum_{n=1}^N \left[k^2 ((\mathbf{n}_n \times \mathbf{p}_n) \times \mathbf{n}_n) \frac{e^{ikR_n}}{R_n} + (3\mathbf{n}_n(\mathbf{n}_n \cdot \mathbf{p}_n) - \mathbf{p}_n) \left(\frac{-ik}{R_n} + \frac{1}{R_n^2} \right) \frac{e^{ikR_n}}{R_n} \right] \quad (1)$$

where $\mathbf{p}_n$ describes n -th complex electric dipole moment vector and k is the wavenumber [2]. $\mathbf{n}_n$ represents the unit vector from n -th ID at $\mathbf{r}'_n$ to the observation point $\mathbf{r}$ and R_n is $|\mathbf{r} - \mathbf{r}'_n|$. As the IDM demands significant computational resources, applying matrix calculation is necessary for efficiency. The measured E-field can be represented as $\mathbf{E}_{meas} = \mathbf{G}\mathbf{p}$ where $\mathbf{G}$ is the dyadic green's function (DGF) matrix and $\mathbf{p}$ is the matrix of electric dipoles at fixed positions, respectively. Here, the optimized electric dipoles can be extracted as $\tilde{\mathbf{p}} = \argmin (|\mathbf{E}_{opt} - \mathbf{E}_{meas}|^2)$ using the least square method where $\mathbf{E}_{opt} = \mathbf{G}\tilde{\mathbf{p}}$. Then, $\mathbf{E}_a$ radiated by the AUT at an arbitrary observation point can be calculated as $\mathbf{E}_a = \mathbf{G}_a\tilde{\mathbf{p}}$ where $\mathbf{G}_a$ represents the DGF between the observation point and the IDs.

We assume that the antenna aperture lies on the center of the xy plane where the antenna is heading towards the z axis. Fig. 1 shows the x component's normalized magnitude of $\mathbf{E}_{meas}$ and extracted $\tilde{\mathbf{p}}$ for a sample HPEM source antenna, where only the x -component of the E-field ($\mathbf{E}_{meas,x}$) was originally measured. The diameter of the antenna is about 28λ where λ is the operating wavelength. The correlation coefficient between $\mathbf{E}_{meas}$ and $\mathbf{E}_{opt}$ is verified as 0.995.

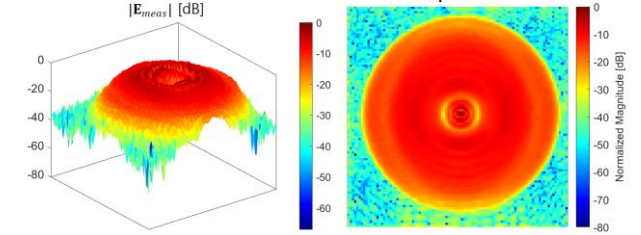


Fig. 1. $|\mathbf{E}_{meas}|$ and $|\tilde{\mathbf{p}}_x|$ of the sample HPEM source antenna.

Fig. 2 describes the calculated $|\mathbf{E}_x|$ along the xz plane towards the propagating direction where the transmitting power of the HPEM source is 1 GW. Visualizing the near-field distribution of the HPEM source and can be helpful in setting design parameters such as beamwidth of the HPEM source and predicting its effective range.

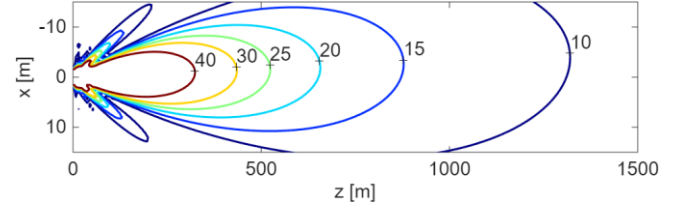


Fig. 2. Contour of $|\mathbf{E}_x|$ [kV/m] along the xz plane.

IV. CONCLUSION

In this paper, near-field radiation characteristics of an HPEM source is analyzed based on the IDM. The IDs are extracted from an E-field measurement result where they are used to re-radiate the E-field at an arbitrary observation point. Calculation results show that through the IDM, the near-field radiation characteristics can be accurately analyzed beyond the spatial range and resolution where the conventional FFT-based analysis is applicable. The IDM can be helpful in analyzing the near-field of an HPEM source and its effective range.

ACKNOWLEDGMENT

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Propagation Loss Prediction Using the 3D Parabolic Equation Method

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Abstract— We present a 3D parabolic equation (3DPE) model for modeling electromagnetic (EM) propagation. The model structure is based on the Split Step Fourier Transform (SSFT). It is compared to a full-wave Finite Integral Technique (FIT) simulation and Intelligent Ray Tracing (IRT) to identify its benefits and limitations. The 3DPE model is shown promising for accurate and efficient simulation of propagation in complex (urban/indoor) scenarios. It outperforms IRT in both accuracy and efficiency, and generates path gain predictions over 100s of meters in just 10s of seconds. The model is particularly well suited for lower frequencies, up to about 2 GHz, as it does not scale well with increasing frequencies.

Keywords— propagation modeling, 3D parabolic equation, split step Fourier transform

I. INTRODUCTION

Numerical models of electromagnetic (EM) propagation give insight in the efficient deployment of high power EM (HPEM) effectors, or in nuclear EM pulse (NEMP) propagation inside protective structures, such as bunkers. However, accurate and efficient numerical modeling of EM propagation in a complex (urban) environment is a computationally demanding task.

Several approaches exist, each with their own (dis)advantages. The parabolic equation (PE) model balances complexity and accuracy, although commonly only applicable in 2D situations. At the cost of increased complexity, this approach can also be used to model lateral effects, allowing applicability in complex urban scenarios [1]. However, because of fundamental limitations, a 3D version of the PE model (3DPE) is still restricted to paraxial propagation, limiting the accuracy in scenarios with a high number of reflections. In this paper, a 3DPE model specifically built for such a complex urban scenario is proposed, implemented such that the intrinsic limitations of the PE approximation are counteracted.

II. METHODOLOGY

The propagation problem under consideration consists of an EM field propagating in the x -direction, and is discretized in steps of length Δx . The method starts by marching in positive x -direction until the end of the domain. At each discretized x -point, the field is calculated over the corresponding yz -plane, taking the effects of terrain, objects, and environmental conditions into account.

At every interaction with the environment, the incident field at that point is stored. At the end of the domain, the propagation direction reverses, marching the field backward. The stored incident fields from before are now used as initial field distributions for the backward propagation, after multiplication with the corresponding reflection coefficients. This procedure repeats until a user defined error threshold is met, yielding the final total field distribution.

The initial field is provided by enforcing a field distribution in the yz -plane, derived from a mathematical expression of the source, a detailed numerical simulation, or simply a plane wave (for NEMP). To solve the discretized propagation equation, the split step Fourier transform (SSFT) method was chosen, which offers a highly efficient solution to PE by using powerful Fourier algorithms. The

PE is rewritten such that all refractive effects and obstacle reflections are computed in the spatial domain, and all diffractive effects and boundary conditions in Fourier domain. These operations are applied separately, at half step interval, which results in the propagation geometry as given in Fig. 1.

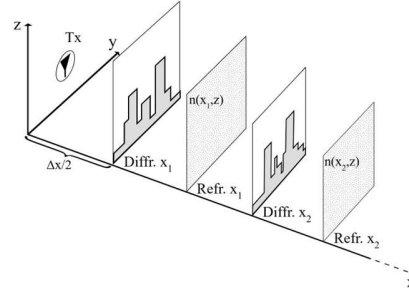


Fig. 1. Schematic depiction of the propagation geometry used for the SSFT method. Refractive and diffractive effects are split into separate (half) steps.

III. RESULTS AND DISCUSSION

Modeled propagation in a $285 \times 270 \text{ m}^2$ urban environment at 600 MHz is shown in Fig. 2. The result was generated in 95.4 seconds, which is over 20 times faster than an intelligent ray tracing (IRT) algorithm. 3DPE offers superior modeling of diffractive effects, and comparably accurate reflection and refraction. We found that IRT scales better with frequency. This makes 3DPE especially useful in applications where fast calculation is preferred at the cost of a slight decrease in accuracy, at frequencies up to about 2 GHz.

In general, accurate results are achieved already after 3–5 forward-backward iteration cycles, as discussed earlier.

The source is modeled as a two-sector antenna, to bypass the paraxial limitation of 3DPE. Full 360° coverage would involve iterating over multiple paraxial directions, and combining the results afterwards.

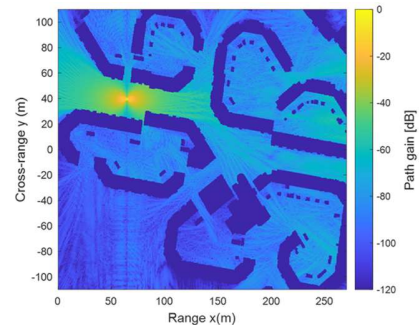


Fig. 2. : Modeled path gain using the 3DPE model, in an OpenStreetMap environment, using $H_{Rx} = H_{Tx} = 1 \text{ m}$, at 600 MHz. The resolution is $0.5 \times 0.5 \times 0.5 \text{ m}^3$. The source is modeled as a two-sector antenna.

IV. CONCLUSION

A 3DPE propagation model was developed and tested. In an urban environment, at 600 MHz, it executes 20 times faster than IRT, in 10s of seconds. It is especially useful when speed is more important than accuracy, at frequencies up to about 2 GHz.

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AI-Assisted Design and Optimization of Thin Broadband Electromagnetic Absorbing Coatings for EMC Applications

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Abstract – This work presents an AI-assisted approach for designing thin broadband electromagnetic absorbing coatings for electromagnetic compatibility (EMC) applications. Composite materials containing various ferromagnetic fillers dispersed in polymer matrices were experimentally characterized using coaxial measurements, and the Nicolson–Ross–Weir method was applied to extract complex permittivity and permeability. The measured parameters were incorporated into a surrogate-model-based optimization framework to determine optimal layer thicknesses and material combinations. The resulting three-layer coating (total thickness 3.38 mm) achieves a reflection loss better than -5.10 dB across the 2–18 GHz range, with a minimum of approximately -12 dB near 4 GHz.

Keywords – electromagnetic absorbers, EMC, AI-assisted optimization, ferromagnetic composites, multilayer coatings

I. INTRODUCTION

Unwanted electromagnetic reflections and resonances inside compact electronic systems can degrade electromagnetic compatibility and increase susceptibility to interference [1],[2]. Thin magnetically lossy coatings are widely used to mitigate such effects by attenuating reflected electromagnetic waves [3],[4]. However, achieving broadband performance under strict thickness constraints remains a challenging design task. This study proposes an AI-assisted framework for optimizing multilayer absorbing coatings using experimentally measured material parameters.

II. METHODOLOGY

Composite samples containing various ferromagnetic fillers dispersed in polymer matrices were fabricated and characterized. Frequency-dependent electromagnetic properties were measured using coaxial transmission line techniques. The Nicolson–Ross–Weir method was applied to extract complex permittivity and permeability from measured S-parameters. The obtained material parameters were used to calculate reflection coefficients of multilayer structures. A Python-based surrogate model and optimization routine were implemented to efficiently explore layer thickness combinations and material selections under a total thickness constraint.

III. RESULTS AND DISCUSSION

The optimization procedure identified a three-layer configuration with a total thickness of 3.38 mm. The optimized structure achieves a reflection loss better than -5.10 dB across the entire 2–18 GHz

frequency range, while the minimum reflection reaches approximately -12 dB near 4 GHz, as shown in Fig. 1.

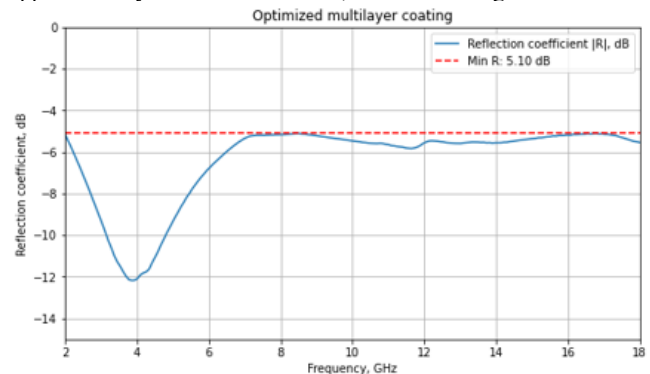


Fig. 1. Simulated reflection coefficient of the optimized three-layer absorbing coating.

These results demonstrate that combining measured electromagnetic material parameters with AI-assisted optimization enables efficient design of thin broadband absorbing coatings suitable for EMC mitigation.

IV. CONCLUSION

This study demonstrates an AI-assisted methodology for designing thin multilayer electromagnetic absorbing coatings using experimentally characterized material parameters. The optimized three-layer structure satisfies the target requirement of reflection loss below -5 dB across 2–18 GHz while maintaining a total thickness below 5 mm. The proposed approach provides a practical framework for developing compact absorbing materials for EMC applications.

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Resonant Path Statistics of Pre-Percolated Random Fiber Networks and Implications for E1 EMP Shielding Bands

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Abstract– We investigate sub-percolated steel fiber reinforced concrete (SFRC) as a topology-defined resonant medium relevant to E1 EMP attenuation. A 3D random fiber microstructure is mapped to a graph representation, from which resonance path lengths (L) and contact-hop counts (N) are statistically extracted. The results reveal a consistent structural scaling between L and N , alongside a systematic shift of the L distribution with varying fiber volume fractions (V_f). By mapping these topological statistics to the Lorentz oscillator model, we provide a structure-derived estimate of resonance-band relocation.

Keywords– Electromagnetic Pulse (EMP), Steel Fiber Reinforced Concrete (SFRC), Lorentz Oscillator Model.

I. INTRODUCTION

Mitigating high-frequency electromagnetic pulses, particularly the fast-rise E1 component, requires understanding broadband attenuation mechanisms in structural materials.[1] Steel Fiber Reinforced Concrete (SFRC) below the geometric percolation threshold ($V_c \approx 0.3\%$ [2]) does not form a continuous conductive network. In this pre-percolation regime, disconnected fiber clusters behave as localized dipole-like resonant elements rather than bulk conductors. Since the E1 EMP spectrum extends into the GHz range, identifying the structural parameters that determine resonance-band location is important for frequency-selective attenuation design.

II. METHODOLOGY

Randomly distributed continuous fibers (length 19.5 mm, diameter 0.1 mm) within a 100^3 mm^3 host matrix are converted into a 3D node-edge graph. For each isolated cluster, the resonance path length (L) is defined as the longest conductive path, representing the dominant electrical length of a dipole-like segment. The number of contact hops (N) within the path is also extracted. Data were aggregated from 30 independent structural realizations for $V_f = 0.05\%$ and 0.10% . For a finite conductive segment embedded in a dielectric host, the fundamental resonance frequency scales inversely with its effective electrical length. Accordingly, the macroscopic response may be described using the Lorentz oscillator formulation [3]:

$$\epsilon_{eff}(\omega) = \epsilon_{\infty} + \sum \frac{\omega_p^2}{\omega_0^2 - \omega^2 + j\omega\gamma} \quad (1)$$

Within this topology-based interpretation, the structural parameters provide physical insight into the Lorentz terms: the path length determines the fundamental resonance frequency ($\omega_0 \propto 1/L$), while the number of contact hops contributes to resistive loss and thus influences the damping factor (γ).

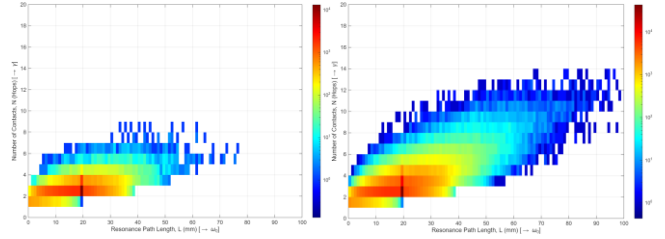


Fig. 1. 2D histograms of resonance path length (L) versus contact hops (N) extracted from 30 realizations. (Left) $V_f = 0.05\%$. (Right) $V_f = 0.10\%$. The statistical expansion of L and N suggests a downward shift of ω_0 and increased damping tendency.

III. RESULTS AND DISCUSSION

The statistical extraction reveals a clear structural scaling between topology and resonance potential. Fig. 1 presents the 2D histograms of L and N for $V_f = 0.05\%$ and 0.10% . At $V_f = 0.05\%$, clusters are predominantly short with fewer hops. As V_f increases to 0.10% , the probability density broadens and shifts toward longer L and higher N . According to the inverse-length scaling of ω_0 , this structural shift suggests a downward movement of the resonance band toward lower GHz frequencies, overlapping with portions of the E1 EMP spectrum. The increase in N indicates enhanced resistive dissipation, which may contribute to greater attenuation near the resonance band. Furthermore, the increase in contact hops (N) translates to higher contact resistance, providing a direct physical basis for the damping factor (γ) in the Lorentz model. This topology-to-resonance mapping allows us to extrapolate the macroscopic shielding behavior of massive SFRC structures without relying on computationally explosive full-wave electromagnetic simulations.

IV. CONCLUSION

Pre-percolated SFRC networks exhibit statistically defined resonant scaling governed by topology. The proposed topology-to-resonance mapping provides a compact predictive framework for estimating EMP-relevant frequency band behavior without requiring full-wave electromagnetic simulations.

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Contact-Resistance-Dependent Conductivity Scaling in 3D Random Fiber Networks for EMP Shielding Applications

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Abstract— The electrical conductivity of random fiber networks is fundamentally constrained by the intrinsic fiber resistance and the inter-fiber contact resistance. We investigate the macroscopic DC conductivity (σ_{eff}) of 3D steel fiber reinforced concrete (SFRC) networks by treating the microscopic contact resistance (R_c) as an independent variable. Our computational results reveal a distinct saturation-to-inverse-linear crossover behavior. Based on this observation, we propose a compact mathematical model predicting the entire conductivity response from a single structural calculation. This structurally derived σ_{eff} provides a reliable parameterization for the Drude model, essential for evaluating broadband shielding effectiveness against electromagnetic pulse (EMP) threats.

Keywords— Electromagnetic Pulse (EMP), Steel Fiber Reinforced Concrete (SFRC), Contact Resistance, Effective DC Conductivity, Drude Model, Resistor Network.

I. INTRODUCTION

Steel Fiber Reinforced Concrete (SFRC) is widely utilized for electromagnetic pulse (EMP) shielding [1]. Because EMPs, such as High-Altitude EMP (HEMP), exhibit broad high-frequency spectrum, analyzing the broadband electromagnetic response of SFRC is crucial [2]. Above the percolation threshold, a continuous conductive network forms, whose macroscopic DC conductivity—governed by intrinsic fiber resistance and junction contact resistance (R_c)—dictates this broadband shielding performance. Because evaluating complex 3D random networks requires extensive computational resources, we propose a compact, predictive analytical model based on R_c . This approach significantly reduces the computational burden while directly providing a crucial parameter for effective medium theories such as the Drude model.

II. CONDUCTIVITY MODELING

To extract effective conductivity, randomly distributed steel fibers (length $L_f = 19.5$ mm, diameter $D_f = 0.1$ mm) are mapped onto a 3D node-edge resistor network within a $(100 \text{ mm})^3$ size host matrix [3]. By treating R_c as a variable while fixing the structural topology, we observed that effective conductivity follows a distinct inverse-saturation curve. We model this behavior with a compact analytical equation:

$$\sigma_{eff}(V_f, R_c) = \frac{1}{\alpha(V_f)R_{fiber} + \beta(V_f)R_c} \quad (1)$$

where $\alpha(V_f)$ and $\beta(V_f)$ are structure-dependent scaling factors, and R_{fiber} is the resistance of a fiber.

The key advantage of Eq. 1 is its predictive capability. Once the geometric factors α and β are determined from a single computational run of a given network structure, the entire conductivity response across any arbitrary R_c values can be analytically predicted without requiring repeated 3D simulations.

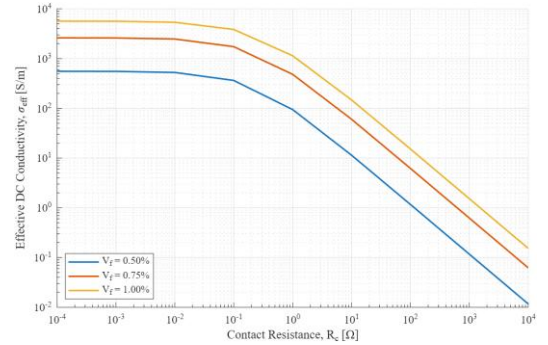


Fig. 1. Effective conductivity versus contact resistance, validating the proposed predictive model across varying V_f .

III. RESULTS AND DISCUSSION

The proposed model was validated through network simulations across different volume fractions. As shown in Fig. 1, the normalized conductivity demonstrates universal crossover behavior. At low R_c , the conductivity saturates due to the intrinsic limit of R_{fiber} . As R_c increases, the overall network conductivity becomes contact-resistance-limited, exhibiting a linear decay with a slope of -1 on a log-log scale. This compact scaling model elucidates the physical bounds of network conductivity and provides a crucial link to macroscopic material modeling. For EMP hardening, predicting broadband performance is paramount. The analytically predicted σ_{eff} can be directly utilized as the primary DC parameter in the Drude formula ($\epsilon_{eff} = \epsilon_\infty - j\sigma_{eff}/\omega\epsilon_0(1 + j\omega\tau)$) to efficiently evaluate the high-frequency shielding properties of the composite.

