# Phase Compensation and Waveform Reshaping of Picosecond Electrical Pulses Using Dispersive Microwave Transmission Lines

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*Abstract*—**In this paper we study the phase compensation effect of microwave transmission line dispersion and propose a simple, effective method for reshaping and compressing picosecond electrical pulses generated from photoconductive switches. We show that a piece of a dispersive strip transmission line can be used as a "phase equalizer" to compensate the phase distortion included in asymmetric pulses, resulting in effective reshaping and compression of these ultrashort pulses. Initial design formulas of the strip transmission lines for this purpose are presented, together with computer simulation results which confirm the theoretical predictions. Finally we present experimental results to show the substantial pulse reshaping effect, as well as a comparison between theory and measurement.**

## I. Introduction

ULTRAFAST electrical pulses with picosecond and subpicosecond durations are finding steadily increasing applications, such as microwave and millimeter-wave generation, time-domain network analyzers [1], and coherent microwave transient spectroscopy [2]. To generate these extremely short electrical pulses, high-speed photoconductive switches are widely used, although novel methods for generating electrical pulses with femtosecond durations have been reported recently [3]. One of the problems with the photoconductive switch is that the fall time of the electrical pulses generated is strongly dependent on the carrier lifetime, which is usually several hundred picoseconds for many intrinsic photoconductive materials. To reduce the pulse width, ion implantation techniques are commonly employed. The introduction of high defect density by such techniques, however, usually results in low mobility, poor stability, and a lack of compatibility with the fabrication processes of other devices [4].

Because of the abovementioned limitations of the ion implantation technique, there has been increasing interest in obtaining ultrashort electrical pulses which are independent of the carrier lifetime of photoconductive materials. In an earlier work, Li *et al.* studied a number of pulse-forming devices based on the charged line concept, but no rigorous analysis of the pulse waveform was offered [5]. Frankel *et al.* reported the formation of picosecond electrical pulses by inductive shunt and series capacitance discontinuities embedded in a coplanar transmission line [6]. Other pulse-forming methods, such as introducing impedance mismatch into the pulse-generating structure [7] and asymmetric excitation of charged transmission lines [8], have also been proposed recently.

In contrast to the previous pulse-forming methods, which unexceptionally make use of transmission line discontinuities, we have proposed a new method of pulse reshaping and compression using dispersive transmission lines [9]. The dispersion of picosecond electrical pulses on microwave transmission lines has been studied intensively during the past few years [10]–[12], mainly from the viewpoint of how these ultrafast signals are distorted as they travel along the transmission lines. Positive utilization of the dispersion properties in the transmission and control of these ultrashort pulses, on the other hand, has hardly been studied closely. Li *et al.* had noted in their paper [10] the sharpening effect of a single-sided exponential pulse traveling along microstrip lines. A rigorous analysis of this phenomenon, however, has never been obtained.

In this paper we investigate in detail the propagation of such asymmetric pulses, with special attention to the influence of dispersion on the phase distortion of the signal pulses. It is found that microwave transmission lines, such as microstrip lines and coplanar waveguides, possess properties similar to those of a phase equalizer [13] and can be used to correct the phase distortion included in an asymmetric electrical pulse. At a certain distance along the transmission line, the pulse becomes almost symmetric in waveform, and its FWHM (full width half maximum) is several times smaller than that of the original pulse. This provides a simple and effective method for obtaining electrical pulses with good symmetry and a short falling edge.

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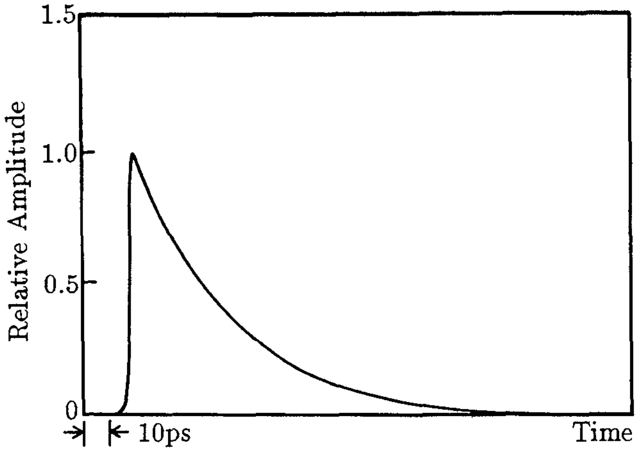


Fig. 1. Waveform of a single-sided exponential pulse.

