Physical Mechanisms and Design Principles in Mode Filters for Oversized Rectangular Waveguides

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Abstract—Mode filters, based on oversized rectangular waveguides with periodic symmetrical junctions, are addressed through analytical models under a small coupling approximation to explain the physical mechanisms of their behaviour, and provide guidelines for their design. In the junctions, which are partially filled with a lossy dielectric, both propagating and evanescent modes play a role. Filter parameters are studied with respect to their impact on the excitation, propagation, and absorption of such modes in the junctions. Particular attention is paid to practical choices a designer has to face in real applications. The performance of single versus multiple lengthwise slots with the same total length is compared, and high-power matters like voltage and thermal breakdown are discussed. A novel junction with two absorbing layers is proposed to help designers in maximizing absorption, while fulfilling thermal constraints.

Index Terms—high-power microwaves, modal coupling and interaction, passive microwave components, waveguide filters, guided-wave structures.

I. INTRODUCTION

Oversized waveguides are an effective transmission means for large amounts of microwave power (and energy) over medium/long distances, and are largely employed in diverse applications within the branch of the so-called high-power microwaves. Examples include particle accelerators and colliders [1], space propulsion and power beaming [2], radars [3], directed energy weapons [4], wireless power transfer [5], lower hybrid [6] and electron cyclotron [7] systems for nuclear fusion. Oversized waveguides combine high power capability with low ohmic attenuation, but they are subject to unwanted conversion losses into spurious modes, i.e., propagating modes different from the working one, mostly at discontinuities such as flanges or bends. Mode filters are passive microwave components designed to absorb these unwanted modes so as to avoid dangerous trapped-mode resonances [8].

Rectangular and circular waveguides are mostly used in the aforementioned application fields. In particular the former are employed for reflectometry measurements in tokamak plasmas [9] and ultra-high-power systems at the Stanford linear accelerator center [10]. Their use was proposed in the International Tokamak Experimental Reactor (ITER) as regards both plasma diagnostics [11] and heating [12] systems. Berdnik et al. designed a slot antenna based on a multi-mode rectangular waveguide [13], and modal fields in all-walls longitudinally corrugated waveguides have been recently studied [14].

Mode filters for oversized rectangular waveguides can be realized, without relying on inner septa, through symmetrical slots connected to junctions, which are partially filled with an absorbing material. According to the first papers, which briefly addressed these components in the 1960s [15], [16], the unwanted modes can be grouped in two classes. The class A comprises the $TE_{m0}$ and $TM_{m0}$ modes with $n \neq 0$; the class B includes the $TE_{mn}$ modes with $m \neq 1$, which can be further divided depending on whether $m$ is even or odd. The first type of modes can be attenuated with the structure of Fig. 1a, whereas the filter of Fig. 1b absorbs the second class with $m$ even and the $TE_{0n}$.

The study of these structures was not pursued for several years following the fact that optical fibers were preferred to waveguides for long-distance communications. In recent years, it has been resumed due to the renewed interest in oversized waveguide transmissions. Advancements in computational electromagnetics now enable the analysis of mode filters with full-wave commercial simulators [17], [18], but general-purpose solvers are not efficient for multi-mode, electrically large structures. Moreover these approaches provide little insight on the functional dependences between filter parameters and performance. A full-wave modal approach was proposed in [19], and partially overcomes these issues, but physical mechanisms are just briefly sketched.

By relying on a modal formalism and the perturbation approximation, the present work sets up plain physical models for mode filters and apply them to real design matters. The calculation of power absorption is more accurate than in [19] and benchmarked against three different full-wave solvers. In comparison with [20], junctions with TM modes, magnetic conductors and practical constraints are taken into account, and model limitations are discussed. Furthermore the electric field amplitude along the junction is derived, and a novel geometry with two absorbing layers is studied with an extended version of the model.

