

Supplementary Information

On-chip nanoplasma for adaptive electromagnetic protection

Ruiqi Huang, Hanqing Liu, Jibin Liu, Yanlin Xu, Chenxi Liu, Song Zha, Peiguo Liu

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Note 1 Experiment on an embodied intelligent robot irradiated by HPM

Initially, the embodied intelligent robot successfully powers on and runs properly (Fig. S1a), while performing interactive greeting gestures (Fig. S1b). Nevertheless, upon exposure to HPM radiation, the robot collapses instantly as if being struck, falling to the ground and suffering permanent functional paralysis (Fig. S1c).



Figure S1: Embodied intelligent robot in **a**, the initial power-on state, **b**, the normal operational state, and **c**, the inoperable state following HPM radiation.

Note 2 Limitations of semiconductor diodes

The core trade-off for diodes used in high-frequency, high-power radio-frequency (RF) front-end protection lies in the difficulty of simultaneously achieving low cut-off capacitance, high power-handling capability, and fast response. Taking the widely used PIN diode in high-power electromagnetic protection as an example: the PIN diode consists of a heavily doped P^+ region, an intrinsic or lightly doped I region, and a heavily doped N^+ region (Fig. S2a). Under zero DC bias, when a high-power microwave signal is injected, carriers are injected from the P^+ and N^+ boundaries into the I region during the positive half-cycle of the signal. During the negative half-cycle, not all the injected carriers can return; some remain in the I region. After several cycles, a stable non-equilibrium carrier accumulation builds up in the I region, triggering conductivity modulation, which converts the I layer from a high-resistance state to a low-resistance state, thus achieving self-turn-on. For applications at high frequencies, diodes must exhibit low cut-off capacitance. Reducing the junction area decreases cut-off capacitance and improves high-frequency isolation, but compromises current capacity and power handling capability; increasing the junction area enhances power capacity but increases cut-off capacitance. Thickening the I layer reduces capacitance but raises on-resistance and response time.

At even higher frequencies, performance limitations shift progressively from intrinsic chip parameters to packaging and interconnect parasitic effects—parasitic capacitance and inductance can cause direct conduction in the cut-off state (Fig. S2b). Taking the SMP1320 PIN diode manufactured by Skyworks Solutions, Inc. as an example, the impedance curve of the diode in the off state shown in Fig. S2c exhibits high impedance at low frequencies. As frequency increases, the impedance begins to decrease to varying degrees until reaching the resonant frequency, after which it starts increasing again. At the resonant frequency, the cut-off impedance reaches its minimum due to the parasitic effects, and the impedance of the diode is comparable in the OFF and ON states (Fig. S2d). which implies that the diode loses its switching capability.

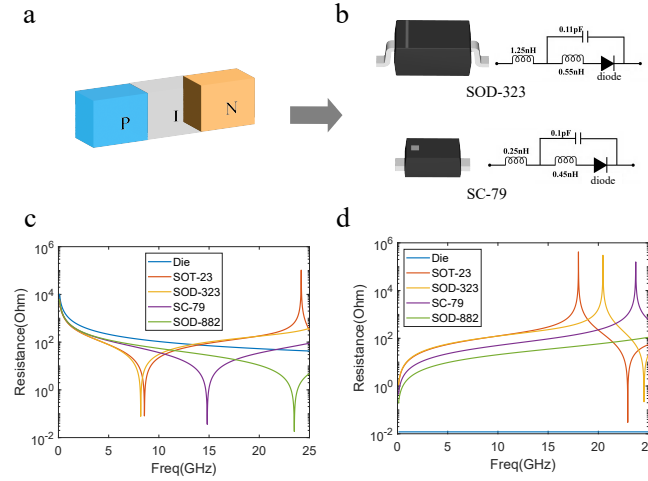


Figure S2: **a**, Geometric diagram of a PIN diode. **b**, Packaging and interconnect parasitic effects. **c**, OFF-state and **d**, ON-state impedance curves of SMP1320 diodes for various packages.

Fig. S3 illustrates the intersection points of on-state and off-state impedance curves of SMP1320 diodes corresponding to different package types. The impedance curves of the diode with Die package exhibit no intersection point (Fig. S3a). At 100 GHz, the impedance difference between the ON and OFF states still reaches 10 Ω . Nevertheless, owing to package parasitics, the impedance difference diminishes to zero (i.e., the intersection point) at a substantially lower frequency. At the intersection frequency, the diode loses its switching functionality. The

intersection points of impedance curves occur at lower frequencies of 5.56 GHz (Fig. S3b) and 5.45 GHz (Fig. S3c) for the larger-sized SOT-23 and SOD-323 packages, respectively. This indicates that they can only be applied to frequency bands below the corresponding intersection frequencies. The smaller SC-79 (Fig. S3d) and SOD-882 (Fig. S3e) packages exhibit higher intersection frequencies of 9.16 GHz and 14.06 GHz, respectively, implying that the switch is applicable to higher operating frequency bands. However, there is also a conflict between compact, low-parasitic packages and high heat-dissipation capability. Consequently, in adaptive high-power electromagnetic protection scenarios that demand high frequency, high tolerance, and fast response, the diode exhibits clear limitations. By comparison, no intersection point exists on the impedance traces of the nanoplasma switch (NPMS) shown in Fig. S3f. Even at 400 GHz, a 20 Ω impedance difference is sustained between ON and OFF states, which reveals that the NPMS supports operation at frequencies substantially above the feasible frequency range of diodes.

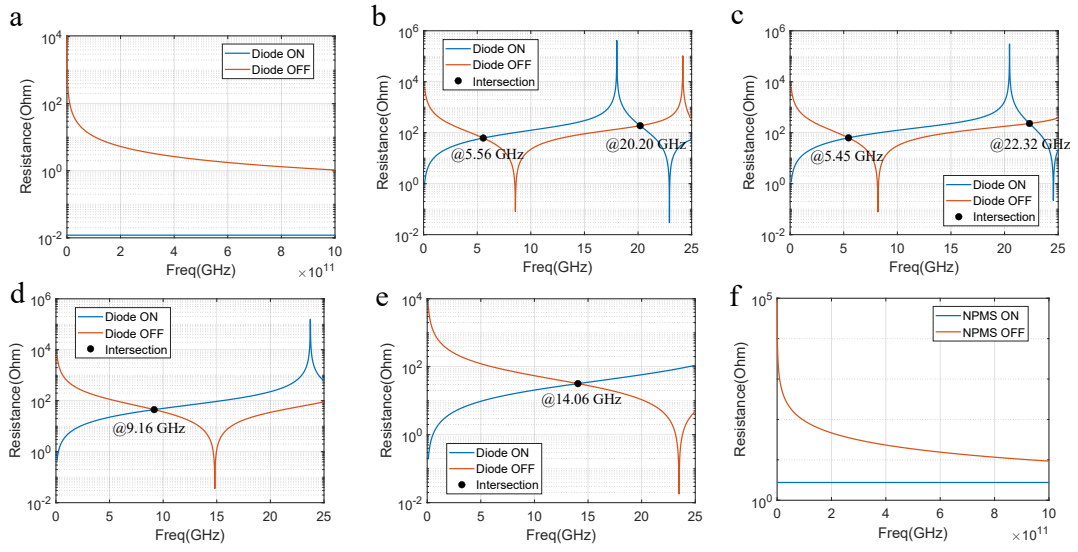


Figure S3: The intersection of ON-state and OFF-state impedance curves of SMP1320 diodes for **a**, Die package, **b**, SOT-23 package, **c**, SOD-323 package, **d**, SC-79 package, and **e**, SOD-882 package. **f**, The ON-state and OFF-state impedance curves of the NPMS.

The heat dissipation of NPMS devices and PIN diodes can be quantitatively described by a typical thermal time constant τ , which is attributed to the time required for heat to diffuse through the entire switch and substrate to increase temperature. Following a thermal model that describes the heat equation in a circular thin plate[1], the temperature of the entire system T at the frequencies f far below resonance is given by:

$$T = \frac{T_{th}}{i2\pi f\tau + 1}, \quad (S1)$$

where T_{th} is the temperature at the heat source.

