Attenuation in Lossy Circular Waveguides

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Abstract ─ We present a simple closed-form approach to calculate the attenuation of waves in lossy circular waveguides. A set of characteristic equations is first derived by matching the tangential fields at the wall boundary with the constitutive properties of the conducting wall material. In order to represent fields' penetration into the lossy wall, a perturbation term is then introduced into the equation. We apply the Finite Difference Method to derive the closed-form expression of the perturbation terms for TE and TM modes. The propagation constant can be found by incorporating the perturbation term into the dispersion relation. Our results show good agreement with those obtained from the rigorous transcendental equations. However, unlike the transcendental approach which is usually laborious in solving, our closed-form approach leads to simpler analysis and, therefore, allows the attenuation to be easily computed.

Index Terms ─ Attenuation constant, circular waveguide, propagation constant, tangential fields, TE modes and TM modes.

I. INTRODUCTION

Analysis of loss in a circular waveguide has been widely performed using the rigorous transcendental formulation developed by Stratton [1]. Literature which implements Stratton's approach in their analysis includes those in hollow waveguides, dielectric rods, multilayered coated waveguides, as well as, lossy and superconducting waveguides [2-8]. In reality, the tangential fields in a waveguide are continuous across the boundary of the wall. In Stratton's approach, the tangential fields which describe the propagation in the waveguide are matched at the boundary with those penetrated into the lossy conducting wall. A transcendental equation is then derived by finding the determinant of the coefficients. Although the loss computed using Stratton's approach shows high accuracy, the roots of the solution can only be achieved via a root-finding algorithm. This means that an effective compiler tool, an efficient algorithm and appropriate initial guesses are necessary in order to allow the solution to converge. Hence, the process of solving numerically for the roots of a transcendental equation is usually laborious.

Unlike the transcendental equations, closed-form equations lead to much simpler analysis. The results of the closed-form solutions can easily be obtained in a straight-forward manner and they give more intuitive insights into the inherent behavior of the variable to be solved [9, 10]. However, due to the assumptions made while simplifying the equations to their closed-form expressions, the simplicity found in these equations usually comes at the expense of accuracy. Take for example, the closed-form power-loss method adopted by most textbooks to illustrate the loss in waveguides [11-13]. This method is only valid for certain modes and at a certain range of frequencies f above its cutoff f_c [14, 15]. When deriving for its mathematical expressions, the power-loss method assumes wave propagation in a lossless waveguide. Since a lossless waveguide behaves like an ideal high pass filter, it gives infinite attenuation at f below f_c. Also, the modes in a lossless waveguide are orthogonal to each other, i.e., each mode can exist separately in the waveguide. In reality, however, the modes in a practical waveguide may co-exist at the same time. Hence, the power-loss method fails to account for the loss arises from the concurrent existence of multiple modes in the waveguide. Clearly, the power-loss method may only be good enough for finding the initial approximation of loss in a waveguide. It may not, however, be appropriate when loss in the waveguide is a

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critical factor and accurate prediction of it is necessary.

In [10], we have developed a novel closed-form approach to calculate the loss in both rectangular and circular waveguides with finite conducting wall. The loss in the waveguides is found by solving for the root of a quadratic equation, derived by matching the tangential fields at the boundary with the electrical properties of the wall material. Although the results were found to be accurate and that they agree very well with those obtained from the rigorous approaches (such as Stratton's formulation), the mathematical expressions are long and cumbersome. Here, we extend further the closed-form solution in [10] for the case of a circular waveguide. We will demonstrate that by removing the redundant higher-order variables in the equations, a much simpler set of equations, which give equally accurate results, can be derived.

II. FORMULATION

Figure 1 shows the geometry of a circular waveguide. At the boundary of the wall, the constitutive properties can be related to the tangential electric fields E_t and tangential magnetic fields H_t as [14, 15]:

$$\frac{E_t}{H_t} = \sqrt{\frac{\mu_w}{\varepsilon_w}}, \tag{1}$$

where μ_w and ε_w are the permeability and the permittivity of the conducting wall material, respectively. The permittivity ε_w is complex and is given by [12]:

$$\varepsilon_w = \varepsilon - j\frac{\sigma_w}{\omega}, \tag{2}$$

where ω is the angular frequency, σ_w the conductivity of the waveguide wall, and ε is the permittivity of free space. The equations obtained from (1) admit non-trivial solutions only when the determinant vanishes. This yields the following characteristic equation for a circular waveguide [14]:

