Slot waveguide-based splitters for broadband terahertz radiation

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Abstract: We demonstrate a slot waveguide-based splitter for broadband terahertz (THz) radiation using a T-shaped waveguide structure. The structure consists of a fixed-width input waveguide and variable-width output waveguides. We experimentally measure and numerically simulate the THz transmission and reflection properties as a function of the output waveguide width and show that a transmission line model can effectively describe the observations. Based on the high degree of agreement between the experimental results, numerical simulations and the model, we infer the optimal waveguide parameters. The device structure offers new possibilities in designing compact THz devices.

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References and links

1. Introduction

The ability to create sharp right angle bends in electronic interconnects has allowed for the creation of compact circuit layouts. There is great interest in creating analogous capabilities for optical circuits, which could be designed to utilize different regions of the electromagnetic spectrum. However, for such geometries, greater losses are typically observed as one moves to higher and higher frequencies. As an example, in typical optical waveguides, where the refractive index contrast between the core and cladding tends to be somewhat small, radiative losses typically increase as the bend radii decrease [1]. This problem can be minimized, to some extent, by increasing the refractive index contrast between the core and cladding, as has been nicely demonstrated in the area of silicon photonics [2]. An alternative approach that has been developed utilizes photonic crystal waveguides [3]. While the approach allows for sharp right angle bends, in principle, a variety of technical issues have limited the practical realization of low loss bends and splitters [4].

In the far-infrared, the use of metal-dielectric-metal waveguide structures, such as parallel plate [5] and slot [6,7] waveguides, have been shown to allow for low loss, low dispersion propagation of broadband terahertz (THz) radiation. The latter structure is particularly appealing, since it has the potential to allow for the fabrication of a variety of guided-wave devices. Identical to conventional parallel plate waveguides, the lowest order mode of a slot waveguide does not exhibit a cutoff frequency and has recently been shown to guide broadband THz radiation in straight and s-bend geometries [6,7]. It is worth noting that slot waveguides have generated significant recent interest for guiding, splitting, and filtering at optical frequencies. Although there have been several experimental studies based on this waveguide geometry [8–10], there have significantly more publications based on theory and simulation [11–15].

In this submission, we experimentally demonstrate a high transmittance slot waveguide-based splitter that incorporates sharp right angle bends (i.e. a T-waveguide splitter) and supports the propagation of broadband THz radiation. By fixing the width of the input waveguide and varying the width of the output waveguides, we find that there exists an optimal ratio for the waveguide widths in order to maximize the THz throughput. We explain this using conventional transmission line theory.

2. Experimental details

We fabricated the slot waveguide-based splitter using three aluminum sheets, each with a thickness, w, of 1 mm. The edges were polished using diamond machining, in order to minimize propagation losses via scattering. The input slot waveguide, shown schematically in Fig. 1(a), was 5 cm long with an input gap, d1, set to 100 µm. The inclusion of the third metal sheet, shown in Fig. 1(b), formed the two 2.5 cm long output waveguides. This third plate was placed on a translation stage, so that the width of the output waveguides, d2, could be varied. We used conventional THz time-domain spectroscopy to characterize the device. In the experimental setup, a photoconductive device launched broadband THz pulses into the waveguide that were linearly polarized perpendicular to the long axis of the waveguide gap and parallel to the metal surface. Hyper-hemispherical silicon lenses were used to couple the broadband THz pulses into and out of the waveguides. A photoconductive detector was oriented to measure the same polarization of the radiated THz pulses. Both figures show locations of the THz electric field that will be discussed in the text below.

Numerical simulations for the propagation properties of the waveguide structures examined here were performed using 3D finite-difference time-domain (FDTD) simulations. The metal was modeled as a perfect electrical conductor, which is a reasonable approximation for real metals in the THz regime, surrounded by air. We used a spatial grid size of 10 µm, which was sufficient to ensure convergence of the numerical calculations, and perfectly matched layer absorbing boundary conditions for all boundaries.
Fig. 1. Schematic diagram of two different slot waveguide geometries examined. (a) Two metal plates separated by a gap spacing of $d_1 = 100 \, \mu m$ forms the input waveguide. (b) A third metal sheet was included to form the two 2.5 cm long output waveguides. This third plate was placed on a translation stage, so that the width of the output waveguides, $d_2$, could be varied between 100 and 300 $\mu m$. The thickness of the metal plates, $w$, was 1 mm. The values of $E_i$, where $i = 0 \ldots 6$, and the associated dots correspond to points where the THz electric field could be measured. In all cases, the THz electric field was measured in the far-field. The double-sided red arrow shows the polarization of the input electric field.

