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Efficient time-domain numerical analysis of waveguides with tailored wideband pulses

Jonathan Viquerat

Abstract—A simple procedure for the generation of accurate polychromatic sources in waveguides taking into account mode dispersion is presented. It allows for an efficient use of time-domain solvers in the analysis of guided modes, and can be used wether the mode dispersion and field distribution are known analytically or numerically. This method is implemented in the DIGENES Discontinuous Galerkin Time-Domain (DGTD) solver (http://diogenes.inria.fr), and is validated on a waveguide mode converter.

Index Terms—Discontinuous Galerkin Time-Domain (DGTD) methods, optical waveguides, mode dispersion, wide-band source.

I. INTRODUCTION

The numerical analysis of optical devices with time-domain methods usually allows the use of polychromatic sources (dipole, plane wave, etc), thus characterizing the structure response in a single simulation. For waveguides, however, the existence of modes brings the task of building wide-band sources to a higher level of complexity [2]. More precisely, two main obstacles must be overcome: (i) for a given mode, the effective index \( n_{\text{eff}} \) depends on \( \omega \), and (ii) the frequency-domain 2D mode field distribution \( \hat{E}_{2D} \) also depends on \( \omega \).

The bootstrapping method, proposed in [2], is an effective yet computationally-expensive way to overcome these problems. An erroneous source is injected and propagated in an extended version of the considered waveguide. Along the propagation, the exact mode establishes in the waveguide, while the remaining energy is scattered towards the surrounding PML regions. At the edge of the waveguide, a probe plane is used to record passing fields, providing an accurate source for future computations.

In [2], the authors propose another workaround, in which a wide-band source made from the sum of several polychromatic pulses. Each of them is centered around a given frequency and holds an appropriate 2D field distribution associated to this frequency. In this method, the sampling of the 2D field distribution and that of the effective index are concurrently made at the selected frequencies, which are most often limited to a handful.

In this paper, we propose a straightforward polychromatic source formulation for waveguide analysis based on an inverse Fourier transform of the frequency-domain solution, incorporating both the variations of the 2D field distribution and that of the mode index. The knowledge of both these quantities can be either analytic or numeric (i.e. computed in a third-party solver or read from a file). Additionally, we describe the way to impose of such a source through a Total-Field/Scattered-Field (TF/SF) interface in the DGTD framework, as well as how to compute figures of merit such as transmission, overlap or mode conversion efficiency. The accuracy of this approach is first tested on a textbook case, before being applied to a waveguide mode converter.

II. DESCRIPTION OF THE SOURCE

A. Source construction

Assuming that \( \hat{E}_{2D} \) and \( n_{\text{eff}} \) are known, we remind that the corresponding monochromatic time-domain source propagating in direction \( z_+ \) at angular frequency \( \omega_0 \) is obtained as follows:

\[
E(t) = \mathbb{R} \left[ \hat{E}_{2D}(\omega_0) e^{-j\beta(\omega_0)z} e^{j\omega_0 t} \right],
\]

where \( \beta(\omega) \) is the mode propagation constant:

\[
\beta(\omega) = \frac{n_{\text{eff}}(\omega)}{c_0}.
\]

For plane wave analysis, polychromatic signals are most often obtained by modulating a sine or a cosine function at angular frequency \( \omega_0 \) with the derivative of a gaussian bump centered on \( \omega_0 \) and with well-suited variance parameter \( \sigma \). Then, taking the Inverse Fourier Transform (IFT) of the frequency-domain signal yields the desired time-domain pulse. One can follow the same procedure here, taking into account the effective index of the mode:

\[
E(t) = \mathbb{R} \left[ \int_{-\infty}^{+\infty} \hat{E}_{2D}(\omega) \sigma e^{-\frac{(\omega-\omega_0)^2}{2\sigma^2}} e^{-j\beta(\omega)z-\omega t} \right].
\]

Most often, the latter IFT is not analytic and must be numerically computed using an Inverse Discrete Fourier Transform (IDFT):

\[
\phi(t) \simeq \sum_{k=1}^{n_{\omega}} \hat{\phi}(\omega_k) e^{j\omega_k t} \Delta\omega,
\]

where \( \Delta\omega \) is the chosen frequency step:

\[
\omega_k = \omega_{\text{min}} + (k - 1) \Delta\omega,
\]

and \( n_{\omega} \) is the number of frequency samples:

\[
n_{\omega} = \frac{\omega_{\text{max}} - \omega_{\text{min}}}{\Delta\omega} + 1.
\]
B. Sampling and the reciprocal Shannon theorem