$$\begin{aligned}&\left[jk_r^2\sqrt{\frac{\mu_w}{\varepsilon_w}} + \omega\mu_d k_r \frac{J_n'(u)}{J_n(u)}\right]\times\\&\left[jk_r^2\sqrt{\frac{\varepsilon_w}{\mu_w}} + \omega\varepsilon_d k_r \frac{J_n'(u)}{J_n(u)}\right] = \left[\frac{nk_z}{a_r}\right]^2\end{aligned}, \tag{3}$$

where $k_r = \sqrt{k_d^2 - k_z^2}$ is the dispersion relation of the circular waveguide, k_z the propagation constant, $J_n(u)$ denotes the Bessel function of the first kind, $J_n'(u)$ its derivative, n the order of the Bessel function, k_d, μ_d and ε_d are respectively the wavenumber, permeability and permittivity of the dielectric core material and a_r is the radius of the circular waveguide. The argument of the Bessel function u is given as:

$$u = \sqrt{a_r^2\left(k_d^2 - k_z^2\right)}, \tag{4}$$

Since TE and TM modes in a lossless circular waveguide are determined by the roots of $J_n'(u_{nm}) = 0$ and $J_n(u_{nm}) = 0$, respectively [10], (3) can be expanded into the form of a quadratic equation, with $\frac{J_n'(u)}{J_n(u)}$ or $\frac{J_n(u)}{J_n'(u)}$ as the variables to be solved for. Here, the n and m subscripts denote the n-th order and m-th zero of $J_n(u)$, respectively. By convention, the n subscript always represents the number of half-wave field variations in the ϕ-direction; whereas, the m subscript denotes the number of half-wave field variations in the r-direction [12]. Hence, different combinations of n and m variables produce different TE and TM modes in the waveguide. By expanding (3) and substituting (4) for the lossless case (i.e., $u = u_{nm}$) into k_z (i.e., $k_z = \sqrt{k_d^2 - \left(\frac{u_{nm}}{a_r}\right)^2}$) and k_r (i.e., $k_r = \frac{u_{nm}}{a_r}$), the quadratic equations for TE and TM modes can be expressed respectively as (5a) and (5b) below:

$$\begin{aligned}&k_d^2\left(\frac{u_{nm}}{a_r}\right)^2\left[\frac{J_n'(u)}{J_n(u)}\right]^2 + j\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s} + \varepsilon_d Z_s\right)\left[\frac{J_n'(u)}{J_n(u)}\right]\\&-\left[\left(\frac{u_{nm}}{a_r}\right)^4 + \left(\frac{n}{a_r}\right)^2\left(k_d^2 - \frac{u_{nm}^2}{a_r^2}\right)\right] = 0\end{aligned}, \tag{5a}$$

$$\begin{aligned}&\left[\left(\frac{u_{nm}}{a_r}\right)^4 + \left(\frac{n}{a_r}\right)^2\left(k_d^2 - \frac{u_{nm}^2}{a_r^2}\right)\right]\left[\frac{J_n(u)}{J_n'(u)}\right]^2 -\\&j\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s} + \varepsilon_d Z_s\right)\left[\frac{J_n(u)}{J_n'(u)}\right] - k_d^2\left(\frac{u_{nm}}{a_r}\right)^2 = 0\end{aligned}, \tag{5b}$$

where $Z_s = \sqrt{\frac{\mu_w}{\varepsilon_w}}$ is the surface impedance.

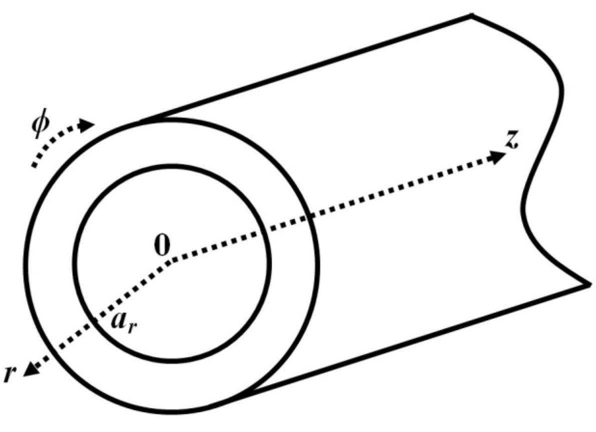

Fig. 1. A circular waveguide.