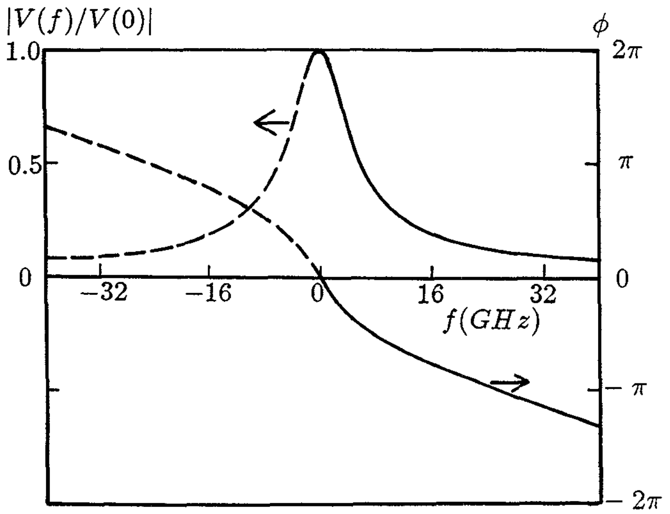


Fig. 2. Fourier spectrum of the single-sided exponential pulse shown in Fig. 1.

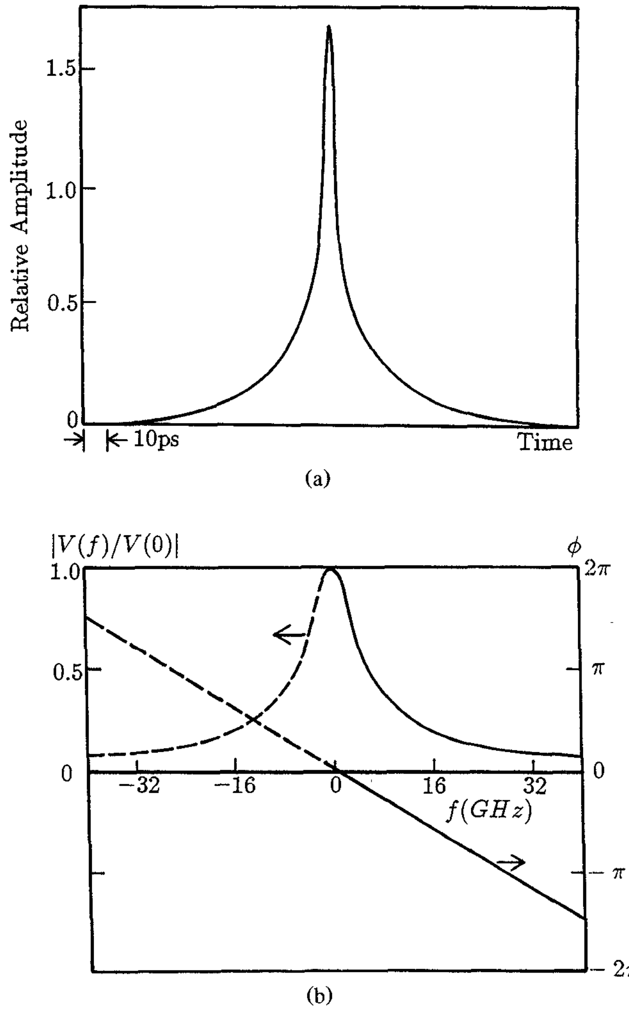


Fig. 3. (a) Transform limit pulse with the same spectral envelope as that of the pulse shown in Fig. 1 and (b) its Fourier spectrum.

## II. Theoretical Analysis

When a photoconductive switch is excited by ultrashort laser pulses, an electrical pulse with approximately the same rise time as that of the optical pulse will be generated. The fall time of the pulse, however, is determined primarily by the carrier lifetime, which may extend to a few hundred picoseconds for certain photoconductive materials. As a result, the generated electrical pulse is usually of an asymmetric waveform, which can be approximately described by a Gaussian rising edge followed by an exponentially decaying tail [6]. These asymmetric pulses are called single-sided exponential pulses, as shown in Fig. 1, and are the typical output pulse forms of a number of photoconductive switches.