The "heat transport in thin shells" module of COMSOL Multiphysics allows us to simulate the thermal transport in NPMS devices and diodes loaded on the substrate. We build a simple 3D domain as illustrated in Figs. S4a and S4b, where heat sources are set on the NPMS nanogap and top surface of PIN, respectively. We define the thermal insulation condition on the bottom of the substrate, with an initial temperature $T_0 = 273.15$ K. The mesh size is using "Extra Fine" to ensure the simulation accuracy of thermal transport. We then define a harmonic perturbation of heating with a frequency that varies from 10 to 10^6 Hz (NPMS) and 10^{-2} to 10^3 Hz (diode), and thus to obtain the average temperature T_{ave} of the membrane as a function

of heating frequency f , as shown in Figs. S4c and S4d. The heat capacity, mass density and thermal conductivity for NPMS devices (SiC substrate and GaN electrode materials as used in our work) and diodes (FR4 substrate, Au electrode, GaAs and Si diode materials as commonly used before) are substituted into the model. By fitting Eq. S1 to the simulated T_{ave} versus f , we extract the thermal time constant τ about 2.366×10^{-3} and 2.554×10^{-3} s for GaAs- and Si-based diodes, respectively, and 3.489×10^{-6} s for NPMS devices. Therefore, we confirm a much enhanced heat dissipation in the proposed on-chip nanoplasma compared with that in conventional devices.

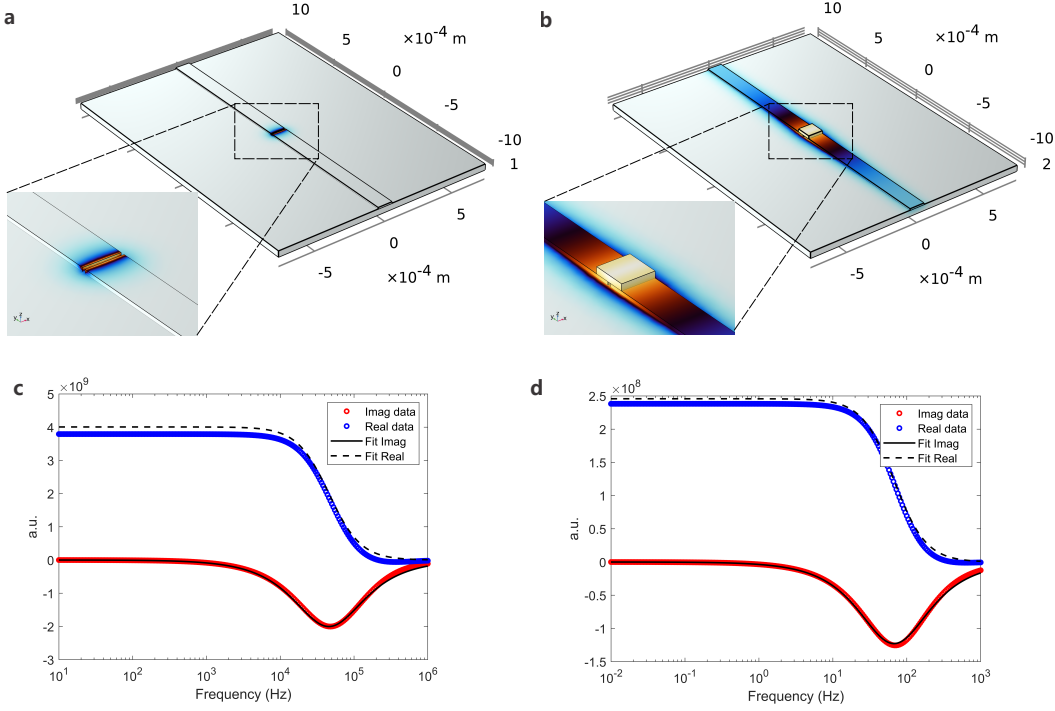


Figure S4: COMSOL simulations of thermal transport in NPMS and diode-based switch. **a** and **b**, Schematic diagram and temperature distribution on COMSOL for NPMS and diode-based switch, respectively. **c** and **d**, Simulation results of average temperature T_{ave} versus heating frequency f , which are fitted with Eq. S1 to extract the thermal time constant τ of the NPMS and diode-based switch.

Note 3 Local short-circuiting failure of all-metal nanoplasma switch

The NPMS is capable of carrying ampere-level on-state currents and has a high power-handling capacity. Nevertheless, this imposes extremely high current density and power density loads on the switch electrodes. For NPMS devices equipped with metal electrodes, localized electrode temperatures under the on-state operation can reach the melting point of the metal, triggering melting of the metal over a certain area. Once the molten metal migrates into the nanoscale electrode gap, there is an extremely high risk of gap short-circuiting and subsequent switch failure. As an example, Fig. S5 presents the molten metal-induced gap short-circuit of an all-metal NPMS. The metal electrode consists of a Ti adhesion layer and a Pt functional layer with thicknesses of 20 nm and 500 nm, respectively.

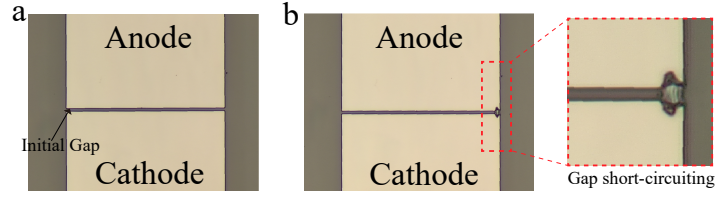


Figure S5: **a**, Initial state of the metal electrode gap. **b**, Molten-metal-induced gap short circuit.

Note 4 Preparation of the GaN-based nanoplasma switch

A 900 nm thick GaN film with a 50 nm AlN buffer layer was epitaxially grown using MOCVD on semi-insulating SiC substrates (Fig. S6a). The GaN layer is n-type doped with a doping concentration of $3 \times 10^{18} \text{ cm}^{-3}$, and the resulting GaN film achieves a sheet resistance of $47 \Omega/\text{sq}$.

The NPMS was fabricated using a two-step photolithography process. The first photolithography and ion-beam etching steps were used to define the GaN structures (Fig. S6b), whereas the second photolithography, metal deposition, and lift-off steps were employed to form the Ti/Au (20 nm/100 nm) electrodes (Fig. S6c). The detailed fabrication procedure is described below.

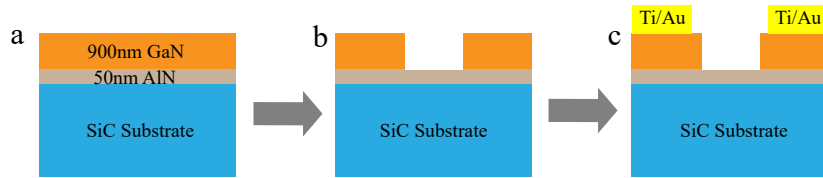


Figure S6: Schematic of the fabrication process. **a**, Layer schematic of GaN epitaxial wafer. **b**, Cross-sectional schematic of the GaN electrode gap. **c**, Schematic of the metal contact electrodes on GaN.

Patterning and ion-beam etching of GaN

1. **Surface pretreatment and photoresist coating.** The wafer was first treated in an HMDS system to improve the adhesion between the substrate and the photoresist. AR80 photoresist was then dispensed onto the sample and spin-coated at 500 rpm for 5 s, followed by 3000 rpm for 30 s. The coated sample was soft-baked at 95°C for 90 s.
2. **First photolithographic exposure.** The desired pattern was transferred into the photoresist using a Nikon I12 stepper with an exposure time of 300 ms.
3. **Development and inspection.** The exposed photoresist was developed in a 2.38% TMAH aqueous developer for 45 s. The developed pattern was subsequently inspected using a high-magnification optical microscope to verify the pattern fidelity and identify possible lithographic defects.
4. **Photoresist hard bake.** After development, the sample was baked at 100°C for 2 min. This step hardened the photoresist and removed residual solvent and moisture, thereby improving the resistance of the photoresist mask during the subsequent etching process.
5. **Oxygen-plasma descum.** Residual photoresist at the bottom of the developed openings was removed by oxygen-plasma cleaning using an M4L system. The plasma power and treatment duration were 200 W and 2 min, respectively.