For a lossy but highly conducting waveguide, $J_n'(u)$ for TE modes and $J_n(u)$ for TM modes are close to zero. Hence, the second order of these functions can be ignored. The solutions to the quadratic equations in (5) can then be found as:

$$\frac{J_n'(u)}{J_n(u)} = \frac{\left(\frac{u_{nm}}{a_r}\right)^4 + \left(\frac{n}{a_r}\right)^2\left[k_d^2 - \left(\frac{u_{nm}}{a_r}\right)^2\right]}{j\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s} + \varepsilon_d Z_s\right)}, \qquad (6a)$$

and,

$$\frac{J_n(u)}{J_n'(u)} = \frac{jk_d^2\left(\frac{u_{nm}}{a_r}\right)^2}{\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s} + \varepsilon_d Z_s\right)}, \qquad (6b)$$

for TE and TM modes, respectively.

The argument of the Bessel function of a lossy waveguide u is assumed to be perturbed from that of the lossless case u_{nm}, i.e.,

$$u = u_{nm} + \delta_u, \qquad (7)$$

where δ_u is a perturbation term. Substituting (7) into $J_n'(u)$ for TE modes and $J_n(u)$ for TM modes, (6) can then be expressed as:

$$J_n'(u_{nm} + \delta_u) = \frac{\left\{\left(\frac{u_{nm}}{a_r}\right)^4 + \left(\frac{n}{a_r}\right)^2\left[k_d^2 - \left(\frac{u_{nm}}{a_r}\right)^2\right]\right\}J_n(u)}{j\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s} + \varepsilon_d Z_s\right)}, \qquad (8a)$$

and,

$$J_n(u_{nm} + \delta_u) = \frac{jk_d^2\left(\frac{u_{nm}}{a_r}\right)^2 J_n'(u)}{\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s} + \varepsilon_d Z_s\right)}. \qquad (8b)$$

Using the Finite Difference Method (FDM), the first and second derivatives of the Bessel function for a lossless waveguide with argument u_{nm} can be approximated as follows:

$$J_n(u_{nm} + \delta_u) = J_n'(u_{nm})\delta_u + J_n(u_{nm}), \qquad (9a)$$

and,

$$J_n'(u_{nm} + \delta_u) = J_n''(u_{nm})\delta_u + J_n'(u_{nm}). \qquad (9b)$$

Substituting (9b) into (8a) and (9a) into (8b) and solving for δ_u, we obtain:

$$\delta_{uTE} = \frac{\left\{\left(\frac{u_{nm}}{a_r}\right)^4 + \left(\frac{n}{a_r}\right)^2\left[k_d^2 - \left(\frac{u_{nm}}{a_r}\right)^2\right]\right\}J_n(u)}{j\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s} + \varepsilon_d Z_s\right)J_n''(u_{nm})} - \frac{J_n'(u_{nm})}{J_n''(u_{nm})}, \qquad (10a)$$

and,

$$\delta_{uTM} = \frac{jk_d^2\left(\frac{u_{nm}}{a_r}\right)^2 J_n'(u)}{\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s} + \varepsilon_d Z_s\right)J_n'(u_{nm})} - \frac{J_n(u_{nm})}{J_n'(u_{nm})}, \qquad (10b)$$

where δ_{uTE} and δ_{uTM} denote respectively the perturbation term δ_u for TE and TM modes. Since $J_n'(u_{nm}) = 0$ for TE modes, $J_n(u_{nm}) = 0$ for TM modes, and $J_n''(x) = \left[\left(\frac{n}{x}\right)^2 - 1\right]J_n(x)$ [16], by approximating $J_n(u) \approx J_n(u_{nm})$ and $J_n'(u) \approx J_n'(u_{nm})$, the perturbation terms in (10) can then be written as:

$$\delta_{uTE} = \frac{\left(\frac{u_{nm}}{a_r}\right)^4 + \left(\frac{n}{a_r}\right)^2\left[k_d^2 - \left(\frac{u_{nm}}{a_r}\right)^2\right]}{j\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s} + \varepsilon_d Z_s\right)\left[\left(\frac{n}{u_{nm}}\right)^2 - 1\right]}, \qquad (11a)$$

and,

$$\delta_{uTM} = \frac{jk_d^2\left(\frac{u_{nm}}{a_r}\right)^2}{\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s} + \varepsilon_d Z_s\right)}. \qquad (11b)$$