We begin our analysis by investigating the Fourier spectra of these asymmetric pulses. Fig. 2 shows both spectral amplitude and phase of the single-sided exponential pulse given in Fig. 1. It is found that the phase of the signal, $\phi$, is a nonlinear function of frequency, $f$, which means that there is a phase distortion existing in the signal pulse. If we define $f_{0.1}$ as the frequency where the spectral amplitude is 10% of the peak value, we can approximately express $\phi$ with the polynomial

$$\phi = a_1 \frac{f}{f_{0.1}} + a_2 \left( \frac{f}{f_{0.1}} \right)^2 + a_3 \left( \frac{f}{f_{0.1}} \right)^3 + 0(f^4) \qquad (1)$$

for $f < f_{0.1}$. If terms of higher order are negligible, the nonlinear part, or phase distortion of $\phi$, can be attributed mainly to the second- and third-order terms in the above polynomial.

When a pulse travels along a transmission line, it will receive a continuous phase delay. The amplitude of the spectrum, however, will remain the same as long as the loss and nonlinearity of the transmission line are negligible. It is interesting to see how the phase delay affects the signal phase, $\phi$, and results in a change in the pulse shape. A special case is where $\phi$ becomes a linear function of frequency, $f$, as shown in Fig. 3(b). By an inverse Fourier transform we obtain in the time domain a pulse of the type shown in Fig. 3(a). This is called a Fourier transform limit (TL) pulse, which is the narrowest pulse possible for a given spectral amplitude distribution.

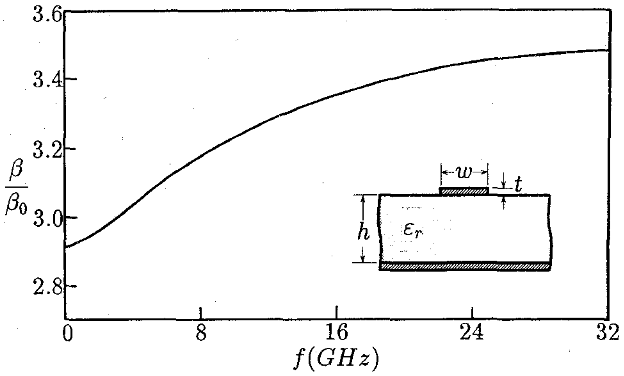


Fig. 4. Frequency dependence of the propagation constant, $\beta/\beta_0$, of a 50 Ω microstrip line ($\epsilon_r = 12.9$, $h = 2$ mm).

Since the TL pulse has a very narrow pulse width as well as good symmetry, it is interesting to find a way of obtaining such pulses by eliminating the nonlinear part in the signal phase. We will show below that a piece of a dispersive transmission line can be used to realize this phase compensation, resulting in effective reshaping and compression of pulses with asymmetric waveforms.

The dispersion characteristics of various types of transmission lines, among them microstrip lines, coplanar waveguides, and coplanar strips, have been studied in great detail by many authors. For computer analysis, a simple approximation formula developed by Yamashita *et al.* [14] has been widely used. We rewrite the dispersion formula here in the following general form, which is valid for all the abovementioned types of transmission lines:

$$\frac{\beta}{\beta_0} = \frac{\sqrt{\epsilon_r} - \frac{\beta_{\mathrm{TEM}}}{\beta_0}}{1 + aF^{-b}} + \frac{\beta_{\mathrm{TEM}}}{\beta_0}. \tag{2}$$

Here $F = f/f_{\mathrm{TE}}$ is the normalized frequency, $f_{\mathrm{TE}} = c/4h\sqrt{\epsilon_r - 1}$ is the cutoff frequency for the lowest order TE mode, $\beta_{\mathrm{TEM}}$ is the propagation constant assuming the quasi-TEM approximation, and $a$ and $b$ are constants which depend on the type and dimensions of the transmission line and can be obtained by curve-fitting the dispersion data calculated with numerical methods. For a 50 Ω microstrip line on GaAs substrate with a thickness $h = 2$ mm, the propagation constant, $\beta/\beta_0$, within the frequency range of our interest is shown in Fig. 4.

