Modal Dispersion Control and Distortion Suppression of Picosecond Pulses in Suspended Coplanar Waveguides

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***Abstract*—In this paper we investigate the propagation and dispersion characteristics of picosecond electrical pulses in a suspended coplanar waveguide (SCPW) and show that the SCPW is a very promising transmission structure for ultrashort pulses with picosecond and subpicosecond durations, compared with conventional coplanar waveguides (CPWs). Numerical results of modal dispersion of the SCPW are presented and compared to those of the conventional CPW, and a field coupling theory is employed to explain the evolution in the dispersion behavior. An evaluation based on the numerical analysis shows that the SCPW with properly controlled dispersion can result in an improvement of about 5 times in pulse transmission capability than the conventional CPW. The propagation of picosecond pulses along the SCPW has been studied by both computer simulations and experimental measurements, both showing a substantial suppression in pulse distortion compared to the case of conventional coplanar waveguides.**

I. Introduction

THE STEADFAST development of millimeter-wave integrated circuits and high-speed digital circuits has led to great advances in the field of ultrafast, very high frequency electronics. With the combination of photoconductive switches and ultrashort lasers, electrical pulses with picosecond and subpicosecond durations are now routinely available, and can be accurately characterized using electro-optic (EO) or photoconductive (PC) sampling techniques [1], [2]. Because of their extremely wide frequency bandwidth, these ultrashort electrical pulses are finding steadily increasing applications such as microwave and millimeter wave generation, time-domain network analyzers [1] and coherent microwave transient spectroscopy [3].

Compared to the generation and characterization of picosecond electrical pulses, which have been an almost matured technology, the development of transmission structures capable of handling the extremely wide bandwidth of these signals still remains an important issue so far. This would not be possible without a full understanding of the mechanisms of pulse distortion, including modal dispersion, conductor and dielectric losses, material resonances and other phenomena in the transmission line. Whitaker *et al.* have studied the propagation of ultrafast signals on both normal and superconducting strip transmission lines [4], [5]. A more accurate model for pulse propagation along coplanar striplines has been developed by Phatak *et al.*, where the influence of the substrate modes on the transmission line modes as well as a few other factors has been taken into consideration [6]. The effect of optical phonons on femtosecond pulse propagation has also been demonstrated by Hasnain *et al.* in a recent paper [7].

For electrical pulses with several ten picoseconds durations, the modal dispersion due to geometrical dimensions of the transmission line is the dominant factor of pulse distortion. This modal dispersion can usually be reduced by making the dielectric substrate thinner to increase the cutoff frequency of the dominant mode in the transmission line [8]. Nearly dispersion-free propagation of pulses of a few picoseconds has been reported very recently by Goossen *et al.*, where a microstrip with buried silicide groundplane was utilized [9], [10]. In order to be compatible with the fabrication process of other circuit components, however, the thickness of the dielectric substrate cannot be reduced without limitations, and is usually around 100 ~ 200 μm for most GaAs MMIC's [11]. It is thus worthwhile to investigate the mechanisms of modal dispersion in transmission lines compatible with the present MMIC technology and find a way to control their dispersion behavior, realizing distortion-free propagation of ultrafast signals in high-speed circuits.

In a previous paper [12] we proposed the structural control of the dispersion characteristics of microstrip lines by adding two parallel conductors on both sides of the strip conductor. The field coupling effect results in a change in the dispersion behavior of the transmission line, which can be used for pulse shaping in some applications [4]. We have made further investigation into this dispersion control effect and the properties of pulse propagation in these dispersion-controlled transmission lines. In particular, the modal dispersion of a suspended coplanar waveguide (SCPW) has been studied and found that it has a very high cutoff frequency compared with conventional microstrip lines (MS) or coplanar waveguides (CPWs) fabricated on the same dielectric substrate [13]. It is be-

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lieved that the SCPW should be a very promising waveguiding structure for interconnection of devices in high-speed circuits. We have investigated the propagation of picosecond pulses along the SCPW by computer simulations as well as experimental measurements, both showing a substantial suppression in pulse distortion compared to the case of conventional coplanar waveguides (CPWs).