The propagation constant of the lossy waveguide k_z can be computed by substituting the perturbation terms in (11) and k_z in (4) into (7), i.e.,

$$k_z = \sqrt{k_d^2 - \left(\frac{u_{nm} + \delta_u}{a_r}\right)}. \qquad (12)$$

The propagation constant k_z is a complex variable which consists of the phase constant β_z and attenuation constant α_z, as shown in (13) below:

$$k_z = \beta_z - j\alpha_z. \qquad (13)$$

Hence, by extracting the imaginary part of the propagation constant k_z, the attenuation in the waveguide can be obtained. For the convenience of casual readers, we outline the final expressions of the attenuation constant for TE modes α_{zTE} and TM modes α_{zTM} here. It is worthwhile noting that u_{nm} for TE and TM modes can be found respectively in Tables 9.1 and 9.2 of [17]:

$$\alpha_{zTE} = \mathrm{Im}\left|\sqrt{k_d^2 - \frac{1}{a_r^2}\left\{u_{nm} - \frac{j\left[\left(\frac{u_{nm}}{a_r}\right)^4 + \left(\frac{n}{a_r}\right)^2\left(k_d^2 - \frac{u_{nm}^2}{a_r^2}\right)\right]}{\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s} + \varepsilon_d Z_s\right)\left[\left(\frac{n}{u_{nm}}\right)^2 - 1\right]}\right\}^2}\right|, \qquad (14a)$$

and,

$$\alpha_{zTM} = \text{Im}\left|\sqrt{k_d^2 - \frac{1}{a_r^2}\left\{u_{nm} + \frac{jk_d^2\left(\frac{u_{nm}}{a_r}\right)^2}{\omega\left(\frac{u_{nm}}{a_r}\right)^3\left(\frac{\mu_d}{Z_s}+\varepsilon_d Z_s\right)}\right\}^2}\right|. \quad (14b)$$

III. RESULTS AND DISCUSSION

To verify our formulations, we compute and analyze the loss in a hollow circular waveguide with copper wall. The radius of the waveguide is a_r = 8.1 mm. The attenuation constants of the dominant TE11 mode below and above cutoff f_c are depicted, respectively in Figs. 2 and 3. As can be seen from the figures, the attenuations predicted by our closed-form approach agree very closely with those by Stratton's rigorous equation. Indeed, it could be observed from Fig. 3 that the attenuations below millimeter wavelengths computed using both methods are almost indistinguishable. Figures 4 and 5 show the attenuation constants of the TM11 mode. Like the case of the dominant mode, the attenuations below and above cutoff f_c of the TM11 mode, obtained from Stratton's and our methods are in very good agreement. It is worthwhile noting that, we have applied the Powell-hybrid root-finding algorithm to solve for Stratton's transcendental equation. The process has been lengthy since the initial guesses were to be constantly refined in order to ensure convergence to the appropriate solution. Unlike, Stratton's approach, however, solutions can be easily found in a straight-forward manner using our closed-form method.

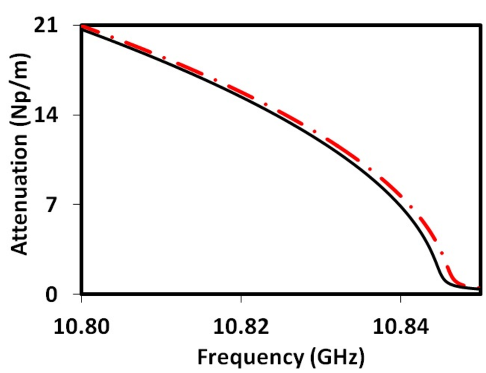

Fig. 2. Attenuation of TE11 mode below cutoff in an 8.1 mm radius, hollow waveguide with copper wall. The attenuations are computed using our method (solid line) and Stratton's (dashed-dotted line) method.

In order to show that our formulations work equally well in waveguides with different sizes, we compare the attenuation computed from both methods with the waveguide radius a_r varying from 5 mm to 55 mm. The attenuation of TE11 and TM11 modes with respect to the size of the waveguide are depicted, respectively, in Figs. 6 and 7. Since the operating frequency f = 100 GHz is above the cutoff frequencies f_c, all waves are in propagating modes. As can be observed from both figures, the attenuation constants obtained from both Stratton's and our method agree very well and are almost indistinguishable. Indeed, the maximum discrepancy found from both results is less than 1.25×10^{-4} Np/m.

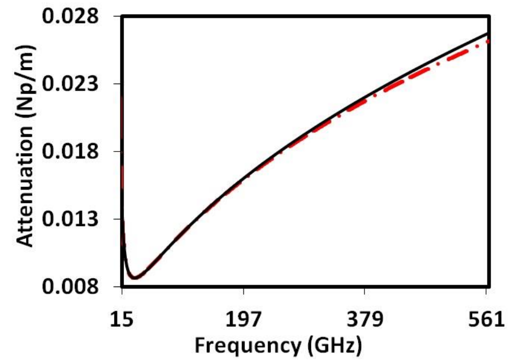

Fig. 3. Attenuation of TE11 mode above cutoff in an 8.1 mm radius, hollow waveguide with copper wall. The attenuations are computed using our method (solid line) and Stratton's (dashed-dotted line) method.