IV. CONCLUSION

We demonstrate that the macroscopic DC conductivity of 3D random fiber networks exhibits a distinct saturation-to-inverse-linear crossover dependent on contact resistance. This compact scaling model directly parameterizes the Drude formula, enabling highly efficient broadband EMP shielding evaluations for SFRC infrastructures.

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A MV-level Pulsed Current Injection Testing Platform for Overhead Power Lines

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Abstract—Many researchers have done conducted immunity tests for EUT on overhead power lines with different testing circuits, which differ in involving a pulser with the parallel impedance or with the series impedance. This paper presents the pulsed current injection method with a long coupling cable, which reduce the waveform distortion caused by EUT and obtain more accurate results. A MV-level PCI testing platform is built to validate the method.

Keywords— pulse current injection, HEMP

I. INTRODUCTION

Pulsed current injection (PCI) technique is commonly used for conducted immunity testing of electrical equipment and electronic systems under the high-altitude electromagnetic pulse (HEMP). Overhead power line systems are proved to be severely threatened by HEMP^[1]. For testing the conducted immunity of EUT along overhead power lines, many researchers directly used the PCI method by connecting the EUT in series to a lumped source impedance. Many devices in practical power line systems are installed without protective devices. If this PCI method is directly applied, the injected pulse waveform may be distorted due to the impedance of the EUT. For high-impedance devices like insulators, it may further alter the operating state of pulsers and even render the tests unfeasible. Furthermore, the characteristic impedance of overhead power lines can be estimated as about 400 Ω . Since the induced current is up to several kA, the EUT on the overhead lines is threaten by the MV-level conducted HEMP^[2].

This paper presents a MV-level PCI testing platform based on a long- coupling cable. Output current waveforms of the system are measured, and the experimental results of 10kV EUT are studied.

II. METHODOLOGY

The PCI test circuit is shown in Fig. 1. The series impedance of the pulser is replaced by a long cable with the characteristic impedance Z_c equal to the irradiated cable. If the length of the coupling cable is long enough compared with the width of the injected pulsed current, the reflection from the terminal EUT will be very small. It can reduce the waveform distortion. And the pulsers would not be affected by high-impedance EUT.

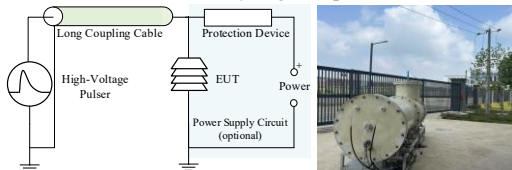


Fig. 1. Circuit and the picture of the PCI test platform with the long coupling cable.

A MV-level PCI testing platform is built with 200m-long coupling cables. The pulser is composed of a multi-stage Marx

generator, DC sources and a pulse sharpener. After the reshaping by the pulse sharpener, which is composed of a small capacitor and a peaking switch, the output voltage can have a nanosecond-level rising edge. The pulser can generate the MV-level double-exponential voltage. The source impedance is designed as 400 Ω , corresponding to the impedance of the 10-kV overhead line.

The output current waveforms of the pulser with a 400 Ω series resistor and with a long coupling cable are measured and compared, as shown in Fig.3. The rise time of the current waveform with the series impedance is 15 ns, and the pulse width is 510 ns. As a contrast, the rise time of the waveform with the long coupling cable is 20 ns, and the pulse width is 542 ns. The rise time and the pulse width still meet the requirements. The responses of electrical equipment such as insulators and surge arresters are experimentally investigated based the platform. The impact on EUT is analyzed by comparing different injection methods and voltage levels.

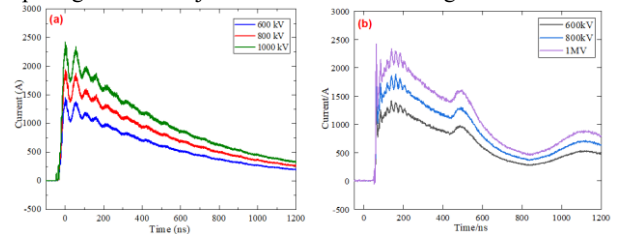


Fig. 2. Output current waveforms of the pulser (a) with a 400 Ω series lumped resistor and (b) with a long coupling cable.

IV. CONCLUSION

PCI technique has been introduced in several standards as a conducted immunity testing method. This paper investigates the MV-level PCI testing platform with long coupling cables. This test method enables simultaneous voltage/ current studies on EUT with arbitrary impedance. In contrast, the PCI method with the series resistor can only be used to obtain current effect results when the short-circuit current equals to the injected current through EUT. If the output voltage of the pulser is not high enough, the method will be limited for the high-impedance EUT.

ACKNOWLEDGMENT

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Equivalent Circuit Modeling of a Differential Switched Oscillator Based on Frequency Response Measurements

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Abstract– This paper presents a circuit-level model for the simulation of Differential Switched Oscillators (DSWO) using LT-SPICE. The proposed framework integrates an established nonlinear spark gap model from the literature with distributed transmission line parameters and an RLC-equivalent network for antenna representation. The model was validated through a direct comparison between simulation results and experimental measurements. A high degree of agreement was achieved, with a resonant frequency deviation of only 3% between measured and simulated data. These results confirm the model's effectiveness as a low-computational cost, reliable tool for the design and performance estimation of High-Power Electromagnetic (HPEM) sources.

Keywords– Differential Switched Oscillator, Frequency, Monopole Antenna

I. INTRODUCTION

High-Power Electromagnetic (HPEM) sources, particularly Switched Oscillators (SWO), are widely used in intentional EMI studies due to their capability to generate high electric field strengths. The Differential SWO (DSWO) provides higher radiated power, improved stability, and reduced dielectric stress compared to monopolar designs, since the charging voltage to ground is halved.

Although introduced by Carl E. Baum [1], DSWO designs still rely heavily on computationally intensive full-wave simulations. This paper presents a validated circuit-level model of a DSWO integrated with a redesigned monopole antenna in the lower UHF band, incorporating a nonlinear spark-gap and RLC equivalent network to accurately predict transient and frequency-domain behaviour for faster system optimization.

II. DSWO + ANTENNA MODEL

The DSWO was evaluated using both experiments and simulations. Experimentally, a ± 20 kV source powered the oscillator, and radiated fields were measured with a 10 GHz D-dot sensor at 3.40 m. Numerically, a circuit-level LTSPICE model extended a validated single-ended design to a differential topology, modelling the monopole antenna as an RLC network based on measured Q-factor and resonance, and including electrode coupling via capacitance across the spark gaps [2]. A spark gap switch model utilizing sulfur hexafluoride (SF₆) as an insulating gas at varying pressure levels is integrated into the proposed model. The schematics of circuit model is depicted in Fig. 1. Each electrode branch follows the configuration reported in [3], with additional coupling elements to account for inter-electrode interaction. Simulated resonance frequencies were then compared with experimental results for validation.

III. RESULTS AND DISCUSSION

The DSWO frequency response was obtained by applying a Fast Fourier Transform (FFT) to measured and simulated transients. Experimental resonance occurred at 473 MHz, while the LTSPICE model predicted 460 MHz, a $\sim 3\%$ difference likely due to parasitic inductances and environmental reflections. The close agreement of peaks confirms that the RLC-equivalent network and nonlinear spark-gap accurately capture the system's dominant oscillation mode.

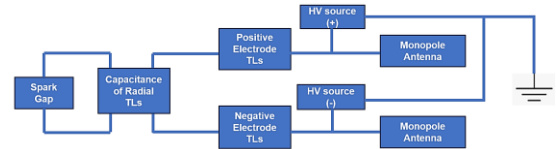


Fig. 1. The schematics of circuit model of the Differential SWO (DSWO)

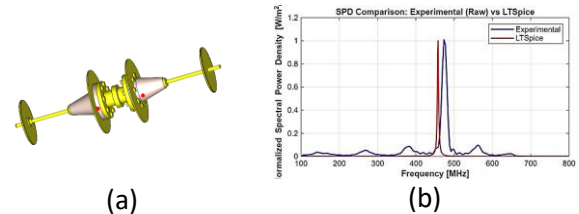


Fig. 2. a) The geometry of DSWO b) Radiated spectral density for both experimental setup and equivalent circuit.

IV. CONCLUSION

This work validates a circuit-level model of a Differential Switched Oscillator with a monopole antenna. Measured and simulated results show strong agreement, with only a 3% error in resonant frequency. The discrepancy in the minor frequencies is attributed to reflections occurring during chamber testing. Additionally, radiation loss is not accounted for in the proposed model. The LT-SPICE framework provides a computationally efficient and reliable tool for rapid design and optimization of HPEM sources, offering a practical alternative to full-wave simulations. Future work will involve improving the circuit model and conducting further tests.

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Experimental Validation of a 142/208 GHz Dual-Frequency Gyrotron

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Abstract– This paper presents the development and experimental characterization of a dual-frequency, kW-class gyrotron for electron cyclotron resonance heating and current drive (ECRH/ECCD). The system operates at 142 GHz (TE_{7,2} mode) and 208 GHz (TE_{9,3} mode). The research followed a systematic verification process, integrating internal mode converter design, low-power cold testing, and pulsed high-power hot tests. Experimental results yielded output powers of ~42 kW at 142 GHz and ~34 kW at 208 GHz. The spatial evolution of the output beam radius was analyzed along the axial distance z , showing that the manufactured quasi-optical system effectively produces high-quality Gaussian beams suitable for fusion plasma applications.

Keywords– dual-frequency gyrotron; ECRH/ECCD; quasi-optical mode converter

I. INTRODUCTION

To achieve efficient plasma heating and current drive in next-generation fusion devices like K-DEMO, ECRH/ECCD performance is typically assessed over a broad frequency range, motivating microwave sources with frequency flexibility [1]. Dual-frequency gyrotrons address this need by supporting multiple resonance conditions within a single device. This study focuses on the experimental validation of a 142/208 GHz laboratory-scale gyrotron. The primary objective is to confirm that the cavity and quasi-optical mode-converter designs are faithfully realized in the prototype, delivering consistent beam quality and power performance at both operating frequencies.

II. METHODOLOGY

The development of the dual-frequency gyrotron involved three integrated stages: numerical design, hardware verification, and operational testing. The development of the dual-frequency gyrotron involved three integrated stages: numerical design, hardware verification, and operational testing. First, the cavity was optimized to support TE_{7,2} and TE_{9,3} modes using the UNIST Gyrotron Development Tool (UGDT) [2]. A Denisov-type launcher with specific surface perturbations and a three-mirror quasi-optical system were designed and optimized using the Surf3D code [3]. Surf3D is a specialized numerical tool used for the synthesis and optimization of quasi-optical reflectors. For low-power hardware verification (cold test), a custom high-order mode generator was fabricated to replicate the cavity modes. This allowed for the precise alignment and performance check of the launcher and mirrors without thermal complexities. Following cold-test validation, the system was integrated into a high-power test stand for hot testing. The gyrotron operated under pulsed conditions (20 μ s). Output power was calculated by applying the Friis transmission equation to signal levels measured via a Schottky diode. Beam profile measurements were captured using Liquid Crystal Paper (LCP) and diode scanning at varying axial positions to evaluate propagation characteristics.

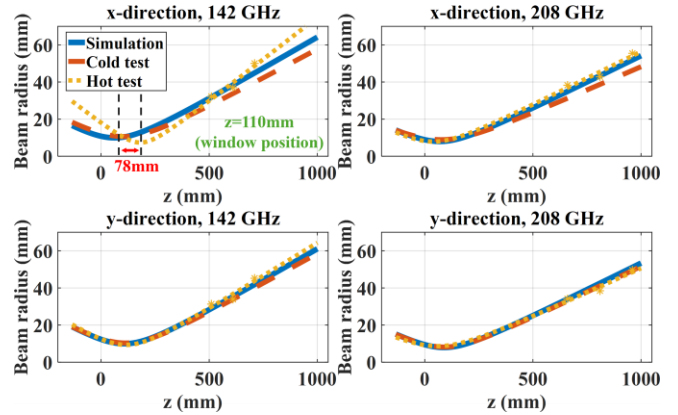


Fig. 1. Comparison of experimental and simulated beam radius propagation for 142 GHz and 208 GHz.

III. RESULTS AND DISCUSSION

The gyrotron achieved stable microwave oscillations at both frequencies. Under a cathode voltage of 34 kV and a beam current of 7 A, stable operation was obtained at 5.43 T for 142 GHz and 7.78 T for 208 GHz, yielding measured output powers of ~42 kW at 142 GHz and ~34 kW at 208 GHz, respectively, which are within the expected range for the present operating conditions. Beam propagation was evaluated by extracting the beam radius as a function of the axial distance z , as shown in Fig. 1, and the measured beam-radius curves follow the simulated Gaussian expansion trends. The 208 GHz results show close agreement with the design in both axes, while the 142 GHz dataset exhibits a limited deviation, most noticeably in the x-direction at larger z . This discrepancy may arise from alignment-driven operating-point shifts and unintended competing-mode oscillations; nevertheless, the overall propagation trend supports proper operation of the quasi-optical mode converter at both frequencies.

IV. CONCLUSION

The 142/208 GHz dual-frequency gyrotron was validated through pulsed high-power experiments. The measured power levels and beam-radius evolution confirm that the quasi-optical mode converter provides consistent beam formation at both operating frequencies.

ACKNOWLEDGMENT

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Generation of High-voltage Sub-nanosecond Pulses Based on Circuit Power Synthesis

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Abstract– High-voltage sub-nanosecond pulses are widely used due to the high amplitude, narrow pulse width and ultra-wide band. To significantly improve the amplitude limit of the output of avalanche transistorized sub-nanosecond pulsers, a 32-channel circuit power synthesis network composed of Wilkinson power combiner (WPC) and tapered transmission line (TTL) to synthesize the output of 32 positive avalanche transistorized sub-nanosecond pulse modules is developed. Each module generates pulses with a peak of 8.5 kV, a rising edge of 120 ps and a pulse width of 500 ps. The synthesis yields pulses with a peak of 25.5 kV, a rising edge of 410 ps and a pulse width of 660 ps.

Keywords– high-voltage sub-nanosecond pulse, circuit power synthesis, WPC, TTL, avalanche transistorized pulser

I. INTRODUCTION

Based on the impedance transformer theory, a 32-channel circuit power synthesis network composed of WPC and TTL that satisfy high efficiency synthesis of sub-nanosecond pulses is designed^{[1]-[3]}. The proposed network has expanded the operating bandwidth and power capacity. Compared with the limitation of output amplitude (less than 10 kV) of a pulser, the synthesis output have significantly raised the amplitude upper limit of sub-nanosecond pulses^[4].

II. METHODOLOGY

A 2-channel WPC module with seven isolation resistors and quarter-wavelength transmission line stubs, as shown in Fig. 1, is designed with wide operating bandwidth, high isolation and high synthesis efficiency, based on the principle of odd-mode and even-mode equivalent theory [1], [2]. The 4-channel WPC is achieved by cascading three 2-channel modules. Large surface mount resistors are adopted to improve the insulation threshold of WPC.

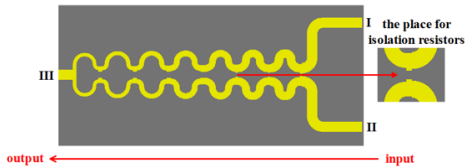


Fig. 1. The top view of proposed 2-channel WPC module.

A 4-channel TTL, as shown in Fig. 2, is designed based on the small-reflection theory [3]. To improve the efficiency, bandwidth and power capacity, the length is 1 m and the thickness is 10 mm. No isolation resistors makes TTL bear higher voltage than WPC.

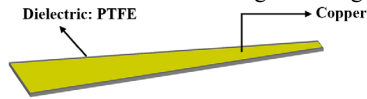


Fig. 2. The model of proposed 4-channel TTL.

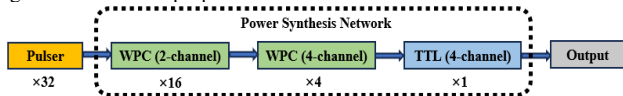


Fig. 3. The structure of proposed power synthesis network.

The structure of proposed 32-channel power synthesis network is

illustrated in Fig. 3. The combiners are interconnected via cables.

III. RESULTS AND DISCUSSION

The key parameters of designed WPC and TTL are simulated first, as illustrated in Table I. Then the measured result of output high-voltage sub-nanosecond pulse of 32-channel power synthesis is tested, as shown in Fig. 4. The output pulses have a peak of 25.5 kV, a rising edge of 410 ps and a pulse width of 660 ps. Due to the insertion loss and pulse jitter, there is still a gap compared to the ideal superposition which is $8.5 \times \sqrt{32} \times 0.95^3 \times 0.85 \approx 34.7$ kV.

TABLE I. KEY PARAMETERS OF POWER SYNTHESIS NETWORK

Table Head	Key Parameter		
	Combiner	Working band	Efficiency
Power Synthesis Network	WPC	0.1 GHz ~ 3.5 GHz	0.95(2-channel)
	TTL	0.5 GHz ~ 3.0 GHz	0.85(4-channel)

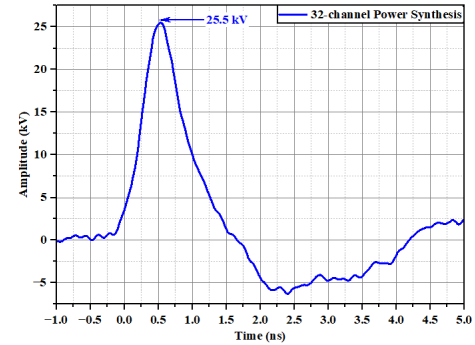


Fig. 4. Time domain result of output pulse from 32-channel power synthesis.

IV. CONCLUSION

The designed 32-channel power synthesis network can generate sub-nanosecond pulses with a peak of 25.5 kV with wide operating bandwidth and high synthesis efficiency. The measured results offer some references to high-voltage sub-nanosecond pulse generation.

ACKNOWLEDGMENT

The authors thank Yinhui Cheng for the technical support and helpful discussions.

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Finite Interaction-Length Effects on Amplification Stability in a W-Band Planar Meander-Line TWT

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Abstract— A W-band planar meander-line traveling-wave tube (TWT) was studied for different interaction lengths. While dispersion analysis predicts beam-wave synchronism near 100 GHz, the amplification behavior of the finite structure also depends on the interaction length. In 3D particle-in-cell (PIC) simulations, a 37-period structure showed stable amplification near 99 GHz with about 14 dB gain, whereas a 50-period structure did not provide further gain increase and showed multiple frequency components with strong harmonic peaks in the output spectrum. This result shows that interaction-length effects are important in planar TWT design, beyond the synchronism condition obtained from periodic dispersion.

Keywords— planar traveling-wave tube (TWT), meander-line slow-wave structure, interaction length, Particle-in-cell (PIC) simulation

I. INTRODUCTION

Planar traveling-wave tubes (TWTs) using meander-line slow-wave structures (SWSs) are promising for compact millimeter-wave vacuum electronic devices [1]. The operating frequency is usually determined from the synchronism condition between the electron beam and the periodic SWS dispersion [1], [2]. However, in practical planar TWTs, the beam-wave interaction is also affected by the finite interaction length and the waveguide-to-SWS coupling structure [3]. Therefore, the amplification condition cannot be determined only from the ideal periodic dispersion. In this work, the effect of interaction length on the amplification behavior and stability of a W-band planar meander-line TWT is numerically investigated.

II. METHODOLOGY

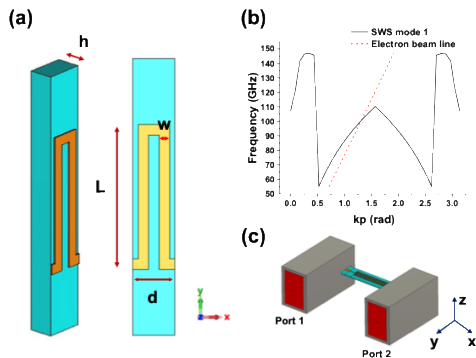


Fig. 1. (a) Unit cell of the planar meander-line slow-wave structure (SWS). (b) Dispersion curve of the SWS and the electron beam line, showing the beam-wave synchronism near 100 GHz. (c) Finite waveguide-coupled SWS structure.

Figure 1 shows the planar meander-line SWS and the

corresponding finite waveguide-coupled configuration. The unit-cell dimensions are $L=0.45$ mm, $d=0.13$ mm (2p), $w=0.0325$ mm, $h=0.08$ mm, and copper thickness $t=0.005$ mm [4]. The eigenmode dispersion indicates a beam-wave synchronism near 100 GHz for a beam voltage of 2.8 kV, and Fig. 1(c) shows the finite waveguide-coupled structure used in the PIC simulations.