6. **Ion-beam etching.** The exposed GaN layer was etched using an ion-beam etching (IBE) system. The ion energy, beam current, and etching time were 400 eV, 100 mA, and 90 min, respectively. The patterned photoresist served as the etching mask, allowing the unprotected GaN regions to be removed.
7. **Photoresist removal.** Following ion-beam etching, the remaining photoresist was removed by ultrasonic cleaning in acetone and isopropyl alcohol. The sample was subsequently rinsed with deionized water for 2 min and dried using a nitrogen gun.

Fabrication of the Ti/Au electrodes

1. **Surface pretreatment and second photoresist coating.** The etched sample was treated again in an HMDS system. A layer of 1303 photoresist was then spin-coated at 500 rpm for 5 s, followed by 3000 rpm for 30 s. The sample was soft-baked at 110 °C for 120 s.
2. **Second photolithographic exposure.** The electrode pattern was defined using an MA6 mask aligner with an exposure time of 4 s. After exposure, the sample was baked at 110 °C for 2 min.
3. **Development and inspection.** The exposed photoresist was developed in a 2.38% TMAH aqueous developer for 40 s. The resulting pattern was inspected using a high-magnification optical microscope before metal deposition.
4. **Ti/Au deposition.** A Ti/Au metal stack was deposited using a magnetron sputtering system. The thicknesses of the Ti adhesion layer and the Au electrode layer were 20 nm and 100 nm, respectively.
5. **Lift-off.** The sample was immersed in acetone and subsequently ultrasonically cleaned in acetone and isopropyl alcohol to remove the photoresist and the metal deposited on top of it. The sample was then rinsed with deionized water and dried using a nitrogen gun, leaving the patterned Ti/Au electrodes on the substrate.
6. **Device dicing.** Finally, the fabricated wafer was diced along the predefined alignment marks using a DISCO dicing system to obtain individual micro/nanoscale plasma-switch devices.

In summary, the fabrication process consisted of HMDS surface treatment, photoresist coating, photolithographic pattern transfer, oxygen-plasma descumming, ion-beam etching of the GaN layer, a second photolithography process, Ti/Au magnetron sputtering, metal lift-off, and device dicing.

Fig. S7 illustrates five NPMS devices with different dimensions and structures. Devices 1, 1-2, 1-3, and 1-4 feature parallel edge-to-edge electrode structures. Device 1 has an electrode gap of 800 nm and an electrode width of 100 μm . Device 1-2 has an electrode gap of 800 nm and an electrode width of 150 μm . Device 1-3 has an electrode gap of 1000 nm and an electrode width of 100 μm . Device 1-4 has an electrode gap of 2000 nm and an electrode width of 100 μm . Device 2 features a tip-to-edge electrode structure with an electrode gap of 800 nm and an electrode width of 100 μm .

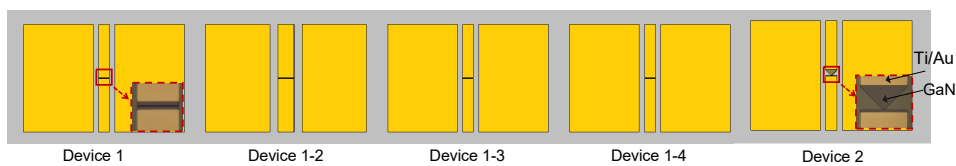


Figure S7: NPMS devices with different dimensions and structures.

Note 5 Conduction response simulation of the nanoplasma switch

The NPMS turns on and off through the formation and disappearance of plasma inside its nanoscale electrodes. When a voltage exceeding the threshold is applied, gas discharge occurs in the gap, forming a highly conductive plasma that turns on the NPMS. In general, the gas discharge process can be quantitatively described by a coupled system of transport equations for charged particles, centered on a fully coupled solution framework consisting of the continuity equation (S2), poisson equation (S3), and charge-density relation (S4)[2]:

$$\frac{\partial n_i}{\partial t} + \nabla \cdot (w_i n_i - D_i \nabla n_i) = R_i \quad (\text{S2})$$

$$\nabla \cdot (\varepsilon_r \varepsilon_0 \mathbf{E}) = \rho \quad (\text{S3})$$

$$\rho = e \sum_i z_i n_i \quad (\text{S4})$$

where the subscripts $i = e, p, n$ denote electrons, positive ions and negative ions, respectively; n_i is the number density of charged particles, $\mathbf{E}$ is the electric field, z_i is the charge number of carriers, w_i is the drift velocity in the electric field, D_i is the diffusion coefficient, R_i is the reaction rate and e is the elementary charge. Equations (S2) – (S4) describe the drift–diffusion–reaction transport of charged particles and the modulation of the electric field by space charge. To initiate and sustain the discharge process, the cathode must continuously supply seed electrons.

The turn-on triggering mechanism of NPMS has been primarily interpreted using microscale gas-breakdown models derived from the modified Paschen framework, in which cathode field emission (FE) supplies seed electrons and facilitates gap breakdown[3]. The FE process is classically described by Fowler-Nordheim (FN) theory[4, 5], giving the FE current density J_{FE} as below:

$$J_{\text{FE}} = A_{\text{FN}} \frac{F^2}{\phi} \exp \left(-\frac{B_{\text{FN}} \phi^{3/2}}{F} \right) \quad (\text{S5})$$

where ϕ (unit: eV) is the work function of cathode material, F (unit: V/m) is the local electric field at the cathode, A_{FN} and B_{FN} are constants equal to $1.54 \times 10^{-6} \text{ A V}^{-2} \text{ eV}$ and $6.83 \times 10^9 \text{ eV}^{-3/2} \text{ m}^{-1} \text{ V}$, respectively. The cathode local electric field F is typically enhanced from the applied field E . The field enhancement effect is generally described by a field enhancement factor β ($F = \beta E$), which represents the geometric effect of the electrode surface, usually depending on the local roughness and defects.

We employ the "electrical discharge" and "circuit" module of COMSOL Multiphysics to establish a gas discharge numerical model driven by the FE mechanism. Fig. S8a illustrates the complete external circuit model diagram, consisting of a voltage source, a current-limiting resistor of $50 \text{ } \Omega$, and an NPMS. The NPMS is simplified to a two-dimensional model in an air environment. The high electric field in the nanoscale gap enhances the FE effect, which is essential for seed electron generation. The electric field across the entire gap accelerates the electrons and ions to high energy, leading to the gas breakdown. The air domain bounded by the electrodes constitutes the computational region of the model. The FE process of the cathode is incorporated into the simulations by converting the FE current density J_{FE} into the electron emission flux $\mathbf{\Gamma}_e$ at the cathode boundary using Eq. (S6). Here, $\mathbf{n}$ denotes the unit normal vector at the cathode interface pointing from the gap to the cathode, and e represents the elementary charge.

$$-\mathbf{n} \cdot \mathbf{\Gamma}_e = \frac{J_{\text{FE}}}{e} \quad (\text{S6})$$

With the structural parameters and electrode material fixed, the conduction response characteristics of the NPMS devices are predominantly governed by the field enhancement factor

β . For example, we set the gap distance d between electrodes to 800 nm, the width w of the electrode is 100 μm , the electrode thickness is 900 nm, and the work function of the electrode material is 4 eV[6]. The simulation results show that an increase in the field enhancement factor β leads to a reduction in the switching threshold voltage (Fig. S8b and c).

