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Effects on electronics exposed to high-power microwaves on the basis of a relativistic backward-wave oscillator operating on the X-band

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Abstract
An analysis of the effects of electromagnetic pulses from a high-power microwave (HPM) radiation technique is conducted using a relativistic backward-wave oscillator (RBWO) which uses relativistic electron beams in vacuum circuits. The application described here is based on a relativistic electron device and uses relativistic electron beams to generate high-power electromagnetic radiation. The RBWO was fabricated to operate in a relativistic region with a gamma factor ($\gamma$) of 2 at an acceleration voltage of 500 kV. A mode-converted relativistic back-wave oscillator with an antenna that converts TM\textsubscript{01} to TE\textsubscript{11} was designed and fabricated because the electric field of the center in the RBWO circuit is null. The effects on electronic devices by HPM radiation and exposure were assessed. The effects on electronic devices exposed to HPMs, the failure of information equipment, and modulation of and interference with the received signal through a theoretical model of the threshold power relative to the influence on the target were confirmed in a high-output microwave exposure environment. Particularly, information devices containing semiconductors can undergo serious failures and breakdowns due to the thermal secondary breakdown caused by the high-output transient electromagnetic waves, and it is a theoretical consideration that reverse voltage occurs due to the generation of surge current when caught in the PN-junction region. Finally, the range of power regarding the effectiveness of the electromagnetic coupling of electronics exposed to HPM radiation was estimated.

Keywords
High power microwave (HPM); relativistic backward wave oscillator (RBWO); electromagnetic pulse (EMP); effects on electronics; electromagnetic coupling; high power microwave radiation

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1. Introduction

In recent years, the development of the technology of electromagnetic wave-generating devices using electron beams in a vacuum has been active. Specifically, the main breakthrough in technology of the microwave oscillator circuit has been used in a variety of electromagnetic-wave radiation applications, and active development is ongoing. It is essential to develop new direct-energy weapons because conventional non-electromagnet systems are no longer feasible for use with the complex electronic equipment and advanced technology at the center of high-power microwave (HPM) sources. This was one of the major challenges in relation to the application of high-power electromagnetic (HPEM) and HPM technology during the last half of the recent century [1,2]. HPEM refers to an artificial electromagnetic environment that can harm electronic devices through the generation of electromagnetic pulses (EMPs). The HPEM environment is intended to paralyze electronic and communication devices such as semiconductors, computers, and mobile phones intentionally by artificially creating ultra-wideband or HPM charges [3]. The main characteristics of a HPM source are as follows. First, radiated electromagnetic waves can be exposed to various targets. This not only damages electronic devices, but also causes their permanent failure. Second, such a source can be used for a variety of purposes. A direct energy transfer cannot only accurately target an object with electromagnetic waves but also destroy entire systems of electronic devices. Third, it can be used in all weather conditions. It can spread through the air even in a sky full of clouds or in dust, fog or rain. Fourth, it is possible to propagate to the inside of devices, and all exposed electronic devices can be neutralized. Fifth, it can be performed on multiple platforms. Cruise missiles, unmanned airplanes, as delivery systems and other devices can be built or installed on fixed platforms [4].

In this study, a relativistic backward-wave oscillator (RBWO) using a relativistic electron beam as a HPM generation source is designed, fabricated and tested. It is a typical oscillator capable of generating a high-output EMP wave of several hundred megawatts through the interaction of a backward wave in a corrugated low-wave structure in a cylindrical waveguide. In this case, an radio frequency (RF) coupler including a low-speed wave structure part and capable of measuring RF power inside a cylindrical waveguide was designed and fabricated using a coaxial beam-rotating antenna (COBRA) connected to a cylindrical waveguide. In addition, it was designed and manufactured to be able to radiate into the air. The RF output radiated into the air was measured through a receiving antenna. The fundamental mode of the electromagnetic waves of this oscillation device is TM_{01}. Since the electric field value in the center of the RBWO circuit is 0, it is impossible to radiate electromagnetic waves onto a targeted object. Therefore, a COBRA capable of radiating after mode conversion with TE_{11} is required. Hence, a mode-converted relativistic back-wave oscillator was designed and fabricated, and the effects on electronic devices by HPM radiation and exposure were assessed [5,6]. After these confirmation experiments, additional experiments were carried out to determine the effects of the EMPs after irradiation affected various electronic devices. By doing so, it was possible to confirm the effects on electronic devices exposed to HPMs via a theoretical model of the threshold power relative to the influence on the target under a high-output microwave exposure environment. Particularly, information devices containing semiconductors can undergo serious failures and breakdowns due to thermal secondary breakdowns caused by high-output transient electromagnetic waves. In relation to this, a theoretical approach holds that reverse voltage arises due to the generation of strong current...
flows when the waves are caught in the PN-junction region. On the basis of this research, microwave-hardening methods can be developed to minimize the effects of transient electromagnetic waves on semiconductors [7–11].

2. Specifications and experimental results with a RBWO to assess the effects of EMP generation

Figure 1 shows a number of desirable features for hypothetical HPM sources which can be feasibly applied to the development of an RF-directed energy weapon (DEW). This figure presents the RF power versus the frequency of solid-state devices and a HPM vacuum electronic devices (VED). The dashed blue and thick red lines are the high-power research frontier and the average-power frontier, respectively. Thus far, we have been using the terms “directed (to point or move a thing toward a place, and aim), energy (the capacity to do work (=force x distance)), and weapon (any means of attack or defense)” the way most people do with only a vague understanding of what they actually mean, as mainly through their perceived applications. Some clarification may be necessary with a few definitions from basic physics, and then perhaps some expansion of the terms. Current DEWs normally include only sources that are electromagnetic in origin, including lasers, charged particle beams, radio frequency devices, and HPMs. However, directed-energy weapons (DEWs) are devices which destroy and defeat targets using radiated waves or beams of microscopic particles. Future DEWs may include sources other than electromagnetic types, such as acoustic waves.