3. Experimental results, simulations and discussion

We initially measured the time-domain and frequency-domain properties of the THz radiation, both in the absence of the slot waveguides and at the output of the two different slot waveguide configurations, corresponding to $E_2$ and $E_4$, as shown in Fig. 1. Figure 2 shows the measured time-domain THz waveforms in all three experimental configurations with $d_1 = 100 \, \mu m$ and $d_2 = 200 \, \mu m$, demonstrating that no significant dispersion or pulse reshaping processes occurred. Although all three time-domain waveforms, shown in Fig. 2, are plotted on the same graph, it is important to note that no inference of coupling efficiency can be made, because different experimental geometries were used in measuring the different waveforms. We note that the measured time-domain waveform at the position corresponding to $E_6$ was nearly identical in all respects to $E_4$, demonstrating true splitting capability.

Figure 3 shows the corresponding normalized amplitude spectra. The spectra associated with the outputs of the two slot waveguide configurations appears to exhibit somewhat greater high frequency content than the incident THz beam. This arises from the fact that the frequency content of the incident THz beam is spatially dependent, with the high frequency content more concentrated near the beam center. The spectra associated with $E_2$ and $E_4$ look very similar, although the latter spectrum is smaller in amplitude than the former. We attribute the difference largely to frequency-independent loss mechanisms associated with the longer propagation length. We attribute the very similar looking oscillations in the two spectra to the identical input coupling conditions in the two waveguide geometries, since the output coupling conditions were different. It is clear that there is no cutoff frequency associated with the outputs of the two different waveguide configurations, as would be expected for the lowest order $TM_0$ (TEM) mode [16]. The next higher order $TM_1$ mode is characterized by a cutoff frequency, $\nu_c = c/2d$, where $c$ is the speed of light in vacuum and $d$ is the gap spacing in the slot waveguide. For the slot waveguides considered here, the smallest cutoff frequency occurs...
when \( d_2 = 300 \, \mu\text{m} \) (i.e. \( \nu_c = 0.5 \text{ THz} \)). Thus, the lowest order TEM mode is the dominant mode.

![Graph](image1)

**Fig. 2.** Measured time-domain waveforms in the absence of the waveguide structure (red trace), at the output of the structure shown in Fig. 1(a) (black trace) and at the output of the structure shown in Fig. 1(b) (green trace).

![Graph](image2)

**Fig. 3.** The normalized amplitude spectra corresponding to the waveforms shown in Fig. 2.

![Graph](image3)

**Fig. 4.** Measured ratio of \( E_4/E_2 \) (filled circles), calculated as the average value between 0.1 and 0.8 THz from Fig. 2, as a function of \( d_2/d_1 \). (Inset) Measured ratio of \( E_4/E_2 \) for \( d_1 = 100 \, \mu\text{m} \) and \( d_2 = 200 \, \mu\text{m} \) as a function of frequency.

In Fig. 4, we show the measured ratio of \( E_4/E_2 \) as a function of \( d_2/d_1 \). In order to obtain these values, we computed the average value of \( E_4/E_2 \) between 0.1 and 0.8 THz, for each value of \( d_2/d_1 \). Since THz time-domain spectroscopy measures the electric field, we note that
$E_1/E_2$ can be greater than 0.5. The inset shows the ratio as a function of frequency for $d_1 = 100 \mu m$ and $d_2 = 200 \mu m$. Aside from the relatively small oscillations that appear as a function of frequency, the ratio is relatively frequency independent. This is equally true for data obtained with other values of $d_2$. The fact that this ratio is essentially frequency independent over the frequency range considered here suggests a relatively simple model may be used to interpret the data.

In order to understand this data and predict the properties of other analogous waveguide geometries, we use numerical finite-difference time-domain (FDTD) simulations to demonstrate that a simple transmission line model accurately describes the data. To demonstrate this, we model the slot T-waveguide as the junction of three transmission lines, as shown in Fig. 5(a), where the input waveguide has a characteristic impedance given by $Z_1 = Z_0 d_1/w$, each of the output waveguides have a characteristic impedance given by $Z_2 = Z_0 d_2/w$, and $Z_0$ is the characteristic impedance of free space [17]. The effective load impedance, $Z_L$, seen by the input waveguide consists of the parallel combination of the two output waveguides (i.e. $Z_L = Z_2/2$). Using these definitions, the amplitude reflection coefficient is given by

$$r = \frac{Z_L - Z_1}{Z_L + Z_1} = \frac{Z_2 - 2Z_1}{Z_2 + 2Z_1}.$$  \hfill (1)

In order to minimize the amplitude reflection, we want $Z_1 = Z_L = Z_2/2$ and, therefore, $d_1 = d_2/2$. We note that this differs from the analysis given in [12].