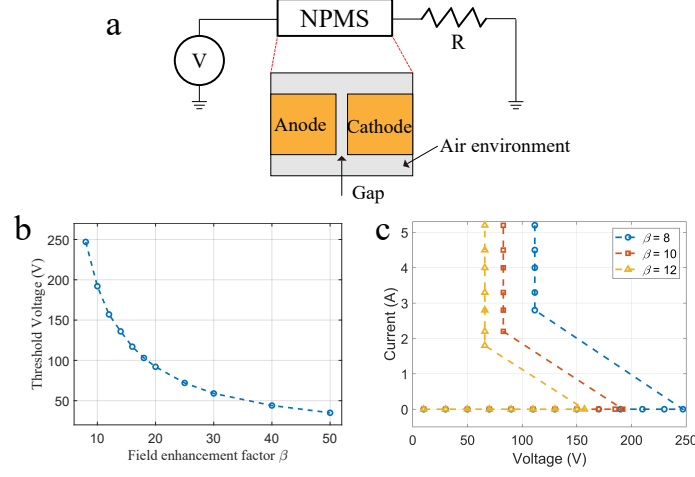


Figure S8: Simulation of the NPMS conduction response. **a**, Complete external circuit model diagram. **b**, Threshold voltage versus the field enhancement factor β . **c**, Simulated current-voltage curves corresponding to $\beta = 8, 10$, and 12 .

We can see from Fig. S9 that despite differences in the switch threshold voltage, the turn-on transition time remains nearly identical (around 15 ps). This indicates that a higher threshold voltage corresponds to a faster turn-on speed of the switch. In detail, the turn-on speeds of 7.5 V/ps, 11 V/ps and 13 V/ps are associated with threshold voltages of 157 V, 192 V and 247 V in sequence.

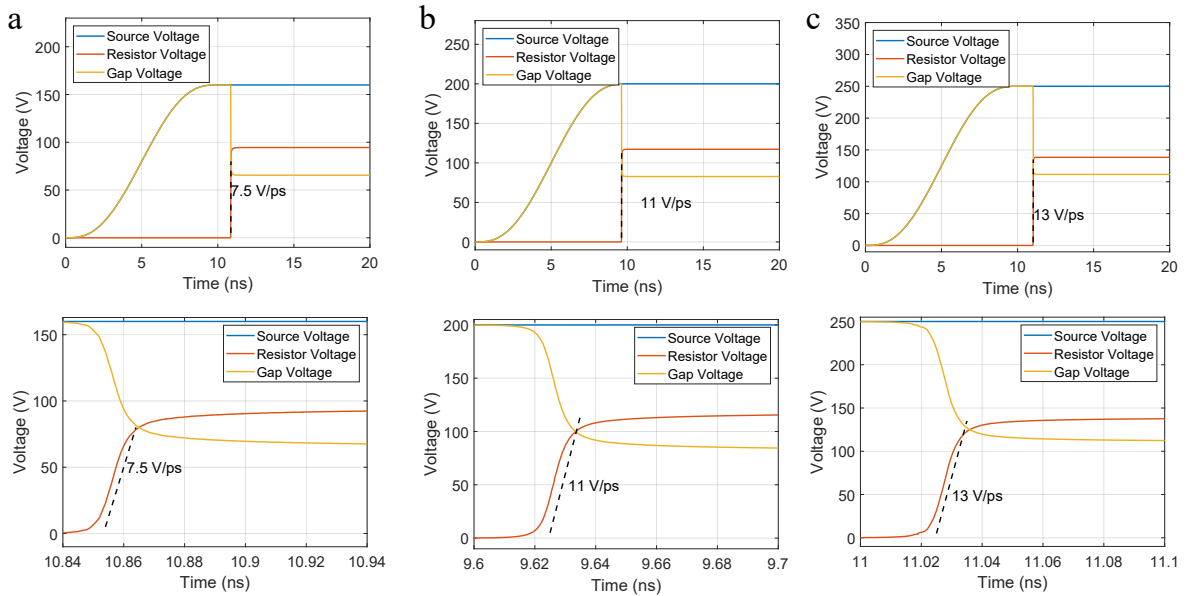


Figure S9: Simulated transient voltage responses at threshold voltages of **a**, 157 V, **b**, 192 V, and **c**, 247 V.

Note 6 Measurement of current-voltage characteristics of the nanoplasma switch

The current-voltage (I - V) characteristics of the five fabricated NPMS devices are measured using the standard transmission-line pulse (TLP) method. A HPPI-TLP-4010C pulse generator was employed to apply voltage pulses to the device under test, while the corresponding transient voltage and current waveforms were recorded using a Tektronix MOS64 digital oscilloscope with a sampling rate of 25 GS/s and a bandwidth of 1 GHz. The applied pulses had a width of 100 ns and a rise time of 10 ns. The pulsed measurement minimized the influence of long-term Joule heating and enabled characterization of the transient electrical response of the device. For each pulse, the voltage and current values are extracted from the flat-top region of the recorded waveforms and averaged to obtain a single data point on the I - V curve. The pulse amplitude is then increased in set steps, and the measurement sequence is repeated to obtain the desired I - V curve.

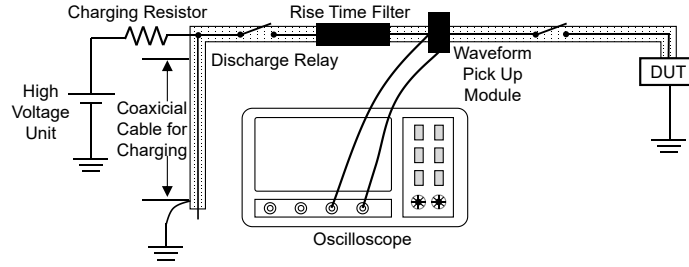


Figure S10: TLP test setup diagram.

Fig. S11 shows the I - V curves and off-state transmission coefficients of NPMS devices illustrated in Fig. S7. As can be seen, these devices demonstrate high on-state currents exceeding 3 A and low on-state dynamic resistances R_{ON} lower than 5 Ω . They also exhibit small transmission coefficients even in the Ku band, indicating their small cut-off capacitances C_{OFF} . The corresponding fitting R_{ON} and C_{OFF} are listed in Table S1. Table S2 lists the R_{ON} and C_{OFF} of commercial diodes according to their official datasheets. Their cut-off frequencies and absolute maximum currents (termed rated operational current in the main text) are also provided in the Tables. The measured NPMS devices exhibit a rated operational current versus cut-off frequency product ($I_{ro} \times f_c$) that is nearly two orders of magnitude higher than that of commercial diodes (Fig. S12).