II. Dispersion Control of Coplanar Waveguides

Fig. 1 shows the cross-sectional view of a suspended coplanar waveguide (SCPW) with a metallic shielding enclosure. This structure differs from the conventional CPW in that an air layer is introduced beneath the dielectric substrate in order to reduce dispersion resulted from the inhomogeneity of the transmission line structure. In practical applications, a side-groove or pedestal is usually added to the metal shielding enclosure so that the dielectric substrate can be fixed mechanically. The dispersion characteristics of such coplanar waveguides have been recently analyzed by Alessandri *et al.* using the generalized transverse resonance technique [14].

To obtain the dispersion characteristics of the SCPW structure as shown in Fig. 1, we have used the spectral domain approach [15]. The strip conductors are assumed to be negligibly thin and lossless. We are not repeating details of the numerical method since it has been introduced extensively in the literature. Rather, we have attempted to find out the mechanism of modal dispersion in these waveguides and explain the behavior of dispersion curves following changes in the geometry and dimensions of the transmission line. We can then expect to find a way of controlling the dispersion properties intentionally, and reducing signal distortion along the transmission line to the minimum.

The general behavior of the propagation constant of a coplanar waveguide with increased frequency is plotted as curve A in Fig. 2. To explain this dispersion characteristic we introduced a field coupling theory in a previous paper [12]. The CPW structure as shown in Fig. 1 ($h_1 = 0$) can be considered as a shielded slotline added later by a central conductor. The dominant mode of the slotline structure is represented by the curve B, with a cutoff frequency of f_c. The introduction of the central strip, which forms a two-conductor structure, results in the existence of a quasi-TEM mode as the dominant propagation mode in the CPW. At low frequencies, its propagation constant is represented by β_{TEM}, the result when quasi-TEM approximation is assumed (curve C). At extremely high frequencies, the field energy is almost totally concentrated in the dielectric substrate, and the normalized propagation constant eventually converges to $\sqrt{\epsilon_r}\beta_0$. Between these two extremes, the behavior of the propagation constant with increased frequency, as shown by curve A, looks like an asymptote of the two curves B and C. This might be the result of field coupling when the two modes corresponding to B and C become phase matched and interact strongly near the intersection of the two curves.

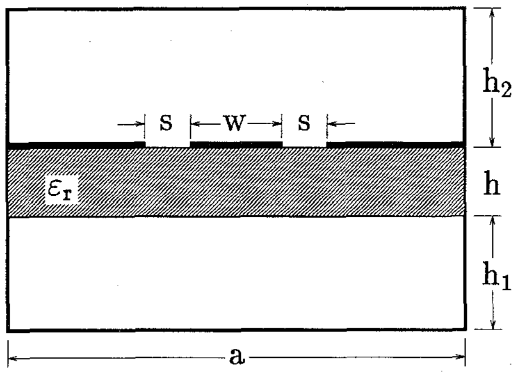

Fig. 1. Cross-sectional view and geometry of the suspended coplanar waveguide (SCPW).

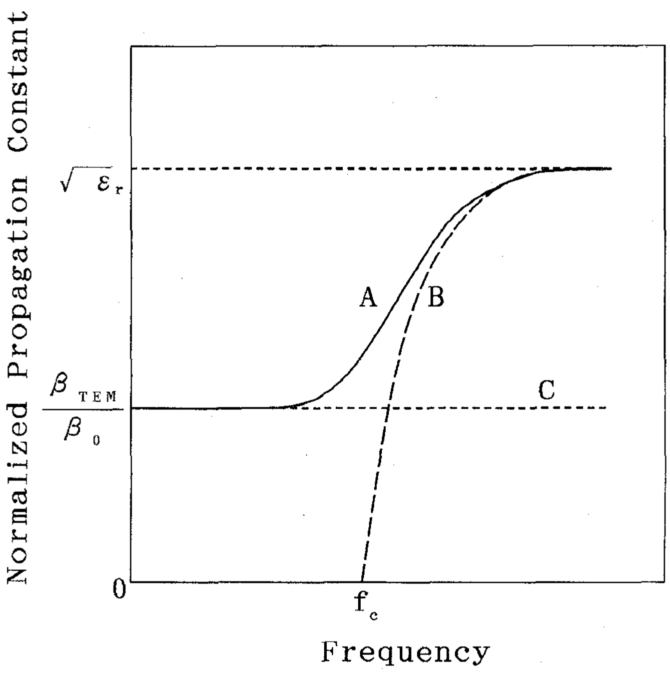

Fig. 2. General behavior of frequency dispersion of the dominant mode of a coplanar waveguide.