![Diagram](image)

**Fig. 5.** (a) The equivalent transmission line model for the waveguide geometry shown in Fig. 1(b). (b) Numerically calculated values of the amplitude reflection coefficient, $r$, as a function of $d_2/d_1$. The filled circles correspond to results from FDTD simulations with $w = 1$ mm and $d_1 = 100 \mu m$, while the solid line corresponds to the fit using Eq. (1).

In Fig. 5(b), we show the results for FDTD simulations of the amplitude reflection coefficient along with the predictions based on the transmission line model Eq. (1). In general, the excellent agreement suggests that the simple model is sufficient for the geometry and dimensions described here. It is worth noting that for $d_2/d_1 = 2$, the amplitude reflection value is not exactly equal to zero. We do not believe that this is a numerical artifact. Rather, we believe it arises from the fact that the transmission line model is only a (good) approximation. Finally, a negative value of $r$ simply implies that the broadband pulse encounters a $\pi$ phase shift upon reflection.

In an analogous fashion, the internal amplitude (electric field) transmission coefficient, $t$, is given by

$$t = \frac{E_3}{E_i} = \frac{\sqrt{Z_2}}{(2Z_1 + Z_2)}.$$  \hfill (2)
As with the amplitude reflection coefficient, numerical simulations agree well with predictions based on Eq. (2) (not shown). Based on conservation of energy considerations, it is clear that $|r|^2 + 2 \left( \frac{Z_1}{Z_L} \right) |t|^2 = 1$. Given that the numerical FDTD simulations and the transmission line model agree well for the experimental conditions discussed above, we now apply it the measured transmission properties, $E_4/E_2$. It is straightforward to show that

$$\frac{E_4}{E_2} = \xi \frac{\sqrt{2Z_1(Z_o + Z_s)}}{(2Z_1 + Z_s)(Z_o + Z_s)},$$

where $\xi$ is a constant that accounts for a number of non-idealities that occur in experiments, including propagation losses due to the finite conductivity of metals at THz frequencies as well as scattering from surface imperfections, differences in detection efficiencies for $E_4$ and $E_2$ and other possible loss mechanisms. In Fig. 6, we show $E_4/E_2$ as a function of $d_2/d_1$ obtained from numerical simulations, experimental results (taken from Fig. 4) and Eq. (3). In the figure, the best fit of Eq. (3) to the experimental data occurs with $\xi = 0.82$ and no other free parameters. We also scaled the FDTD simulation results by the same factor of 0.82. The excellent agreement, within a scaling factor, between the experimental and numerical results suggests that the transmission line model may be used to compute other parameters relevant to the current geometry and may be extended to other slot waveguide-based device geometries.

![Graph showing the ratio of $E_4/E_2$ as a function of $d_2/d_1$.](image)

Fig. 6. Ratio of $E_4/E_2$ as a function of $d_2/d_1$. The filled (black) circles correspond to experimental data, the filled (red) triangles correspond to results obtained from FDTD simulations scaled by a factor of 0.82, and the solid line corresponds to the best fit to Eq. (2) with $\xi = 0.82$ and no other free parameters.

4. Conclusion

In conclusion, we have demonstrated a high transmittance slot waveguide-based splitter that incorporates right angle bends. Within a multiplicative factor close to 1, we demonstrate that a transmission line model accurately models the data, which is further validated by numerical simulations. The fact that this factor, $\xi$, is not equal to 1 may arise, in part, from the right angle geometry of the structure. The use of mitered bends has been shown previously to reduce the potential for parasitic discontinuity capacitances at microwave frequencies [18]. Finally, we couple broadband THz radiation into the structure from an external source. We expect that by embedding a nonlinear optical medium into the input slot waveguide and directly generating broadband THz radiation within the device, both the signal-to-noise and bandwidth of the guide-wave mode may be dramatically improved [19,20]. Such capabilities offer new opportunities in developing compact broadband THz devices and circuits.

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