The paper is organized as follows: the excitation of the modes in a junction and their absorption are dealt in Section...
II, where the analytical models are discussed and validated. Section III explains the function of mode filters in high-power systems, and focuses on improving their absorption both in general and in compliance with some design constraints. Conclusions are drawn in Section IV.

II. ANALYTICAL MODEL

A. Coupling

Given an infinitely long cylindrical waveguide with cross-section $S$ in the $x'y'$ plane, and a current source $J$ localized in a volume $V$ of the waveguide, a set of TE and TM modes is excited. The amplitude of the generic $i$-th forward wave from this set can be derived as [21]:

$$A_i^+ = \frac{1}{2} \int_S J \cdot E_i^- dx'dy'dz'$$

being $E_i^-(x', y', z')$ the E-field of the $i$-th backward wave and $E_{1x}(x', y')$ and $h_{tx}(x', y')$ its transverse eigenfunctions.

Assuming a small coupling regime, we can apply (1) to each junction of the rectangular waveguide in Fig. 1, by replacing $J$ with the surface currents on an equivalent unslotted wall. Such impressed currents only depend on the unperturbed magnetic field distribution on that wall if the Schelkunoff’s field equivalence principle is used, as depicted in Fig. 2. After normalizing the eigenfunctions, so that the integral at the denominator of (1) is unity, the $i$-th wave amplitude excited at the junction input becomes

$$A_i^{(m)} = \int_{\text{slot}} H \times \mathbf{e}_{i}^{(m)} \cdot \hat{z}'dS$$

This expression can be thought as a projection integral of waveguide surface currents on junction eigenmodes: the larger the match of a current distribution with a mode pattern, the larger the coupling.

For the geometry and dimensions of Fig. 2, the analytical solutions of (2) have been already given [19], but their implications have not been completely addressed. We will analyse them considering the $i$-th junction mode with indices $mn$ and the $j$-th waveguide mode with indices $pq$. The ratio between their wave amplitudes, namely the coupling coefficient, has the following main functional dependence

$$K_{i \rightarrow j} \propto \sqrt{Z_{mn}} \cdot \left\{ \sin \left[ \frac{p\pi(a_w - b_c)}{2a_w} \right] - (-1)^n \sin \left[ \frac{p\pi(a_w + b_c)}{2a_w} \right] \right\}$$

where $Z_{mn}$ is the wave amplitude and the following definitions are adopted for longitudinal and transverse wavenumbers:

$$\gamma_{i,j} = \sqrt{u_i^2 - k_0^2}$$

$$u_i = u_{mn} = \sqrt{(m\pi/a_c)^2 + (n\pi/b_c)^2}$$

$$u_j = u_{pq} = \sqrt{(p\pi/a_w)^2 + (q\pi/b_w)^2}$$

with $k_0 = \omega\sqrt{\varepsilon_0\mu_0} = 2\pi/\lambda_0$. The coupling in filters with slots on the side walls is directly deduced from the previous one by considering a 90 deg flipped waveguide.

The function of (3) in braces gives useful guidelines for slot design. Using prostaferesi formulas its form becomes

$$\begin{align*}
2\sin \left[ -p\pi b_c/(2a_w) \right] \cos(p\pi/2), & \text{ for } n \text{ even } \\
2\cos \left[ -p\pi b_c/(2a_w) \right] \sin(p\pi/2), & \text{ for } n \text{ odd }
\end{align*}$$

showing that it vanishes when $n + p$ is odd. An important consequence of this condition is that the fundamental mode of the waveguide TE$_{10}$ cannot couple to TE$_{(jm)}^{(jun)}$ junction modes, being $p = 1$ and $n = 0$. The TE$_{(01)}^{(jun)}$ mode is the lowest-order junction mode the TE$_{10}$ can interact with. Its absorption has to be minimized because the TE$_{10}$ is the operational mode of the waveguide and is supposed to pass through the filter with minimal insertion losses. If the slot has $b_c < \lambda_0/2$, according to (4a) the TE$_{(01)}^{(jun)}$ mode is evanescent in the vacuum section of the junction (see Fig. 2). The fraction of coupled power reaching the absorbing material is thus

$$|A_{10}^{(m)}|^2 e^{-2h_w \sqrt{(\pi/b_c)^2 - k_0^2}}$$

and can be decreased by reducing $b_c$ or increasing $h_w$. Modes with higher cutoff frequency exhibit faster exponential decay.