The phase delay of a pulse propagating along the transmission line, $\psi$, is expressed as follows:

$$\psi = \frac{2\pi f L}{c} \cdot \frac{\beta}{\beta_0} \tag{3}$$

where $L$ is the propagation distance, and $\beta/\beta_0$ is given by (2). Because of the frequency dependence of $\beta/\beta_0$, $\psi$ is also a nonlinear function of frequency. If this phase delay can be used to correct the nonlinear part in the signal phase, $\phi$, so that $\phi - \psi$ becomes a linear function of the frequency, $f$, we can expect to obtain a pulse which is similar to the TL pulse shown in Fig. 3(a).

To describe the predicted phase compensation effect, it is desirable that $\psi$ be expanded in a way similar to that of (1). Since in (2) the dispersion at $f = 0$ is not defined, we take the Taylor series expansion at $F = 1$ as follows:

$$\frac{\beta}{\beta_0} = \left(\sqrt{\epsilon_r} - \frac{\beta_{\mathrm{TEM}}}{\beta_0}\right)\left\{\frac{1}{a+1} + \frac{ab}{(a+1)^2}(F-1) + \frac{ab(ab-a-b-1)}{2(a+1)^3}(F-1)^2 + 0\left[(F-1)^3\right]\right\} + \frac{\beta_{\mathrm{TEM}}}{\beta_0} \tag{4}$$

for $|F-1| < 1$. Consequently,

$$\psi = b_1 f + b_2 f^2 + b_3 f^3 + 0(f^4) \tag{5}$$

for $0 < f < 2f_{\mathrm{TE}}$, where

$$b_1 = \frac{2\pi L}{c}\left[\frac{\beta_{\mathrm{TEM}}}{\beta_0} + \left(\sqrt{\epsilon_r} - \frac{\beta_{\mathrm{TEM}}}{\beta_0}\right) \cdot \frac{2(a+1)^2 - 3ab(a+1) + ab^2(a-1)}{2(a+1)^3}\right] \tag{5a}$$

$$b_2 = \frac{2\pi Lab[2(a+1) - b(a-1)]}{cf_{\mathrm{TE}}(a+1)^3}\left(\sqrt{\epsilon_r} - \frac{\beta_{\mathrm{TEM}}}{\beta_0}\right) \tag{5b}$$

and

$$b_3 = \frac{\pi Lab(ab-a-b-1)}{cf_{\mathrm{TE}}^2(a+1)^3}\left(\sqrt{\epsilon_r} - \frac{\beta_{\mathrm{TEM}}}{\beta_0}\right). \tag{5c}$$

Comparing (5) with (1), we find that if the second- and third-order terms in the two polynomials are equal to each other, the resultant phase of the signal after propagation becomes

$$\phi - \psi = \left(\frac{a_1}{f_{0.1}} - b_1\right) f + 0(f^4) \tag{6}$$

which is a linear function of $f$ when higher order terms are neglected. Taking an inverse Fourier transform, we can expect to obtain a pulse which is close to the TL pulse shown in Fig. 3(a). By equating $b_2$ and $b_3$ in (5b) and (5c) to their corresponding coefficients in (1), we obtain the cutoff frequency, $f_{\mathrm{TE}}$, and the optimum length, $L_{\mathrm{opt}}$, of the transmission line:

$$f_{\mathrm{TE}} = \frac{f_{0.1}a_2[b(a-1) - (a+1)]}{2a_3[2(a+1) - b(a-1)]} \tag{7a}$$

and

$$L_{\mathrm{opt}} = \frac{ca_2 f_{\mathrm{TE}}(a+1)^3}{2\pi abf_{0.1}^2\left(\sqrt{\epsilon_r} - \frac{\beta_{\mathrm{TEM}}}{\beta_0}\right)[2(a+1) - b(a-1)]}. \tag{7b}$$

Once $f_{\mathrm{TE}}$ is known, the dimensions of the transmission line, mainly the substrate thickness, $h$, can be easily determined from the relation $f_{\mathrm{TE}} = c/4h\sqrt{\epsilon_r - 1}$. By selecting the constants, $a$ and $b$, the above design formulas can be used for a number of transmission lines, including microstrip lines, coplanar waveguides, and coplanar striplines [11].