Figure 1. Domains of applications of peak-power performance limits for HPM and SSD VED, *(A) in the figure is the RBWO assessed here.
(from infrared to ultrasonic waves) or other fluid and particle structures (such as vortex rings). From the above definitions, further differentiation among DEW types on the basis of HPM technologies follows. High-energy laser weapons operate via electromagnetic radiation with wavelengths usually in the infrared range. HPM weapons radiate electromagnetic energy in the high RF spectrum. Charged-particle beam weapons can project energetic charged atomic or sub-atomic particles, usually electrons [7–18]. Because the main purpose of an effectiveness analysis and EMP generation is related to DEWs in the military arena, HPM VEDs such as a magnetically insulated line oscillator, relativistic magnetron, relativistic klystron, vircator, gyrotron, and RBWO which use accelerated relativistic electron beams are commonly studied [5,6,19–32]. HPM VEDs for experimental studies pertaining to an effectiveness analysis with EMPs and the effectiveness of the electromagnetic coupling of electronics exposed to HPM radiation rely on a RBWO, as highlighted by the red dot in Figure 1. Research activities pertaining to the generation of HPM and millimeter-wave radiation have been conducted for nearly six decades. Many recent studies have been prompted by the availability of high-power, pulsed-power accelerators capable of producing intense, relativistic electron devices which drive RF structures [33–36].

One of the most successful examples of a HPM source utilizing high-current relativistic electron beams is the RBWO [37]. The proposed HPM VED is based on this type of device.

The RBWO is well known as a HPEM wave (HPM) device where a slow-wave structure (SWS) supports Cerenkov synchronism between the cylindrical TM$_{01}$ in the fundamental mode and a hollow electron beam, as shown in Figure 2.

![Image of RBWO and effectiveness](image.png)

**Figure 2.** The relativistic backward-wave oscillator (RBWO) and effectiveness according to the electric field.
To realize a device which generates hundreds of MW or a more powerful GW-level microwave device, the RBWO circuit was devised using the Cherenkov radiation that occurs when a linear relativistic electron beam interacts with a SWS. The RBWO circuit consists of SWSs, a resonant reflector, and a collector, as shown in Figure 3. In this dissertation, a 500 kV–5 kA pulsed electron beam at 100 ns is tested. It is intended to use more than 0.5 GW from a 10 GHz operating frequency.

The characteristics of the RBWO are shown in Figure 4. It is possible to improve the efficiency of the device and increase the regime of stable single-frequency operation. Previously discussed models assume that the backward propagating wave, after interacting with the beam, leaves the system without reflection. In practice, the backward wave undergoes reflection at the beam entrance of the cavity (where it is cut off by a drift tube) and becomes a forward wave. The forward wave does not interact with the beam, but it may be reflected at the beam exit of the cavity owing to poor coupling with the output horn. The reflected forward wave becomes a backward wave which can interfere constructively or destructively with an amplified backward wave. This is illustrated in Figure 4. In practice, the total round-trip amplitude reflection coefficient can be in excess of 70% and thus can be expected to influence the operation of the device. In addition, $\omega(k)$ is in the TM$_{01}$ mode in an infinitely long periodic structure. The straight line which intersects the cold structure dispersion curve is the electron beam Doppler line, $\omega = kv_{\text{beam}}$, where $v_{\text{beam}}$ is the beam initial velocity. The wave number $k$ of a point of intersection between the beam line and a sinusoidal line denotes the backward wave; while the frequencies of these two waves are identical, the operating frequency is 10 GHz. The RBWO circuit was designed to be operated at a level of 500 kV at 10 GHz according to a study of the dispersion relation.

Figure 5 shows an image and the specifications of the RBWO system. The RBWO was designed for optimized generation at 0.5 GW–10 GHz in the range of 500 kV–5 kA on the basis of the dispersion relationship shown in Figure 4. The pulse width of the electron beam voltage and current depending on the Blumlein pulse forming line is approximately 100 ns. The duration of the RF pulse is from 20 to 30 ns at this time. The maximum magnetic field generated by a pulsed magnet power supply consisting of an R–L–C circuit is 3.4 T at 10 ms in order to focus the electron beams. The power efficiency in terms of the ratio of the output

![Figure 3. RBWO circuit.](image-url)
The RF output power of the radiated microwave is measured using a crystal detector, and the frequency is measured by the heterodyne method, as shown in Figure 6. The receiving antenna can receive the radiated EM-wave pulse, after which the signal goes to an oscilloscope. In addition, the RF coupler used with the device can directly send the signal to the

Figure 4. Dispersion relation of RBWO.

Figure 5. The features and specifications of the RBWO system.
oscilloscope and measure the RF power before radiation using the crystal detector. Therefore, the RF power is calculated in reference to calibrated materials. The receiving antenna consists of an X-band waveguide adapter with RF absorbers. Moreover, Figure 6 shows the actual composition used in the RF power measurement experiment with the RBWO system.