Previous remarks also affect slot position and orientation. Currents that are widthwise and lengthwise to the slot respectively excite the TE$_{(jm)}^{(jun)}$ and TE$_{(mn)}^{(jun)}$ modes, whose $e_t$ in (2) has either $y$ or $x$ components. Only the former are above cutoff when $b_c < \lambda_0/2$, and they are not excited by the dominant waveguide mode if its surface currents flow parallel to the rectangular apertures. Different arrangements from Fig. 1 would extract active power from the TE$_{10}$.

The plots of wall currents provide a clear insight of previous concepts. Here we rely on this visual approach for two representative cases shown in Fig. 3. The first one depicts the TM$_{11}$, i.e., a class A mode that only affects apertures on the lateral walls of the waveguide. This is true for any TM mode: the fact of being transverse magnetic ($H_z = 0$) implies that the wall currents are always parallel to the longitudinal axis of the waveguide. The second case is an example of class B modes with $m$ odd: this subset, which includes the TE$_{10}$ too, cannot be absorbed with junction-based mode filters. Most of their surface currents are lengthwise to the slots, and

Fig. 2. Junction on the top wall seen as a separate waveguide, fed by an equivalent current distribution $J_s = -\hat{z}' \times H_{\text{wall}}$ over a perfect magnetic conductor (PMC). The structure can be terminated with a perfect electric conductor (PEC), and the modal amplitude of an evanescent mode is sketched.
the small transverse components on the top and bottom wall are symmetrical and averaged to zero in the integral (2). In authors’ opinion, this drawback has to be considered in the design of components like bends, which are the main source of spurious modes. The conversion into TE\textsubscript{p0} modes with odd \( p \) should be minimized even at the expenses of other unwanted modes. Otherwise different devices have to be used, e.g., those based on directional coupling [15], which are more cumbersome and suppress a narrower modal spectrum than junction-based mode filters.

The rectangular apertures on the waveguide walls are defined by the lengths of their short and long sides, named as \( b_c \) and \( a_c \), respectively (see Fig. 3). According to (6), the former has to be short to reduce insertion losses and, in any case, less than half a vacuum wavelength. For junction modes with \( n = 0 \), the dependence of coupling coefficients on \( b_c \) is

\[
K_{\text{TE}_p \rightarrow \text{TE}_m} \propto \frac{1}{p\sqrt{b_c}} \sin \left( \frac{\pi r b_c}{2a_w} \right)
\]

(7)

for slots on the top and bottom walls, and

\[
K_{j \rightarrow \text{TE}_m} \propto \frac{1 - e^{-\gamma_{\text{m0}} b_c}}{\sqrt{b_c}}
\]

(8)

for slots on the side walls. If sine and exponential functions are replaced with the 0\textsuperscript{th} and 1\textsuperscript{st} terms of their Taylor’s expansion for small arguments, both previous expressions reduce to \( K \propto \sqrt{b_c} \). Such result means that large values of \( b_c \) increase the power coupled to propagating junction modes from unwanted waveguide modes and so the absorption of the latter. The choice of \( b_c \) is thus a compromise between the absorption performance of the mode filter and its figure of insertion loss.

As far as the longest dimension of the slot is concerned, the behaviour of the squared coupling coefficient is depicted in Fig. 4 for a longitudinal junction on the top or bottom wall of a waveguide. Such test case will be also used later on in the paper because it is a practical example of the 5 GHz system proposed for ITER [12]. The transmission line is based on the WR-430 (\( a_w \times b_w = 109.22 \times 54.61 \text{ mm}^2 \)) with ten propagating modes at 5 GHz, and the spurious waveguide mode under investigation is the TE\textsubscript{01}. The latter is indeed easily generated in some H-plane trapezoidal bends [22].