Figs. 3 and 4 show some of the numerical results of the dispersion characteristics of CPWs with different geometries and dimensions. The curves A_1 and A_2 in Fig. 3 are the propagation constant of two conventional CPWs ($h_1 = 0$) with different slot widths, $s = 280$ μm and $s = 10$ μm, respectively. The horizontal lines, C_1 and C_2, represent the normalized propagation constant in the quasi-TEM approximation, β_{TEM}/β_0, for these two CPWs. When the center conductor in Fig. 1 is removed, the CPW becomes a shielded slotline structure, and curves B_1 and B_2 represent the dominant modes of the slotlines for the above two cases.

The differences of the dispersion behavior between curves A_1 and A_2 in Fig. 3 can be explained as follows:

1) When the slot width between the center conductor and side ground planes in Fig. 1, s, is large, there is a relatively tight coupling between the quasi-TEM mode and slotline mode, since the field distribution of the two modes resemble each other to a consid-

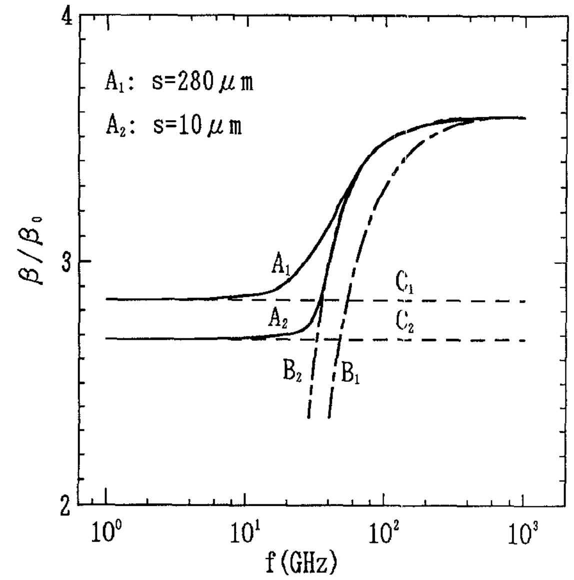

Fig. 3. Comparison of the dispersion properties of two shielded coplanar waveguides with different slot widths. (Other parameters: $w = 0.14$ mm, $a = 2$ mm, $h_1 = 0$, $h = 0.2$ mm, $h_2 = 2$ mm, $\epsilon_r = 12.9$)

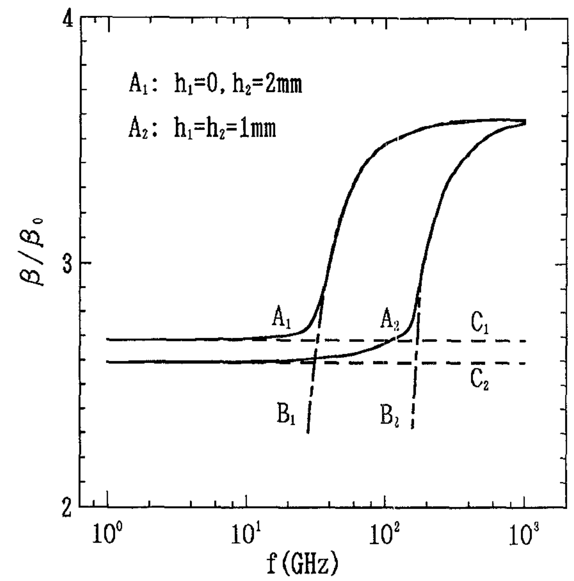

Fig. 4. Comparison of the dispersion properties of a conventional CPW (A_1) and a SCPW (A_2). (Other parameters: $w = 0.14$ mm, $a = 2$ mm, $h = 0.2$ mm, $s = 10$ μm, $\epsilon_r = 12.9$)

erable extent. The result is a gradual rising in the curve of the propagation constant (A_1).

2) When the slot width, s, is so narrow that the energy is highly concentrated in the two gap regions between the center strip and side ground conductors, the field patterns corresponding to the quasi-TEM mode and the dominant slotline mode are almost orthogonal to each other. This is the case of weak coupling between the two modes. The dispersion effects are not appreciable up to the vicinity of the intersection point of B_2 and C_2, where the propagation constant begins to rise rapidly with increased frequency (A_2).

