Subpicosecond Pulse Propagation on Coplanar Waveguides: Experiment and Simulation

S. Gupta, J. F. Whitaker, and G. A. Mourou

Abstract—**Experimental results are presented for subpicosecond pulse propagation on normal-metal, coplanar-transmission-line structures. The pulse distortion that occurs is modelled using a semiempirical curve fit to the fullwave analysis for the modal dispersion and quasi-static approximations for the conductor and radiation loss. Without using any adjustable parameters, very good agreement is obtained for the delay, rise time and amplitude of the pulse for various propagation distances. For terahertz-bandwidth pulses on lines similar to the one studied, the modal dispersion and radiation losses are the dominant pulse-shaping mechanisms.**

I. Introduction

PICOSECOND pulses propagating on normal-metal, coplanar-circuit interconnects are degraded by the dispersion and attenuation characteristics of these guided-wave structures [1]–[3]. The dispersion of a coplanar transmission line arises primarily due to the dielectric inhomogeneity at the surface [4] and the complex surface impedance of the electrodes. The attenuation is due to the conductor losses, substrate conductivity losses and radiation or surface wave losses [5]. Modeling of picosecond pulse propagation on coplanar lines needs to incorporate the previous effects, while trying to retain simplicity in computation through the use of approximations verified via simulations and experiments.

In this letter, we present experimental results on subpicosecond pulse propagation on normal-metal coplanar-waveguide (CPW) transmission lines defined on a semi-insulating GaAs substrate. We have considerably extended the bandwidth as compared to previously published results [1]–[3], and data for various propagated distances are presented. Simulation is done using a semiempirical curve fit to the fullwave analysis for modal dispersion [4], and quasi-static approximation for the conductor [1] and radiation loss [6]. Excellent agreement is obtained for the important pulse parameters—propagation delay, rise time and pulse amplitude—without the use of any adjustable parameters. For the typical dimensions of the coplanar lines and the frequency bandwidth of the pulse considered here, the distortion of the pulse shape is dominated by the modal dispersion (caused by the dielectric mismatch at the surface), and the loss is dominated by the radiation loss [7].

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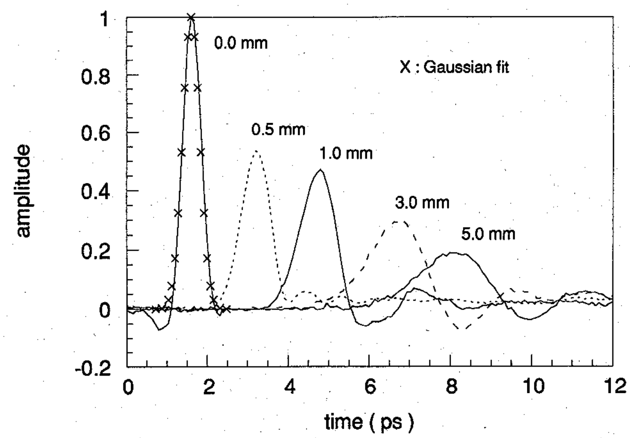

Fig. 1. Subpicosecond pulse propagation data on coplanar-waveguide transmission line on GaAs substrate. (Propagation delay is shown in Fig. 3.)

II. Experiment

Coplanar-waveguide (CPW) and coplanar-strip (CPS) transmission lines were photolithographically defined on a 2μm-thick layer of low-temperature-grown (200°C by molecular beam epitaxy) GaAs on a semi-insulating GaAs substrate. A 10μm-wide gap in one of the conductors is used as a photoconductive switch in order to launch an electrical pulse by exciting it with a short optical pulse. A balanced, colliding-pulse mode-locked (CPM) laser is used to produce 100-fs optical pulses at 100-MHz repetition rate at $\lambda = 620$ nm for this purpose. The low-temperature-grown GaAs layer has a fast carrier recombination time and allows the generation of sub-picosecond electrical pulses [8]. An example is the waveform labelled 0.0 mm in Fig. 1; it has $\sim$ 0.52-ps full-width-at-half-maxima (FWHM) duration, and it serves as our input signal. The 10-dB cutoff point in the spectrum of this input transient is about 1.6 THz. The propagated electrical pulse is measured by external electrooptic sampling—using a movable $LiTaO_3$ microprobe [9]—at propagation distances $z = 0.0$, 0.5, 1.0, 3.0, and 5.0 mm. This electrooptic measurement technique has a temporal resolution of less than 0.3 ps [9]. The excitation or switching beam is fixed, and the signal at each propagated distance is sensed by the remote field-sensitive probe and then calibrated by applying a known voltage to the coplanar lines. The transmission lines are long enough so that reflections from the terminations do not appear in the time window of interest.