In the slots considered for Fig. 4, the shortest side \( b_c \) is \( \lambda_0/12 \), a value that fulfils the small coupling approximation and gives negligible insertion losses for the working TE\textsubscript{10} mode. The peaks in correspondence of integer multiples of 30 mm are due to \( Z_{m0} \) in (3) that diverges when \( a_w = m\lambda_0/2 \), i.e., when the cutoff frequency of a junction mode is equal to the operational frequency. This is a limit of the present model, which is thus not reliable very close to the mode cutoff: the actual behaviour, as predicted by a full-wave code [19], is shown in the inset of the same figure for the TE\textsubscript{20} mode. However junction modes with \( \gamma_{m0} \approx 0 \) lead to negligible absorption, so the behaviour in correspondence of their cutoff is of no concern.

Regarding the openings on the side walls, Fig. 5 shows the power coupled from different waveguide modes to the dominant junction mode. Since slot length \( a_c \) is restricted by the height of the waveguide, only a few junction modes, often a single one as in the present case, are above cutoff.

### B. Absorption

The junctions are multi-layered loaded waveguides with the same cross-section as the slots (see Fig. 2). They are usually terminated with a PEC, but different loads can be used. Other parameters are the depths of their vacuum (\( h_v \)) and absorber (\( h_a \)) sections and the complex dielectric constant...
of the absorbing material. Next examples will use doped Silicon Carbide (SiC) with relative permittivity $\epsilon_r = 13.5$ and electrical conductivity $\sigma = 1.65$ S/m at 5 GHz.

Such structures behave like cavities: power absorption is maximized at the resonance, i.e., for constructive interference between forward and backward waves. In terms of a simple resonator model, the field intensity in the vacuum section of the junction is

$$W = \frac{1}{2} \int \int |E |^2 dV$$

where $R_m$ and $\Gamma_m$ are the reflection coefficients seen by the backward and forward $mn$-th wave at $z = -h_v$ and $z = 0$, respectively, according to Fig. 6. The resonant condition is satisfied for

$$h_v = \frac{\lambda_{mn}}{4\pi} (\angle R_m + \angle \Gamma_m + 2\pi \ell)$$

with $\ell = 0, \pm 1, \pm 2, \ldots$, and $\lambda_{mn}$ is the mode wavelength. A proper choice for $\ell$ considers its impact on both insertion losses and filter size. On one side, according to (6), the $TE_{01}$ mode loses more than 99.8% of its power before reaching the absorber if $h_v > \lambda_{01}/2$. On the other side, in practical applications, several waveguides may be closely arranged along the transmission path, constraining the acceptable size of the device.

For a rigorous calculation of the power absorption of junction modes, the wave amplitudes in the lossy dielectric must be derived. We equate transverse field expressions at layer boundaries and exploit mode orthogonality to obtain an independent system of equations for each mode. With $k_m = \alpha_m + j\beta_m$ indicating the propagation constant in the absorbing material, we find the following linear system

$$B^+_{mn} = B^-_{mn} e^{2\beta_m h_a}$$  \hspace{1cm} (11a)

$$\sqrt{Z_{mn}} (A^+_{mn} + A^-_{mn}) = \sqrt{Z_{mn}} (B^+_{mn} + B^-_{mn})$$  \hspace{1cm} (11b)

$$\sqrt{Z_{mn}} (A^+_{mn} - A^-_{mn}) = \sqrt{Z_{mn}} (B^+_{mn} - B^-_{mn})$$  \hspace{1cm} (11c)

$$A^+_{mn} e^{\gamma_m h_v} = A^+_{mn} + R_m A^-_{mn} e^{-\gamma_m h_v}$$  \hspace{1cm} (11d)

where (11a) comes from the junction end, (11b) and (11c) from the vacuum-absorber interface and (11d) from the slot input. In the first condition, $\pm$ signs apply to PEC and PMC terminations, respectively.