Electrical breakdown is an important and general problem in many HPM experiments. Microwave anechoic chambers of the type used to identify RF radiation patterns are currently in use for a variety of indoor antenna measurements, EM-interference measurements, and EM-compatibility measurements [38–42]. The waveforms of the diode voltage, beam current, and microwave RF powers measured by the receiving antenna and RF coupler are shown in Figure 6. The mixed signals in Figure 6 exist because the voltage signal from the local signal generator in Figure 6 was used to calculate the frequency by the heterodyne method. The waveforms between the RF signal from the receiving antenna and the mixed signal from

Figure 6. Experimental setup and experimental results of the mode-converted RBWO.
the mixer in Figure 6 are similar to each other. The diagnostics of the RF power were measured by the RF coupler receiving antenna. The RF coupler was used to diagnose the RF power before radiation in the cylindrical waveguide. The receiving antenna detected the pulsed microwave radiation, and the radiated RF output power could be calculated using calibrating materials. Shown in Figure 6 are the results of a comparison between the RF power from the RF coupler and the radiated power from the receiving antenna. The figure shows that the RF power measured by the RF coupler is nearly identical to that at the receiving antenna. At this time, the operating frequency was 10 GHz (operating frequency = local signal generator frequency − FFT mixed signal frequency = 10.3 GHz − 300 MHz = 10 GHz), and the pulse width was approximately 30 ns (FWHM). The forms of the pulses are similar to each other.

3. Interactions with EMP under HPM propagation

Several types of targeted objects exposed to an HPM environment for a promising weapon incorporate various relevant parameters for those parts to achieve a particular degree of lethality. These experimental conditions will extend coverage of external interactions to include the target and the environment between the target electronics and the HPM source (RBWO), as indicated by the diagram in Figure 7.

The HPM source produces the desired fields and the antenna shapes the fields. The propagation transfer function represents the attenuation of the fields given the distance and sometimes from the atmosphere. Different frequencies react with the target surface differently, and the best coupling is achieved using the appropriate resonant frequencies. To access the interior electronics, the fields (and currents) must penetrate into the system, and the transfer function(s) also depends strongly on the frequency and local resonances. Finally, the fields and currents enter the target electronics and interact with the target. The process in Figure 7 is very complex and may be nonlinear with high-power fields, but it is still essential to our analysis process to establish the relationship between the electromagnetic environment and the target response. Typically this relationship is established empirically, but there are many variables (knobs) in the test design. Many organizations and countries consider effects data sensitive; accordingly, the range of data-sets available to researchers is very

![Figure 7. Process from the HPM source to the effect.](image-url)
limited. In the effects section of this paper, we will derive our effect relationships from some of the open literature papers available. Finally, we conclude this set of papers with a set of vignettes where we consider several different tactical scenarios, using actual sources to draw some conclusions about the effectiveness of these applications in a tactical environment. Owing to the limited effects data, the errors in our conclusions will be large, but overall it will be useful to explain the process. There are considerations of the transfer functions from the HPM source to the target with the goal of predicting the effectiveness of certain weapon types. The analysis outlined in Figure 7 serves to help us predict the effects RF illumination from the HPM weapon will have on the target. No strictly analytical prediction techniques have been successful; therefore, effect predictions are derived from effects data. In this section, we outline a number of general analytical techniques to estimate or maximize the effects, but most of our predictions will come from the somewhat sparse data in readily available sources.

The HPM source transmits the radio frequency fields. For this section of the paper, the source fields will be derived from the selected HPM weapon applications and combined with the antenna on the application to predict the field at the assumed target.

The COBRA used here, which consists of a mode-converting antenna lens with three sectors and a three-dimensional hyperbola convex lens, shown in Figure 8, creates an appropriate aperture field distribution for radiation using the phase differences of each sector of a multi-stepped dielectric structure. Each ray which passes through each sector experiences different electrical path lengths due to the different thicknesses of the dielectric material. These different electrical path lengths cause phase differences at the aperture plane and change the aperture field distribution.

As shown in Figure 8, the phase center of the reference point RBWO electromagnetic waves propagating through a circular waveguide is the end point of the center of the collector aperture size of the antenna, and the antenna length and thickness of the lens are used in a theoretical formula to determine this. The antenna is shown below, and it can be designed considering these values.

The electric field passed through the COBRA on a basis of the formula of radiated power can be estimated as follows,

$$ W_{\text{rad}} = \frac{P_{\text{output}}}{4\pi r^2} \quad G = \frac{1}{2} \frac{E^2}{Z_0} $$

$$ E_{W_{\text{peak}}} = \sqrt{\frac{2P_{\text{peak}}}{\pi a^2}} $$

$$ E_{\text{aperture, peak}} = E_{W_{\text{peak}}} \times \frac{a'}{a} $$

$$ E_{W_{\text{peak}}} = 19.98 \text{ MV/m} $$

$$ E_{\text{aperture, peak}} = 1.43 \text{ MV/m} $$

$$ Z_{101} = 100 \text{ohm} $$

$$ a' = 18 \text{mm} $$

$$ a = 250 \text{mm} $$

Figure 8. Design result of mode converting antenna with convex lens.
where $W_{\text{rad}}$, $P_{\text{output}}$, $G$, $r$, $E$, and $Z_0$ are the radiated power, RF output power, antenna gain ($= 27.5$ dBi), the distance between the COBRA lens and the targeted object, the radiated electric field and the wave impedance of free space ($= 377$ ohms), respectively.

If $P_{\text{output}}$ and $r$ are 0.5 GW and 2 m, respectively, the value of the radiated electric field is about 0.5 MV/m. This level is the range of the HPM field under the extreme environment according to Figure 2.