## III. Computer Simulations

In order to confirm the pulse-reshaping and compression effects discussed in the previous section, a computer program has been written to simulate pulse propagation along strip transmission lines. We use an algorithm similar to that given by Li *et al.* [10]. The input signal, a single-sided exponential pulse of the type shown in Fig. 1, is expressed as follows:

$$V(0,t)=\begin{cases} V_0 e^{-4\ln 2 (t/\tau_1)^2}, & t<0 \\ V_0 e^{-t/\tau_2}, & t>0 \end{cases} \tag{8}$$

where $\tau_1$ and $\tau_2$ are time constants which determine the rise and fall times of the input pulse, respectively. A forward Fourier transform gives the spectrum of the input pulse, $\mathscr{F}[V(0,t)]$. Multiplying this by the propagation factor, $e^{-\gamma(f)L}$, and taking an inverse Fourier transform will result in $V(L,t)$, the pulse waveform at a propagation distance, $L$. The above procedure can be expressed as

$$V(L,t)=\mathscr{F}^{-1}\{\mathscr{F}[V(0,t)]\cdot e^{-\gamma(f)L}\} \tag{9}$$

where $\mathscr{F}$ denotes the Fourier transform, and the complex propagation constant, $\gamma(f)$, is given by

$$\gamma(f)=\alpha(f)+j\beta(f) \tag{10}$$

where $\alpha(f)$ and $\beta(f)$ are the attenuation constant and phase constant of the transmission line, respectively.

In analyzing the phase compensation effect in the previous section, we have neglected the frequency-dependent attenuation of the transmission line, $\alpha(f)$. Since the propagation distance of interest is usually within a few centimeters in most of the cases considered here, this should not become a serious problem [15]. To prove the validity of this assumption and to compare the simulation results with experimental data, however, we have included the effect of propagation attenuation in the simulation algorithm. The dielectric loss of a microstrip line is given by, e.g., [16]:

$$\alpha_d(f)=27.3\cdot\frac{\epsilon_r}{\epsilon_r-1}\cdot\frac{\epsilon_{\text{eff}}(f)-1}{\sqrt{\epsilon_{\text{eff}}(f)}}\cdot\frac{\tan\delta}{\lambda_0}\quad(\text{dB/m}) \tag{11}$$

where $\lambda_0$ is the free-space wavelength, $\tan\delta$ is the loss tangent, and $\epsilon_{\text{eff}}(f)=(\beta/\beta_0)^2$ is the effective dielectric constant at frequency $f$. The conductor loss is calculated as follows [16]:

$$\alpha_c(f)=1.38\cdot\frac{R_s}{hZ_0}\left[1+\frac{h}{w_e}\left(1+\frac{1.25}{\pi}\ln\frac{2h}{t}\right)\right]\cdot\frac{32-(w_e/h)^2}{32+(w_e/h)^2}\quad(\text{dB/m}) \tag{12}$$

where the conductor surface resistance, $R_s$, the characteristic impedance of the microstrip line, $Z_0$, and the effective width of the conductor strip, $w_e$, are given by

$$R_s=\sqrt{\pi f\mu_0\rho} \tag{12a}$$

$$Z_0=\frac{60}{\sqrt{\epsilon_{\text{eff}}(f)}}\ln\left[\frac{8h}{w_e}+0.25\frac{w_e}{h}\right] \tag{12b}$$

and

$$w_e=w+\frac{1.25t}{\pi}\left\{1+\ln\frac{2h}{t}\right\} \tag{12c}$$

respectively. Here $\rho$ is the resistivity of the strip conductor, $h$ is the thickness of the substrate, and $w$ and $t$ are the width and thickness of the strip conductor, respectively. Finally, the attenuation constant, $\alpha(f)$, when assuming no radiation losses, can be written as