The choice of \(n_\omega\) (or equivalently of \(\Delta \omega\)) will have a major impact on the accuracy and efficiency of the procedure. Indeed, if it is chosen too high, the frequency-domain expression under the integral in (3) will be over-sampled, and its evaluation will be uselessly expensive. On the opposite, if it is chosen too low, the obtained time-domain signal will suffer from aliasing, resulting in overlapping incident wave trains. In order to select an adequate value for \(n_\omega\), we make use of the reciprocal Shannon theorem [3], stating that the sampling rate \(\Delta \omega\) used to compute (3) should meet the following criterion:

\[
\Delta \omega \leq \frac{2\pi}{T},
\]

where \(T\) is the total duration of the transformed time-domain signal. Although in most cases, \(T\) might not be known, in our case it can be well approximated thanks to the Gaussian envelope of the frequency-domain expression (2). To do so, we introduce a parameter \(\epsilon\), such that fields with amplitudes smaller than \(\epsilon\) are considered to be zero. Then, we seek the width of the gaussian envelope for which the signal level is equal to \(\epsilon\). A few lines of calculus yield:

\[
\Delta \omega = \frac{2\pi}{\sigma \sqrt{8 \ln(\frac{1}{\epsilon})}}. \tag{4}
\]

In the following, we choose \(\epsilon = 1 \times 10^{-3}\), and therefore:

\[
\Delta \omega \approx 0.845 \frac{\pi}{\sigma}. \tag{5}
\]

C. Total-Field/Scattered-Field injection plane

The discretization of the electric and magnetic fields in the DGTD framework is naturally discontinuous, meaning that the unknowns are duplicated on each face joining two cells of the mesh [4]. The consistency of the approximation is re-established through the imposition of a numerical flux \((E_\star, H_\star)\) at the interface, which handles the information exchange between cells (the reader is referred to [4] and [5] for a thorough development of the DGTD method). As a consequence, it is possible to define separate areas in the mesh, in which the numerical unknowns either hold the total fields \((E_{\text{tot}}, H_{\text{tot}})\) or the scattered fields \((E_{\text{sca}}, H_{\text{sca}})\). The interface between these regions is the so-called TF/SF interface. At this interface, the fluxes are modified in order to inject the incident field in the TF region, and to subtract it in the SF region (see figure 1). With this method:

- It is possible to access both the TF and the SF fields in their regions of interest simultaneously;
- Incident fields can be imposed virtually anywhere in the mesh;
- Accessing both the TF and SF fields at the same place (i.e. on the TF/SF interface) allows for the computation of multiple figures of merit with a single computation;
- In the present case, if a mode is properly imposed and propagated through the TF region of an unperturbed waveguide, the fields in the SF regions should be identically zero, making it a reliable method to test the source described in the last section.

D. Figures of merit

A brief word must be said about the figures of merit that will be used in this paper, namely the transmission \(T\), the overlap \(O_M\) and the mode conversion efficiency \(\eta_{M_\star}\). These quantities are based on the computation of the averaged Poynting vector \(\pi\):

\[
\pi(x, \omega) = \frac{1}{2} \mathbb{R} \left( E(x, \omega) \times H^*(x, \omega) \right). \tag{6}
\]

In expression (6), the frequency-domain fields are evaluated on control surfaces (here, the TF/SF surfaces) with an on-the-fly Fourier transform during the time-domain computation. Thanks to the discontinuous nature of the DG field discretization, these quantities can be evaluated either on the total-field or the scattered-field side of the TF/SF faces, allowing to compute the transmission as:

\[
T(\omega) = \frac{\int_{S_\text{tot}} \pi_{\text{tot}} \cdot n}{\int_{S_i} \pi_{\text{inc}} \cdot n}, \tag{7}
\]

where \(\pi_{\text{tot}}\) is computed on the total-field side of the exiting TF/SF surface \(S_\text{tot}\), \(\pi_{\text{inc}}\) is computed using the incident field only on the total-field side of the inputting TF/SF surface \(S_i\), and \(n\) is the unit normal. As such, the transmission computes the normalized transmitted power from the input to the output plane, ranging from 0 to 1. The overlap, also ranging from 0 to 1, measures the similarity of the transmitted fields \(\hat{E}\) and \(\hat{H}^*\) to that of a given mode \(\hat{E}_M\) and \(\hat{H}_M^*\):

\[
O_M(\omega) = \frac{\mathbb{R} \left( \Pi_{S_\text{tot}} \left( \hat{E}_M, \hat{H}_M^* \right) \right)}{\mathbb{R} \left( \Pi_{S_i} \left( \hat{E}, \hat{H}^* \right) \right)}, \tag{8}
\]

with:

\[
\Pi_S \left( \hat{E}, \hat{H}^* \right) = \int_S \hat{E} \times \hat{H}^* \cdot n.
\]
Finally, the mode conversion efficiency, which computes the normalized transmitted power that has been coupled to a given mode $M$, is easily defined as:

$$
\eta_M(\omega) = T(\omega)O_M(\omega).
$$

### III. Numerical Experiments

#### A. Straight rectangular waveguide

We consider a rectangular waveguide of lateral dimensions $a = 10.16$ mm and $b = 22.86$ mm with PEC (Perfect Electric Conductor) walls. The goal of this section is to propagate a TE$_{01}$ mode in it, and to assess the efficiency and accuracy of the procedure described above. The operating frequency range is [8, 10] GHz. We exploit a high-order DGTD method [5] on an adequately-refined tetrahedral mesh. The incident field is imposed on a Total-Field/Scattered-Field interface, separating a total field region (in the center of the numerical domain) from a scattered field region (before the input and after the output). The mesh, shown in figure 2 is terminated on both ends with a Perfectly-Matched Layer (PML) region to absorb outgoing field.