For the antenna transfer function, an antenna is needed to radiate via an optimal method. Assuming that the antenna size is many wavelengths, the signal can be radiated in a highly directive manner. Maximizing the fields to illuminate an object of interest leads to a focused aperture antenna [43]. In this case, the fields incident on the system are

$$E_{\text{rad}}(\vec{r}, S) = \sqrt{\xi} W_{\text{rad}}$$

(3)

where $W_{\text{rad}}$ is the radiated power in air, $s$ is the Laplace-transform (two-sided) variable ($= $ complex frequency), and $\xi (= s/c)$ is the free-space propagation constant. $c$ is the speed of light. Therefore, along the boresight,

$$W_{\text{rad}} = \frac{A_{\text{eff}}}{2\pi r} \left( \frac{s}{c} \right)^2 P_{\text{output}}$$

(4)

$W_{\text{rad}}$ is the radiated power density. Here, $A_{\text{eff}}$ is the maximum effective antenna area, and $P_{\text{output}}$ is the output power.

For a highly oscillatory waveform, the frequency-domain concepts for approximations in the time domain can be used. This approach is also valid for chip pulse. The output power is, accordingly,

$$P_s(s) \equiv V_s(s)I_s(-s)$$

$$P_{\text{output}} \approx V^2 Y_s$$

$$Y_s \equiv \text{Re} \left[ Y_s(s_\omega) \right] = \text{Re} \left[ Y_s(j\omega_s) \right]$$

$$s_\omega \approx j\omega_s$$

(5)

Combining Equations (3) and (4), with Equation (5) yields

$$E_{\text{rad}}(\vec{r}, s) = \frac{1}{2\pi r} \left( \frac{s}{c} \right) \sqrt{\xi A_{\text{eff}} \text{Re} \left[ Y_s \right]} V$$

(6)

The transfer function for the antenna is defined as follows:

$$F_a = \frac{E_{\text{rad}}}{V} = \frac{s \sqrt{\xi A_{\text{eff}} \text{Re} \left[ Y_s \right]}}{2\pi rc}$$

(7)

The transfer function clearly increases linearly with chip pulse frequency; $\omega = -js$.

For field propagation in free space, the $E$-field and $B$-field are orthogonal and the fields fall off according to $1/r$. Target upset usually scales with the peak field; hence, the upset
scales with $1/r$ as well. The power density therefore decays via $1/r^2$, as do the thermal effects on the target.

Another factor is the possible attenuation of the signal between the antenna and the illumination system. For example, if the anticipated propagation path involves significant loss mechanisms, one may wish to introduce a separate instance of attenuation. Therefore, a transfer function can be assigned to this propagation, allowing for any attenuation beyond the antenna factor. $F_p$ is strictly a function of the complex frequency.

$$F_p = \frac{s \sqrt{\zeta A_{eff} \text{Re}[Y_s]}}{2\pi rc}$$  \hspace{1cm} (8)

However, a highly resonant pulse can be considered as a function of $\omega_s$ provided there is not too much dispersion, such as one may experience at certain frequencies in the ionosphere. For the present purposes,

$$F_p \approx 1$$  \hspace{1cm} (9)

It should also be noted that the breakdown electric field strength can vary along the propagation path. For instance, this can be associated with variations in the air pressure. Therefore, there may be limitations on the signal imposed due to the nonlinear characteristics of the propagation path.

When analyzing the interaction with the system exterior under an electromagnetic wave with a complex system, it is often possible to identify some outer surface $S$ of the system which serves as a crude conducting field. The first-order manner of investigating this concerns the general subject of electromagnetic topologies. The first step in this approach is to find the exterior response of the system. Most important, this involves finding the short-circuit surface current and charge densities of $S$; these are cases in which there are large appendages such as power or communication lines, for which the analysis can be generalized to include corresponding appropriate equivalent circuits from an external interaction problem.

In the analysis of the external interactions, there are three basic frequency ranges of concern. For wavelengths large relative to the largest characteristic dimension ($l$) of the scatter, there is the quasi-static regime. In this regime, the surface current density is proportional to the incident magnetic field, and the surface charge density is proportional to the incident electric field. These relationships are frequency-dependent but also depend on the geometry, position of observation, and polarization of the incident electric and magnetic fields. The surface fields can be some factor multiplied by the incident fields, which can easily be one order of magnitude (10 times) higher or lower, with even more extreme variation possible.

In the resonance regime, the basic concept for analysis is the singularity expansion method. The natural frequencies (in the left half of the plane) of the exterior represent damped sinusoids in the time domain or poles on the complex frequency plane. This frequency region extends from a low frequency where $l$ is approximately a half of a wavelength to somewhat higher frequencies, both in the sense of multiple half-wave resonances and in the sense of the half or quarter wavelengths corresponding to various protrusions of $S$.

In the high-frequency (or optical) regime, the basic concept for analysis is the geometric theory of diffraction and its variants. On the illuminated side of the system, the first term,
often referred to as physical optics, can be used. The surface fields are simply twice the relevant components of the incident fields. Near the edges and exterior points or the corners, these can be even larger. On the shadow side, a more detailed analysis can estimate the fields but, in general, the fields are smaller and less important there. Summarizing, the exterior transfer function has

$$F_{\text{ext}} = \frac{\rho \sqrt{\zeta A_{\text{eff}} \Re\{Y_s\}}}{2\pi c} \approx 1 \text{ (order of magnitude sense)}$$  \hspace{1cm} (10)$$

except in the resonance region, where it can have significant peaks which could be matched to $\omega_s$ in Equation (5) to increase the response. For low frequencies, there is some variation of this over the body. For high frequencies, this applies primarily the illuminated side of the body.

The next step in the topological decomposition of the system response is the consideration of the internal interaction. This is lumped as a transfer function from the surface current and charge densities on the exterior of S to some interior port of interest, producing voltage and current waveforms there.