III. RESULTS AND DISCUSSION

The PIC simulation results show that the beam-wave interaction strongly depends on the interaction length. For a 37-period structure, stable amplification is obtained near 99 GHz, with a gain of about 14 dB under the present condition. In contrast, when the interaction length is increased to 50 periods, the gain does not increase further. In this case, the output spectrum shows multiple frequency components and strong harmonic peaks. These results indicate that the interaction length significantly influences the amplification behavior of planar TWTs, even when the synchronism condition is satisfied [5].

IV. CONCLUSION

These results indicate that planar TWT performance cannot be determined solely from the synchronism condition obtained from periodic dispersion. Instead, the interaction length of the finite structure plays an important role in determining the amplification behavior. Therefore, interaction-length effects should be carefully considered in the design of planar TWTs. This consideration will become increasingly important for future implementations employing non-ideal electron sources such as field-emission cathodes.

ACKNOWLEDGMENT

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Power Delivery Network Design and Analysis Focusing on EM Information Leakage

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Abstract—In this paper, the design and analysis of the power delivery network (PDN) in semiconductor systems focusing on electromagnetic (EM) information leakage, are conducted. EM fields are measured at the various locations in the PDN to evaluate the leakage. Filtering conditions are also altered to increase the efficiency. Lastly, the PDN design guide is provided, including shielding methods to prevent the EM information leakage.

Keywords— EM Information Leakage, Power Delivery Network

I. INTRODUCTION

Electromagnetic (EM) field radiation from a cryptographic device in semiconductor systems can reveal secret information. Even though tamper-resistant designs are applied to the cryptographic IC, the secret information inside the IC may leak from interconnections in the PCB/Package of the devices via EM field radiation. Among various interconnections, the cryptographic core-PDN is most likely to leak secret information [1]. It is crucial to properly evaluate the EM information leakage from the PDN and provide design guidance.

II. METHODOLOGY

To evaluate the EM information leakage, various devices are fabricated, including PCBs and FPGAs [2]. Correlation electromagnetic analysis (CEMA) has been applied to various locations in the devices, focusing on cryptographic core PDNs (IC-Package-PCB; hierarchical structure), vias, pads, and even decoupling capacitors. For a more accurate evaluation, various filtering window functions are adopted during the CEMA to suppress the irrelevant noise impacts.

It has been verified that exposed PDN leaks information via EM radiation. Noise simulation based on PWL current and hierarchical PDN has been conducted to suppress the EM information leakage. For some cases were found to be difficult to suppress the EM information leakage, and, in these cases, it is better to allocate the PDN to internal layers of the PCB or shield the exposed parts and interconnects.

III. RESULTS AND DISCUSSION

Fig.1 shows EM information leakage evaluation results. Decoupling capacitors are inevitable for the power integrity and EMC design. However, they are mostly placed on the top/bottom of the package and PCB unless we adopt special shields or embedded substrate decoupling capacitors. It is verified that at some decoupling capacitors, measuring the EM field radiation (or even direct contact voltage or current measurements) could retrieve the whole secret key information being processed inside the cryptographic IC. For the cryptographic device, additional evaluation should be conducted, and proper countermeasures must be adopted to in the core PDN.

After various simulations and design modifications, a design that delivers a low correlation coefficient value at the target location in

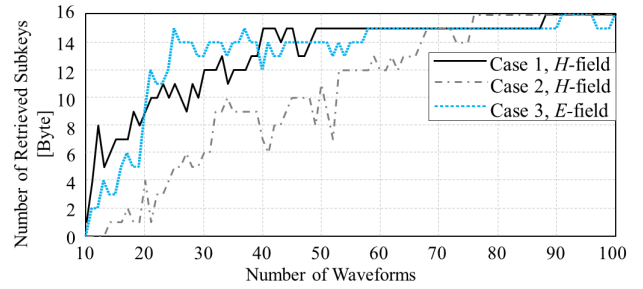


Fig. 1. Evaluation of EM information leakage at various locations in the PDN

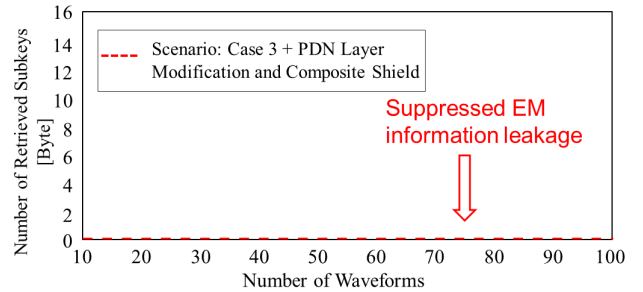


Fig. 2. EM information leakage suppression with the proposed PDN simulation, analysis, and design guidance

the PDN can suppress the information leakage. However, all modifications cannot ensure the leakages. In a such case, additional shielding techniques or changing the PCB layout and material/process should be considered, which is summarized in Fig. 2.

IV. CONCLUSION

It has been validated that the exposed interconnects in the PDN, especially in PCB-level PDN is vulnerable to side-channel attacks using EM fields. The proposed design successfully suppressed the EM information leakage.

ACKNOWLEDGMENT

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Robust Hardware Trojan Detection Against Temperature Variation Using Electromagnetic Side-Channel Signals

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Abstract– This study analyzes the impact of temperature variation on EM signal-based hardware Trojan (HT) detection and proposes a robust detection method. EM signals collected from an kintex 7 fpga across 20°C to 40°C reveal an inverse relationship between temperature and signal amplitude. While temperature-specific models suffer significant performance degradation under mismatched conditions, a Deep SVDD model trained across the full temperature range achieves 100% accuracy and F1-score under all test conditions.

Keywords– Hardware Trojan, Electromagnetic Side-Channel, Temperature Variation, Deep SVDD

I. INTRODUCTION

Hardware Trojans (HTs) are malicious circuits embedded in hardware that pose severe threats to safety-critical systems through globalized supply chains [1]. EM signal-based side-channel analysis has emerged as a promising non-destructive approach for HT detection, and Deep SVDD has demonstrated high detection accuracy even against unknown HTs by learning solely from normal data [2], [3]. However, existing studies largely assume fixed temperature conditions, despite temperature variation directly affecting CMOS device characteristics and EM signal amplitude [4]. This study analyzes the impact of temperature variation on EM-based HT detection and proposes an all-temperature Deep SVDD training strategy for robust detection across temperature-varying environments.

II. METHODOLOGY

Experiments were conducted on a Kintex-7 FPGA, where a self-developed RC car control module was implemented as the golden model (GM) and HTs were constructed from the Trust-HUB benchmark. EM signals were measured using a Langer near-field probe at 50 MHz center frequency, while temperature was varied from 20°C to 40°C in 5°C increments. Signals were transformed to the frequency domain via FFT, and peak values around the operating clock were extracted as features. Two training strategies were compared: a temperature-specific model, in which a separate Deep SVDD model is trained on GM data at each temperature, and an all-temperature model (ALL), which trains a single Deep SVDD on GM data across the full temperature range.

III. RESULTS AND DISCUSSION

As shown in Fig. 1, signal amplitude decreases with increasing temperature for both GM and HT, while a consistent amplitude gap between them is maintained across all conditions. As shown in Table I, temperature-specific models suffer severe performance degradation under mismatched temperature conditions, dropping to Acc. 50.00% and F1 33.33%. In contrast, the ALL model achieves 100% Accuracy and F1-score across all test conditions, demonstrating robust HT detection in temperature-varying environments.

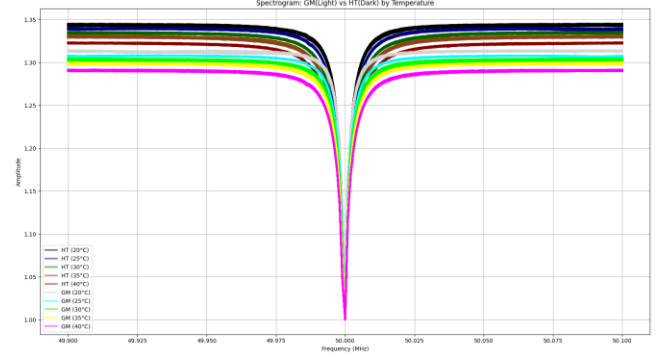


Fig 1. EM signal amplitude of GM and HT across temperatures (20°C–40°C).

Train Temp.	Test 25°C (Acc / F1)	Test 29°C (Acc / F1)	Test 30°C (Acc / F1)
20°C	100 / 100	100 / 100	86.63 / 86.38
25°C	77.00 / 75.72	50.00 / 33.33	50.00 / 33.33
30°C	50.00 / 33.33	100 / 100	100 / 100
35°C	50.00 / 33.33	100 / 100	87.50 / 87.30
40°C	50.00 / 33.33	50.00 / 33.33	75.00 / 73.33
ALL	100 / 100	100 / 100	100 / 100

TABLE I. DETECTION ACCURACY (%) OF EACH MODEL ACROSS TEST CONDITIONS

IV. CONCLUSION

This study demonstrated that temperature variation causes an inverse change in EM signal amplitude, and that a Deep SVDD model trained across the full temperature range achieves robust HT detection with 100% Accuracy and F1-score under all tested conditions, outperforming temperature-specific models.

ACKNOWLEDGMENT

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Efficient Estimation of Irradiation Frequency Pairs for Frequency-Conversion-Based Echo TEMPEST Using Harmonic Responses

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Abstract—In frequency-conversion-based Echo TEMPEST, the search cost for optimal irradiation frequency pairs is a significant challenge. Focusing on the correlation between frequency conversion and harmonic responses caused by equipment nonlinearities, this study proposes an efficient method to estimate frequency pairs from the harmonic characteristics observed during a single-frequency sweep. Fundamental experiments using direct injections demonstrated that the proposed method is in good agreement with the conventional exhaustive search, verifying its effectiveness.

Keywords—electromagnetic information leakage, self-interference.

I. INTRODUCTION

Echo TEMPEST is a threat in which intentional electromagnetic interference (IEMI) causes leakage of internal equipment signals, compromising confidentiality [1]. A major challenge is self-interference between the incident and reflected waves at the target equipment. To mitigate this, a method has been proposed that irradiates the equipment with two frequencies and suppresses self-interference by converting the reflected wave's frequency via internal nonlinear components [2]. However, the effective frequencies depend on transfer functions and nonlinear characteristics, and have conventionally been determined by exhaustively sweeping the two frequencies. In this study, we propose an efficient method to estimate the irradiation frequency pairs that effectively establish frequency-conversion-based Echo TEMPEST, focusing on harmonic responses observed during a single-frequency sweep.

II. FREQUENCY SEARCH USING HARMONIC RESPONSES

Frequency-conversion-based Echo TEMPEST utilizes second-order conversion components, $|f_1 \pm f_2|$, generated by the equipment's nonlinear characteristics when two frequencies, f_1 and f_2 , are applied [2]. The proposed method exploits the fact that frequency conversion under multi-frequency excitation and harmonics under single-frequency excitation originate from the same nonlinear characteristics [3], and estimates the irradiation frequency pairs from the harmonic characteristics observed during a single-frequency sweep.

A frequency f_r with high nonlinear conversion and coupling efficiencies is extracted from the harmonic responses. A frequency spacing Δ (larger than the resolution bandwidth, RBW) is centered at $\frac{f_r}{2}$. The frequency pair is selected as $f_1 = \frac{f_r - \Delta}{2}$ and $f_2 = \frac{f_r + \Delta}{2}$.

To eliminate free-space propagation effects and verify the correlation between frequency conversion and harmonic responses based solely on the equipment's nonlinear characteristics, the evaluation was conducted via direct injection through a coaxial cable.

III. EVALUATION RESULTS AND CONCLUSION

This experiment used a USB module outputting a UART signal as the target device. An SMA connector was soldered to the TX port and connected to a Raspberry Pi 5 in an anechoic chamber. Fig. 1(a) and Fig. 1(b) show the setups for the conventional and proposed methods. The spectrum analyzer measured f_r in zero-span mode (RBW: 3 MHz, 10001 points). The signal generator (SG) swept

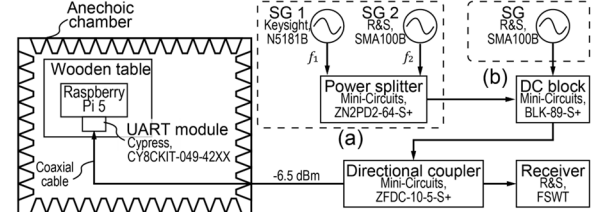


Fig. 1. Experimental setup. (a) Setup for the conventional method. (b) Setup for the proposed method.

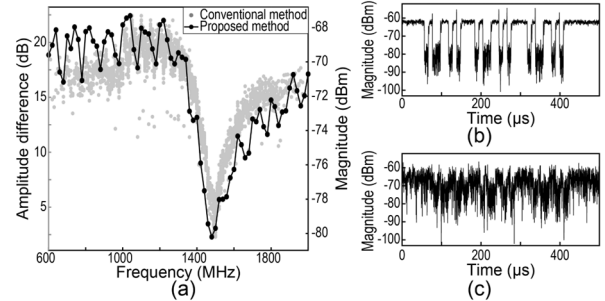


Fig. 2. Evaluation results. (a) Frequency characteristics. (b) Echo TEMPEST result at 1040 MHz (f_1 : 510 MHz, f_2 : 530 MHz). (c) Echo TEMPEST result at 1480 MHz (f_1 : 730 MHz, f_2 : 750 MHz).

300–1000 MHz with a 10 MHz step. The amplitude difference between the UART High and Low levels was calculated. The SG output was calibrated to -6.5 dBm at the directional coupler output.

Fig. 2(a) presents the evaluation results. In this experiment, assuming a nonlinear component where even-order harmonics are strongly observed as a representative element causing frequency conversion, we focused on the second-order harmonic under single-frequency excitation. The frequency characteristics from the conventional exhaustive search agreed well with the second-order harmonic characteristics in the proposed method. Figs. 2(b) and 2(c) show the results of frequency-conversion-based Echo TEMPEST at $f_r = 1040$ MHz and 1480 MHz, where the second-order harmonic characteristics were maximized and minimized, respectively. Echo TEMPEST was efficiently established under the condition where the characteristics were maximized.

These results confirm the proposed method can efficiently estimate the frequency pairs that effectively establish frequency-conversion-based Echo TEMPEST. Building on these findings, future work will evaluate the proposed method's effectiveness in a more realistic free-space propagation environment using antennas.

ACKNOWLEDGMENT

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Design-Oriented Vulnerability Evaluation of Cryptographic Hardware in Side-Channel Attacks

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Abstract– We propose a design-oriented method for evaluating and improving the resistance of cryptographic hardware to side-channel attacks. By deriving correlation-based secure regions and checking whether transfer characteristics stay within them across frequency, the method identifies vulnerable bands and determines allowable impedance and required attenuation. Unlike conventional analysis, which only detects leakage, it provides quantitative design targets. Validation on an FPGA-based AES board confirms its practicality for early-stage, SCA-resilient PCB design.

Keywords– EM information leakage, Cryptographic hardware, Side-channel attack, design-oriented vulnerability evaluation

I. INTRODUCTION

Side-channel attacks (SCAs) pose significant risks to cryptographic hardware, yet evaluating their impact through physical measurements is costly and time-consuming^[1]. Conventional simulation-based approaches mainly identify vulnerable frequency bands but do not provide quantitative design guidance for improving the leakage path. Our recent work^[2] demonstrated that simulated switching-current waveforms and power-distribution-network impedance can reliably predict leakage behavior, enabling vulnerability assessment without physical attacks. Building on this foundation, we present a design-oriented method that uses the required attack success rate to derive correlation thresholds, construct secure regions for transfer characteristics, and determine both allowable impedance and the attenuation needed to meet a specified security level.

II. METHODOLOGY

We propose a design-oriented method that provides a target attenuation for the transfer impedance of the side-channel leakage path, based on the EMC macro-model^[3]. Following the flow illustrated in Fig. 1, the vulnerability of a printed circuit board implementing cryptographic functions is evaluated, and the corresponding attenuation target is determined.

In the conventional approach, the simulated switching current $I(f)$ and transfer function $Z(f)$ are combined to compute the leakage spectrum $V(f)$, and frequency-domain correlation power analysis is used merely to determine whether vulnerable frequency bands exist. In contrast, the proposed method derives a correlation threshold ρ_{req} from the required success rate and constructs a secure region of the transfer characteristics on the complex plane. By checking whether the actual or simulated transfer function lies within this region across frequency, designers can obtain not only the vulnerable bands but also the target attenuation and the maximum allowable impedance, $|Z(f)| - |Z_m(f)|$, determined by the minimum radius of the secure-region boundary. This demonstrates that the proposed flow delivers concrete and quantitative design guidance that conventional approaches cannot provide.

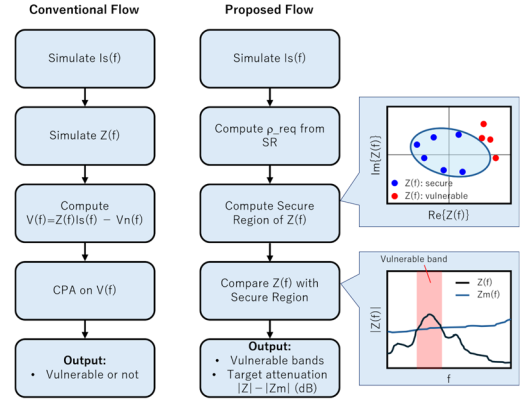


Fig. 1. Comparison between conventional and proposed flows.

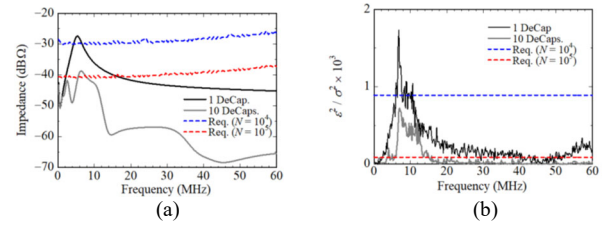


Fig. 2. (a) The transfer impedance of the leakage path exceeded both required levels under the initial condition with one decoupling capacitor (black). It was redesigned below the upper required level (blue) with ten capacitors (gray). (b) The signal-to-noise ratio of the measured leakage also reduced with ten capacitors below the upper level.

III. RESULTS

To validate the effectiveness of the proposed method, we applied it to a printed circuit board on which the Advanced Encryption Standard with a 128-bit key (AES-128) was implemented using an FPGA. The evaluation confirmed that the required security level was achieved when the attenuation target derived by the proposed method was satisfied. Conversely, the device exhibited vulnerability in frequency bands where the target attenuation was not met. These results demonstrate that the proposed method can accurately determine the attenuation target for the leakage path's transfer characteristics and guide designers in improving the robustness of the PDN against side-channel attacks.

ACKNOWLEDGMENT

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Modeling of HEMP-Induced Currents and Voltages in Buried Shielded Cables

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Abstract—High-altitude electromagnetic pulse (HEMP) can induce hazardous transient currents and voltages in buried infrastructure, even when shielded cables are used. We present a validated modeling framework for predicting HEMP-induced responses in shielded cables buried in lossy ground. The method is based on transmission line theory and accounts for frequency-dependent soil parameters and shield transfer functions. Terminal responses are obtained using the Baum-Liu-Tesche (BLT) formulation and validated against reference solutions. The proposed methodology could support vulnerability assessment and electromagnetic protection design.

Keywords—HEMP, buried shielded cables, transfer impedance, transfer admittance, field-to-line coupling, BLT equations

I. INTRODUCTION

High-altitude electromagnetic pulses (HEMP) generate intense, wide-area electromagnetic fields capable of indirectly coupling into extended conductors such as power and communication cables. Underground routing and shielding reduce exposure, but induced transients may still threaten connected equipment. Accurate prediction is therefore essential for electromagnetic hardening of critical infrastructure (e.g., [1]–[4]). This paper focuses on the practical modeling of HEMP coupling to shielded cables buried in lossy ground [2] and [3], with emphasis on configurations representative of real installations. Intrinsic cable parameters are assumed to be excitation-independent; incidence-angle effects appear only in the distributed source terms.

II. MODELING METHODOLOGY

Fig. 1(a) shows a shielded cable buried at depth d in a homogeneous lossy half-space (conductivity σ , relative permittivity ϵ_r) and illuminated by an incident HEMP plane wave. The tangential electric field at the cable depth, obtained from Fresnel transmission in the half-space, serves as distributed excitation to the external circuit formed by the cable shield and ground return. The internal response is determined using through the cable transfer impedance and admittance [2]. Ground effects are included through frequency-dependent per-unit-length impedance and admittance. The system is solved using the BLT formulation [2] for arbitrary terminations.

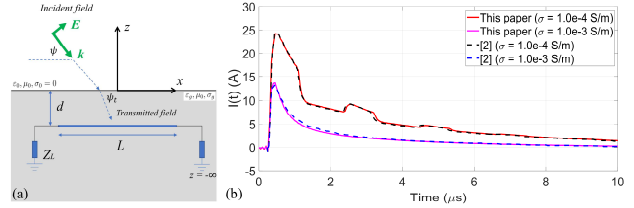


Fig. 1. (a) Geometry of a buried shielded cable under plane-wave excitation; (b) induced load voltage for different soil conductivities (solid: this work; dashed: adapted from [2]).