From (11), the wave amplitudes in the absorber can be expressed as a function of the power coupled through the slot.

Then the damping $W$ of a junction mode with unitary input amplitude ($A^+_{mn} = 1$) can be calculated as

$$W_{mn} = \frac{1}{2} \int \int |E |^2 dV = \frac{\sigma |Z_a|^2}{2|\mathcal{G}_m|^2} \left( F_0 + \delta_{TM} \left| \frac{u_{mn}}{\kappa_{mn}} \right|^2 F_1 \right)$$

with

$$\mathcal{F}_s = \sinh(2\alpha_{mn} h_a) / \alpha_{mn} \mp (-1)^s \sin(2\beta_{mn} h_a) / \beta_{mn}$$

$$\mathcal{G}_i = \frac{e^{k_i h_a}}{2 \sqrt{Z_{i}^* Z_{i}'}} \left[ e^{\gamma_i h_v} (Z_{i}^* + Z_{i}') - R_i e^{-\gamma_i h_v} (Z_{i}^* - Z_{i}') \right]$$

$$\mp \frac{e^{-k_i h_a}}{2 \sqrt{Z_{i}^* Z_{i}'}} \left[ e^{\gamma_i h_v} (Z_{i}^* - Z_{i}') - R_i e^{-\gamma_i h_v} (Z_{i}^* + Z_{i}') \right]$$

where $s = 0, 1$, the subscript $i$ stands for $mn$, and $\delta_{TM}$ equals to 1 and 0 for TM and TE junction modes, respectively.

Although the term $R$ in $\mathcal{G}_m$ has to be computed numerically, the closed-form expression (12) represents a powerful tool to understand the junction behaviour, allowing the analytical study of functional dependences on geometrical and electromagnetic parameters. The variation of (12) with the length of the vacuum section and that of the absorber section is shown in Fig. 7 for a PEC-ended junction with two propagating modes: $TE_{10}$ and $TE_{20}$. The first plot has a periodicity of $\lambda_{10}/2$ (≈ 34 mm) with respect to the length of the vacuum section, this is consistent with (9). For the $TE_{20}$ the periodicity is larger (≈ 118 mm) because it is closer to its cutoff frequency: as the latter is approached, the optimal values of $h_v$ become unfeasible because the guided wavelength tends to infinity. In addition, Fig. 7 suggests that the absorption is weaker for higher-order modes. The parameter $h_v$ plays the major role in maximizing the absorption, while $h_a$ allows a
finer optimization. Some plots versus \( h_a \) with the optimal \( h_v \) are given in Fig. 8 for the TE\(_{20}\) mode by varying the properties of the absorber. The real part of its permittivity determines a shift of the optimal \( h_a \), while the imaginary parts affects more the amplitude of the curve oscillations.

C. Validation

Given an input waveguide mode carrying a power \( P_j \), and \( M \) junction modes with the lowest cutoff frequency among the excited ones, the absorbed power is approximately

\[
P_A \approx P_j \sum_{i} |K_{j\rightarrow i}|^2 W_i
\]

Such expression has been applied to the TE\(_{01}\) in a WR-430 with a short junction, and results have been compared with the calculations of three full-wave solvers based on different numerical methods. Fig. 9 reports a case of benchmark showing good agreement.

We opted for a numerical validation because direct experimental characterization of oversized structures like mode filters cannot be performed with standard equipment. Spectrum and (vector or scalar) network analysers with their adapters are conceived to measure the dominant mode of a rectangular waveguide; mode converters are thus required, but their design and characterization is a topic of research by itself.