At low frequencies, apertures and small antennas are usually differentiators. Consider a small aperture with a wire behind it. This is usually modeled as a transmission line with a series voltage source proportional to the exterior B-dot and a transverse current source proportional to the exterior surface D-dot. Provided the wire has terminated infinite, nonzero load impedances at low frequencies, the transfer function $F_i$ in this low-frequency range is proportional to $\omega$. Similar comments apply to electrically small antennas on S; the low-frequency behavior is usually proportional to $\omega$ or an even a higher power of $\omega$ if special designs are used to reduce the low-frequency response further.

As the frequency is increased, interior resonances appear. These can be associated with internal wires which may be as long as $l$ (the largest characteristic dimension of the object) or perhaps somewhat longer. The first resonance may then occur at $\omega_{l/2}$ where this is half-wave or even quarter-wave resonant. Various lumped elements on these wires can even lower the first resonance. As the frequency is increased, shorter wires can become more important. Of moderate interest are the smallest dimensions that are important for resonances. These can include aspects such as small apertures, i.e. windows in S, or even the dimensions of boxes. In any event, this establishes some high-resonant frequency $\omega_h$ of interest. Between $\omega_l$ and $\omega_h$, the transfer function $F_i$ exhibits resonant behavior with order-of-magnitude variations as a function of the frequency.

Above $\omega_h$, the transfer function can be estimated in a different way. Considering the various wires carrying electrical signals between the various boxes, note that for wavelengths less than the distance of these wires from a ground plane provided by the local structure, the signals induced are proportional to $\omega^{-1}$; i.e. the wires act as integrators. At these high frequencies, apertures in S allow fields in a roughly frequency-independent manner such that the transfer function $F_i$ of the exterior fields to the wire signals (voltage and current) should be roughly $\omega^{-1}$. This neglects factors such as special filters at the box inputs. One may estimate $\omega_h$ based on the smallest typical resonant dimensions on the rough order of a GHz.

A canonical system response can now be defined, as illustrated in Figure 9. Here, $F_{\text{ext}}$ is combined with $F_i$ to form a composite transfer function from the incident fields to the response box inputs. This is a canonical transfer function in that it applies to typical systems,
but note that there can be exceptions for special system features. We divide the frequency spectrum as follows:

\[
\begin{align*}
\text{Band 1: } & f \leq \frac{c}{4l} \approx \frac{c}{4l} \text{ (aperture and small antenna coupling region)} \\
\text{Band 2: } & f \leq f_i \leq f_h \approx \frac{5c}{l} \text{ (resonance region, external and internal)} \\
\text{Band 3: } & f_h \leq f_i \text{ (integration region)}
\end{align*}
\]

(11)

Here, \( l \) is the characteristic dimension of the internal object.

The effectiveness of the chosen field waveform in the generated currents on and in the target is strongly dependent on the geometry of the target, as expressed in wavelengths. For large wavelengths relative to the target, the coupling is quasi-static or near static. The coupling efficiency in the quasi-static regime increases with the frequency and is low compared to the near-resonance efficiency. For wavelengths small compared to the target, the coupling becomes optical at the limit and decreases with the frequency. At target lengths near the wavelength, the coupling becomes complex; it is maximum but may vary over an order of magnitude over very narrow resonance regions.

Various techniques can be used to estimate the external currents and fields for a target. Some simple approaches were described in the previous section. These currents and fields only cause effects when they penetrate the target to the target electronics. Coupling paths to critical electronics are vulnerable to the various ways fields and currents can penetrate a target from a remote HPM source.

4. Experimental setup, results, and effectiveness analyses of exposure to EMPs from HPMs

Microwave anechoic chambers are currently in use for a variety of indoor antenna measurements, electromagnetic interference measurements, and electromagnetic compatibility measurements. The prime requirement is that a transmitting antenna located within the chamber generates a known field throughout a volume (of the chamber) of a sufficient size to perform antenna measurements. This volume is frequently referred to as a “quiet zone,” and the level of interference between direct and reflected waves within it determines the performance of the anechoic chamber [38–42].

![Figure 9. Canonical system response as a function of the frequency.](image)
The setup of the experimental system for the effectiveness analysis by EMPs from HPMs on a basis of a RBWO operating on the X-band is shown in Figure 10. This is set for an EMP effectiveness analysis at the near-field (the distance between the COBRA (antenna) and an X-band adaptor measuring the radiated RF output power in air = 2 m). The attenuation antenna to be aimed for the reference also measures the radiated RF output power in air (Figure 10(a)). In order to analyze the detailed data under the condition of HPM radiation, LED circuit electronics (samples) were installed in the anechoic chambers.

This setup was used to analyze the electrical effectiveness pertaining to the breakdown RF power level and the threshold electric field under the radiated HPM environment.

The electric field and RF output power of the radiated HPM from the RBWO with a comparison with the results of a numerical simulation are measured to use the parameter sweep results for the effectiveness analysis of the EMPs from the HPM shown in Figure 11.

Figure 10. The setup of the experimental system for the EMP effectiveness analysis: (a) Measurement system with an X-band adaptor with an attenuation antenna, (b) LED sample for the electrical analysis of exposure to EMPs with a video device equipped with a shielding box.
The electric field and RF output power were measured through a D-dot probe and an X-band receiving antenna, respectively. The LED sample for the electrical analysis exposed to EMPs in Figure 12 is the indicator estimating the critical range between a soft kill and a hard kill. In the case of a soft kill, none of the electronics are permanently damaged, but all electronics including the LED sample are rendered forever unusable. Typically, they cannot be repaired. The red dashed line indicates the critical point concerning the boundary between a soft kill and a hard kill. Measured values of the electric field expressing 5% error bars are shown in the dotted lines in comparison with the results of the numerical simulation.