III. RESULTS AND PRACTICAL IMPLICATIONS

The model was validated against published reference cases, showing good agreement in both frequency and time domains [2]. Fig. 1(b) compares the induced load voltage for different soil conductivities (solid: this work, dashed: from [2]). Increasing soil conductivity mainly increases attenuation and dispersion of the ground return, while fast rise-time components can still reach the load, making termination conditions and shield transfer characteristics critical for protection design.

IV. CONCLUSION

A validated modeling framework for HEMP-induced coupling to buried shielded cables has been presented. The approach supports vulnerability assessment and supports the design of effective electromagnetic protection strategies for underground cable systems in critical infrastructure. Future work will consider coupling to a network of underground cables with arbitrary geometries and taking into account frequency-dependent termination models.

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Vulnerability Assessment Using Single-Point Attenuation Measurement with External Stirring

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Abstract– This paper proposes an inside-to-outside single-point attenuation measurement method using an external mechanical stirrer to assess electromagnetic attenuation of unshielded facilities. By placing the stirrer outside near the receiving antenna, indoor space constraints are avoided while maintaining regulatory compliance. Validation using a reconfigurable testbed shows good agreement with the standard IEC multi-point method, with a mean absolute error of 1.86 dB.

Keywords–attenuation, unshielded facilities, single-point measurement, inside-to-outside propagation, external stirring.

I. INTRODUCTION

Assessing the electromagnetic (EM) vulnerability of critical infrastructure against intentional electromagnetic interference (IEMI) is essential. While IEC 61000-4-23 specifies the outside-to-inside (O2I) propagation method [1], it can interfere with nearby civilian electronics. Inside-to-outside (I2O) propagation-based methods mitigate this by confining the source [2]. However, internal stirring requires significant indoor clearance. This paper proposes an I2O method with external stirring, placing the stirrer and receiver outside to overcome spatial and regulatory constraints while maintaining statistical validity.

II. MEASUREMENT METHOD

The validation was conducted using a reconfigurable testbed ($3.0 \times 3.0 \times 2.5 \text{ m}^3$) mimicking reinforced concrete with tempered glass windows [3]. We compared the standard IEC 61000-4-23 multi-point method based on continuous-wave immersion (CWI) with the proposed I2O single-point method using an external stirrer. As shown in Fig. 1, the measurement configuration consists of a transmitter inside the facility and a receiver with an external stirrer positioned outside.

To prevent interference, sensitive internal electronics should be powered off during I2O measurements. The receiving antenna is located in the shadow region behind the stirrer (reflector dimensions: $0.4 \times 0.6 \text{ m}^2$) at a distance of 0.6 m (2λ at the lowest frequency). The stirrer rotates at a fixed position without spatial translation.

In this configuration, the stirrer blocks the pseudo-line-of-sight path and induces power fluctuations through scattering and diffraction. The peak received power obtained during rotation provides a statistical estimate comparable to the spatial average in standard multi-point methods.

III. RESULTS AND CONCLUSION

Fig. 2 shows the attenuation from 1 to 6 GHz, while Table I summarizes the corresponding statistics. The proposed I2O method with external stirring yielded a mean attenuation of 11.72 dB, showing a mean absolute error (MAE) of 1.86 dB relative to the multi-point method. This is comparable to the inside-stirrer method

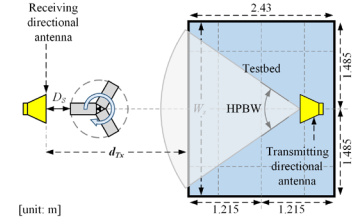


Fig. 1. Measurement configuration of the inside-to-outside single-point method with external stirring.

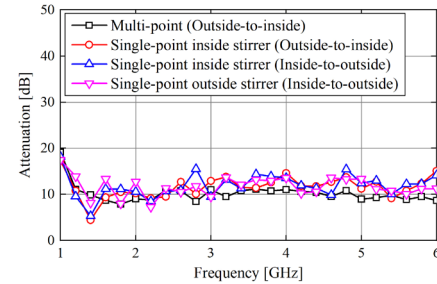


Fig. 2. Comparison of attenuation between measurement methods.

TABLE I. THE MEASUREMENT CHARACTERISTICS

Propagation	Outside-to-inside		Inside-to-outside	
Assessment method	Multi-point	Single-point	Single-point	Single-point
Stirrer position	-	Inside	Inside	Outside
Mean attenuation [dB]	10.20	11.61	11.92	11.72
Mean standard deviation [dB]	1.50	2.17	2.22	1.82
Mean range (max-min) [dB]	6.21	11.95	11.93	9.00
Mean absolute attenuation error [dB]	-	1.84	2.09	1.86

(MAE: 2.09 dB), confirming that external stirring in the shadow region is a valid alternative.

This approach provides a practical and regulatory-compliant solution for IEMI vulnerability assessment.

ACKNOWLEDGMENT

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Shielding Effectiveness Reconsidered:

The Need for Alternative Attenuation Metrics in Complex Enclosures

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Abstract– This paper will precis a series of part published papers by the authors which seek to illustrate why the traditional Shielding Effectiveness (SE) specification and measurement approach often fails to deliver the intended outcome for large and complex shielded enclosures. Occasionally, large enclosures or structures are required to provide attenuation of internal and external electromagnetic disturbances. In this paper we will discuss some potential flaws in using the traditional SE methods and then present alternative approaches.

Keywords– Shielding Effectiveness (SE), Shielded Enclosure Attenuation, HEMP, IEMI, Radio Astronomy

I. INTRODUCTION

Throughout 2026 the Authors plan to publish a series of papers in the IEEE EMC Society Magazine discussing issues with and proposing solutions for Shielding Effectiveness (SE) specification setting and measurement methods. The need to minimise Electromagnetic Interference (EMI), both radiated and conducted, has been a known requirement since the invention of Electrical machines and the dawn of Radio. The authors of this paper are particularly interested in the protection of internal sensitive electronics from high-power electromagnetic environments such as High-Altitude Electromagnetic Pulse (HEMP) and Intentional Electromagnetic Interference (IEMI). We are equally interested in the minimization of internally-generated emissions to prevent them from interfering with sensitive external equipment, as in the case of radio astronomy. The traditional method used to mitigate EMI is that of shielding for which very many standards exist, the most common ones being IEEE 299 [1]; IEC 61000-4-23 [2]; IEC 61000-5-6 [3] and Mil-Std-188-125 [4]. A thorough review of these standards is included in [5]. However, our collective experience in a mandated shielding specification is that the final structure, once built, often fails to achieve the specified performance.

II. DISCUSSION – THE ISSUES

In the practical implementation of shielded enclosures, indeed in many applied EM problems, small changes, small violations or breaches, can have a profound effect on shielding performance. Areas that are arguably most apparent include: Conductive penetrations; Conductor placement and routing; Seams; Interfaces; and Apertures.

There are issues with the traditional SE test method too, for example, the calibration phase, the consequence of cavities and modal effects, of antenna loading, of appropriate reference levels for comparison, of net enclosure radiation pattern, of polarisation properties, and non-reciprocity.

III. DISCUSSION – SOME POTENTIAL ALTERNATIVES

Throughout our investigation we suggest several changes in the general approach for consideration as follows:

- The main goal of a shielded enclosure is to provide attenuation of radiated and conducted EM signals. SE measures the performance of the shield, whilst this is an important indicator it is not the goal
- It is necessary to embrace the fact that conductive penetrations are a significant contributor to the attenuation of the shielding enclosure and so conducted attenuation should be considered at the outset – in the specification of the shield.
- We believe that reciprocity does not hold in practice so Shielded Enclosure performance should be measured inside-to-outside or outside-to-inside depending on the application.
- We believe that cavity resonance and polarization effects are important and should be incorporated in formulating the measurement approach.

IV. SUMMARY

In our paper we will attempt to elaborate on the issues and new approaches summarized above. We recognize that there is no perfect method to derive a universally-objective description of shielded enclosure performance. There are practicalities that will always make this goal unattainable. However, we hope that these suggestions are taken forward by others into the development of a new standard for the specification and measurement of high-quality shielded enclosures.

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Fully Optical Sensor For Absolute, Vectorial and Real-Time Magnetic Field Measurement

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Abstract— *Magnetic field measurement is usually performed using a loop-shaped metallic antenna whose size is matched to the wavelength of the signal to be analyzed. Here we present an innovative and complementary alternative based on the Faraday effect, enabling comprehensive characterization of the magnetic field. The developed fingertip optical probe allows the real time measurements of one component of the magnetic field vector with a sub-millimeter spatial resolution. The probe is temperature insensitive and the minimum detectable magnetic field reaches a few $\text{mA}\cdot\text{m}^{-1}\cdot\text{Hz}^{-1/2}$.*

Keywords— electromagnetic field measurement, vectorial characterization, non invasive sensor, ultra wide band analysis.

I. INTRODUCTION

Electromagnetic (EM) field metrology involves the use of reference sensors delivering an analog signal that represents the wave to be measured. Ideally, this signal is provided in real time and the sensor response is linear with the amplitude of the analyzed EM component. Finally, the measurement probe must not interfere with the quantity being measured. Regarding the electric component E of the wave, numerous studies have led to vector optical sensors with ultra-wide bandwidth (10 Hz–THz), sub-millimeter spatial resolution, dynamic range exceeding 120 dB, from fields below V/m up to air breakdown [1]–[3]. These probes enable *in-situ* specific absorption rate measurements, characterization of high power microwaves, or near field antenna mapping. However, concerning the magnetic (H) component of the EM wave, such probes did not exist. At high frequencies, H -field measurements are typically carried out using commercially available magnetic loops. Their bandwidth remains limited and intrinsically limited by their geometry. Moreover, the delivered signal corresponds to the derivative of the magnetic flux. Their metallic composition and galvanic connection are source of EM interferences, especially in near-field measurements. We here present, to our knowledge, the first pigtailed magneto-optic (MO) probe working in a reflection scheme and suitable for measurement in harsh environment.

II. THE MAGNETO-OPTICAL PROBE

A schematic and a photography of the probe are proposed in Fig. 1.

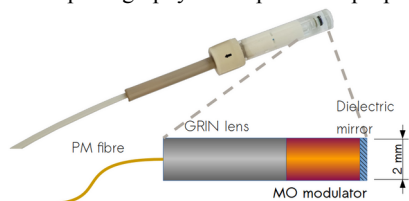


Fig. 1. Photography and schematic of the MO probe.

The probe includes a MO crystal that converts the H -field to be measured into a modulation of the optical polarization state. Appropriate optical processing within the MO modulator converts this polarization modulation into an optical power modulation, which is finally converted into current thanks to a photodiode. The

shaping of the beam is ensured by a gradient-index (GRIN) lens. The useful optical signal is reflected by a dielectric mirror and collected in a return fiber. A fully dielectric packaging ensures the mechanical and optical integrity of the probe. The probe is connected to a commercial optoelectronic instrument (eoSense 4th Gen.) including an ultra-low noise laser and a fast photodiode. The system delivers a voltage proportional to one component of the H -field vector, colinear with the probe symmetry axis.

III. PERFORMANCES OF THE MO SYSTEM

In order to perform the linearity assessment of the response, the probe is placed in a calibrated TEM cell fed by a function generator. Fig. 2 shows the obtained experimental result. The measurement dynamics exceeds 50 dB.

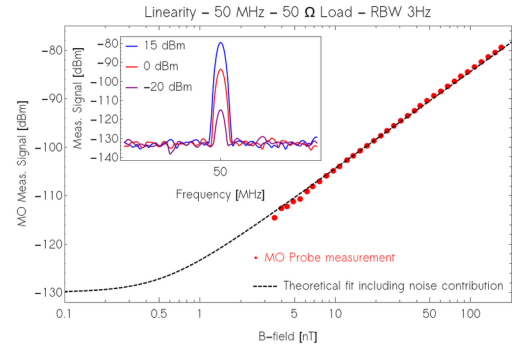


Fig. 2. Linearity of the MO response as a function of the applied H -field. Some raw signals, measured with a spectrum analyzer are given in the inset for three powers at the input of the TEM cell.

The temperature dependence has also been characterized in a climatic chamber. A weak dispersion of 0.4 dB has been observed on the H -field signal with 20°C temperature variations.

IV. CONCLUSION

We demonstrate for the first time the possibility of performing real-time and non-invasive H -field measurements using a fully optical sensor. The measurement sensitivity is greater than 4 mArms/m and the theoretical bandwidth exceeds several GHz. The presented system will enable innovative measurements in many fields such as electromagnetic compatibility (EMC), near-field mapping, and biomedical applications.

ACKNOWLEDGMENT

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Analysis of Output-Characteristic Measurement of a 100 MHz Oscilloscope for HPEM Systems

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Abstract—This paper analyzes the output characteristics of a High-Power Electromagnetic Pulse (HPEM) system using a 100MHz oscilloscope and a power detector. Conventional multi-GHz oscilloscopes are unsuitable for integrated HPEM radiation systems due to their bulkiness and excessive specifications for simple monitoring. Test results using a source with a 20ns pulse width revealed a 6dB discrepancy between the measured and actual outputs. This attenuation occurs because the 100MHz oscilloscope cannot fully capture the rise and fall times of the pulses. Based on these findings, further analysis of test data will be conducted to ensure measurement reliability for compact HPEM systems.

Keywords— HPEM, HPEM measurement, Waveguide Directional Coupler, 100MHz Oscilloscope, Power Detector

I. INTRODUCTION

In High-Power Electromagnetic Pulse (HPEM) radiation systems, accurate measurement and monitoring of the output generated by the HPEM source are essential to predict the impact of pulses on target objects. Typically, the output of an HPEM source is measured using high-power waveguide directional couplers [1]. These outputs are traditionally monitored using multi-GHz oscilloscopes. However, multi-GHz oscilloscopes are often unsuitable for integrated HPEM radiation systems due to their bulky size and unnecessarily high specifications for simple monitoring purposes [2]. This paper analyzes the output characteristics of an HPEM system using a compact 100MHz oscilloscope and a power detector.

II. METHODOLOGY

Figure 1(a) illustrates the configuration and results of the HPEM measurement calibration test using a signal generator and power amplifier. The signal from the high-power waveguide directional coupler was split using a power divider and connected to a 2.5GHz oscilloscope for comparison. The input for the power detector (Herotek DTM 180NB), which is connected before the 100MHz oscilloscope, was controlled within the linear operation range of -15dBm to 5dBm, considering the expected output of the HPEM source. Tests confirmed that as the pulse width from the signal generator decreased, the error in the measured results increased [2]. Figure 1(b) shows the configuration for an HPEM measurement test using an actual HPEM source. The source consists of a Marx generator and a Relativistic Magnetron (RM), producing a signal with a pulse width of approximately 20ns. The output signal is delivered to the power detector after being attenuated through the high-power waveguide directional coupler.

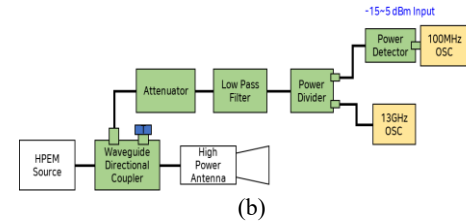
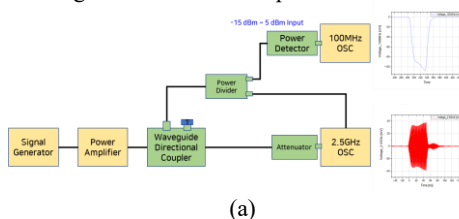


Fig. 1. Experimental setup for HPEM measurement. ((a) cold test, (b) hot test)

III. RESULTS AND DISCUSSION

The graph in Figure 2 plots the results of the test shown in Figure 1(b). The blue line represents the calibration test data, while the magenta mark represents the actual measured output of the HPEM source. A difference of approximately 6dB was observed in the output detected by the power detector. The result after applying a 6dB compensation is shown in red mark. This confirms that the 100MHz oscilloscope outputs an attenuated value because it cannot fully capture the rise and fall times of the output pulse.

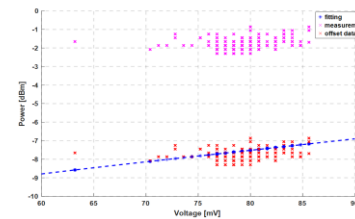


Fig. 2. Experimental result of HPEM measurement.(magenta : measured, red : offset, blue : cold test)

IV. CONCLUSION

Through this study, it was confirmed that a difference of approximately 6dB occurs between the measured characteristics and the actual output at the specific pulse width of the HPEM source. Based on these analytical results, we plan to secure further HPEM system test data to enhance reliability, enabling the measurement of high-power electromagnetic pulse system outputs using a compact configuration.

ACKNOWLEDGMENT

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Non-Contact Measurement of Noise Source Impedance of Switched Mode Power Supply

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Abstract– High-power external events such as electromagnetic pulses (EMP) and surges can affect the operation of important electronic equipment in facility environments. To understand how these disturbances interact with the power input of such devices, it is necessary to accurately identify their input noise source impedance under real operating conditions. This study introduces a non-contact measurement method that allows input noise source impedance to be measured in place, depending on the operating states of facility equipment, without requiring any intrusive access to the power lines. The measured impedance examined to show how the impedance changes with different operating states across a range of frequencies. These impedance characteristics offer useful clues about frequency regions where equipment behavior may change when exposed to high-power transient events. The proposed non-contact method provides a practical way to characterize input noise source impedance in actual facility environments.

Keywords– Electromagnetic compatibility (EMC), Electromagnetic interference (EMI), Electromagnetic pulse (EMP), Inductive coupling, Noise source impedance.

I. INTRODUCTION

Electrical equipment in facility environments can be affected by external electromagnetic disturbances coupled through power lines. High-power transient events such as electromagnetic pulses (EMP), lightning, and surge phenomena may propagate through the power input path and influence internal circuits. Therefore, understanding the electrical characteristics of the power input path, particularly the input impedance, is important for analyzing the response of equipment to such disturbances [1].

Inductive coupling–based impedance measurement using current probes enables in-circuit impedance observation without direct electrical contact with the power line, making it suitable for measurements on operating systems.

In this paper, a non-contact measurement approach using current probes is introduced to observe the input noise source impedance of electrical equipment under different operating states [2].

II. PREDICTION OF COMMON-MODE CURRENT IN NON-LISN CONDITIONS

The noise source impedance of electrical equipment was measured using a non-contact inductive coupling method with two current probes and a vector network analyzer (VNA). One current probe injects the test signal generated by the VNA into the power line through inductive coupling, while the other probe measures the response signal from the same line. Since the current probes are clamped on the power cable, the noise source impedance can be measured without direct electrical contact with the circuit, allowing measurements to be performed while the equipment is operating. In addition, to ensure sufficient coupling in the low-frequency region, measurements were performed by adjusting the number of turns of the current probe. Fig. 1 shows the non-contact noise source impedance measurement setup using the current probes and the

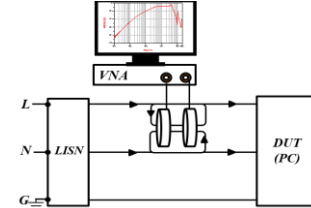


Fig. 1. Measurement setup.

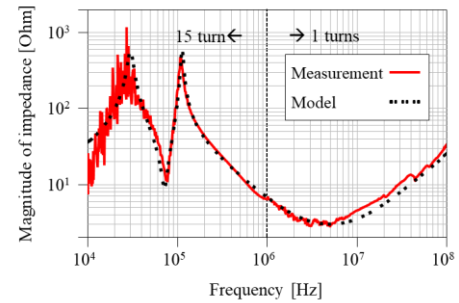


Fig. 2. Noise Source impedance experimental result of SMPS.

VNA [1]-[2].

Based on the measured noise source impedance characteristics, a passive equivalent circuit model was constructed. Fig. 2 shows the measured impedance and the calculated results of the passive equivalent circuit overlaid on the same plot for comparison. This equivalent circuit model can be used in future analyses to investigate the noise behavior of the equipment when high-power signals are applied.

III. CONCLUSION

In this paper, the noise source impedance of electrical equipment was measured using a non-contact inductive coupling method with current probes and a vector network analyzer (VNA). A passive equivalent circuit model was derived based on the measured impedance characteristics and compared with the measurement results. The proposed model can be used for future analysis of noise characteristics under high-power signal conditions.

ACKNOWLEDGMENT

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Analysis of Common-Mode vs. Differential-Mode Three-Phase Power Filters for HEMP Mitigation

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Abstract— This paper examines the performance limitations of common-mode (CM) three-phase power filters compared with differential-mode (DM) filters when exposed to high-energy electromagnetic disturbances such as High-Altitude Electromagnetic Pulse (HEMP). Simulation results show that CM filters provide minimal attenuation when subjected to differential mode line-to-ground impulses, resulting in high residual currents. DM filters consistently offer superior attenuation and reliable impedance behavior under both RF and fast-pulse stimuli. Recommendations for HEMP-resilient filter architectures are presented.

Keywords— Filter, HEMP, EMP, ESA, Common Mode, Differential Mode, Mitigation.