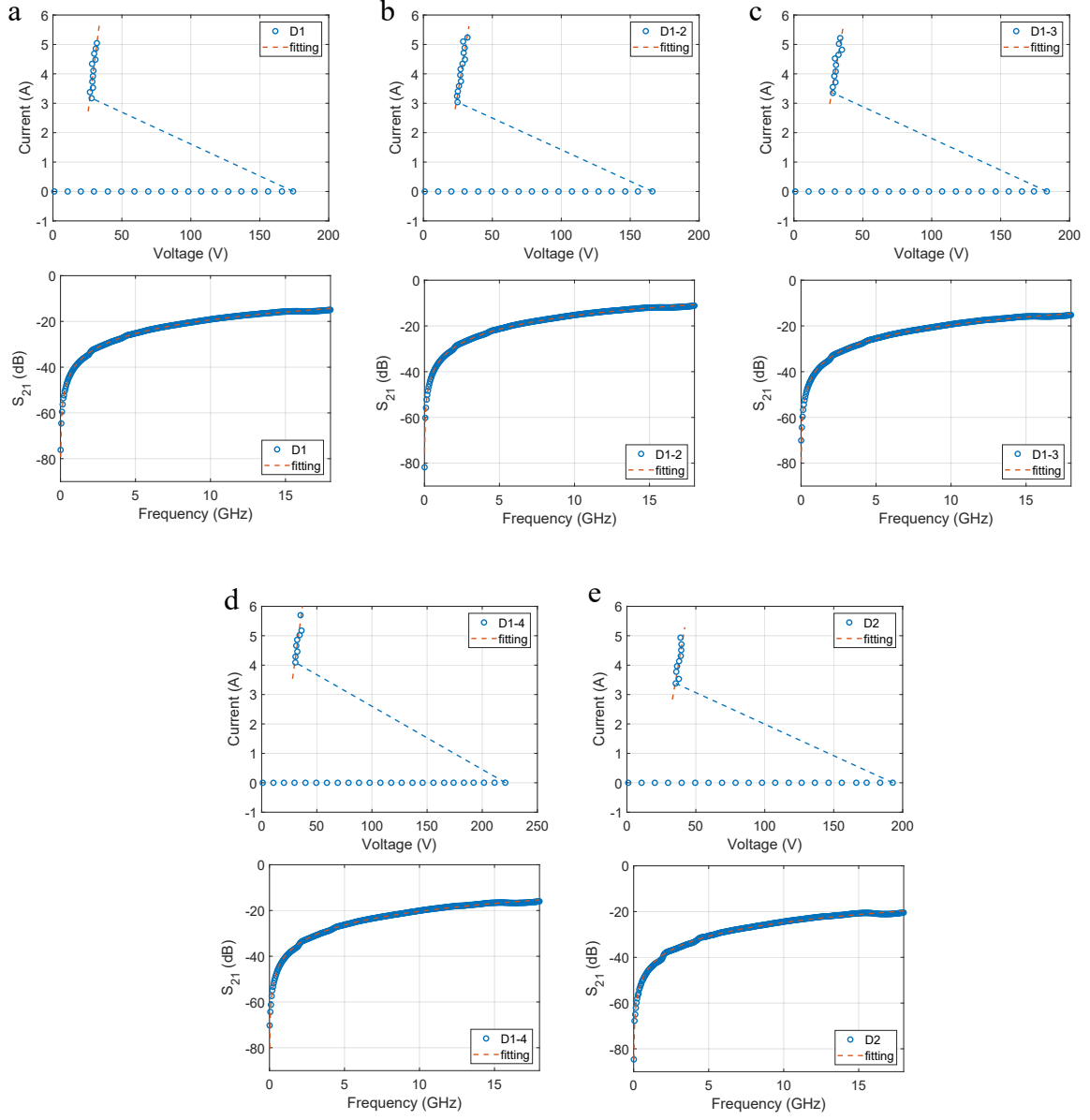


Figure S11: I - V curves and transmission coefficients of NPMS devices. Subfigures **a-e** correspond to devices 1, 1-2, 1-3, 1-4, and 2, respectively.

Table S1: Electrical characteristics of NPMS devices.

	On-state resistance R_{ON} (Ω)	Cut-off capacitance C_{OFF} (fF)	Cut-off Frequency (GHz)	Current (A)
D1	2.8	17.3	3349.2	5.1
D1-2	3.5	27.1	1660.8	5.2
D1-3	3.7	16.8	2551.9	5.2
D1-4	3.7	15.3	2830.7	5.7
D2	3.7	9.2	4731.8	4.9

Note: R_{ON} and C_{OFF} are obtained by fitting the I - V curves and transmission coefficients of the NPMS devices. The cut-off frequency is calculated as $1/(2\pi R_{\text{ON}} C_{\text{OFF}})$. The current corresponds to the maximum measured point on the I - V curve.

Table S2: Electrical characteristics of the commercial diodes.

	On-state resistance R_{on} (Ω)	Cut-off capacitance C_{OFF} (fF)	Cut-off Frequency (GHz)	Current (mA)
BAP51-02	1.5	400	248.8	50
NSR201MX-D	14.0	150	63.2	50
MA4AGFCP910	5.2	18	637.6	/
SMP1345	1.5	190	482.3	100
MMP4401	0.8	350	523.5	100
MA4L011-134	2.1	130	473.7	100
NSVP249SDSF3	4.5	230	136.1	50
MA4SPS402	5.0	45	424.4	250

Note: R_{on} and C_{OFF} are provided in the commercial-diode datasheets. The cut-off frequency is calculated as $1/(2\pi R_{\text{on}} C_{\text{OFF}})$, with an additional general parasitic capacitance of 30 fF taken into account. The reported current corresponds to the absolute maximum current specified in the datasheets.

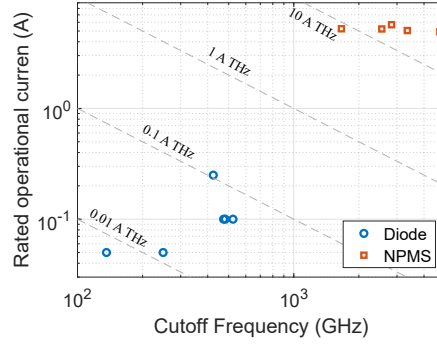


Figure S12: Rated operational current versus cut-off frequency for diodes and NPMS devices.

Note 7 Structural parameters of the proposed designs

SPPM

We design two SPPMs with similar structures, as shown in Fig. S13 and Fig. S14 (termed SPPM-1 and SPPM-2). Their detailed structural parameters are listed in Table S3 and Table S4, respectively.

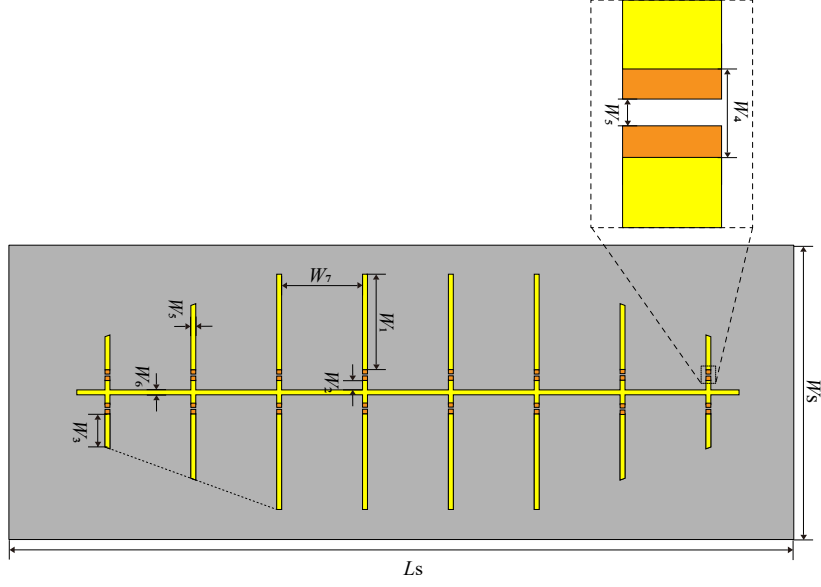


Figure S13: Structural parameters of the proposed SPPM-1.

Table S3: Detailed structural parameters of the designed SPPM-1

Param.	L_s	W_s	W_1	W_2	W_3	W_4	W_5	W_6	W_7
Value	16 mm	6 mm	1.95 mm	0.19 mm	0.67 mm	10 μm	800 nm	0.1 mm	1.65 mm

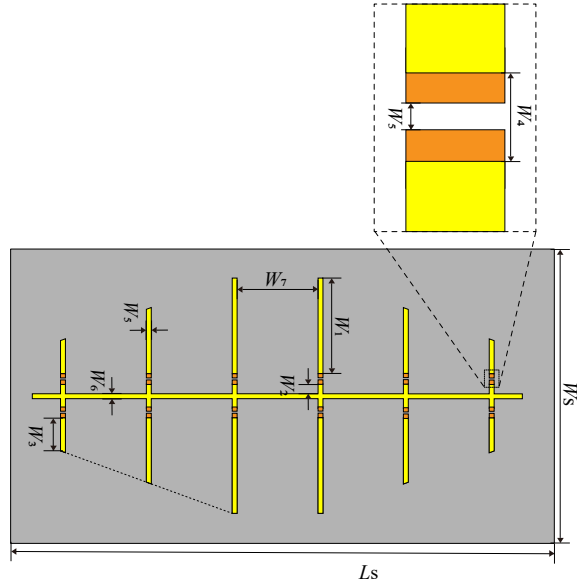


Figure S14: Structural parameters of the proposed SPPM-2.