$$\alpha(f)=\alpha_c(f)+\alpha_d(f). \tag{13}$$

As an example, a single-sided exponential pulse with a 2 ps rise time and a total FWHM of 35 ps is used as the input signal. Taking a forward FFT (fast Fourier transform) we obtain the phase, $\phi$, as well as the amplitude, $V(f)$, of its spectrum as shown in Fig. 2. The coefficients $a_1$, $a_2$ and $a_3$ in (1) can be derived by least-square curve-fitting of the FFT data and are found to be $-7.29$, 6.96, and $-3.37$, respectively. Assuming a 50 Ω microstrip line on GaAs substrate, we have $\epsilon_r=12.9$, $\beta_{\text{TEM}}/\beta_0=2.91$, $a=0.94$, and $b=1.5$ [14]. Using the formula (7a), the cutoff frequency, $f_{\text{TE}}$, is calculated to be 16.9 GHz, which corresponds to a substrate thickness of 1.3 mm. Applying formula (7b), we obtain the optimum length of the microstrip line: $L_{\text{opt}}=11$ mm. Computer simulation results of pulse propagation along the designed microstrip line are shown in Fig. 5. It is seen that near the optimum length, $L_{\text{opt}}$, the pulse is closest in waveform to that of the TL pulse shown in Fig. 3(a). The FWHM of the compressed pulse is 9 ps, which is about four times narrower than that of the original pulse. Further propagation along the microstrip line will again distort and broaden the pulse.

## IV. Experiments

The experimental system for studying pulse propagation is shown in Fig. 6. The optical source is a picosecond laser pulser producing 50 ps (FWHM) pulses at a center wavelength of 820 nm. The repetition rate of the pulse is 10 MHz and its peak power is 240 mW. This laser pulse is used to excite our InGaAsP switch, which has a gap length of 5 μm. The output waveform of the photoconductive switch, as shown in Fig. 7, is an asymmetric pulse with a rise time of 60 ps and a falling edge which extends to several hundred picoseconds.

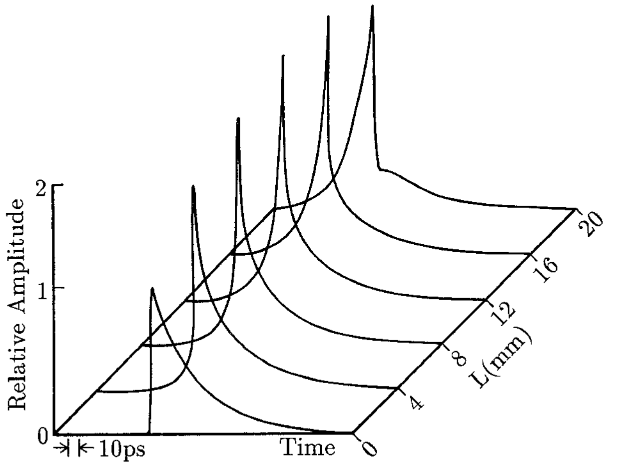


Fig. 5. Computer simulation results of the propagation of a 35 ps single-sided exponential pulse along 50 Ω microstrip lines on GaAs substrate ($\epsilon_r = 12.9$, $h = 1.3$ mm).

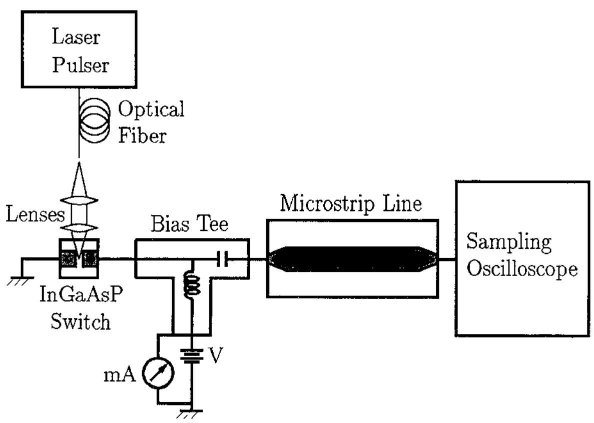


Fig. 6. Schematic of the experimental arrangement for studying pulse propagation.