I. DEFINITIONS AND FUNCTIONAL DIFFERENCES

Common-mode three-phase power filters are widely deployed to suppress conducted electromagnetic interference in industrial, commercial, and critical-infrastructure electrical networks. While their performance under conventional RF noise conditions is well documented, their behavior under extremely fast, high-energy transients—particularly those originating from a high-altitude electromagnetic pulse—is less understood. This paper reviews fundamental definitions, structural differences, and simulated pulsed response characteristics of common-mode and differential-mode filters. DM filters attenuate noise appearing between conductors and ground. This noise may be of same amplitude and phase or different on each conductor. CM filters target noise appearing simultaneously on all conductors with respect to ground as shown in figure 1.

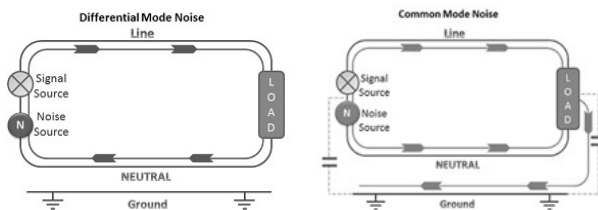


Figure 1. DM and CM current paths

However, CM filters depend on magnetic flux cancellation, which is easily disrupted under unbalanced conditions and disturbances.

II. HEMP MITIGATION LIMITATIONS OF COMMON-MODE FILTERS

Simulations show that CM filters do not provide effective protection in tested parameters when a pulsed disturbance appears on their lines. These parameters include residual current, rate of rise,

and the action integral response (root action). The purpose of these metrics is to ensure that any signals penetrating inside the shielded building do not perturb the electronics operating inside. The residual peak current is the current let through that will still pass on to the load after being clamped down by the Electronic Surge Arrestor (ESA) and attenuated by the filter. The rate of rise measures the maximum absolute value of the first derivative of a current waveform $I(t)$ with respect to time, di/dt , expressed in units of amperes per second. The last parameter (root action) measures how much power is delivered to the load by the pulse after passing through all the mitigation measures [1]. CM filters exhibit highly reduced line impedance when pulsed, such as during line-to-ground testing. Residual currents remain high because the CM choke core becomes saturated or fails to develop the required counteracting magnetic flux, which then allows pulse energy to flow to the load affecting other tested parameters as well.

HEMP standards such as MIL-STD-188-125-1 and IEC 61000-4-24 [1] [2] inject extremely fast transient currents on each power line of the protection device, such as a filter. This is done to simulate an actual HEMP event. Under these conditions CM filters fail to generate significant impedance, allowing fast transients to bypass filtering almost entirely. Thus line-to-ground pulses produce minimal attenuation in CM filter structures. DM filters, by contrast, remain effective because their operation does not depend on magnetic cancellation. An alternative CM topology would include DM inductors on each line, but this is not always practical as the size of these inductors can be significant especially when filters are in service feeds greater than 100 amperes. Simulations demonstrate that adding this inductance can significantly reduce current let throughs and reduce other critical parameters like rate of rise. But the best approach would be to have the filter optimized to work with discrete lines in DM.

III. CONCLUSIONS

Based on the simulations and established circuit behavior DM-dominant filter architectures offer superior response to HEMP conditions. For critical infrastructure requiring protection against HEMP, filters that incorporate strong DM impedance and minimize reliance on magnetic cancellation should be preferred.

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Application of a Supercapacitor-Based Primary Power Supply Under Low-Temperature Conditions

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Abstract—This paper investigates the limitations of lithium-ion batteries as energy sources for high-voltage primary power supply (PPS) operating at temperatures below 0 °C and evaluates the applicability of a supercapacitor-based configuration. Lithium-ion batteries exhibit reduced output current capability at low temperatures due to increased internal resistance [1], which may degrade performance in pulsed high-voltage applications. Compared to lithium-ion batteries, supercapacitors exhibit relatively stable power characteristics under low-temperature conditions[2]. In this study, the temperature-dependent charge–discharge characteristics of a supercapacitor were experimentally evaluated. A supercapacitor-based PPS was implemented, and its high-voltage operation was verified. The results confirm stable capacitance characteristics under low-temperature conditions and reliable PPS operation at room-temperature.

Keywords— High-Voltage Primary Power Supply, Supercapacitor, Lithium-ion Battery, Low-Temperature Operation, Pulsed Power

I. INTRODUCTION

High-voltage PPS used in pulsed power applications require reliable energy delivery within a limited charging interval. Lithium-ion batteries are widely used due to their high energy density; however, their output current capability decreases at temperatures below 0 °C as internal resistance increases [1]. This temperature-dependent degradation may affect system stability in low-temperature conditions. This effect can result in charging delay and reduced reliability of high-voltage PPS operation. Supercapacitors maintain stable electrical characteristics under low-temperature conditions [2]. Based on these characteristics, this study evaluates the applicability of a supercapacitor as a primary energy source for high-voltage PPS operation in low-temperature conditions.

II. METHODOLOGY

Two experimental investigations were conducted to evaluate the applicability of a supercapacitor as an energy source for PPS. First, charge–discharge tests were performed to evaluate the electrical characteristics of the supercapacitor under low-temperature conditions. The supercapacitor bank was tested while the temperature inside the environmental chamber was varied stepwise over a low-temperature range from −30 °C to 50 °C, and the charge and discharge behaviors were measured. The capacitance variation was calculated from the constant-current discharge voltage curve between 80% to 40% of the maximum charge voltage to evaluate the temperature-dependent stability of the supercapacitor. Second, a high-voltage output operation test was conducted with the supercapacitor applied as the primary energy source of the PPS. The PPS was configured to generate both DC high-voltage output and a trigger pulse output based on the input voltage supplied by the energy source. The output voltage of the supercapacitor was applied to the PPS input, and the system operation was evaluated at room-

temperature. The test confirmed that the PPS operated as intended when powered by the supercapacitor, generating the required output functions for pulsed power operation.

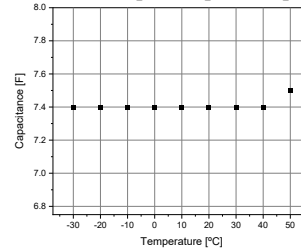


Fig. 1.

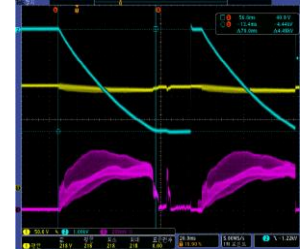


Fig. 2.

Fig. 1. Calculated Capacitance of the Supercapacitor as a function of Temperature obtained from Charge-Discharge Tests.

Fig. 2. The Output Voltage Waveform of the PPS.

III. RESULTS AND DISCUSSION

Fig. 1. shows that the calculated capacitance of the supercapacitor exhibits minimal variation over the tested temperature range. This result indicates that the electrochemical characteristics of the supercapacitor are relatively insensitive to temperature changes within the tested range. Fig. 2. confirms that the supercapacitor-based PPS operates normally at room-temperature, generating the required DC high-voltage and trigger pulse outputs. These results demonstrate that the supercapacitor can serve as a stable voltage source for PPS operation under low-temperature conditions.

IV. CONCLUSION

This study investigated the impact of low-temperature conditions on energy source performance in high-voltage PPS and evaluated the applicability of a supercapacitor as an alternative to lithium-ion batteries. Experimental results demonstrated that the supercapacitor maintained stable capacitance characteristics over a temperature range from −30 °C to 50 °C and enabled normal PPS operation when applied as the Primary voltage source. These finding indicate that the use of supercapacitors can effectively mitigate temperature-dependent performance degradation associated with battery-based systems. In this regard, Supercapacitors offer a feasible system-level approach for enhancing the operational stability of high-voltage PPS in low-temperature conditions. Further investigation of high-voltage PPS operation with a supercapacitor under low-temperature conditions will be conducted in future work..

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Lightning, HEMP and Generalised Pulse Mitigation Strategies.

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Abstract— This paper reviews generalised mitigation strategies for lightning and high-altitude electromagnetic pulse (HEMP) environments for control lines. A theoretical framework is developed and applied to practical approaches ranging from clamping devices to multi-stage filters. A three-layer mitigation concept is proposed, and the efficacy and limitations of the approach are discussed.

I. INTRODUCTION

Shielded facilities must transmit and receive signals to and from external transducers, yet the copper cables required for this interface can inadvertently conduct conducted or radiated intentional electromagnetic interference (IEMI) into the facility. These interference mechanisms frequently appear as fast transient pulses capable of damaging equipment or corrupting sensitive measurements.

This work evaluates a three-layer mitigation architecture—the environmental layer, interface layer, and measurement layer. The strengths and weaknesses of each layer are discussed for EMI and IEMI scenarios, along with exceptions to the three-layer rule.

II. METHODOLOGY

We first assess various clamping devices and compare their operational characteristics. These include metal oxide varistors (MOVs), gas discharge tubes (GDTs), and semiconductor crowbar devices. We then develop example low-pass filters to support these clamping components. Three filter formats are examined:

- High-power external modular filters
- PCB-based internal component filters
- Digital filter implementations

Laplace-domain representations are converted to discrete-time equivalents using:

$$H(s) = \frac{\omega^2}{s^2 + \frac{\omega}{Q}s + \omega^2} \quad (1)^1$$

$$s \rightarrow \frac{(1 - Z^{-1})}{(1 + Z^{-1})} \quad (2)^1$$

$$H(Z) = \frac{b_2 Z^{-2} + b_1 Z^{-1} + b_0}{a_2 Z^{-2} + a_1 Z^{-1} + a_0} \quad (3)$$

These equations enable the construction of a third, digital-filter variant. The merits, limitations, and appropriate placement of all

three filter types are compared, supported with examples.

III. RESULTS AND DISCUSSION

MOVs, spark gaps, and crowbar circuits effectively limit surge amplitudes and protect downstream electronics from overload. However, they must be paired with appropriately designed low-pass filters to achieve effective suppression of fast transients such as HEMP or pulse-based IEMI.

Filters alone often struggle to dissipate or absorb large transient energy while maintaining efficient signal handling. Conversely, clamping components without filtering allow excessive high-frequency energy to propagate into sensitive circuitry.

Digital filtering offers advantageous flexibility, enabling the implementation of transfer functions impractical in physical circuits, and providing dynamic signal conditioning upstream or downstream of physical filter stages.

IV. CONCLUSION

The results indicate that an optimal mitigation strategy requires multiple complementary filters distributed across system layers. Digital filters enhance the achievable response and compensate for the physical limitations of analogue components. When properly integrated, the three-layer approach offers robust and scalable protection for lightning, HEMP, and IEMI scenarios.

ACKNOWLEDGMENT

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Pulsed Electromagnetic Field Generation Using Timed Array Structures

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Abstract– The bounded-wave topology for NEMP hardness validation testing is unfeasible for larger platforms, such as warships. Several alternatives are available, each with specific advantages and disadvantages. In this paper we explore the use of NEMP illumination using multiple antennas in a time domain array to generate the high field levels over a section of the total device under test. Sectional testing is specifically chosen over full illumination to ensure parameters such as high field strength and homogeneity are obtained over a limited volume.

Keywords– NEMP, hardness validation testing, time domain array, impulse radiating antenna

I. INTRODUCTION

A NEMP hardness validation method that instills sufficient confidence must reflect realistic field strength, homogeneity, polarity and time domain fidelity [1]. Traditionally, NEMP hardness validation testing is performed by using a bounded wave simulator, hence a waveguide-like topology within which the device under test (DUT) is placed. In such a topology the required field parameters are generally met. However, bounded-wave simulators become impractical for DUT dimensions beyond tens of meters due to the exponential growth in volume and power requirements [2]. Consequently, full-platform testing of large platforms, such as warships, is considered infeasible for a bounded wave test; moreover, full-threat testing in a functional setting (e.g. ship on water), escalates difficulties rapidly.

In previous attempts to validate the NEMP hardness of larger devices radiated tests were conducted, in which the DUT is placed at a distance from the simulator and illuminated. This means that a portion of NEMP is no longer simulated which contains significant low-frequency content that is inefficient to radiate in practical free-space simulators. In past simulators, neither the required field strength nor field homogeneity was achieved in the testing volume [2]. The EMIS-III-VPD is used as the baseline reference for this research. In the nineties, this vertically polarized dipole simulator was the standard for NEMP hardness validation testing of Dutch Navy ships.

Achieving the required 50 kV/m homogeneously over a large volume at 100 m is considered an unrealistic specification. Therefore, in this paper the choice is consciously made to illuminate the DUT sectionally instead. This approach prioritizes local field compliance at the expense of full-platform illumination. We are aware that this introduces a limitation with respect to the standards, as the very low frequency components of NEMP are no longer effectively simulated. This work will explore if the use of a time domain array of directed elements could be used to achieve this objective. Furthermore, a practical perspective is presented where the advantages and limitations of our proposed approach will be quantified.

This research is funded by TNO research program V2430.

II. METHODOLOGY

The first step is to select which types of antennas are suitable to transmit ultra-wideband signals in a low-dispersion manner. For the selection of the antenna element, a system-oriented prioritization as follows: 1. Construction and design simplicity, 2. Time domain fidelity, 3. Polarity. The 4th dominant parameter identified is the antenna directivity as it directly increases achievable field strength, improves spatial homogeneity over the test zone, and reduces unintended illumination outside the facility. The antenna of choice is the reflector impulse radiating antenna (IRA) [3]. This choice is motivated by the points scored across the identified metrics as shown in table 1. These are weighted such that the first row scores points up to 20, the second, third and fourth row score points up to 10, and the last two rows score points up to 5.

TABLE I – ANTENNA CHOICE MOTIVATION

Metric	Biconical	TEM Horn	Vivaldi	Reflector IRA	Lens IRA
Radiation Pattern	2	12	10	20	16
Time Domain Fidelity	6	8	8	10	8
Design Simplicity	6	10	8	4	1
Construction Practicality	4	10	8	4	2
Polarization	1	2	2	5	5
Literature Availability	2	3	2	5	3
Total	23	45	38	48	35

The IRA will be simulated in CST Microwave Studio. Subsequently, numerical results will be used to optimize the NEMP-performance of the IRA. The optimized IRA is then used as basic radiation element to design the time domain array.

To maintain the desired NEMP-signal fidelity in the testing volume, the placement of the IRAs will be calculated using an optimization algorithm developed to optimize IRA placement based on IRA-element far field characteristics in time domain and desired testing volume field parameters as input. This research paper is still being worked on, and the results will be presented at the conference. This paper forms a part of a larger project aimed at the future of NEMP testing of warships, specifically addressing a potential method to generate an adequate NEMP-like pulse.

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Simulation-Based Study on High-Power Electromagnetic Pulse (EMP) Protection for Mobility Platforms

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Abstract—This paper presents a simulation-based evaluation of high-power electromagnetic pulse (EMP) effects on a mobility platform. Measured meso-band and hyper-band EMP signals were modeled and applied to a full-vehicle electromagnetic simulation. Electric-field distributions were analyzed under different ground conditions, polarization angles, and source distances. The results identify EMP-vulnerable regions and provide guidance for shielding enhancement and system-level hardening.

Keywords—EMP, HPEM, mobility platform, shielding effectiveness, electromagnetic simulation

I. INTRODUCTION

High-power electromagnetic (HPEM) environments, including EMP, HPM, and UWB sources, can disrupt or permanently damage modern vehicles containing dense electronic systems. With the proliferation of electrified and autonomous platforms, electromagnetic resilience has become a critical design requirement. This work investigates system-level coupling characteristics of a mobility platform under meso-band (2.5 MHz–1 GHz) and hyper-band (up to 5 GHz) EMP excitation. The objective is to identify leakage mechanisms and vulnerable regions using realistic measured-source modeling and full-wave simulation.

II. METHODOLOGY

A three-dimensional vehicle model was constructed using a full-wave electromagnetic solver. Mesh refinement and error correction were performed to ensure numerical stability and computational efficiency. Measured EMP waveforms were transformed into frequency-domain representations and applied as dipole and plane-wave sources at separation distances of 5 m and 10 m. Ground conditions were modeled as metallic ground and asphalt. Field polarization (vertical/horizontal) and incidence angle were varied. Electric-field distributions were evaluated using near-field probes. The electromagnetic response follows Maxwell's equation,

$$\nabla \times \mathbf{H} = \mathbf{J} + \partial \mathbf{D} / \partial t \quad (1)$$

where $\mathbf{H}$ is magnetic field intensity, $\mathbf{J}$ current density, and $\mathbf{D}$ electric flux density.

III. RESULTS AND DISCUSSION

Ground conditions significantly influenced field coupling. Metallic ground produced stronger reflections; however, asphalt provided more realistic operational conditions and was used for primary assessment.

Representative meso-band simulation results under DS_10m excitation ($\theta = 85^\circ$, vertical polarization) are shown in Fig. 1. Strong electric-field concentrations appear near the front-end and lower-body regions of the vehicle. Peak field levels occur around 50 MHz, and metallic ground conditions yield higher maximum values compared with asphalt.

Reducing the source distance from 10 m to 5 m substantially increased peak electric-field intensity, confirming strong near-field coupling effects. Field polarization also affected coupling paths: vertical polarization enhanced roof and pillar regions, while horizontal polarization increased side-body coupling. These results indicate that structural discontinuities, lower-body regions, and aperture areas are primary EMP-vulnerable locations in mobility platforms.

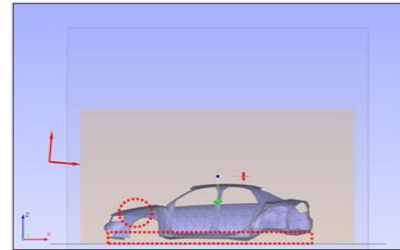


Fig. 1. Electric-field distribution of the mobility platform under meso-band (DS_10m) EMP excitation ($\theta = 85^\circ$, vertical pol.). Field concentration is observed near the front-end and lower-body regions.

IV. CONCLUSION

Simulation results based on measured EMP sources show that electric-field distribution in mobility platforms strongly depends on frequency band, ground condition, polarization, and source distance. Asphalt modeling provides realistic evaluation conditions, while reduced separation distance significantly increases coupling intensity. The proposed approach enables systematic identification of EMP-vulnerable regions and supports practical shielding and electromagnetic hardening strategies for advanced mobility systems.

ACKNOWLEDGMENT

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Degrading Leakage Signal Reconstruction Quality via Font Boundary Noise Injection in TMDS Interfaces

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Abstract— With the continuous advancement of eavesdropping technologies, electromagnetic video reconstruction techniques such as TEMPEST pose a significant threat. This risk is further amplified by AI-driven recognition tools that can automatically extract text from reconstructed images. Conventional hardware countermeasures are often cost-prohibitive. This paper proposes a cost-effective software-based defense based on readability-preserving boundary perturbation. We introduce a novel noise injection method that specifically disrupts font boundary continuity while keeping the text visually readable. Experimental results demonstrate that the proposed approach effectively degrades the quality and machine-readability (e.g., character error rate) of reconstructed images from Video Display Units (VDU) utilizing TMDS.

Keywords— TEMPEST, TMDS, Software-based Countermeasure

I. INTRODUCTION

Electromagnetic (EM) emanations from digital display links can enable remote reconstruction of on-screen content, which modern attackers further enhance via phase-aligned coherent averaging and automated text recognition [1]. Our experiments show that reconstructed text readability is largely governed by stroke boundaries; interior-only perturbations often remain legible when edge transitions and connectivity are preserved, as boundary structures survive reconstruction as high-frequency components. We therefore propose a rendering-level boundary-band defense that preserves a solid black text core while perturbing the font boundary zone. This demonstrates that disrupting edge continuity is the decisive factor in degrading reconstruction quality under modern averaging- and recognition-based attacks.

II. METHODOLOGY

We evaluate a rendering-level defense by comparing two noise-injection strategies across font sizes of 40, 32, and 28 pt. For each text sample, we generate a strict binary glyph mask via hard thresholding to eliminate anti-aliasing, while preserving a solid black text core to maintain legitimate on-screen readability.

The Interior-Noise baseline injects high-contrast binary (0/1) noise into the glyph interior while leaving a spatial margin to ensure the outer stroke boundary remains unchanged. In contrast, the proposed Boundary-Band Noise method perturbs a narrow character-background transition zone—defined via morphological erosion and dilation of the glyph mask—using patterned masking, as illustrated in Fig. 1.

Reconstructed outputs are obtained using a fixed attacker pipeline applying 40-frame phase-aligned coherent averaging. To quantify the security-usability trade-off, the Character Error Rate (CER) is computed for the original rendered images and the reconstructed leakage signals from both normal and security fonts, as summarized in Table I. Text extraction via the Qwen 2.5-VL 3B vision-language model (VLM) is used to assess if boundary-band disruption more effectively degrades machine-readability compared to interior-only distortion while preserving user legibility, as illustrated in Fig. 2.

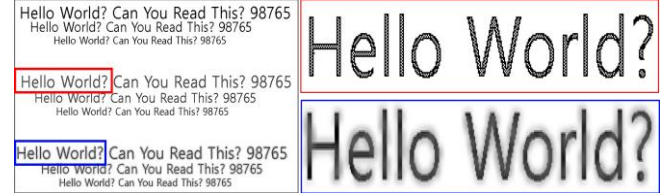


Fig. 1. Noise strategies: input samples (left) and visual comparison of Interior Noise (top right) vs. Boundary-Band Noise (bottom right).