Fig. 2 shows the experimental data for subpicosecond pulse propagation on CPW lines having a center-conductor with $s = 15$ μm and slot width $w = 10$ μm (nominal wave impedance 50 Ω). The substrate is a 500μm-thick GaAs

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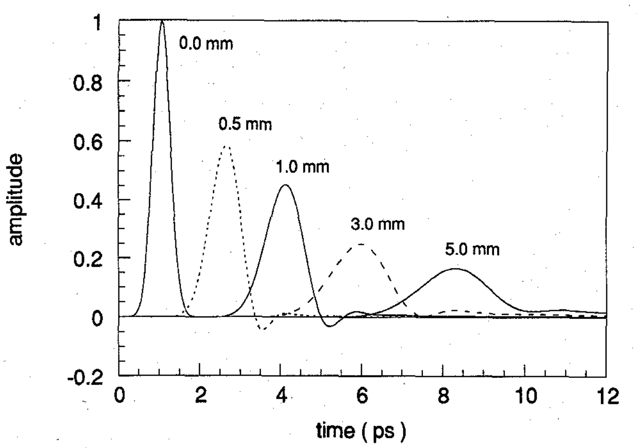

Fig. 2. Simulation of subpicosecond pulse propagation including modal dispersion and conductor and radiation losses. (Propagation delay is shown in Fig. 3.)

wafer ($\varepsilon_r = 13.2$). The metallization used was 0.05/0.40 μm Ti/Au, having a measured dc resistivity of $\sim 2.6 \times 10^{-6}$ Ω-cm. The propagated pulses at various distances have been shifted in time to fit properly in the plot. (Propagation delays are shown in Fig. 3.) Note that there are several important features of the pulse evolution besides the propagation delay. The amplitude decreases drastically, the rise time lengthens considerably, the falling edge steepens somewhat and oscillations develop at the trailing side of the pulse. These are discussed in the following section.

III. Simulation and Discussion

We have modeled the modal dispersion of short electrical pulses by a semiempirical curve fit to the fullwave analysis for CPW lines obtained in [4]. Due to the very large aspect ratio of the substrate thickness to CPW transverse dimensions, we treat the substrate as semi-infinite. The frequency-dependent effective permittivity can be written as [4], [7]

$$\sqrt{\varepsilon_{\text{eff}}(f)} = \sqrt{\varepsilon_{\text{eff}}(0)} + \left[\frac{\sqrt{\varepsilon_{\text{eff}}(\infty)} - \sqrt{\varepsilon_{\text{eff}}(0)}}{1 + af^{-b}}\right] \quad (1)$$

where $\varepsilon_{\text{eff}}(0)$ is the quasi-static value given in [10], and $\varepsilon_{\text{eff}}(\infty) \approx \varepsilon_r$. From [4] we take $b = 1.8$ for a coplanar waveguide and calculate $a = 6.366$ (f in THz).

The frequency-dependent complex surface impedance of the coplanar waveguide with electrodes of thickness t and conductivity σ is given by [1]

$$Z_s(f) = g\left(\frac{i2\pi f\mu_0}{\sigma}\right)^{1/2} \coth\left[(i2\pi f\mu_0\sigma)^{1/2} t\right], \quad (2)$$

where g is a geometrical factor which is constant for the given coplanar waveguide [10]. The dispersion due to the conductors is modeled mainly by the imaginary part of the surface impedance and the losses (skin-effect loss) mainly by the real part.

In addition, radiation loss into the semi-infinite substrate occurs due to the fact that the pulse source, acting as a propagating dipole has a higher phase velocity at the surface than the phase velocity of the field in the substrate [5]. This generates radiation at a cone angle ψ into the substrate. The

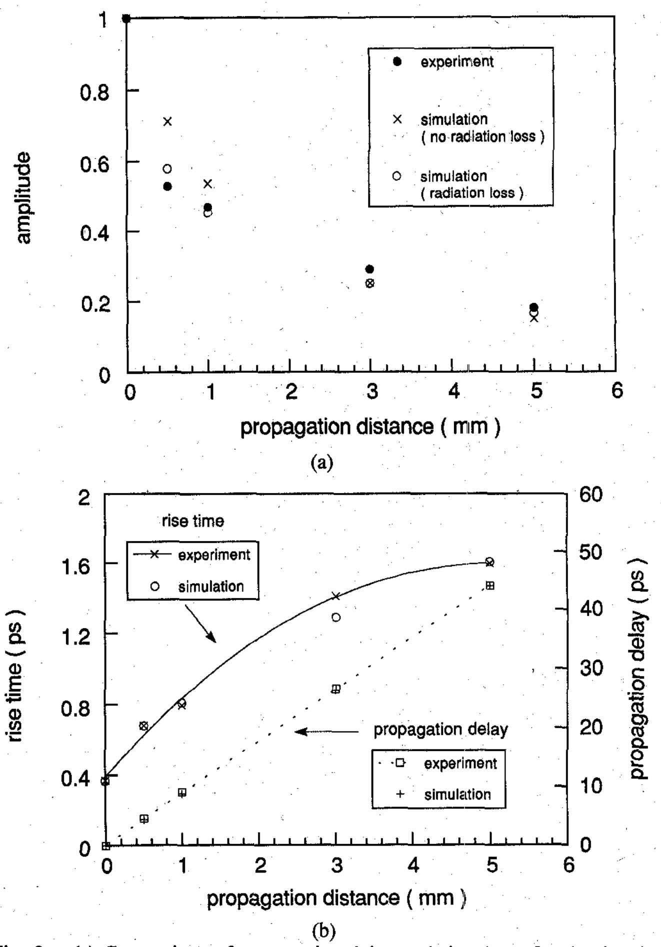

Fig. 3. (a) Comparison of propagation delay and rise times for the data in Fig. 1 and Fig. 2. Curves are intended only as a guide for the eye. (b) Comparison of pulse amplitude for the data in Fig. 1, Fig. 2, and a simulation that disregards radiation losses.