The abovementioned fact that the dispersion behavior of the CPW can be controlled from gradual dispersion to abrupt dispersion can be used for pulse shaping in some applications [4]. On the other hand, however, there is little improvement in the useful frequency bandwidth from the viewpoint of transmission lines, since the cutoff frequency of the slotline mode decreases as the slot width becomes narrower. In Fig. 3, the coplanar waveguide with $s = 280$ μm can be used in a frequency range up to about 20 GHz (when β is 3 percent larger than β_{TEM}), while that of the CPW with $s = 10$ μm is approximately 30 GHz.

In order to make the transmission line free of dispersion in a wider range of frequencies, we can introduce an air layer beneath the dielectric substrate to form a suspended coplanar waveguide (SCPW). The structure can be considered as the combination of a unilateral finline and a central strip conductor, and the dispersion behavior can be explained in the same way as above. The advantage of making the CPW suspended is that the corresponding finline structure has a relatively wide singlemode bandwidth as it resembles the ridged waveguide [16]. Fig. 4 shows a comparison of the dispersion properties of a conventional CPW (A_1) and a SCPW (A_2). The dominant mode of the finline (curve B_2) has a much higher cutoff frequency than the case without the under air layer, and the useful frequency bandwidth is greatly increased. Assuming the same dielectric substrate (200 μm GaAs), center strip width (140 μm) and slot width (10 μm), we find that the SCPW can be used in the frequency range near to 100 GHz, about three times broader than that of the conventional CPW (30 GHz).

III. Simulation Results of Pulse Propagation

To compare the performance of the various coplanar waveguides discussed in the previous section, we have carried out computer simulations of pulse propagation using an algorithm described in a previous paper [17]. We used the following Gaussian pulse as the input signal

$$V(0, t) = V_0 e^{-4\ln 2(t/\tau)^2} \tag{1}$$

where τ represents the FWHM (Full Width at Half Maximum) of the pulse. The input signal is Fourier transformed into its frequency spectrum, $\mathfrak{F}[V(0, t)]$. Multiplying this by the propagation factor, $e^{-\gamma(f)L}$, and taking an inverse Fourier transform results in $V(L, t)$, the pulse waveform at a propagation distance L. The above procedure is expressed as follows

$$V(L, t) = \mathfrak{F}^{-1}\{\mathfrak{F}[V(0, t)] \cdot e^{-\gamma(f)L}\} \tag{2}$$

where $\mathfrak{F}$ denotes the Fourier transform and $\gamma(f)$ is the propagation constant of the transmission line.

A complete description of the propagation constant, $\gamma(f)$, should include both the attenuation constant, $\alpha(f)$, and the phase constant, $\beta(f)$, as its real and imaginary part respectively. Since in this paper we are considering pulses of a few tens of picoseconds, whose frequency components are also within a few tens of gigahertz, signal attenuation due to conductor and dielectric losses as well as radiation losses is usually small and can be neglected without much influence on the simulation results of pulse propagation [18], [5]. For simplicity we have considered only the imaginary part of the propagation constant, $\beta(f)$, in our simulation program. The full-wave analysis results of $\beta(f)$ as shown in Section II can be approximately expressed by the following dispersion formula [19].

$$\frac{\beta}{\beta_0} = \frac{\sqrt{\epsilon_r} - \beta_{\text{TEM}}/\beta_0}{1 + aF^{-b}} + \frac{\beta_{\text{TEM}}}{\beta_0} \tag{3}$$

where $F = (f/c)4h\sqrt{\epsilon_r - 1}$ is the normalized frequency, β_{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 from numerical computation. Although this approximate formula was originated for use with microstrip lines, it has also been used extensively to describe the frequency dispersion of coplanar waveguides [20] and coplanar striplines [21], by changing two parameters, because of its simplicity and good accuracy. As an example, for the dispersion curve A_1 shown in Fig. 3, we obtain $a = 0.23$ and $b = 1.85$ after a least-square curve-fitting procedure, and Formula (3) has an accuracy of within 2 percent in the frequency range of from 0 to 100 GHz.