Fig. 2. Reconstructed images: samples (left) and visual comparison of Interior Noise (top right) vs. Boundary-Band Noise (bottom right).

TABLE I. AVERAGE CER COMPARISON BY STRATEGY

Strategy	Character Error Rate (Error/Total)	
	Original Image	Reconstructed Image
Baseline (Clean)	0.0000 (0/93)	0.0000 (0/93)
Interior Noise	0.0000 (0/93)	0.0000 (0/93)
Boundary-Band Noise	0.0000 (0/93)	1.3656 (127/93)

III. RESULTS AND DISCUSSION

Experiments were conducted in an anechoic chamber using a 1920×1080 FHD monitor at a leakage frequency of 742.5 MHz. As summarized in Table I, strategy-wise aggregated CER (Error/Total) shows that all strategies maintained a 0.0000 CER for original images, confirming that legitimate usability is fully preserved.

Significant disparities emerged in reconstruction: Interior Noise failed to disrupt VLMs (0.0000 CER), whereas Boundary-Band Noise achieved 1.3656 (127/93) CER by inducing extraneous character generation. This confirms that disrupting high-frequency edge continuity is the pivotal factor in TEMPEST defense. As illustrated in Fig. 2, even high-intensity interior noise is ineffective if boundaries remain intact, whereas Boundary-Band Noise successfully neutralizes automated extraction.

IV. CONCLUSION

Experiments identify boundary-band disruption as the primary driver of reconstruction resistance. Unlike ineffective Interior Noise (0.0000 CER), Boundary-Band Noise achieved 1.3656 CER by inducing extraneous characters. This confirms that secure-by-design rendering is a cost-effective, software-based TEMPEST defense for TMDS-utilizing VDUs, as it requires no expensive physical shielding.

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Mode-Separated Measurement for Self-Interference Suppression in Active Electromagnetic Analysis

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Abstract—This paper proposes a mode-separated measurement method to suppress self-interference in Active Electromagnetic Analysis (EMA). By irradiating and receiving different propagation modes on transmission paths, direct transmitter–receiver coupling is reduced. Experimental results demonstrate carrier suppression of up to 17 dB and SNR improvements of up to 20 dB, confirming enhanced side-channel information retrieval performance.

Keywords— side-channel analysis, active EMA

I. INTRODUCTION

Active Electromagnetic Analysis (EMA), which actively irradiates electromagnetic (EM) waves and retrieves secret key information of the cryptographic module from the reflected waves, has been reported as a threat in addition to conventional passive side-channel analysis [1]. However, in Active EMA, direct coupling of the transmitted wave to the receiver causes self-interference, degrading key recovery performance. To address this self-interference, this paper proposes a measurement method for Active EMA conducted through transmission paths such as power lines and communication lines, in which different transmission modes are used for transmission and reception to suppress direct coupling and improve key recovery performance.

II. MODE-SEPARATED ACTIVE EMA

In Active EMA, an incident EM wave is reflected after being amplitude-modulated by temporal impedance variations of nonlinear components near a cryptographic IC induced by cryptographic operations. Because of impedance mismatches inside and outside the device, such as cable connections, part of the reflected wave is converted into a different propagation mode.

The proposed method exploits this mode conversion to suppress self-interference by employing different propagation modes for signal injection and reception (Fig. 1). Specifically, the probes are arranged so that the EM waves on the transmission path are excited and detected as either common mode (CM) or differential mode (DM). By observing different modal components, the method mitigates direct transmitter-receiver coupling.

III. EXPERIMENTAL VALIDATION

To emulate side-channel information leakage from a cryptographic module, a microcontroller was configured to continuously output a 31.4-kHz square wave, thereby inducing periodic impedance variations in the target device (Fig. 2). Measurement probes were placed on the device power line to excite and detect either CM or DM. In the experiments, 664 MHz—at which a modulated reflected wave was observed—was selected as the frequency for EM irradiation.

Tables I and II summarize the evaluation results derived from the observed reflected waves. When the measurement configuration was designed so that different propagation modes were injected and received, the received carrier component was reduced by up to 17.47 dB, while the sideband level increased by up to 15.65 dB. Compared with same-mode operation, the use of different propagation modes significantly improved modulation depth and SNR. In particular, SNR improved by

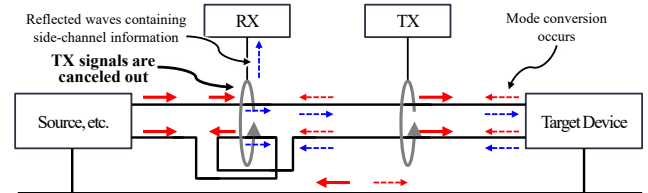
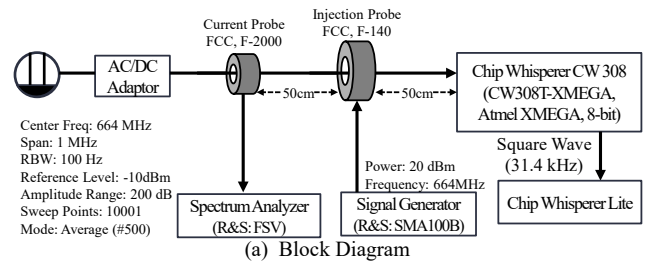
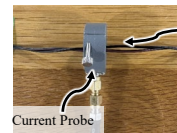


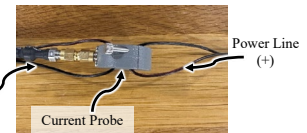
Fig. 1. Measurement method using different transmission modes



(a) Block Diagram



(b) CM clamping method



(c) DM clamping method

Fig. 2. Experimental Setup

TABLE I. EVALUATION RESULTS FOR EM IRRADIATION IN CM

Case	Inj.	Rec.	Carrier [dBm]	Sideband [dBm]	Mod. [%]	CNR [dB]	SNR [dB]
A	CM	CM	-13.58	-88.34	0.0365	88.12	19.37
B	CM	DM	-31.05	-98.06	0.0892	86.46	25.47
Difference (A-B)			-17.47	-9.72	0.0527	-1.66	6.10

TABLE II. EVALUATION RESULTS FOR EM IRRADIATION IN DM

Case	Inj.	Rec.	Carrier [dBm]	Sideband [dBm]	Mod. [%]	CNR [dB]	SNR [dB]
C	DM	DM	-23.26	-110.91	0.0083	87.92	6.29
D	DM	CM	-30.64	-95.26	0.1176	84.99	26.39
Difference (C-D)			-7.38	15.65	0.1093	-2.93	20.10

Inj.: Injection mode, Rec.: Reception mode, Mod.: Modulation index, CNR: Carrier-to-noise ratio, SNR: Signal-to-noise ratio

6.10 dB under CM irradiation and by 20.10 dB under DM irradiation. These results demonstrated that the mode-separated measurement configuration effectively suppressed direct transmitter–receiver coupling and enhanced information reconstruction performance.

ACKNOWLEDGMENT

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Quantitative Analysis of HT-Based Signal Hijacking System

under Self-Interference-Dominated Conditions

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Abstract—This work investigates hardware Trojan-induced backscatter signals from a commercial PCB under self-interference-dominated conditions. Frequency domain analysis is performed to identify harmonic components, and a harmonic-to-floor ratio (HFR) metric is introduced for quantitative evaluation. Measurement results demonstrate observable harmonics under practical experimental conditions.

Keywords—Hardware Trojan, Backscatter, Self Interference, Electromagnetic Side-Channel Attack

I. INTRODUCTION

This paper investigates backscattered signals from a hardware Trojan (HT) embedded in a commercial PCB under self-interference (SI)-dominated conditions [1]. While previous studies used custom-designed PCBs, this work evaluates signal recovery feasibility on a commercial PCB. An RF measurement system with digital signal processing is employed to recover the HT-induced signal, and FFT-based quantitative analysis is performed to evaluate its observability.

II. METHODOLOGY

A. HT-BASED SIGNAL HIJACKING SYSTEM

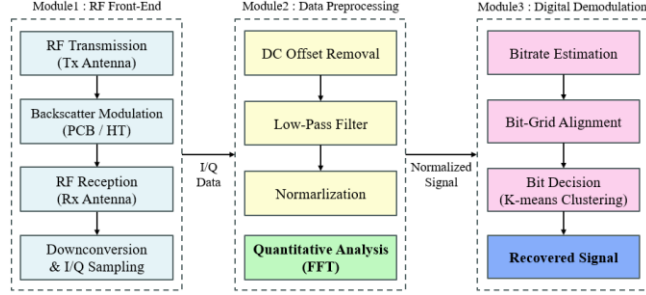


Fig. 1. Overall signal processing architecture for HT-based backscatter signal reception, demodulation, and quantitative analysis

Fig. 1 shows the signal processing architecture consisting of three modules: RF front-end, Data preprocessing, and Digital demodulation. A continuous-wave RF signal illuminates a target PCB containing an HT, which generates amplitude-modulated backscatter signals. The received signal is downconverted and sampled as I/Q data using an SDR.

The preprocessing stage performs DC offset removal, low-pass filtering, and normalization to suppress SI. In the demodulation stage, bitrate estimation, bit-grid alignment, and K-means-based binary decision recover the digital signal.

B. SIGNAL MODEL AND PROCESSING

The received signal, dominated by self-interference from direct coupling between transmit and receive antennas, can be expressed as

$$s(t) = s_{SI}(t) + s_{HT}(t) + n(t), \quad (1)$$

where $s_{SI}(t)$ denotes the self-interference component, $s_{HT}(t)$ represents the HT-induced modulation, and $n(t)$ denotes noise.

A harmonic-to-floor ratio (HFR) metric is defined to evaluate HT-induced spectral observability, calculated as the difference between the harmonic peak and the adjacent spectral floor,

$$HFR = P_{harm} - P_{floor} \text{ (dB)}, \quad (2)$$

where P_{harm} denotes the maximum spectral magnitude within the harmonic region and P_{floor} represents the median spectral magnitude of neighboring regions.

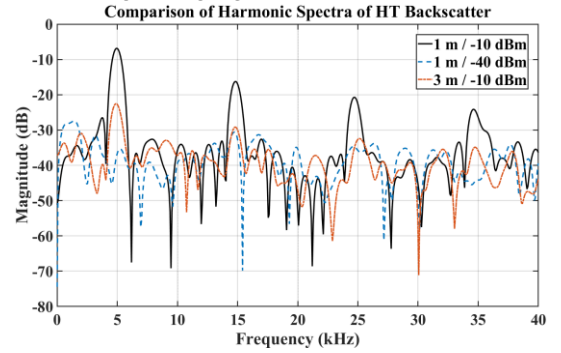


Fig. 2. Comparison of harmonic spectra of HT backscatter for different distances and excitation powers under a 1.8 GHz carrier with a 10kbps alternating bit pattern (1010)s

III. RESULTS AND DISCUSSION

The measured harmonic spectra in Fig. 2 show observable harmonics under SI-dominated conditions. Quantitative HFR comparison indicates that signal recovery is achievable when HFR exceeds approximately 10 dB (28.2 dB at 1m/-10dBm, 13.6 dB at 3m/-10dBm), whereas lower HFR (4.2 dB at 1m/-40dBm) results in unsuccessful recovery.

IV. CONCLUSION

This paper presented a signal processing architecture for HT-induced backscatter analysis from a commercial PCB under SI-dominated conditions. The HFR metric enables quantitative evaluation of spectral observability, and results demonstrate signal recovery feasibility under practical conditions. Future work will extend this analysis to broader experimental scenarios.

ACKNOWLEDGMENT

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Update on IEC SC 77C Activity 2026

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Abstract– The International Electrotechnical Commission (IEC) Subcommittee (SC) 77C develops Electromagnetic Compatibility (EMC) standards for high power transient phenomena. This paper reports on the International Standards that have been published recently, focusing on development of the environment definition for high-altitude electromagnetic pulses (HEMP), as well as test methods for protective devices for HEMP. Additionally, this paper will present an overview of the New Projects that are planned to bring IEC standards into harmony with Korean standards for protection methods, improve the test methods for sensing equipment, and develop a new environment standard for geomagnetic storms.

Keywords– standards, electromagnetic compatibility (EMC), high power transient phenomena, High-altitude Electromagnetic Pulse (HEMP), Intentional Electromagnetic Interference (IEMI).

I. INTRODUCTION to IEC SC 77C

The International Electrotechnical Commission (IEC) develops International Standards to provide definitions and to describe agreed methods for processes and services. Technical Committee (TC) 77 deals with Electromagnetic Compatibility (EMC) standards development. Subcommittee (SC) 77C develops International Standards associated with high power transient phenomena specifically, including High-altitude Electromagnetic Pulses (HEMP), Intentional Electromagnetic Interference (IEMI), and Geomagnetically Induced Currents (GIC) from solar activity. Lightning and other transient phenomena are excluded from the scope of SC 77C.

IEC SC 77C has developed 22 publications including general guides, environment definitions, test methods, and installation and mitigation guidelines. The subcommittee has 20 participating member nations, and a further 20 observer member nations. The work is carried out by subject matter experts from national standards bodies from the participating member nations. The subject matter experts are from a variety of backgrounds including academia and industry.

II. RECENT PUBLICATIONS

In the last two years, IEC SC 77C has developed and published two documents. On May 14th 2025, IEC 61000-2-9 [1] was published. This was a second edition of the document, covering the definition of the electromagnetic environment for the radiated aspects of HEMP. Key updates include an improved definition of the E1 HEMP definition and enhanced guidance for spatial variation. On July 25th 2025, an amendment to IEC 61000-4-23 edition 2 [2] was published, which covers test and measurement techniques for protective devices for HEMP and other radiated disturbances. A major addition to this standard was a new annex describing how to perform quick measurements of coupling into rooms in buildings. This new method makes use of field stirring for quick results.

III. NEW PROJECTS

A. New Work Proposals

At the most recent Plenary Meeting for IEC SC 77C, held in New Delhi, India and online via video-conferencing, in October 2025, some ideas for new projects were presented. The majority of these

projects have since been proposed officially, as described in the following sections.

B. IEC 61000-2-X

This is a new environment standard to describe the electromagnetic environment created by Geomagnetic Induced Currents (GIC) from solar activity. This is in response to the change in scope for IEC SC 77C in 2017, which was updated to include EM disturbances from solar activity. The standard will provide guidance for system manufacturers on the approximate levels to which their systems must be tested to demonstrate immunity.

C. IEC 61000-4-24

This standard describes testing and measurement techniques for protective devices for HEMP conducted disturbances. The planned maintenance work would add an insertion loss measurement method for HEMP protection filters, as well as lifecycle checks.

D. IEC 61000-4-33

This standard provides information on testing and measurement techniques for high power transient parameters. The planned maintenance work will update information on measurement chains, VNA-based vector calibration, consistent FD-TD processing, current measurements, and documentation requirements.

E. IEC 61000-5-5

This standard provides installation and mitigation guidelines for the specification of protective devices for HEMP conducted disturbances. The planned maintenance work will add information on isolating devices and RF limiters for High-Power Electromagnetics (HPEM).

F. IEC 61000-6-6

This is a generic standard that provides information on HEMP immunity for indoor equipment. The planned maintenance work will include specific definitions for EME at each test port, and harmonize the standard with ITU-T SG5 K.78.

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Fiber-Optic Technology For HEMP Protection

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Abstract – High-altitude electromagnetic pulse (HEMP) can severely impact critical infrastructure, requiring effective protection of signal and control lines. While shielding and filtering are widely used, they cannot fully eliminate conductive paths. Fiber-optic links provide inherent decoupling by removing electrical conduction routes, preventing EMP-induced currents from entering protected facilities. This paper briefly reviews fiber-optic-based HEMP protection methods, relevant standards, and key considerations such as system integration, and electromagnetic immunity. Practical implementation issues and future improvement needs are summarized.

Keywords– HEMP, Protection, Fiber Optic, Converter

I. INTRODUCTION

HEMP poses a major threat to military and national infrastructure. Conventional protection relies on shielding for radiated effects and filters for conducted effects, but some systems require complete decoupling to eliminate conductive paths. Fiber-optic transmission using optoelectronic converters provides this capability by removing electrical continuity. As modern systems increasingly depend on high-speed digital communication and satellite-based networks, the need for non-conductive signal interfaces has become more critical. This paper outlines the role of fiber-optic decoupling in HEMP protection and highlights essential technical considerations.

II. OPTOELECTRONIC CONVERTER FOR HEMP PROTECTION

Fiber-optic links are effective for HEMP-hardened facilities because they do not conduct EMP-induced currents. This makes them suitable for preventing electromagnetic energy from entering shielded environments.

A. IEC 61000-5-6

The standard classifies optical links as decoupling devices that remove conductive paths, supporting their use in EMP mitigation.

B. MIL-STD-188-125-1

The standard mandates conversion of all voice and data lines to fiber optics outside the protection boundary and requires non-conductive fiber cables for shield penetration.

C. Fiber-Optic-Based HEMP Protection

In a typical configuration, telephone and data signals are converted to optical signals outside the shielded enclosure. These optical signals are transmitted through fiber-optic cables into the protected area, where they are reconverted

into electrical signals. Multiple electrical signals may be multiplexed into a single optical channel. Optoelectronic converters require dedicated power supplies, and proper design of the fiber-optic penetration point is essential to maintain shielding integrity.

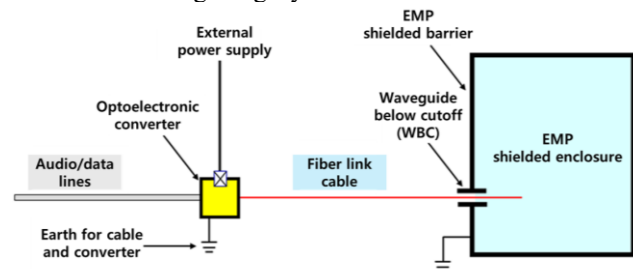


Fig. 1. Fiber Optic Cable Isolation Device

D. Considerations for Fiber-Optic HEMP Protection
Applying fiber-optic technology may alter electromagnetic noise distribution and introduce new failure modes. Key considerations include:

- Electromagnetic effects (EMP/EMI/lightning)
- Functional characteristics of power and signal lines
- Integration with existing equipment and hardening elements
- Reliability, failure analysis, and electromagnetic immunity
- Maintaining electromagnetic topology integrity.

III. CONCLUSION

Fiber-optic decoupling is increasingly important as modern systems require high-speed and satellite communication interfaces. This paper summarized essential standards, implementation methods, and technical considerations for applying optoelectronic converters in HEMP protection. Continued research is needed to improve reliability and electromagnetic immunity in advanced communication environments.

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Feasibility of Pulse Detection Using Frequency-Drift Signatures in Swept-Mode Spectrum Analyzer

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Abstract– This paper presents a simple method to estimate the repetition period and pulse width of RF pulse trains using only a conventional swept-tuned spectrum analyzer. In swept mode, the mixer–LO–IF filter chain creates a deterministic coupling between sweep time and the observed spectral pattern, violating the usual time–frequency independence assumption. By modeling this chain, the method relates sweep settings (sweep time, span, IF bandwidth) to the measured pattern and inverts the relationship to extract pulse period and duty cycle.

Keywords– pulse, spectrum analyzer

I. INTRODUCTION

Quasi-Peak mode of an EMI receiver is essential for the effective analysis of transient pulse signals during Electromagnetic Interference (EMI) measurements, the accurate identification and measurement of pulse signals are critical factors that determine system performance in various RF applications [1]. However, in standard swept-tuned spectrum analyzers, the interaction between the sweep time and the pulse period creates a deterministic coupling, making accurate parameter extraction difficult. This study models this interaction to formulate the relationship between the sweep time and the observed spectral patterns, seeking a solution to precisely estimate the pulse period and duty cycle without the need for separate dedicated equipment.

II. METHODOLOGY

Figure 1 presents the results of leakage electromagnetic waves measured from a 4-cylinder gasoline vehicle using the sweep mode of a spectrum analyzer. Since the vehicle was idling at a constant RPM, the ignition period from the engine spark plugs should remain nearly constant. However, because a swept-mode spectrum analyzer scans frequency sequentially over time, a mismatch between the sweep time and the actual pulse period causes the observed peak position to shift along the frequency axis in each sweep. Assuming a periodic pulse train, this frequency-shift pattern can be mapped onto a 2D coordinate system, where the accumulated timing mismatch forms a tilted diagonal trajectory. This behavior is analogous to sampling oscilloscope, which reconstructs high-speed signals using low-rate sampling. Thus, the peak shift in Fig. 1. Is not random noise but a deterministic result of the difference between the sweep time and pulse period. As illustrated in Fig. 2, this frequency-shift pattern can be mapped onto a 2D coordinate system, where the accumulated timing mismatch forms a tilted diagonal trajectory. The peak shift is therefore not random noise but a deterministic result of the difference between t_{swp} and t_{pls} , analogous to the principle of a sampling oscilloscope. In this study, we estimate the pulse period and width by inversely analyzing this slope, correcting the sweep-time deviation, and realigning the time axis. The results are compared with pulse analyzer measurements to evaluate the feasibility of extracting pulse timing information from

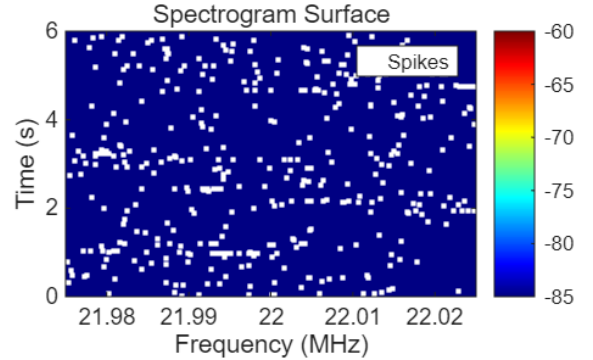


Fig. 1. Frequency drifting effect on spectrogram analyzer by wideband pulse signal.