### III. Design Principles and Enhancements

**A. Scope and Design of Mode Filters**

The performance a mode filter has to fulfill depends on the topology of the microwave system and its components. Bends and converters are the main source of unwanted modes that, when repeatedly reflected inside a section of the transmission line, can set up a standing wave (trapped mode resonance). The effects are an increase of ohmic losses and maximum electric field, leading to a reduced power-handling capability of the transmission line to avoid electrical breakdown.

A practical example is shown in Fig. 10, where the first mode converters inject a spurious mode with electric field amplitude \( E_0 \). This mode resonates along the transmission line with transmissivity \( T \) between the mode converters due to their reflectivities \( R \), but analogous standing patterns can occur along shorter sections of the waveguide path, e.g., between two bends. If the system is modeled as a simple resonator, the maximum electric field of the mode is

\[
E_{\text{max}} = \frac{E_0}{1 - r} \quad (14)
\]

with \( r = \sqrt{TRTR} \). The layout resembles a realistic setup [23], where \( R = -1.1 \text{ dB} \) and \( T \approx 0 \text{ dB} \), resulting in an amplification \( E_{\text{max}}/E_0 \) of 4.47. If a mode filter with attenuation \( A = 0.4 \) is inserted before the last converters, to a first approximation, \( r = \sqrt{TRATRA} \) and the field amplification becomes three times lower.

The example is aimed at clarifying that even the partial absorption of a spurious mode improves the power carrying capacity of a system and may be enough, depending on the specific context. Mode filters are indeed aimed at reducing the electric field below a safety limit rather than achieving a 100% mode purity. For instance Y. Qin and Y. Yang considered satisfactory a reduction of around 2 dB in the trapped-mode resonance spectrum [24].

At any rate, a filter with the absorption of Fig. 9, which is less than 1.5% in correspondence of the highest peak at \( h_a = 3.5 \text{ mm} \), can be hardly satisfactory for an application. An improvement to about 3% can be achieved using the optimal vacuum depth \( h_v \approx 15 \text{ mm} \) that satisfies (10). Besides, according to (7) and (8), the increase of slot width enlarges the coupling. By taking \( b_c = 10 \text{ mm} \), a value still acceptable regarding insertion losses [19], the absorption is...
roughly doubled. A further factor two of enhancement is achieved using two symmetrical junctions as in Fig. 1 rather than a single one. The last knob a designer can play with is the filter length. In the case of filters for modes of class A, an increase of filter length results in cascading junctions on the sides of the waveguide. Regarding filters for modes of class B, the cascade of slots can be replaced also with continuous openings on the top and bottom walls of the waveguide. The best design choice between the two configurations is addressed in the next section. It is worth mentioning that, from a practical viewpoint, the frequency behaviour of mode filters is not a concern because they are mostly employed in microwave systems powered by very narrow-band sources such as high-power klystrons or gyrotrons.

B. Overmoded versus Single-mode Junctions

A key issue in the design of mode filters with junctions on the top and bottom walls is the choice between continuous or segmented grooves. To address this point, two types of mode filters, with total length varying from 60 to 180 mm, have been optimized to absorb the TE$_{01}$ modes. They are based on either a single oversized and several single-mode junctions, respectively. A distance of 2 mm between junctions has been considered to allow for mechanical stiffness, while the constraint $h_a + h_v < b_v/2$ has been enforced to limit filter size. The outcome of this comparison is depicted in Fig. 11, showing that short monomodal junctions are preferable.

A similar comparison for the waveguide mode TE$_{20}$ returns the same result. A study of the modal content in the junction using (3) shows that the coupled power at the slot is much higher for oversized grooves, but the extra power is mostly carried by evanescent modes, whose contribution is negligible because a minimal part of their power reach the absorber. As far as the power coupled to propagating junction modes is only considered, it is comparable, but in the continuous junction it is distributed among different higher-order modes, whose absorption is less efficient as noticed in Fig. 7. Furthermore the vacuum depth of the corrugations can satisfy (10) for a single mode at a time, resulting in suboptimal absorption for the other modes.