Fig. 5 shows the computer simulation results of pulse propagation of a Gaussian pulse along the three coplanar waveguides described in the previous section. The input signal is a 20 ps (FWHM) Gaussian pulse with a rise time (10% to 90%) of 15 ps. The frequency bandwidth of this pulse is approximately 40 GHz, where its spectral amplitude drops down to 10 percent of the peak value. Referring to Fig. 3 we find that the CPW structure with $s = 280$ μm is strongly dispersive within the frequency range between 10 GHz and 40 GHz. The gradual evolution of the pulse waveform along this transmission line is shown in Fig. 5(a). After a propagation distance of 30 mm, the pulsewidth is broadened to about 1.5 times its original width, with a much slower risetime (35 ps) accompanied by a strong ringing tail following the main peak. Accompanying the broadening in pulsewidth, the peak amplitude of the pulse is reduced, and becomes 75 percent that of the original signal after 30 mm propagation. The CPW structure with $s = 10$ μm is less dispersive up to 30 GHz, after which the propagation constant increases rapidly with frequency. Consequently, the pulse waveform after 30 mm propagation has a narrower rise time (29 ps) compared to the case of $s = 280$ μm, but with more oscillation ringings following the main peak due to the stronger dispersion at high frequencies (see Fig. 5(b)). On the other hand, the SCPW structure, as shown by curve A_2 in Fig. 4, is almost dispersion-free up to about 100 GHz. As seen in Fig. 5(c), the 20 ps input pulse remains almost unchanged, with its rise time keeping at 15 ps even after a propagation distance of 30 mm along the transmission line.

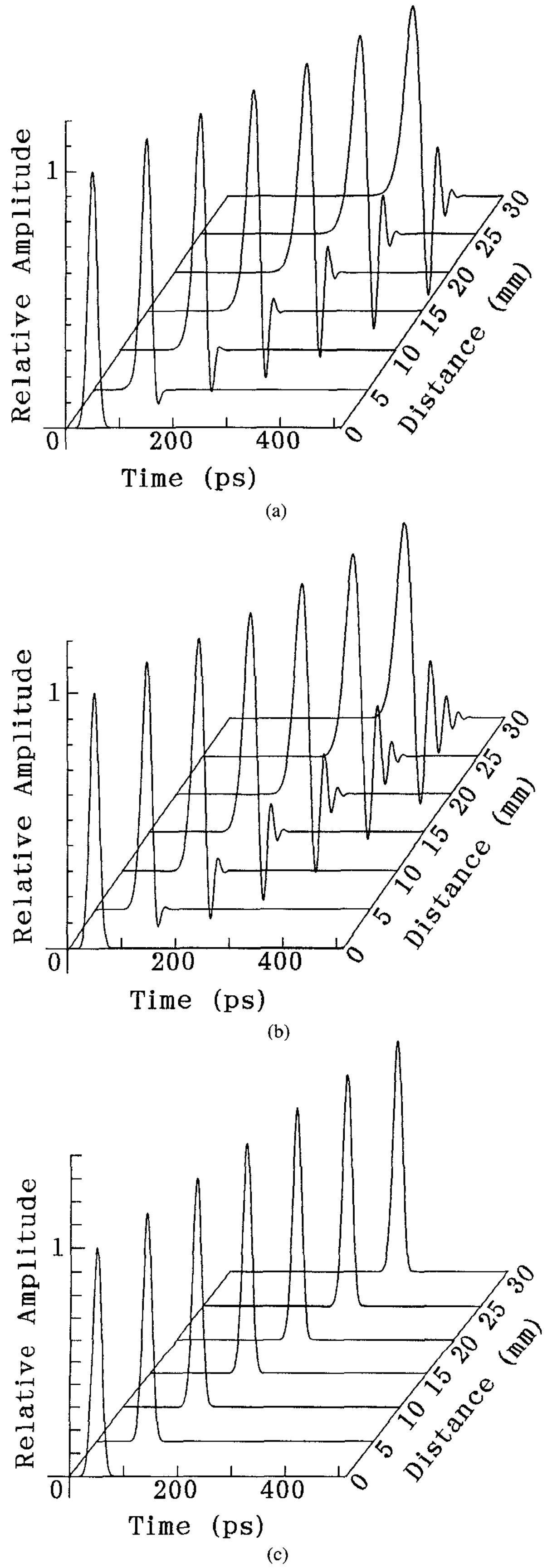

Fig. 5. Computer simulation results of the propagation of a 20 ps Gaussian pulse along three types of transmission lines: (a) a CPW with $s = 280$ μm, (b) a CPW with $s = 10$ μm and (c) a SCPW with $h_1 = h_2 = 1$ mm and $s = 10$ μm. (Other parameters: $w = 0.14$ mm, $a = 2$ mm, $h = 0.2$ mm, $\epsilon_r = 12.9$)