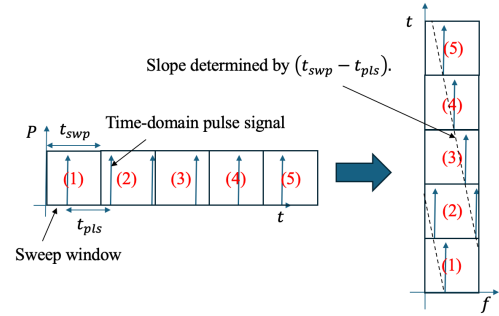


Fig. 2. Conversion between time domain and frequency domain of frequency drifting effect.

swept-mode spectrum analyzer data [2].

III. CONCLUSION

By modeling a spectrum analyzer as a time-frequency sampler, this study enables accurate pulse parameter estimation through the slope analysis of frequency shift patterns. This algorithmic approach eliminates the need for specialized hardware, validating the analyzer's expanded utility as a cost-effective solution for time-domain pulse measurement.

ACKNOWLEDGMENT

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Effective Propagation Constant Characterization of Cable Assemblies from Multi-Length S-Parameter Measurements

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Abstract– This paper presents a measurement-driven method for extracting effective attenuation and phase constants of cable assemblies from single-ended S-parameter data. Transfer matrices are constructed from measured responses, and raw propagation information is obtained from the dominant eigenvalue of a similarity matrix formed from two cable lengths. The phase constant is recovered through phase unwrapping, while the attenuation constant is stabilized using a physics-constrained fitting model. The proposed framework is demonstrated using multi-length HDMI cable assemblies.

Keywords– HDMI, Similarity Matrix, Propagation Constant, De-embedding

I. INTRODUCTION

Practical twisted cable assemblies show frequency-dependent attenuation and phase progression, which are important for signal integrity and electromagnetic propagation analysis. However, direct propagation-constant extraction is difficult because measured assemblies include connector transitions, impedance mismatch, and local discontinuities. Therefore, the extracted values should be regarded as effective propagation constants of the cable assembly, not intrinsic material constants. To overcome this, this paper proposes a stable multi-length measurement-based extraction framework, using HDMI cable assemblies as a representative practical twisted interconnect example.

II. METHODOLOGY

For two cable measurements with different lengths, the corresponding transfer matrices are denoted by T_1 and T_2 , which are obtained from measured S-parameters using standard parameter conversion [1]. The similarity matrix between the two transfer matrices is defined as shown in (1).

$$M = T_2 \cdot T_1^{-1} \quad (1)$$

Let λ denote the dominant eigenvalue of M . The effective propagation constant is then expressed as given in (2) [1][2].

$$\gamma_{\text{ef}} = \alpha + j\beta \approx \frac{1}{\Delta L} \ln |\lambda| \quad (2)$$

Here, ΔL denotes the cable length difference. From (2), the raw attenuation constant is obtained as given in (3).

$$\alpha_{\text{raw}} = \frac{1}{\Delta L} \ln |\lambda| \quad (3)$$

Similarly, the phase constant is determined from the phase of the dominant eigenvalue, as shown in (4).

$$\beta = \frac{1}{\Delta L} \text{unwrap}(\angle \lambda) \quad (4)$$

Since α_{raw} may include local fluctuations caused by measurement noise and residual non-idealities, it is stabilized using the physics-

constrained form given in (5).

$$\alpha_{\text{f}} = c_1 \sqrt{f} + c_2 f \quad (5)$$

Here, the $\sqrt{f}$ and f terms represent the dominant conductor-loss and dielectric-loss trends, respectively.

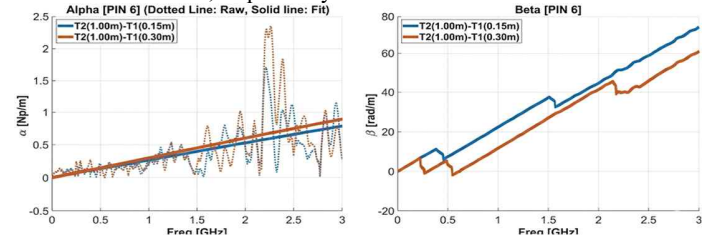


Fig. 1. Extracted effective propagation constants from multiple HDMI cable length-difference pairs: (left) raw and fitted attenuation constant α , and (right) phase constant β .

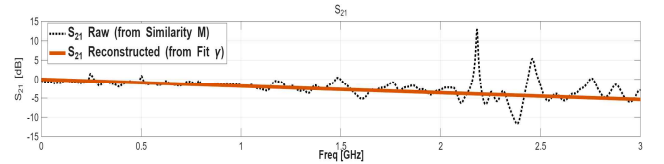


Fig. 2. Comparison of raw and reconstructed S_{21} for the 0.7m HDMI transmission line model

III. RESULTS AND DISCUSSION

Fig. 1 presents the effective propagation constants extracted from multiple HDMI cable length pairs. Compared with the raw attenuation constant, the fitted α exhibits a smoother and more physically consistent trend. The extracted β also maintains stable phase progression after phase unwrapping. Fig. 2 shows that the reconstructed S_{21} better represents the dominant transmission behavior than the raw extracted result, supporting the stability of the proposed method

IV. CONCLUSION

A similarity-matrix-based method for extracting effective attenuation and phase constants of twisted cable assemblies was presented. Using multi-length HDMI measurements, the proposed framework was shown to provide a compact and physically consistent description of cable transmission behavior.

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Design and Analysis of a High-Voltage Broadband Folded Sinuous Antenna

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Abstract– A broadband high-voltage sinuous antenna operating over 100 MHz - 1 GHz is designed and fabricated for time-domain shielding effectiveness (TDSE) measurement of electromagnetic pulse (EMP) protection facilities. The antenna exhibits a reflection coefficient below -10 dB over most of the operating band and achieves a gain greater than 0 dBi at 100 MHz. High-voltage pulse experiments with a maximum voltage of 6 kV and a rise time of 100 ps confirm stable operation without dielectric breakdown. In addition, a dynamic range exceeding 80 dB is achieved over most of the frequency band, demonstrating the feasibility of the proposed antenna for TDSE measurements.

Keywords– sinuous antenna, time-domain shielding effectiveness (TDSE), EMP, broadband antenna, high-voltage antenna,

I. INTRODUCTION

Conventional shielding effectiveness (SE) testing for electromagnetic pulse (EMP) protection facilities, based on MIL-STD-188-125-1 [1], requires excessive time and cost because it relies on single-frequency measurements in the frequency domain. To overcome this limitation, time-domain shielding effectiveness (TDSE) techniques using broadband pulses have been studied [2], [3]. Pulse-based TDSE uses sub-nanosecond impulse signals and broadband antennas to evaluate shielding effectiveness over a wide frequency range.

In this paper, a broadband antenna with a high dynamic range was designed to implement a TDSE measurement method. The operating characteristics of the proposed antenna were experimentally validated.

II. METHODOLOGY

A planar sinuous antenna operating in the 100 MHz - 1 GHz electric-field test band specified by the MIL-STD-188-125-1 was designed. To meet measurement-space requirements, the antenna size was limited to less than 1 m × 1 m with a depth of less than 0.3 m. In addition, the antenna was designed to withstand high-voltage pulse excitation and to achieve a gain greater than 0 dBi at 100 MHz to ensure a sufficient dynamic range. As shown in Fig. 1, edge truncation and a folded structure were applied to improve the low-frequency gain, while sufficient spacing between the arm patterns was maintained to prevent dielectric breakdown. The antenna was simulated using the electromagnetic simulation tool CST.

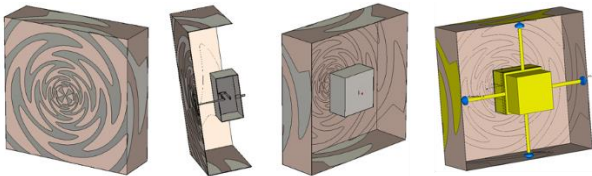


Fig. 1. Three-Dimensional Model of the Designed Sinuous Antenna

A microstrip-type balun was designed for the antenna feed to enable impedance transformation between the 133 Ω planar feed line and the 50 Ω microstrip line. The balun was implemented using an aluminum plate and air dielectric with a total length of 820 mm, and the antenna was miniaturized by applying a bent structure while maintaining the electrical length. In addition, CST simulations

confirmed the backlobe reduction effect achieved by applying a cavity, ferrite, absorber, cross-shaped conductive structure, and 50 Ω terminating resistors.

III. RESULTS AND DISCUSSION

Based on the results of design and simulations, a broadband high-voltage sinuous antenna was fabricated. As shown in Fig. 2, the antenna exhibits S_{11} values below -10 dB over most of the 100 MHz - 1 GHz frequency band. In addition, far-field measurements conducted at Korea Research Institute of Standards and Science confirmed that the antenna achieves a gain greater than 0 dBi at 100 MHz. In addition, a picosecond high-voltage pulse generator was fabricated using the PPG-5/16 picosecond pulse generation module from Megaimpulse, and it was connected to the transmitting antenna to evaluate the operating range of the transmitting and receiving antennas. The pulse generator provides a maximum output voltage of 6 kV with a pulse rise time of 100 ps. Experimental results show that no dielectric breakdown or significant loss occurred in the antenna even when high-voltage broadband pulses were applied. Furthermore, a dynamic range exceeding 80 dB was achieved over most of the frequency band.

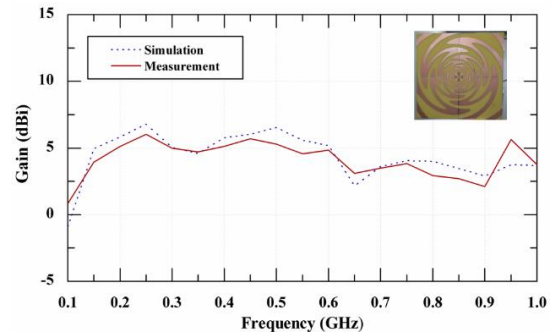


Fig. 2. Measured Gain of the Fabricated Antenna

IV. CONCLUSION

In this paper, a broadband high-voltage sinuous antenna suitable for TDSE measurements was designed and experimentally verified. The proposed antenna demonstrated stable operation under high-voltage pulse excitation up to 6 kV while maintaining broadband characteristics in the 100 MHz - 1 GHz band. In addition, a dynamic range exceeding 80 dB was obtained over most of the operating band. These results indicate that the proposed antenna can serve as an effective candidate for broadband high-voltage measurement systems for EMP protection facilities.

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Mode Competition in Relativistic Magnetron with TE11 Mode Diffraction Output

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Abstract– The relativistic magnetron is a compact and efficient high power microwave source. To match the relativistic magnetron with appropriate antennas, a TE11 mode diffraction output has been developed in S-band. A challenge of relativistic magnetrons is competition between the π mode and the $2\pi/3$ mode. This problem is studied using particle-in-cell simulations. Results show that mode competition is more frequent and complex at high voltages and magnetic field strengths.

Keywords–high power microwaves; relativistic magnetron; diffraction output; particle-in-cell

I. INTRODUCTION

The relativistic magnetron is an efficient high microwave source with an order of 1 GW output power in L- to C-band. Microwaves are produced from electron-wave interaction in a slow-wave structure. The microwaves can be extracted efficiently from the slow-wave structure by using a diffraction output and a circular waveguide output [1]. The common modes in a relativistic magnetron with diffraction output are the $2\pi/3$ mode and π mode. Each mode has a corresponding characteristic electron configuration (Fig. 1). A π mode to TE11 mode diffraction output has been developed to match the relativistic magnetron to antennas for high power microwave applications in S-band [2]. A challenge is to excite the π mode while avoiding sporadic excitation of the $2\pi/3$ mode. To explore causes of mode competition, the relativistic magnetron is simulated using particle-in-cell code at high voltages and magnetic field strengths.

II. SIMULATION RESULTS

The incoming voltage wave amplitude is stepped for different magnetic field strengths (Fig. 2). Each step is simulated separately for 50 ns. Once a mode is formed, the anode-cathode voltage drops because electrons start to cross the anode-cathode gap. The π mode is expected to dominate at lower voltages because the Buneman-Hartree voltage of the π mode is lower than that of the $2\pi/3$ mode [3]. At 0.3 T, the lower voltages produce the highest beam-to-microwave efficiencies and the corresponding electron configurations show that a π mode is formed. Surprisingly, at 0.4 T a $2\pi/3$ mode is formed below its Buneman-Hartree voltage (446 kV) and this phenomenon repeats at 0.5 T and 0.6 T. For each magnetic field strength, at moderate voltages the π mode is dominant, whereas at high voltages the $2\pi/3$ mode appears to compete with the π mode. The higher the voltage, the more frequent is this mode competition.

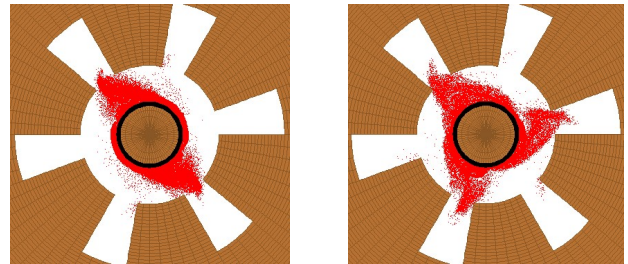


Fig. 1. Example electron configurations at 0.5 T. (Left) $2\pi/3$ mode. (Right) π mode.

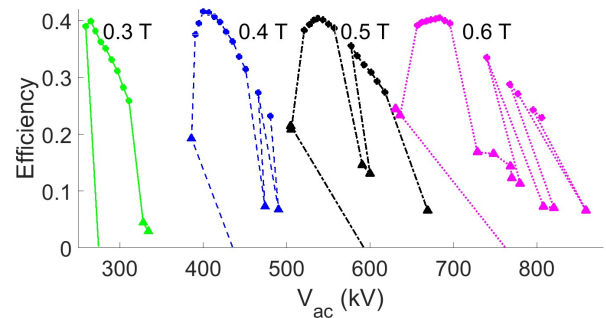


Fig. 2. Beam-to-microwave efficiency at different magnetic field strengths with respect to anode-cathode voltage V_{ac} . Circle and triangle data points correspond to π mode and $2\pi/3$ mode respectively. Lines represent stepping of the incoming voltage wave amplitude.

IV. CONCLUSION

The results of this work show that mode competition may become a greater challenge at high voltages and magnetic field strengths.

ACKNOWLEDGMENT

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setup for Testing the Performance of Low Noise Amplifiers at High Peak Power Pulses

Initial Experiments

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Abstract—We report on experiments where Low Noise Amplifiers (LNA) are subjected to short high peak power pulses while amplifying a usable signal. The behaviour of the LNA is studied by analyzing the impact on the low level ‘usable’ signal when subjected to a short pulse of higher power. It is found that LNAs based on gallium nitride (GaN) can take much higher peak power than those using gallium arsenide (GaAs) technology before they break. However, GaN based LNAs suffer from a much longer pulse recovery time than LNAs based on GaAs-technology. The pulse recovery time for a GaN based device can be several orders of magnitude longer than that of a GaAs based device.

I. INTRODUCTION

Low Noise Amplifiers (LNA) are used in many different types of front ends, from ordinary radio receivers and mobile telephones to advanced radar systems. The behaviour of LNAs when subjected to short pulses of high power is important in applications such as radar receivers and electronic warfare. A long pulse recovery time can have severe implications in radar and electronic warfare applications since the system is essentially ‘blind’ during pulse recovery. In order to study pulse recovery time the amplifier needs to be fed with a continuous low level signal as well as short, repetitive pulses. At very high power pulses the LNAs break and never recover [1].

II. EXPERIMENTAL SETUP

A schematic view of the experimental system can be seen in Figure 1. It consists of two signal generators where generator 1 feeds a 2 GHz signal into a pulsed amplifier capable of up to 5 kW peak power and 1 μ s long pulses. Signal generator 2 produces a continuous signal at 3 GHz and is protected from the pulsed, high power, signal by the use of a 40 dB attenuator and a bi-directional coupler. A power meter measures the power from the pulsed amplifier and the pulse, together with the output signal from the Device under Test (DUT), are monitored in real time using an oscilloscope. The spectrum analyzer is centered on the low level 3 GHz continuous signal after it has been amplified by the DUT.

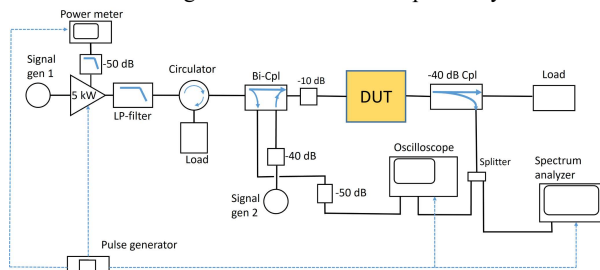


Fig. 1. Schematic picture of test setup used for the experiments reported here.

III. EXPERIMENTAL RESULTS

The amplitude of 3 GHz continuous signal is monitored by means of a spectrum analyzer. In Figure 2 the results from testing a GaAs and a GaN LNA are shown in time domain. The GaAs LNA recover very fast after the 1 μ s pulse until it eventually breaks at 36 dBm. GaN can take much higher power but exhibit a very slow pulse recovery, notice the different times scales.

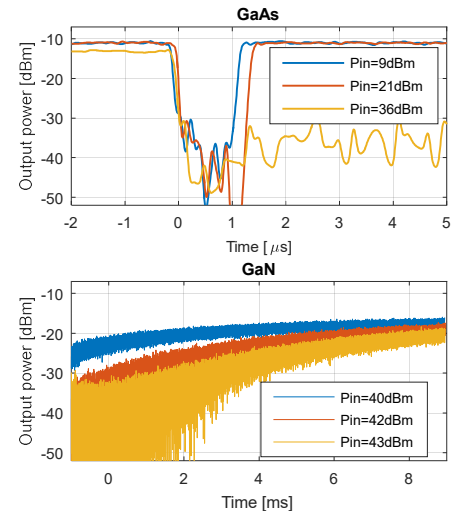


Fig. 2. Results from pulse testing of two different LNAs. The GaAs recover fast after the pulse but breaks at 36 dBm. The GaN can take up to 43 dBm but have a recovery time of milliseconds.

ACKNOWLEDGMENT

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High Power Microwave Mode-Synthesized Horn Array antenna

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Abstract– In this paper, the proposed structure was extended to an 8×8 array-based synthesized-mode horn antenna, achieving a gain of 28.1 dBi. Furthermore, by distributing only one-sixty-fourth ($1/64$) of the power per element compared to a single-source supply, the same performance was maintained, demonstrating the feasibility of a low-power-fed array-based HPM(High Power Microwave) system. **Keywords**– HPM, Mode-Synthesized, Horn antenna, Low power backlobe, Low power supplies

I. INTRODUCTION

Conventional high-power microwave (HPM) systems primarily employ parabolic or flat-reflector antennas [1-2]. However, such HPM systems exhibit strong back radiation, which may cause unintended interference or damage to friendly assets. In addition, because a single high-power source is typically used, the source structure becomes large and heavy.

Therefore, it is effective to employ an array of multiple horn antennas with high power-handling capability, enabling the realization of the performance equivalent to that of a single large high-power source by combining multiple low-power sources.

In this paper, an 8×8 horn antenna arrays operating at 2.45 GHz is designed by extending a previously developed 2×2 array horn antenna with H-plane mode synthesis. To suppress the back lobe, a waveguide short-circuited stub is incorporated into the structure. The design procedure and simulation results are presented and discussed.

II. METHODOLOGY

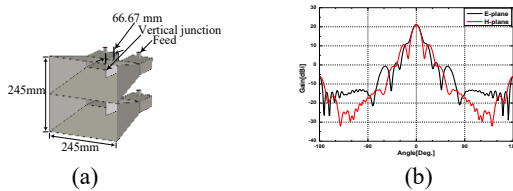


Fig. 1. Characteristics of the 2×2 Array Antenna

Fig. 1 illustrates the structure of a 2×2 mode-synthesized horn antenna array composed of four basic horn antennas, each providing a gain of 10 dBi. The synthesized array aperture was designed to be $245 \text{ mm} \times 245 \text{ mm}$ ($2\lambda \times 2\lambda$). Simulation results show a beamwidth of 25.4° and a realized gain of 16.1 dBi.