Table S4: Detailed structural parameters of the designed SPPM-2

Param.	L_s	W_s	W_1	W_2	W_3	W_4	W_5	W_6	W_7
Value	10 mm	7.8 mm	3.4 mm	0.19 mm	2 mm	10 μm	800 nm	0.1 mm	1.65 mm

The fundamental mechanism for the SPPM metastructure to realize adaptive protection lies in the variation characteristics of its dispersion relation with structural dimensions. Fig.

S15 demonstrates the dispersion curves for different stub heights h of the SPPM unit cell (shown in the inset). The polarization direction of the incident electric field is parallel to the stubs. We can find that the dispersion curves deviate from the light line. At low frequencies, the wave number k of the SPPM unit cell is close to that of the light line; as the frequency increases gradually, it becomes much larger than that of the light line until the k approaches infinity—that is, the frequency approaches the asymptotic frequency. This frequency is regarded as the cut-off frequency f_c of the SPPM, endowing it with low-pass characteristics. It can also be observed that the f_c of the dispersion curve decreases gradually as the value of h increases. This parameter-dependent frequency tuning capability enables flexible regulation of the wave propagation shielding threshold, laying a foundation for adaptive protection across wide frequency bands. Specifically, by incorporating NPMS devices onto the SPPM stubs, the variation of h is equivalently realized through NPMS switching, thereby achieving switching between wave transmission and shielding states in the target operating frequency band.

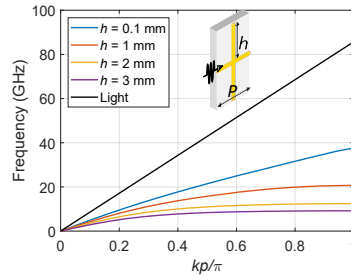


Figure S15: Dispersion curves with different stub height h of the SPPM unit cell.

Protective microstrip patch antenna

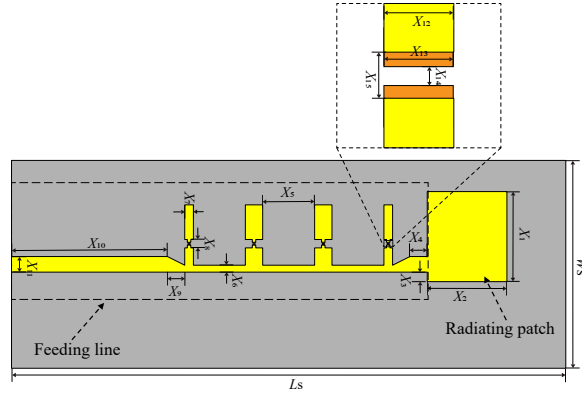


Figure S16: Structural parameters of the proposed protective microstrip patch antenna.

Table S5: Detailed structural parameters of the designed protective microstrip patch antenna

Param.	L_s	W_s	X_1	X_2	X_3	X_4	X_5	X_6	X_7
Value	16 mm	6 mm	2.6 mm	2.3 mm	0.275 mm	0.5 mm	1.5 mm	0.2 mm	0.25 mm
Param.	X_8	X_9	X_{10}	X_{11}	X_{12}	X_{13}	X_{14}	X_{15}	
Value	0.2 mm	0.5 mm	4.5 mm	0.45 mm	0.1 mm	0.1 mm	800 nm	40 μm	

Limiter

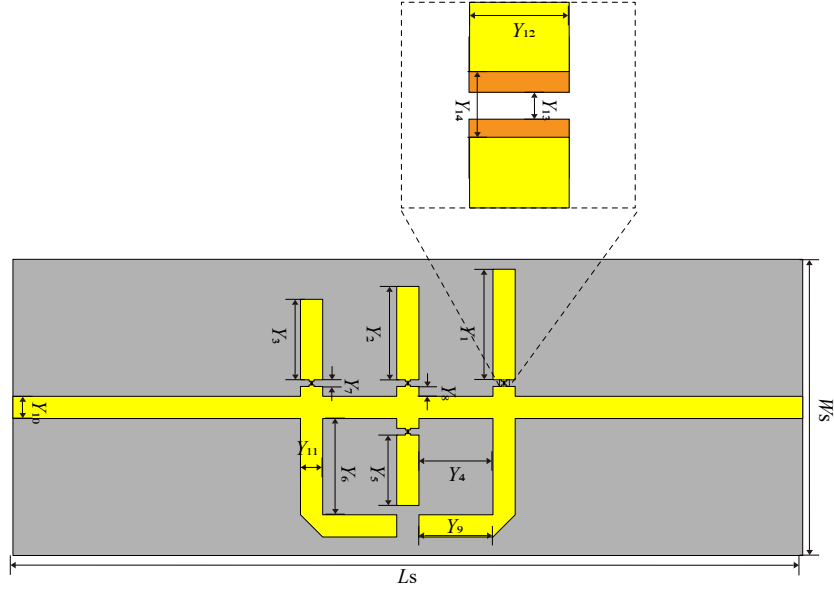


Figure S17: Structural parameters of the proposed Limiter.

Table S6: Detailed structural parameters of the designed limiter

Param.	L_s	W_s	Y_1	Y_2	Y_3	Y_4	Y_5	Y_6
Value	16 mm	6 mm	2.27 mm	1.92 mm	1.65 mm	1.5 mm	1.45 mm	1.95 mm
Param.	Y_7	Y_8	Y_9	Y_{10}	Y_{11}	Y_{12}	Y_{13}	Y_{14}
Value	40 μm	0.23 mm	1.5 mm	0.45 mm	0.45 mm	0.1 mm	800 nm	40 μm

Note 8 HPM excitation experiments

Fig. S18 presents the schematics of the experimental setups for characterizing the responses of the proposed protective designs under HPM excitations. A microwave signal is amplified by a power amplifier and subsequently delivered to the transmitting antenna (Fig. S18a, corresponding to field-level test) or the device under test (DUT) (Fig. S18b, corresponding to circuit-level test) through an isolator. The isolator is employed to suppress reflected power and protect the upstream microwave signal source and power amplifier. The received signal is passed through an attenuator before being recorded using a spectrum analyzer (or oscilloscope). For field-level tests, an extra radiative propagation link is required.

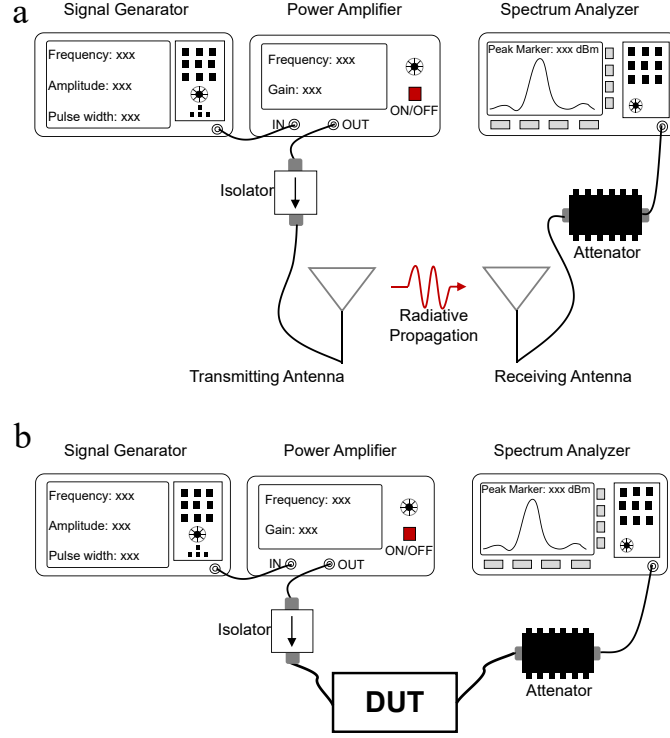


Figure S18: Schematics of experimental setups for HPM excitation. **a**, Field-level test and **b**, circuit-level test.