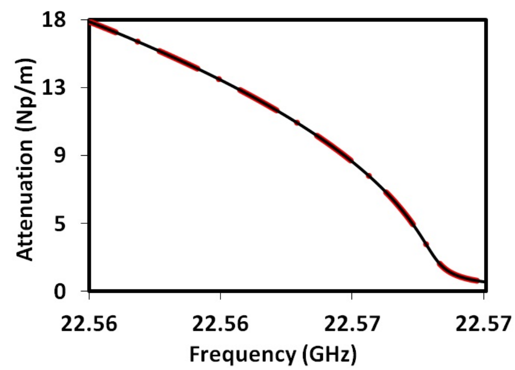

Fig. 4. Attenuation of TM11 mode below cutoff in an 8.1 mm radius, hollow waveguide with copper wall. The attenuations are computed using our method (solid line) and Stratton's (dashed-dotted line) method.

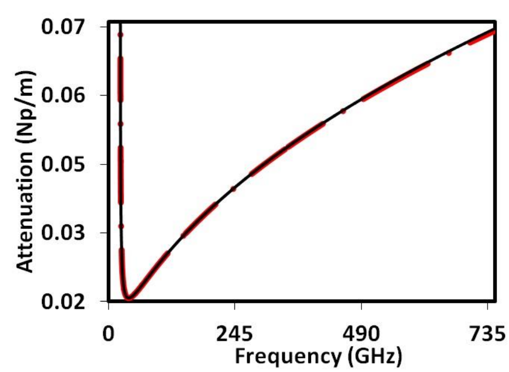

Fig. 5. Attenuation of TM11 mode above cutoff in an 8.1 mm radius, hollow waveguide with copper wall. The attenuations are computed using our method (solid line) and Stratton's (dashed-dotted line) method.

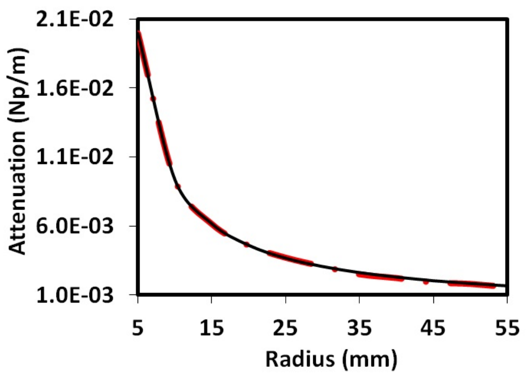

Fig. 6. Attenuation of TE11 mode when a 100 GHz wave propagates in a hollow waveguide with copper wall. The attenuations are computed using our method (solid line) and Stratton's (dashed-dotted line) method.

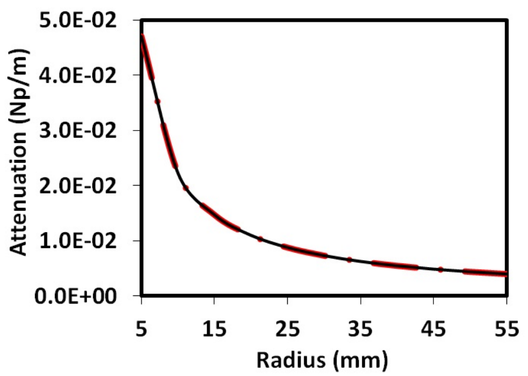

Fig. 7. Attenuation of TM11 mode when a 100 GHz wave propagates in a hollow waveguide with copper wall. The attenuations are computed using our method (solid line) and Stratton's (dashed-dotted line) method.

IV. CONCLUSION

We have developed a set of closed-form formulations for calculating the attenuation of TE and TM modes in a circular waveguide with imperfectly conducting wall. Our approach is based on matching the tangential electric and magnetic fields at the boundary with the electrical properties of the wall material. By neglecting the second-order variables, the roots of the characteristic equation can then be easily expressed in terms of the wavenumbers. Since the behavior of the lossy waveguide is assumed to be perturbed from its lossless case, a perturbation term is introduced into the Bessel function and its derivative. The attenuation constant is found by determining the perturbation terms using the Finite Difference Method and substituting them into the dispersion relation. Our closed-form equations show good agreement with those obtained using Stratton's rigorous approach. Unlike Stratton's transcendental approach, however, our approach leads to simpler and more straight-forward analysis.

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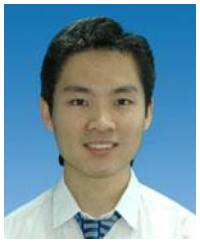

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