corresponding attenuation α for a CPW in the quasi-static limit has a cubic frequency dependence given by [6]

$$\alpha = \left(\frac{\pi}{2}\right)^5 \frac{1}{\sqrt{2}} \frac{\varepsilon_r^{3/2}(1 - 1/\varepsilon_r)^2}{\sqrt{1 + 1/\varepsilon_r}} \frac{(s + 2w)^2}{c^3 K'(k) K(k)} f^3, \quad (3)$$

where $k = s/(s + 2w)$, K and K' are complete elliptic integrals of the first kind, and c is the velocity of light in vacuum. For our transmission line, calculation gives $\alpha = 3.19 \times f^3$ mm^{-1} (f in THz).

The dielectric loss of the GaAs substrate [11], for a terahertz pulse, is about two orders of magnitude smaller than either the skin-effect or radiation loss, and therefore has negligible influence on the pulse evolution.

The full propagation factor, $\gamma(f)$, is computed including all of the previous effects, and then the pulse propagation is modelled in the Fourier domain using [1], [7]

$$V(t, z) = F^{-1}\{F\{V(t, 0)\} e^{-\gamma(f)z}\} \quad (4)$$

where F and F^{-1} represent the direct and inverse Fourier transforms, and $V(t,0)$ and $V(t,z)$ are the input pulse and the transient after a distance z of propagation, respectively.

For $z = 0.0$ mm, the probing electrooptic crystal is seated almost on top of the gap switch and thus picks up a small radiated component in air. This is seen as the negative going prepulse for $z = 0.0$ mm, but it is absent for other distances as it is not a guided wave. Therefore we approximate the input ($z = 0.0$ mm) by a Gaussian pulse of FWHM = 0.52 ps as shown in Fig. 1 and indicated by the X symbols. The simulation results are shown in Fig. 2. In order to quantify the results, the propagation delay, 10–90% rise time, and peak amplitude of the experimental and simulated data are compared in Fig. 3(a) and (b). Note the excellent agreement obtained for all the propagated distances without the use of any adjustable or fitting parameters. In Fig. 3(b) we also show the peak amplitude without radiation loss taken into account. For short distances radiation is seen to be the dominant loss factor, acting before the amplitude decrease becomes dominated by modal dispersion. Similar agreement has also been observed for data on coplanar-strip (CPS) transmission lines.

On a closer look, some discrepancy is observed between experiment (Fig. 1) and simulation (Fig. 2) on the trailing side of the pulse. Firstly the simulated pulse has a slightly slower falling edge, and secondly the ringing at the trailing side of the pulse is much suppressed for the simulated data as compared to the experimental data. This may be because the previous expression for the radiation loss is a quasi-static one and predicts a very strong high-frequency dependence. Due to the modal dispersion as given by (1), $\varepsilon_{\mathrm{eff}}$ is seen to increase with increasing frequency. Hence the high frequency components in the pulse are delayed towards the trailing side. Also, due to this $\varepsilon_{\mathrm{eff}}$ increase, the higher frequencies are more confined to the substrate resulting in a smaller radiation cone angle ψ, and hence smaller radiation loss. This effect is fairly strong and results in a much weaker high-frequency dependence for the radiation loss [12]. This implies that the simulated waveforms using the quasi-static expression (4) would have less high-frequency content than in the experimental data for the same propagation distance. Thus the falling edge would be slower and the ringing would be damped for the previous simulated data, as we observed. A detailed discussion of this can be found in [12]. In spite of this, the important pulse propagation features are predicted very well by the previous simple model.

IV. Conclusion

We have presented experimental data for subpicosecond (terahertz) pulse propagation on coplanar-waveguide transmission lines. The pulse features of practical importance—propagation delay, rise time, and amplitude—are predicted at various propagation distances by the simple model previously described, with very good agreement obtained without the use of any adjustable parameters. Similar agreement is also found for the data on coplanar-strip transmission lines. For the typical dimensions described here, modal dispersion and radiation losses are the dominant pulse-shaping mechanisms for terahertz pulses. As the conductor plays a minor role in the pulse shaping, the use of superconducting electrodes will not reduce the problems of dispersion and losses on these transmission lines. Hence to preserve the pulse integrity, the use of very thin low-dielectric-constant substrates, much smaller transverse-dimension structures, a superstrate or other improvements have to be adopted.

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