SPPM

Characterization of the proposed SPPM structure requires an HPM field environment. Because less power is required to generate the same electric-field strength in a waveguide than in free space, we employ the waveguide injection test method to characterize the response of the SPPM prototype under different field strengths. The selection of placement position takes into account the field distribution characteristics of the TE₁₀ mode inside the waveguide, that is, the field intensity is the highest in the middle of the waveguide cross-section and decreases toward both sides. Therefore, according to the working frequency and structure size, we place one SPPM element at the center of the cross section of a standard WR62 waveguide to test its performance, as shown in Fig. S19. Then, measurements are performed on the waveguide loaded with the SPPM element, which serves as the DUT illustrated in Fig. S18b.

The peak electrical field strength ($E_{\max}$) in waveguide is calculated as follows:

$$E_{\max} = 2\sqrt{\frac{P\eta}{ab}} \frac{1}{\sqrt{1 - \left(\frac{f_c}{f}\right)^2}}, \quad (\text{S7})$$

$$f_c = \frac{3 \times 10^8}{2a}, \quad \eta = \frac{377}{\sqrt{1 - \left(\frac{f_c}{f}\right)^2}}. \quad (\text{S8})$$

where f_c is the TE₁₀-mode cut-off frequency of the waveguide, P is the input power, and a and b are the width and height of the waveguide cross section, respectively. For the standard WR62 waveguide, $a = 15.799$ mm and $b = 7.899$ mm.

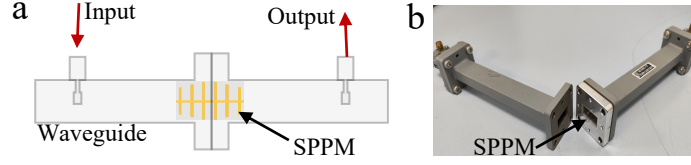


Figure S19: **a**, Schematic and **b**, photograph of an SPPM placed in a WR62 waveguide.

Table S7: Peak electric field strength (Unit: kV/m) at different frequencies and input powers in the WR62 waveguide

Power (dBm)		10	20	30	40	50	60
Freq. (GHz)							
12		0.444	1.403	4.436	14.028	44.360	140.280
13		0.411	1.301	4.113	13.007	41.130	130.070
14		0.396	1.253	3.961	12.525	39.610	125.250
15		0.387	1.225	3.873	12.248	38.730	122.480
16		0.382	1.207	3.817	12.071	38.170	120.710
17		0.378	1.195	3.779	11.949	37.790	119.490
18		0.375	1.187	3.752	11.865	37.520	118.650

The results for SPPM-2 through waveguide injection tests are presented in Fig. S20. The corresponding results for SPPM-1 have already been presented in the main text. As can be seen, the insertion loss is less than 1 dB below 13 GHz and less than 2 dB below 14 GHz, which is consistent with the full-wave simulation results. At 12 GHz and 13 GHz, the output fields were limited to 1.48×10^4 V/m and 1.41×10^4 V/m, respectively, corresponding to shielding effectiveness (SE) values of approximately 18.46 dB and 18.18 dB (Fig. S20b,c).

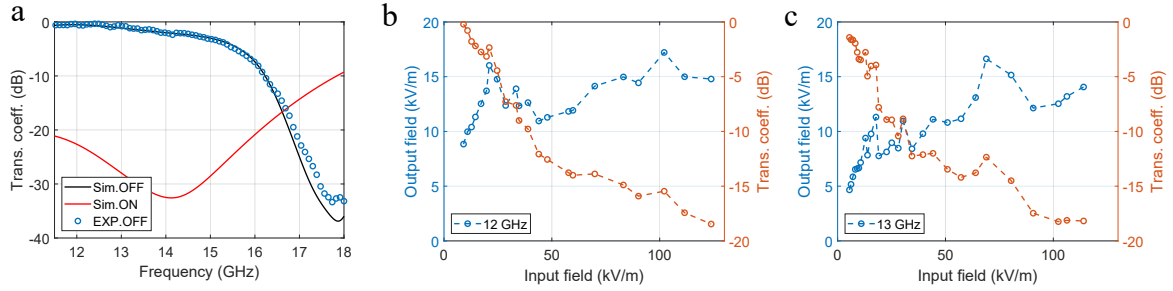


Figure S20: Results for SPPM-2. **a**, Transmission coefficients. Nonlinear responses under HPM field excitations at **b**, 12 GHz and **c**, 13GHz.

We then fabricate a 5×4 array of SPPM-2 elements (Fig. S21a) and characterize its performance under spatial irradiation in a microwave anechoic chamber. The spacing between two adjacent parallel SPPM elements in the array is 10 mm. The entire array is placed in front of the horn antenna, as shown in the Fig. S21. Then, the horn antenna loaded with the array serves as the receiving antenna, as illustrated in Fig. S18a. The results of the array are shown in Fig. S22, and agree well with the waveguide injection test results shown in Fig. S20.

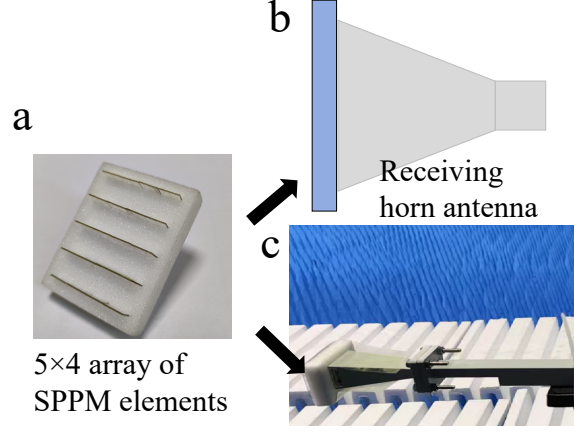


Figure S21: **a**, The 5×4 array of SPPM-2 elements placed in front of the receiving horn antenna. **b**, Placement schematic and **c**, photograph.

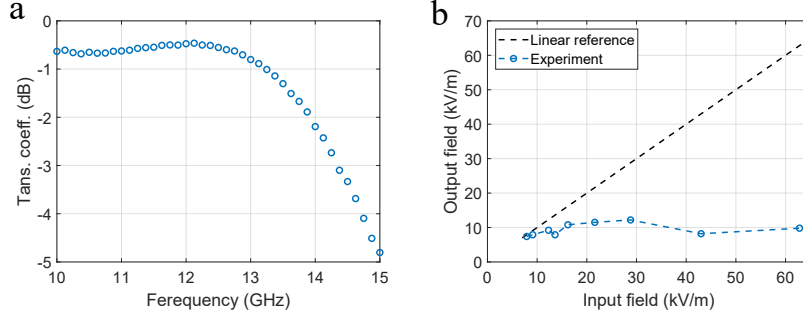


Figure S22: **a**, Insertion loss results for the SPPM array. **b**, Response under HPM field excitation at 14.25 GHz.

Protective microstrip patch antenna

A dedicated high-power radiation source covering the operational frequency band of our designed antenna is unavailable; hence, a spatial field of tens of kV/m strength cannot be constructed to directly excite the designed protective microstrip patch antenna. Therefore, we employ the designed protective antenna as the transmitting antenna[7]. Electromagnetic waves radiated by the proposed antenna are captured by a standard receiving antenna to monitor variations in its radiation capability. By gradually increasing the incident microwave power and monitoring the received signal power, the transmission response of the proposed antenna is characterized under different excitation intensities. The radiation capability is then assessed from the suppression of the received signal at high incident power relative to the low-power transmission response. A decline in the radiation capability indicates that the proposed antenna enters the protection state, and accurate protection effectiveness can be calculated by comparing the radiation performance before and after activation (Fig. S23). Since the antenna is a reciprocal device, the variation in receiving capability can be inferred from the changes in its radiation capability, which enables the evaluation of the protection effectiveness of the proposed antenna.