Moreover, data of the applied voltages versus the radiated RF output power is shown in Figure 13 so that the correlation between the applied voltage and the RF output power can be logically authenticated.

In order to implement the effectiveness analyses of the electronics, an electric watch, a mobile phone, a portable TV, and a laptop computer were used in the experiment, as shown in Figure 14(a)–(d). All of the electronics were permanently damaged when exposed to the radiated HPMs (hard kill). The conditions of all experiments were as follows: 0.5 GW, 10 GHz, and roughly 5 kV/cm (= 0.5 MV/m) (Figure 14(e)).

Electromagnetic energy coupled into a system through deliberate antennas, via penetrations, or directly to the internal circuit wiring due to apertures can degrade the performance of the system. The degree of degradation is a function of many factors depending on the normal operating mode of the system, the mission of the system, and the components utilized in the system. The purpose of this analysis is to determine the mechanisms which cause performance degradation and provide the thresholds at which component degradation occurs (Figure 15).

To assess the impact of an EMP, or any other stimuli, on the performance of a system, the response of the system to the stimulus must be known. This response (in terms of the degradation of a component, the equipment package, or the discrete subsystems or systems) is termed the susceptibility. Susceptibility is defined as the responses of individual components, equipment packages, discrete subsystems, or complete systems to a broad range of

![Figure 11. The parameter sweep results for the effectiveness analysis of the EMP from the high-power microwave.](image)
Figure 12. The setup for the measurements of the electric field and RF output power generated by the radiated high-power microwaves from the relativistic backward-wave oscillator (RBWO).

Figure 13. The relationship between the applied voltage and the RF output power with the simulation results.
electromagnetic waveforms. The conclusion that a system is susceptible does not mean that the system’s performance is deteriorated; it means only that the system responds. At the component level, susceptibility is usually determined empirically and may be expressed in terms of a threshold. For equipment, subsystems, and systems, certain coupling modes and component thresholds are implicit in the determination of the susceptibility characteristics. Knowledge of susceptibility permits the determination of the performance degradation for various conditions of exposure. The reduced capability of a component, equipment, subsystem or system is termed a degradation of performance. The degradation may be determined by jointly considering the susceptibility and environment (stimulus) or more directly by experimental methods.

Figure 14. The experiments of the effectiveness analyses of exposure to radiated high-power microwaves.
<table>
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<tr>
<th>Targeted components</th>
<th>Damage threshold power range (Watts)</th>
</tr>
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<td>High power transistors</td>
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</tr>
</tbody>
</table>

**Figure 15.** Damage threshold power range of the representative components in electronic circuit samples in the effectiveness analyses experiments of exposure to EMPs.

![Image](image12.png)

**Figure 16.** Space charge layer in reverse voltage mode.

![Image](image13.png)

**Figure 17.** Space charge layer with a bevelled P–N junction.
Performance degradation is the deterioration of some feature of a system in response to an undesired electromagnetic environment. In some cases, some performance degradation is tolerable. When the performance degradation exceeds the limits of satisfactory performance due to stress, the system/component is considered vulnerable to that stress. System vulnerability relates the performance of the mission within acceptable limits of degradation.
to magnetic waveforms, certain hostile situations. There are two types of degradation: functional damage and an operational upset. Functional damage refers to permanent damage due to an electrical transient, while operational upset refers to temporary impairment due to an electrical transient.

With regard to semiconductor device failure, the initial understanding of semiconductor device failure is best obtained by considering a single P–N junction. Subsequent extrapolation of the phenomena to multi-junction devices is relatively straightforward. In the case of surface effects, the destruction mechanism of a surface breakdown is usually to establish a leakage path around the junction, thus nullifying the junction action. The junction itself is not necessarily destroyed, and re-etching the surface can return the junction to its normal operation. The problem of theoretically predicting surface breakdown is difficult because it depends on many parameters, such as the geometrical design, doping levels near the surface, lattice discontinuities on the surface, and the general surface conditions. It is well known that the surface of a P–N junction influences the electrical characteristics of the semiconductor device. For a junction perpendicular to the surface, surface breakdown can be explained as a localized avalanche multiplication process caused by the narrowing of the junction space charge layer at the surface (Figure 16).

It has been shown experimentally that the electric field is altered by the contour of the semiconductor surface in the vicinity of the junction, and proper contouring of the surface of a P–N junction results in a lower potential gradient at the surface (Figure 17).

The space charge layer is then spread over a greater surface distance than it would occupy if the surface edge were perpendicular to the junction. For a contoured surface, the maximum electric field at the surface will be less than that within the body of the device. By properly contouring the surface, the peak electric field there can be reduced to a fraction of that in the interior of the device. Hence, it is possible to build junctions which exhibit body breakdown prior to surface breakdown, thus eliminating this type of breakdown as a principal failure mechanism. Given that avalanche breakdowns occur more easily on a surface than within the body of a device, and because they are strongly field-dependent, a reduction in

Figure 20. Typical failure level relationships for a thermal second breakdown in semiconductor junctions.
the electric field on the surface of a P–N junction device is also desirable from this standpoint.

In an internal junction breakdown, the destruction mechanism apparently causes changes in the junction parameters as a result of localized high temperatures within the junction area. These temperatures can be of such magnitudes that alloying or diffusion of impurity atoms occurs to such an extent that the junction is either totally destroyed or its properties drastically changed. The current may be high and localized to cause melting at hot spots within the junction. Such action can result in a resistive path or paths across the junction, developing after re-solidification of the melt at the junction. The primary effect on the device operating characteristics is manifested as a decrease in the diode breakdown voltage and an increase in the leakage current, while for transistors, decreased gain and increased junction leakage currents are observed.