As a first experiment, we consider the propagation of an uncorrected polychromatic pulse, where frequency dependence of $E_{2D}$ and $\beta$ is not taken into account. Both the quantities are approximated by their evaluation at the central frequency $f_c = 9$ GHz. To evaluate the quality of the numerical treatment, we simply consider the evaluation of the $E_y$ component of the electric field at a chosen probe point in the SF region. Thanks to the use of TF/SF interfaces, for an unperturbed waveguide the field should be identically equal to 0 in the SF regions. In figure 3, we show a 3D snapshot of the field during computation: the presence of field in the right SF region indicates a mismatch between the numerical solution and the expected exact solution. Such a result was to be expected, since most frequencies in the pulse were numerically transported with an erroneous propagation constant.

We now consider the $\omega$ dependence of the quantities involved. The effective index of the mode is plotted in figure 4 along with the resulting spectral power of the incident pulse.

In figure 5, we conduct the same experiment as before, but using expression (2) for the omega-dependent quantities. Using expression (4) gives $n_\omega = 24$ for this computation. A 3D snapshot of the computed electric field is shown in figure 5. For a better comparison, we plot in figure 6 a comparative time-domain plot of $E_y$ in the right SF region, with and without the IDFT correction. As can be seen, taking into account the $\omega$ dependence of the incident field and index greatly improves the mode mismatch. The maximal $L^2$ errors on the electric field observed in the right SF region, with and without mode correction, are shown in table 1 along with the corresponding CPU times. The computed error is 60 times lower with the mode correction, while the CPU times is increased by 34%, thus confirming the interest of this method.

#### B. TE$_{01}$-TE$_{02}$ mode converter

We now consider a TE mode converter, as presented in [6]. The converter is composed of a bent rectangular PEC waveguide of dimensions $a = 22.15$ mm and $b = 47.55$ mm. The bending is tailored for a theoretical power conversion of
100% at $f = 8.5$ GHz. The goal of this section is to assess the converter efficiency in the $[7.5, 8.5]$ GHz frequency range in a single run, using the technique presented above. The effective indices of the two modes in the selected frequency range are shown in figure 7. As it shall be useful later, we also plot in the latter figure the effective index of the TE$_{10}$, which is the only other mode which effective index is real in the frequency range of interest. Following the results of the previous section, $n_{\omega}$ is set to 24 for this experiment. The mode conversion efficiency from TE$_{01}$ to TE$_{02}$ is computed following the details of section II-D, and is shown in figure 8 for both pulses. As expected, a peak efficiency of 1 is found for $f = 8.5$ GHz in both cases, while it decays for lower frequencies. However, the uncorrected approach overestimates the conversion rate from roughly 2.5% at the minimal frequency. The obtained results are in line with what was found in 6. Similarly to the previous section, the CPU time is increased by roughly 30.5% when using the mode-corrected source.

Snapshots illustrating the mode conversion are shown in figure 9. The TE$_{01}$ mode pulse is incoming from the left, before being converted in a TE$_{02}$ that leaves the computational domain from the right (full mesh is not shown here). To complete the study, we compute the conversion efficiencies $\eta_{\text{TE}_{01}}$ and $\eta_{\text{TE}_{10}}$, i.e. the normalized amount of energy coupled from the TE$_{01}$ mode at the input to the TE$_{01}$ and TE$_{10}$ modes at the output. Results, presented in figure 10 show that for lower frequencies, a non-negligible part of the pulse is not converted to the TE$_{02}$ mode. On the other hand, the conversion rate to TE$_{10}$ mode is identically 0. It should be noted that a very low level of backscattering was observed in the computations, hence it is not plotted here.

IV. CONCLUSION

In this paper, we provided a simple expression for the computation of accurate polychromatic sources in waveguides. Details about its practical use were also discussed, such as the minimal frequency-domain sampling or the use of TF/SF faces to impose it in the computational domain. The computation of relevant figures of merit was also sketched. The described procedure was then validated on a microwave-regime rectangular waveguide, on which significantly low error levels were observed. Finally, the method was applied to a real-life mode converter, allowing to assess the device response on a large bandwidth in a single computation.
Fig. 9. Successive $E_x$ field snapshots during the conversion of a polychromatic TE$_{01}$ mode to a TE$_{02}$ mode. Full mesh is not shown here.

Fig. 10. Computed mode conversion efficiency from TE$_{01}$ at the input to TE$_{01}$, TE$_{02}$ and TE$_{10}$ modes at the output.

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