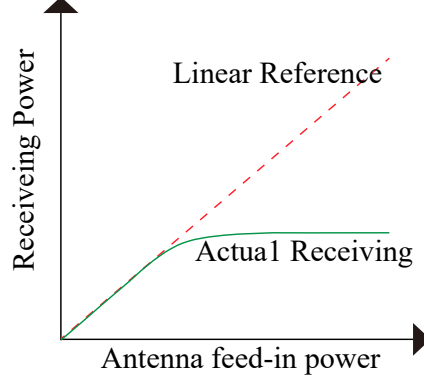


Figure S23: Nonlinear response schematic of the proposed protective antenna.

The relationship between the received power P_r at the antenna terminals and the incident time-averaged power density S_{av} (in W/m^2) is given by

$$P_r = S_{av} A_e \quad (\text{S9})$$

where A_e is the effective aperture of the antenna. For a lossless antenna with gain G , the effective aperture is

$$A_e = \frac{G\lambda^2}{4\pi} \quad (\text{S10})$$

with $\lambda = c/f$ being the free-space wavelength.

The power density of a plane wave is related to the peak electric field strength E_{peak} (in V/m) by

$$S_{av} = \frac{E_{\text{peak}}^2}{2\eta_0} \quad (\text{S11})$$

where $\eta_0 \approx 120\pi \approx 376.73 \Omega$ is the intrinsic impedance of free space.

Combining Eq. (S9) – (S11), we obtain the direct relation

$$E_{\text{peak}} = \sqrt{\frac{8\pi P_r \eta_0}{G \lambda^2}} \quad (\text{S12})$$

For example, for an antenna with a gain of 3.0 dBi operating at 17.5 GHz, a received power of 56 dBm corresponds to an incident peak electric field strength E_{peak} of about 80.2 kV/m. In our measurement of the proposed protective antenna, which has a measured gain of 3.0 dBi, the antenna serves as the transmitting antenna. Based on antenna reciprocity, when the power of 56 dBm is fed into the protective antenna, the excitation borne by the NPMS devices is equivalent to the antenna being subjected to an incident peak field strength E_{peak} of 80.2 kV/m.

Normalized power density figure-of-merit (FOM): Let A_{phys} denote the physical aperture area, A_e the effective aperture (Eq. (S10)), S_{av} the incident time-averaged flux density (Eq. (S11)), and η_{ap} the aperture efficiency. The aperture efficiency is defined as

$$\eta_{\text{ap}} = \frac{A_e}{A_{\text{phys}}}. \quad (\text{S13})$$

It follows from equation(S13) that

$$A_e = \eta_{\text{ap}} A_{\text{phys}}. \quad (\text{S14})$$

For an incident plane wave, the received power is

$$P_r = S_{av} A_e = S_{av} \eta_{\text{ap}} A_{\text{phys}}. \quad (\text{S15})$$

Table S8: Size of diode-based protective antennas and the proposed protective antenna.

Ref.	[8]	[9]	[10]	[11]	[12]	[13]	[14]	This work
Size. (mm ²)	35×50	144×60	$200^2 \times \pi$	$250^2 \times \pi$	180^2	26×30	$250^2 \times \pi$	14×2.6

The normalized power density figure-of-merit (FOM) is defined as

$$\text{FOM} = \frac{P_r \eta_{\text{ap}}}{A_{\text{phys}}}. \quad (\text{S16})$$

Substituting Eq. (S15) into Eq. (S16) gives

$$\begin{aligned} \text{FOM} &= \frac{(S_{\text{av}} \eta_{\text{ap}} A_{\text{phys}}) \eta_{\text{ap}}}{A_{\text{phys}}} \\ &= S_{\text{av}} \eta_{\text{ap}}^2 = \frac{E_{\text{peak}}^2}{2\eta_0} \eta_{\text{ap}}^2. \end{aligned} \quad (\text{S17})$$

For a fixed incident power flux density, the normalized power density FOM is proportional to the square of the aperture efficiency. The physical aperture area is canceled because the received power is proportional to that area.

Note 9 Effect of flexible substrate bending on the electrode gap

Consider two parallel electrodes fabricated on a flexible substrate (Fig. S24). Here the case in which the bending axis is parallel to the electrode edges is considered. In this configuration, the bending-induced tensile or compressive strain acts directly along the gap direction.

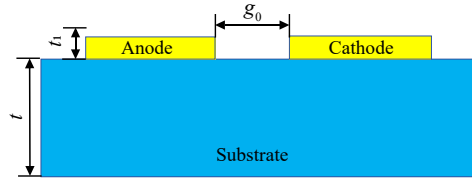


Figure S24: Diagram of parallel electrodes fabricated on a flexible substrate.

Let the initial electrode gap be g_0 , the distance from the neutral mechanical plane to the electrode surface be z , and the bending radius be R (Fig. S25a,b). Under small-deformation pure bending, the strain at the electrode surface is approximately [15, 16]

$$\varepsilon_{\text{gap}} \approx \frac{z}{R}, \quad (\text{S18})$$

where the sign of z/R depends on whether the electrode surface is under tension (Fig. S25c) or compression (Fig. S25d). The deformed gap width is therefore

$$g \approx g_0 (1 + \varepsilon_{\text{gap}}) = g_0 \left(1 + \frac{z}{R}\right). \quad (\text{S19})$$

Accordingly, the change in the gap width is

$$\Delta g = g - g_0 \approx g_0 \frac{z}{R}. \quad (\text{S20})$$

When the electrodes are located on the outer, convex side of the bent substrate, the electrode surface is under tensile strain. Thus,

$$\varepsilon_{\text{gap}} > 0, \quad g > g_0, \quad (\text{S21})$$

and the electrode gap increases.

When the electrodes are located on the inner, concave side of the bent substrate, the electrode surface is under compressive strain. Thus,

$$\varepsilon_{\text{gap}} < 0, \quad g < g_0, \quad (\text{S22})$$

and the electrode gap decreases.

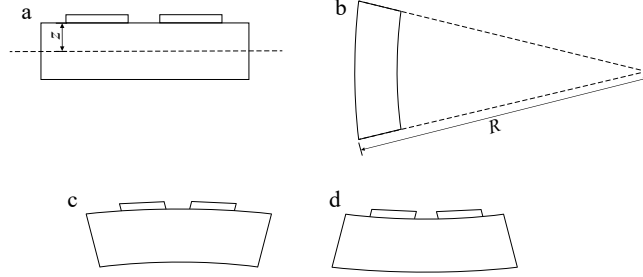


Figure S25: Diagram of a flexible substrate under bending. **a**, Without bending. **b**, Bending radius. **c**, Tensile bending. **d**, Compressive bending.

For a single-layer substrate of thickness t , if the neutral plane is approximately located at the mid-plane of the substrate and the electrode thickness is negligible, then

$$z \approx \frac{t}{2}. \quad (\text{S23})$$

The magnitude of the gap change can consequently be estimated as

$$|\Delta g| \approx g_0 \frac{t}{2R}. \quad (\text{S24})$$

For example, consider

$$t = 20 \text{ } \mu\text{m}, \quad R = 1 \text{ mm}, \quad g_0 = 100 \text{ nm}. \quad (\text{S25})$$

The strain magnitude at the electrode surface is

$$|\varepsilon_{\text{gap}}| \approx \frac{10 \text{ } \mu\text{m}}{1 \text{ mm}} = 0.01, \quad (\text{S26})$$

which corresponds to 1%. The magnitude of the gap variation is therefore

$$|\Delta g| \approx 100 \text{ nm} \times 0.01 = 1 \text{ nm}. \quad (\text{S27})$$

Hence, the gap becomes approximately 101 nm under tensile bending and 99 nm under compressive bending. Although a bending strain of 1% is generally regarded as a significant deformation[17], the resulting geometric change in the electrode gap of the NPMS on the substrate is extremely small, leading to a negligible impact on its electrical performance. Therefore, this switch presents a highly attractive prospect for integration into flexible electronic applications, such as wearable devices and flexible sensors.

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