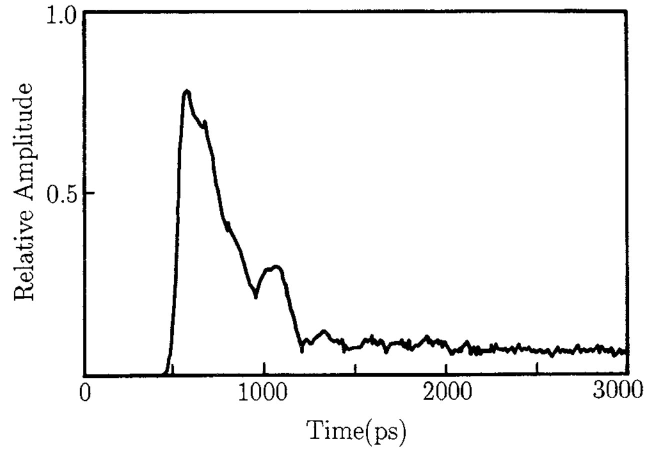


Fig. 7. Output pulse waveform of the InGaAsP photoconductive switch.

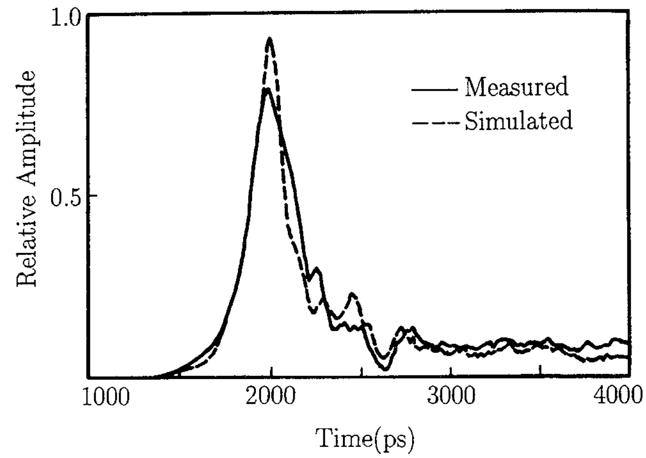


Fig. 8. Experimental (solid line) and simulation (dashed line) results of pulse propagation along the designed microstrip line.

We have designed a microstrip line to reshape this pulse to its Fourier transform limit. Choosing ceramic ($\epsilon_r = 10.3$) as substrate material, we obtained according to (7a) the cutoff frequency of the stripline, $f_{TE}$, as 2.68 GHz, which corresponds to a substrate thickness of 9.2 mm, and a strip width of 8.54 mm in order to maintain the 50 Ω characteristic impedance. The optimum length of the stripline, $L_{opt}$, is calculated to be 11 cm. To facilitate transitions from the stripline to the SMA connectors, a 1-cm-long taper is added to each end of the stripline.

The waveform of the pulse after propagation along the stripline is plotted in Fig. 8 (solid line). For comparison the computer simulation result of the pulse propagation is shown in the same figure (dashed line). It is found that the original asymmetric pulse as shown in Fig. 7 has been reshaped to be close to a symmetrical one. The rising edge of the measured pulse is in excellent agreement with that of computer simulations. The falling edge, however, is less in accordance with the calculations, and the amplitude of the pulse is also somewhat lower than predicted. These discrepancies are mainly due to the introduction of tapers at the two ends of the stripline. The losses at the discontinuities resulted in a lower pulse amplitude, and the degradation of the falling edge was caused by signal reflections at these discontinuities. Otherwise, the main feature of the measured pulse is in reasonable agreement with theoretical predictions.

## V. Conclusions

The propagation of picosecond electrical pulses on strip transmission lines has been investigated, with special attention to the phase compensation effect of the dispersion properties of the transmission lines. It has been shown that by using a piece of a carefully designed strip transmission line, asymmetric electrical pulses can be reformed to be close to their Fourier transform limit. This provides a simple and effective method for reshaping and compressing picosecond electrical pulses generated from photoconductive switches. The pulse-reshaping and compression effect has been confirmed by computer simulations as well as experiments, both of which are in reasonable agreement with theoretical analysis.

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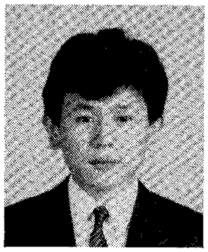

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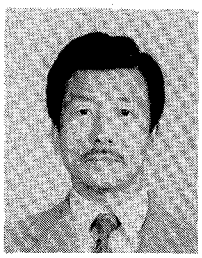

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