The most semiconductor failures occur under reverse bias conditions. This failure mechanism is termed ‘secondary breakdown.’ The voltage-current curve for a P–N junction indicates that for low-reverse voltages, the device conducts only very low current. As the reverse voltage increases, breakdown occurs with a resulting increase in the current flow. Most of the energy during breakdown is dissipated in the junction, as the junction is reverse-biased, resulting in heating of the junction. This results in the type of breakdown known as a ‘thermal second breakdown (Figure 18).’

Thermal second breakdown physically is a local thermal runaway effect at the junction induced by severe current concentrations within the device which are a function of the biasing conditions, excessive junction fields, and material defects. One reference considers a second breakdown as a filamentation phenomenon which occurs in three stages: the nucleation of a broad filament across high resistively, a broad filament across a high-resistivity region, and the growth of a second filament interior to the first, wherein the material is in a stage. The first two are non-destructive. The third involves the formation of a melt channel which results in irreversible device degradation. Under reverse bias, nucleation of a current filament starts at a localized region of high current density in the junction. More than one filament can form depending on the conditions of excitation and on the device geometry.

The voltage across and the current through the junction as a function of time when a high-amplitude transient is applied as reverse bias usually exhibits high-voltage and low-current characteristics. When a thermal second breakdown occurs, a case in which the junction fails, the voltage suddenly decreases and the current rapidly increases. It has been observed that the occurrence of a thermal second breakdown, which is an energy-dependent process, represents the point of incipient permanent damage for semiconductor devices under sub-microsecond pulse conditions. That is, once a thermal second breakdown is initiated with some additional amount of energy being further dissipated in the device, a permanent damage condition results. For most of the semiconductor devices investigated here, device degradation after a second breakdown was a result of junction damage, i.e. re-alloying of the material and the formation of a melt channel. However, for some devices, a low-voltage, high-current mode of operation was observed due to the migration of contact metallization through the junction, thus forming a metallic short. Depending upon the device, either of these phenomena may possibly occur initially, thus precipitating device failure. Another reverse-bias failure mode which has been observed on occasion in transistors is that of a current-mode second breakdown. Essentially, the effect is initiated by relatively high material current densities under the emitter during collector-to-base junction reverse pulsing,
resulting in forward bias on a portion of the emitter. When this bias becomes sufficiently high, the device becomes unstable and switches to a low-impedance, low-sustaining-voltage mode of operation.

The occurrence of the current-mode second breakdown phenomenon in itself generally has not been observed to result in permanent damage. However, if the resulting low-impedance currents are not sufficiently limited, then local hot spots form and the resulting failures are produced in a manner similar to that of a thermal second breakdown.

The forward-biased junction vulnerability to pulsed electrical energy can be understood by considering some of the basic concepts associated with thermal second breakdowns. That is, device degradation is a direct result of essentially similar melting and re-alloying reactions at various current constriction sites within the junction. The significant fact here is that due to the relatively low junction voltage at the forward-bias conditions, a correspondingly high current compared to that for the reverse direction is required to reach the critical failure energy. This larger current, in turn, results in a significant voltage drop being produced in the bulk material. Hence, a higher energy level is generally required with regard to the device terminals, thus producing much of the apparent decrease in the failure sensitivity observed experimentally. Again, because a relatively low initial impedance condition exists, the dramatic switching associated with a thermal second breakdown would not generally be observed.

In addition to single-junction mechanisms, another failure mechanism in transistors is possible due to their multifunction characteristics. This mechanism is known as the punch-through mechanism. The width of the depletion region at a reverse-bias junction will increase as the voltage across the junction increases. Because the collector-base junction of a transistor is usually reverse-biased with a small width, it is possible for the depletion region to extend throughout the width of the base, which effectively results in a short circuit. Under these conditions, the resulting current may be sufficiently large to damage the junction (Figure 19).

Many different microscopic factors may contribute to semiconductor failure. However, the roots of these mechanisms have been found to be linked primarily to the junction temperature. Therefore, in most cases, the treatment of the problem can be reduced to a thermal analysis. The worst case with reference to achieving high temperatures in the junction is when one considers that all of the power dissipated in the device occurs at the junction. This corresponds to the situation in which a high-voltage pulse of reverse polarity is applied to a junction with a high reverse voltage breakdown. When the avalanche breakdown occurs, nearly all of the applied voltage is dropped across the junction and only a small percentage is dropped across the bulk material, except for a very short pulse on the order of 10–100 ns or less where the high current required for failure causes more of a voltage drop across the bulk.