Compared with the E-plane, the H-plane exhibited a 3.5 dB higher side-lobe level of -10.4 dBi .

For comparison with the 4×4 synthesized horn antenna [3], simulations were conducted using an aperture size of $980 \text{ mm} \times 980 \text{ mm}$ ($8\lambda \times 8\lambda$), which is comparable to that of the

8×8 array configuration. The simulation results indicate that a gain of 28.1 dBi can be achieved.

The 8×8 array antenna offers the advantage that the input power required for each feed source can be reduced to $1/64$ of that required for a single-source feed configuration.

Meanwhile, to suppress the back lobe, two stages of waveguide short-circuited stubs were placed along the edges of the array. As shown in Fig. 2, the back-lobe and side-lobe levels were reduced by 6.2 dB and 1.84 dB compare to ref. [3], respectively, resulting in an FBR(Front-to-Back Ratio) improvement of 40.48 dB.

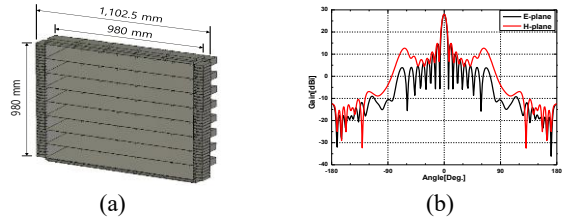


Fig 2. Characteristics of the 8×8 Array Antenna

III. RESULTS AND DISCUSSION

To design an HPM system using a reduced number of sources, the H-plane mode synthesis technique was adopted in a 2×2 horn array configuration. By extending this approach to an 8×8 mode-synthesized horn antenna with an aperture size of $980 \text{ mm} \times 980 \text{ mm}$ ($8\lambda \times 8\lambda$), a gain of 28.1 dBi was achieved. These results confirm the feasibility of realizing an HPM system using multiple low-power sources.

IV. CONCLUSION

By adopting the array mode-synthesis technique, it was verified that an 8×8 mode-synthesized horn antenna can achieve the feeding performance equivalent to that of a large single-source system using multiple small sources. In addition, the attachment of waveguide short-circuited stubs was shown to improve the FBR.

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Field-calibrated Intense Ka-band Pulse Measurements with Electro-optic Sensors

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Abstract– We presents a technique for measuring high-power electromagnetic pulses using electro-optic (EO) probes for real-time absolute electric-field measurement. Electric field pulses up to 100 kV/m generated by a Ka-band high-power amplifier are measured inside and outside a waveguide using two EO probes. A calibration method based on a standard waveguide structure is introduced for accurate sensor characterization. A heterodyne detection scheme is also used to down-convert the pulses, enabling determination of the repetition rate, carrier frequency, and peak electric field strength.

Keywords– High Power Microwave, Electro-optic Sensors, Millimeter-wave Measurement

I. INTRODUCTION

High-power millimeter-wave electromagnetic radiation has become increasingly important in modern electronic warfare. There is growing demand for highly directional, high-power electromagnetic sources for detecting and neutralizing targets such as unmanned aerial vehicles (UAVs). These signals are typically transmitted through waveguides and generated in pulsed form to enhance operational effectiveness. Conventional metallic probes used for field measurements can perturb the field under high-power conditions and may suffer from nonlinearity or damage. To overcome these limitations, the Korea Research Institute of Standards and Science (KRISS) has developed two electro-optic sensors: an ultra-miniaturized all-dielectric probe comparable in size to an optical fiber, and an induced-electrode-type probe integrated on an optical waveguide. These probes are placed inside and outside the waveguide to enable accurate calibration and conversion of the measured signals into absolute electric field values.

II. EXPERIMENTAL METHOD

A Ka-band signal with an output power of 40 dBm is generated using a commercial signal generator and a GaN-based solid-state amplifier and transmitted through a WR-28 waveguide. An ultra-miniaturized all-dielectric EO probe developed at KRISS is positioned at the center of the waveguide aperture, as shown in Fig. 1(a) [1]. The probe can be precisely inserted along the propagation direction using a motor-controlled translation stage.

In an ideal waveguide, the absolute electric field can be analytically determined from the applied microwave power, wavelength, waveguide impedance, and waveguide dimensions [2]. However, reflections at the aperture produce a standing-wave pattern, resulting in a half-wavelength periodic electric-field distribution along the waveguide. By averaging the measured standing-wave field and matching it with the calculated value, the probe can be calibrated.

After calibration, the probe can measure the absolute electric field outside the waveguide. This enables a two-tier calibration method: although the induced-electrode-type EO probe provides higher sensitivity, its larger size prevents insertion into the waveguide. By placing it at the waveguide aperture and referencing the calibrated dielectric probe, the absolute electric field at that location can be determined [2].

III. EXPERIMENTAL RESULTS

The pulsed 30 GHz electric field measured using the induced-electrode-type EO probe placed at the waveguide aperture, where the absolute electric field is known, is shown in Fig. 1(c). Owing to its high sensitivity, the electric-field signal can be detected in real time using a standard oscilloscope. To allow detection with a low-bandwidth oscilloscope, the 30 GHz signal is down-converted using a heterodyne scheme and observed as a 10 MHz carrier, as shown in the inset of Fig. 1(c). The microwave signal was generated as a square pulse with a duration of 10 μ s and measured 20 mm above the waveguide aperture. The measured peak-to-peak voltage of 890 mV corresponds to an electric field of approximately 30 kV/m.

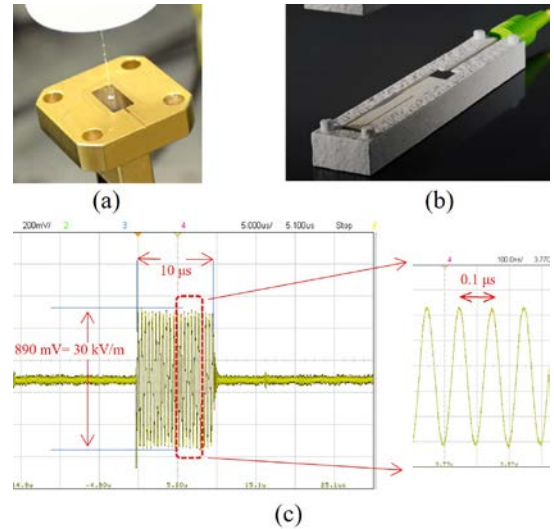


Fig. 1. (a) all-dielectric micro EO probe over WR-28, (b) electrode-coupled waveguide EO probe, (c) measured electric field pulse

IV. CONCLUSION

A technique for measuring high-power millimeter-wave electromagnetic fields based on absolute electric-field evaluation has been presented. Using two electro-optic (EO) probes, a two-tier calibration method was implemented to accurately determine the electric field strength. The approach enables reliable measurement of pulsed high-power millimeter-wave radiation and real-time evaluation of signals such as those used in radar systems, providing a practical solution for electronic warfare applications.

ACKNOWLEDGMENT

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Analysis of the Correlation Between Near-Field Distribution, Radiation Patterns, and Main Reflector Curvature Errors in HPM Reflector Antenna

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Abstract– This paper presents the design and experimental validation of an axially displaced Cassegrain (ADC) dual-reflector antenna for high-power microwave systems, in which the rotation axis of the sub-reflector is displaced. For reliable high-power operation, the effects of curvature errors caused by possible deformation of the main reflector during the fabrication process are quantitatively investigated based on the near-field electric field distribution. Prototypes with different curvature error regions are fabricated, and the resulting radiation characteristics are analyzed and verified through simulations and measurements.

Keywords– ADC dual-reflector antenna, high-power microwave systems, curvature errors, near-field.

I. INTRODUCTION

Dual-reflector antennas are widely used for high-power applications due to their high-gain characteristics. However, as the operating frequency and antenna size increase, deformation of the main reflector during fabrication may occur, leading to degradation in antenna gain and aperture efficiency. In this paper, the effects of main reflector curvature errors are quantitatively analyzed by comparing the near-field electric field distribution and far-field radiation characteristics.

II. ANTENNA DESIGN

The antenna used in this study is an ADC dual-reflector antenna that suppresses re-illumination of the feed antenna, thereby improving aperture efficiency and power handling capacity compared to a conventional Cassegrain reflector antenna. Fig. 1 shows the geometry of the proposed ADC dual-reflector antenna. An RF window was inserted inside the feed antenna to electrically isolate the SF_6 gas from the vacuum [1].

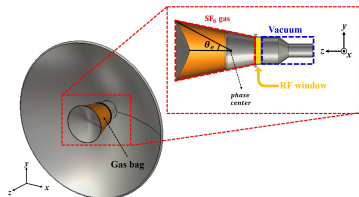


Fig. 1. Proposed ADC dual-reflector antenna [1].

III. MEASUREMENT RESULTS

Fig. 2(a) shows the fabricated prototypes with different curvature error regions in the main reflector. In the first prototype, a curvature error occurred in the radial region $6.4\lambda_0$ – $8\lambda_0$ from the reflector center, while the second prototype eliminated this error. Fig. 2(b-c) present the radiation patterns in the E – and H – plane, respectively.

The second prototype exhibits radiation patterns similar to the simulation results. As shown in the near-field measurements in Fig. 3, it also presents a uniform E-field distribution consistent with the simulation results. In contrast, the first prototype shows a significant decrease in E-field intensity at the location where curvature errors occurred.

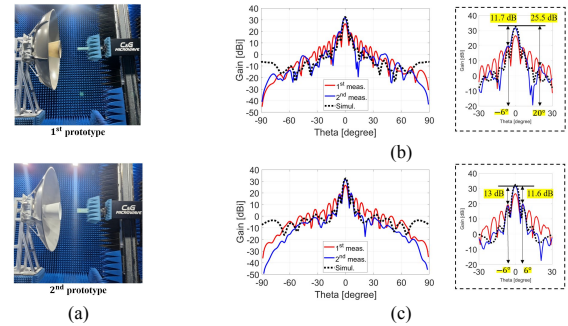


Fig. 2. Simulation and measurement radiation patterns of the ADC dual-reflector antenna. (a) Set up of 1st, 2nd prototype. (b) E – plane. (c) H – plane.

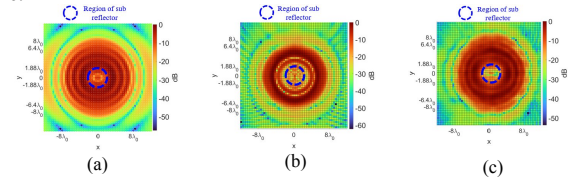


Fig. 3. Simulation and measurement E-field strength distribution. (a) Simulation. (b) 1st prototype. (c) 2nd prototype.

IV. CONCLUSION

This study analyzes the correlation between near-field distribution and antenna radiation characteristics with respect to main reflector curvature errors. Curvature error locations generated during reflector fabrication are quantitatively identified through near-field measurements. Eliminating these errors verified the antenna performance and confirmed the feasibility of high-power operation.

ACKNOWLEDGMENT

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Metasurface-Based Microwave Sensor for Multilayer Skin Phantom Measurement

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Abstract– This study evaluates S_{11} of a metasurface microwave sensor for multilayer skin phantom detection at 10–50 mm using a NanoVNA–Raspberry Pi system. Without tissue, resonance occurs at 2.38 GHz, with S_{11} improving from -15.587 to -21.036 dB as distance increases. At 1.82 GHz, the muscle layer varies from -21.371 to -12.147 dB. The muscle–fat layers shifts resonance to 2.026 GHz (-24.276 to -17.900 dB). The full multilayer structure shows the strongest response at 1.82 GHz (-34.707 to -10.938 dB), confirming effective tissue differentiation.

Keywords– microwave sensor, metasurface, skin layer phantom

I. INTRODUCTION

Microwave sensing is increasingly used for non-invasive biomedical monitoring. Microstrip antenna sensors detect tissue variations via electromagnetic interaction with muscle, fat, and skin layers [1]. Skin phantoms replicate realistic properties for validation without human testing [2]. Low-cost platforms, integrating NanoVNA with Raspberry Pi, provide compact alternatives to conventional systems with sufficient accuracy [3]. This work compares S_{11} responses of muscle, fat, and skin layers at varying distances using a metasurface-based microwave sensor.

II. METHODOLOGY

Fig. 1(a) shows the S_{11} measurement setup consisting of a sensor integrated with a metasurface to sense a multilayer human tissue phantom (muscle, fat, and skin). The sensor is connected to a NanoVNA for measurement, while the data are processed using a Raspberry Pi and displayed via the NanoVNA application. Fig. 1(b) illustrates the sensor–metasurface placement above the human tissue phantom at distances of 10, 20, 30, 40, and 50 mm. Fig. 1(c) presents the fabricated sensor with metasurface layers 1 and 2.

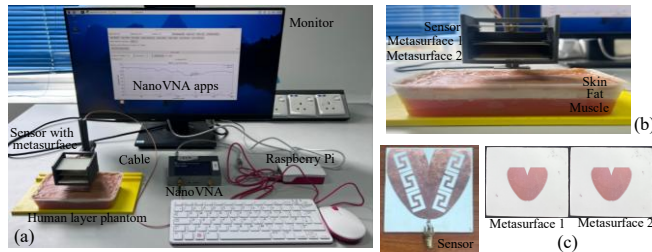


Fig. 1. (a) S_{11} measurement setup, (b) Human tissue layer phantom with sensor and metasurface, (c) Sensor with two layers of metasurface.

III. RESULTS AND DISCUSSION

Fig. 2 shows the S_{11} performance of the microwave sensor at 10–50 mm from the human tissue phantom. Fig. 2(a) presents the reference response without tissue. At 2.38 GHz, S_{11} is -15.587 dB at 10 mm, improving to -19.422 dB and -21.036 dB at 40 mm and 50 mm. Fig. 2(b) shows the muscle layer affects return loss due to electromagnetic interaction. At 1.82 GHz, the S_{11} is -21.371 dB at 10 mm and decreases to -13.248 dB and -12.147 dB at 40 mm and 50

mm, respectively. Fig. 2(c) shows muscle and fat layers causing resonance shifts, with S_{11} decreasing from -24.276 dB at 10 mm to -17.900 dB at 50 mm at 2.026 GHz. Fig. 2(d) shows the full multilayer structure with a similar trend, decreasing from -34.707 dB to -10.938 dB at 1.82 GHz.

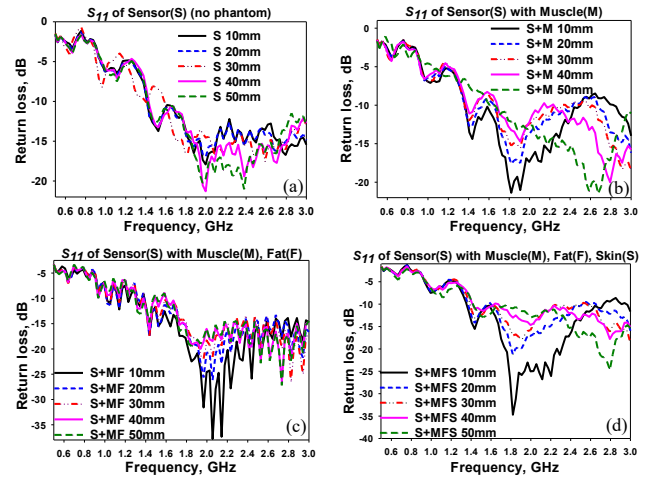


Fig. 2. S_{11} performance of the sensor and human tissue phantom at distances of 10–50 mm: (a) sensor only, (b) muscle, (c) muscle and fat, and (d) muscle, fat, and skin.

IV. CONCLUSION

Increasing sensing distance weakens electromagnetic coupling, altering resonance depth and frequency response. The metasurface-integrated microwave sensor distinguishes tissue layer configurations through distinct S_{11} variations across 10–50 mm. Resonance reached -21.036 dB at 2.38 GHz (no tissue), -12.147 dB at 1.82 GHz (muscle), -17.900 dB at 2.026 GHz (muscle–fat), and -10.938 dB at 1.82 GHz (multilayer). These results confirm its effectiveness for low-cost multilayer tissue differentiation.

ACKNOWLEDGEMENT

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FEM-Based Analysis of Airframe-Induced Radiation Pattern Distortion in UAV-Mounted Log-Periodic Antennas at UHF Frequencies

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Abstract– This paper presents an FEM electromagnetic simulation study assessing how mounting a HyperLOG 4025 antenna on a carbon-fiber hexacopter UAV (DJI S800) affects DVB-T2 coverage measurement. We performed simulations in ANSYS HFSS at 491 MHz, a DVB-T2 frequency in Colombia. We compare 2D and 3D radiation patterns (horizontal and vertical planes) and the S(1,1) reflection coefficient before and after integrating the antenna into the UAV body. Normalized gain figures reveal measurable pattern distortions caused by the carbon-fiber frame, providing insight into the feasibility of UAV-based DVB-T2 field measurements

Keywords– HyperLOG 4025, ANSYS HFSS, FEM, carbon fiber, radiation pattern, DJI S800, UAV.

I. INTRODUCTION

Digital television broadcasting has transformed mass communication worldwide. In Colombia, the rollout of Digital Video Broadcasting – Second Generation Terrestrial (DVB-T2) began in 2009. The initiative targets full population coverage and the eventual end of analog broadcasts. Bogotá, the capital, currently benefits from five DVB-T2 transmitter sites. These are operated by CCNP and RTVC, and the signal extends to nearby municipalities [1].

UAVs have become powerful, cost-effective platforms for in-situ electromagnetic field assessments. This includes antenna measurements and propagation surveys [2-4]. UAVs have been used to verify broadcast antenna patterns under DVB-T conditions in the Andean region [5]. Operating constraints for UAV-mounted measurement systems have also been studied [6]. However, accurate airborne measurements require understanding the perturbations from the UAV platform.

Carbon fiber-reinforced polymer (CFRP) is the preferred material for UAV frames. It acts as a lossy conductor at UHF, causing scattering and near-field coupling that distort radiation patterns and shift antenna resonances [7]. CFRP structures can serve as unintended radiators [7]. This highlights the need for platform-specific electromagnetic characterization. This paper quantifies the effect of the DJI S800 carbon-fiber airframe on the HyperLOG 4025 antenna at 491 MHz, using full-wave FEM simulation in ANSYS HFSS [7]. The goal is to establish a calibration baseline for UAV-based DVB-T2 coverage campaigns.

II. METHODOLOGY

The study uses three main simulation steps: A) 3D CAD Modeling. Build 3D models of the DJI S800 and HyperLOG 4025, referencing datasheets and measurements. Skip small features, such as screws and blades, to reduce mesh size while maintaining electromagnetic accuracy.

B) FEM Electromagnetic Simulation. ANSYS HFSS solves the full-wave EM problem at 491 MHz. HFSS utilizes adaptive mesh

refinement to ensure convergence, eliminating the need for manual mesh tuning. Two simulation scenarios are defined: (i) the isolated HyperLOG 4025 antenna and (ii) the antenna mounted on the DJI S800 airframe, with frame elements using carbon-fiber properties [8].

C) Results Analysis. Extract and compare 2D and 3D radiation patterns (E- and H-plane), normalized gain, and S(1,1) parameters for both scenarios. Quantify and highlight UAV-driven pattern changes.

III. RESULTS AND DISCUSSION

Figure 1 shows the results with the antenna on the DJI S800. Key findings: (a) The azimuth pattern becomes asymmetric due to carbon-fiber arm reflections. (b) Maximum gain drops by about 1.8 dB. (c) The S(1,1) minimum shifts in frequency because of the changed electrical environment. Use calibration to compensate when measuring with UAV-mounted antennas.

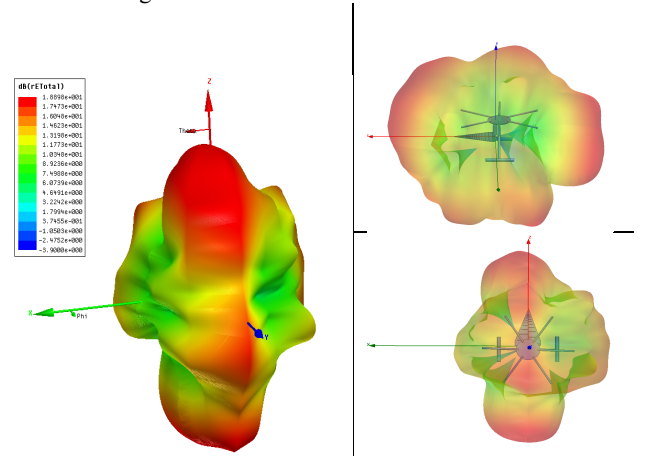


Fig. 1. Simulation of the HyperLOG-4025 antenna in the UAV DJI-S800

IV. CONCLUSION

FEM simulations in ANSYS HFSS show the DJI S800 carbon-fiber structure distorts the HyperLOG 4025 radiation pattern at 491 MHz. The main lobe reshapes, and the peak gain drops. These effects must be characterized and compensated to ensure reliable DVB-T2 coverage measurements. Future work will validate these results with outdoor UAV flights.

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