C. High-power Constraints

In high-power continuous-wave applications, the design of mode filters has to consider voltage breakdown and excessive heating in addition to the absorbing behaviour. The analytical model of section II is a useful tool with respect to such matters too. Furthermore it can be easily extended to study novel junction configurations.

Regarding voltage breakdown, waveguide slots represent the most risky part of the junction because they generally experience the strongest electric field. A constrained optimization of (12) can be quickly carried out with respect to $h_v$ and $h_a$, looking for the highest absorption, given a maximum allowed E-field amplitude at the waveguide opening. The result of such an optimization is reported in Fig. 12, considering an input power of 1 W for the spurious waveguide mode TE$_{01}$. The curve of optimized absorbed power shows that the maximum absorption corresponds to the highest electric field amplitude in the groove and that a reduction of the latter, for a fixed slot size, cannot be achieved without relaxing the filtering performance of the device. In the same figure, there is an inset that depicts the E-field amplitude along the junction for three cases corresponding to $E_{\text{limit}} = 10, 16.5$ and 23 V/m for a spurious mode of 1 W. Unlike Fig. 6, the axis origin ($z' = 0$) is now the junction entrance.
related issue that is worth being addressed is how to distribute the maximum absorption on a larger volume. For the highest $P_A$ of Fig. 12 the power is dissipated inside a very thin layer of Silicon Carbide ($h_a \approx 2.7$ mm) and an increase of the absorber depth would produce suboptimal performance. We propose the use of a multi-layered junction, as the one depicted in Fig. 13 with two layers of SiC. This novel configuration allows a larger design flexibility, so that, for a given value of $h_a$, the set of remaining depths ($h_a$, $h_i$, and $h_l$) can be optimized to achieve absorptions as close as possible to the maximum value of 3.3%. The latter cannot be exceeded because it is the largest value attainable with that coupling. The optimal absorptions achievable with one and two layers of lossy dielectrics are plotted in Fig. 14. A linear system with eight boundary conditions, equivalent to (11) but relevant to the new geometry, was solved and the E-field in the absorbing layers was integrated according to the Joule integral (12).

The largest value of $h_a$ giving the maximum absorption in the two-layers junction is 5 mm. The variation of electric field amplitude along $z'$ for this case is depicted in Fig. 15, from which two remarks can be done. The one is that the power is mostly dissipated in the first layer of Silicon Carbide; the other is that the highest value of the curve is the same as in the one-layer junction. Indeed, also in the multi-layered configuration, the maximum absorption cannot be achieved with an electric field lower than 23 V/m.

It is worth noticing that the optimizations of Figs. 12 and 14 have been carried out in around 20 seconds and 14 minutes respectively, whereas a commercial software on the same computer would require hours and days of computation.

IV. CONCLUSION

Dielectric-loaded junctions in oversized rectangular waveguides can be profitably used to absorb unwanted modes propagating along the transmission line. Their behaviour has been studied through analytical models, whose reliability has been assessed and validated in a regime of small coupling.

To avoid the absorption of the operational mode of the waveguide, junctions must always have a dimension shorter than half a wavelength. Furthermore they can be only positionned longitudinally on top and bottom walls, and transversely on side walls. Power coupling scales almost linearly with the length of the shortest junction edge, while the other size affects the excited modal content. As far as the absorption of junction modes is concerned, a resonator model can be considered, revealing a periodic behaviour with respect to the length of the vacuum section of the groove. Analytical expressions of absorbed power have been instead derived to study the functional relations with the absorber parameters.

Directions to enhance the absorption of mode filters have been thus given. The performance of an oversized groove on the top wall of the waveguide has been compared to the absorption attained by a groove with the same length, but segmented into monomodal cells, demonstrating that the latter choice is preferable. We have shown that the compliance with high-power constraints can entail a suboptimal absorption. As regards thermal issues, multi-layered junctions have been proposed to compensate much part of such reduction of performance.

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