The thermal model for pulses between 100 ns and 1 ms is based on the following assumptions: (1) the heat is generated at the junction, (2) the junction is planar, and (3) the silicon material on either side of the depletion layer of the junction extends out infinitely. A one-dimensional heat equation can then be used to determine the junction temperature.

\[
\frac{\partial}{\partial x} \kappa \left( \frac{\partial T}{\partial x} \right) - \rho C_p \frac{\partial T}{\partial t} = 0
\]  

(12)

Here, the following are defined:

\( K \) thermal conductivity of silicon (W/cm°K)
If a square pulse of electric power is applied to the junction and all of the electrical power is transformed into heat, the maximum temperature of the junction is given by

\[ T_m = T_i + \frac{P}{A} \frac{1}{\sqrt{\pi \kappa P C_p}} t^{\frac{1}{2}} \]

where
\( T_m \) maximum junction temperature \\
\( T_i \) initial junction temperature \\
\( P \) electrical pulse power applied to the junction \\
\( A \) area of the junction \\
\( t \) pulse width

Rewriting the equation and taking the logarithm of both sides yields an equation which plots as a straight line with a \(-1/2\) slope on log-log paper. Equations (14) and (15) show the formula of the thermal failure model.

\[ P_{\text{max}} = A \sqrt{\pi \kappa P C_p} [T_m(0) - T_i(0)] \frac{1}{\sqrt{t}} \]

\[ \log P_{\text{max}} = \log \kappa - 0.5 \log t \]

This model allots the determination of the peak pulse power as a function of the pulse duration if the junction parameters and failure temperature of the junction are known. The theoretical failure curve shown is for a silicon junction with an assumed failure temperature of 675°K.

For short pulse widths (<100 ns), a constant energy condition independent of the pulse width prevails for the initiation of a junction failure. From physical considerations, this is the required energy input to a volume of material (the current constriction site) in order for that volume to achieve an increase in the temperature under adiabatic conditions. One-half power time dependence is obtained for longer pulse widths, which is indicative of the heat loss from the (constriction site) volume to its surrounding medium. Direct time dependence for energy found at longer pulse widths (signifying a constant power input) is indicative of thermal equilibrium resulting in a steady-state temperature at the center of the volume. This constant power level approaches the manufacturer’s CW rating for the device as the pulse width becomes very large (>0.1 ms) (Figure 20).

For failures due to surface break, no \( T \)-dependence ensues, and the failure level is dependent only on the pulse power, or equivalently, the failure level is power or voltage-dependent
and not energy-dependent. For example, the failure of microwave diodes for 2–20 ns pulses has been found to be dependent on the peak power rather than the energy. Other devices have also been found to be sensitive to the voltage; this voltage sensitivity appears to occur occasionally across the surface, i.e. surface flashover, and for higher voltage pulses due to punch through. This voltage sensitivity is a rise-time effect and can occur in some devices with a longer pulse (>30 ns) provided that the mechanism responsible for the voltage sensitivity occurs before any other effect, viz. thermal failure.

5. Summary and discussion

A GW-level RBWO operating on the X-band was fabricated to operate in a relativistic region with a gamma factor ($\gamma$) of 2 at an acceleration voltage of 500 kV–5 kA with a COBRA that converts TM$_{01}$ to TE$_{11}$. A test of the performance of EMP generation for 0.5 GW–10 GHz was successful, indicating that it is possible to analyze the effects on electronics of EMPs generated from HPM radiation and their propagation.

Several failure types have been found through effectiveness analyses of EMPs exposed to HPM environments.

- Resistance value change: Failure is defined as a change in a value beyond normal tolerance. The importance of this change is dependent on the circuit function. This mode of failure can be due to thermal effects (energy dissipation) or the voltage-induced stress.
- Internal breakdown: This breakdown occurs when the resistor under test opens but does not blow apart, with no external evidence of arcing present. These occur due to thermal dissipation with the device.
- Arc across the resistor casing: This type of breakdown is exemplified by an arc across the external surface of the resistor. No damage to the resistor results from this failure.
- Catastrophic breakdown: This type of breakdown occurs when an external arc starts across the resistor, but due to some defect in the ceramic casing, it re-enters the core. The pulse energy is then dissipated in only a small part of the resistor and causes the casing to rupture (blow off) and the resistor to open.
- Synergistic effects: An additional area of concern in the case of a semiconductor failure has been that of synergism between the electrical overstress due to an EMP and the gamma-ionizing dose rate. While exhaustive studies of synergism have not been conducted, four independent and reliable investigations have shown no evidence, or only weak evidence, of synergistic effects on discrete semiconductors or ICs. In complete systems, however, the triggering of a circuit into conduction can cause permanent damage due to electrical pulses.

All of these failure modes have been observed in resistor tests. To define a safe working voltage level for the resistor, it was given as the level where the resistance did not change as a result of the applied pulses.

With regard to microwave hardening, protection of electrical and electronic systems from electromagnetic or induced electrical impulses is certainly not a new problem. The earliest efforts were most likely associated with protection from lightning, which is a natural phenomenon. As the propensity of electronic systems (radar and communications) increased, electromagnetic interference (EMI) became important and protection against its disruptive effects was required. Recent EMPs from a nuclear burst were recognized as yet another
potential disruptive source of electromagnetic energy. There are many ways to protect systems against EMPs. Many of the approaches and concepts were borrowed from EMC (electromagnetic compatibility) and lightning technology. Although this borrowed technology provides guidance for EMP mitigation, EMC and lightning protective techniques, procedures, and devices are not adequate for EMP protection in most cases. There are two types of degradation which must be protected against: (1) functional damage to critical portions or components in the system, and (2) operational upsets of critical portions or circuits within the system. The type of hardening applied and the level of protection provided are functions of the sensitivity of the system to these disruptive effects. Protection is realized by choosing the most appropriate approach or combination of approaches along with the appropriate protection concepts and then implementing these through the design of microwave hardening. The subject of protection approach is concerned primarily with the level at which the required protection is to be achieved (i.e. the system level, circuit level, or component level). When applying protection at the system level, the primary objective is to keep the undesired energy out of the system, or at least the sensitive-mission critical portions of the system. At the circuit level, the primary objectives are to limit the undesired energy reaching the sensitive circuit or components, and to design less sensitive circuits. At the component level, the primary objective is to select the least sensitive component in terms of an undesired response while still preserving the performance criteria.

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References