To offer a further comparison of the performance of the abovementioned coplanar waveguides, we make an approximate evaluation of the minimum pulsewidth which can be supported on these transmission lines. For the Gaussian pulse in Formula (1), we obtain its Fourier transform as follows

$$V(0, \omega) = \sqrt{\frac{\pi}{4 \ln 2}}\, V_0 \tau e^{-\omega^2\tau^2/16\ln 2}. \tag{4}$$

If we define $f_{0.1}$ as the frequency where the spectral amplitude is 10 percent of the peak value, we can derive from (4)

$$f_{0.1} = \frac{2\sqrt{\ln 2 \cdot \ln 10}}{\pi \tau} \tag{5}$$

We further define f_{TEM} as the frequency where β in Formula (3) is 3 percent larger than β_{TEM}, which is given as follows

$$f_{\mathrm{TEM}} = \left(\frac{\sqrt{\epsilon_r} - 1.03\beta_{\mathrm{TEM}}/\beta_0}{0.03a\beta_{\mathrm{TEM}}/\beta_0}\right)^{-1/b} f_{TE} \tag{6}$$

This frequency can be considered as the useful bandwidth of the transmission line, and the dispersion of the propagation constant can be neglected within this frequency range. Since 99 percent of a pulse energy is contained in frequencies less than $f_{0.1}$, we may regard that the pulse will propagate without dispersion if $f_{0.1}$ does not exceed f_{TEM}. We thus obtain the minimum FWHM of the pulse as follows

$$\tau_{\min} = \frac{2\sqrt{\ln 2 \cdot \ln 10}}{\pi f_{\mathrm{TEM}}} \tag{7}$$

The useful frequency bandwidth, f_{TEM}, for the CPW with $s = 280\ \mu$m shown in Fig. 3 is 16.3 GHz, and 76.2 GHz for the SCPW in Fig. 4. The minimum pulsewidth, $\tau_{\min}$, is calculated to be 49.3 ps and 10.6 ps, respectively, for the two cases, which shows an improvement of near to 5 times in the pulse propagation capability of the transmission line.

IV. Experimental Results

Fig. 6 shows the experimental system for studying pulse propagation, which is similar to that described in a previous paper [17]. An optically generated electrical transient, after passing a microstrip gap filter discontinuity to form a nearly Gaussian pulse [22], is coupled into the strip transmission line through coaxial cables and SMA connectors, and the output pulse at the end of the transmission line is connected to the sampling head of a digital oscilloscope (Tektronix 11802). The sampling oscilloscope is controlled via GPIB by a personal computer (NEC PC-9801) for data transference and processing. The sampling head of the oscilloscope (SD-30) has a frequency bandwidth of 40 GHz, which corresponds to a time resolution of 8.8 ps.

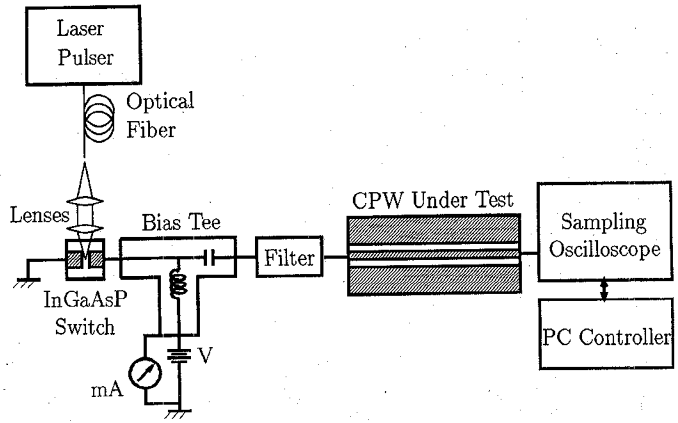

Fig. 6. Schematic of the experimental arrangement for studying pulse propagation.

Figs. 7 and 8 present some experimental as well as simulation results of the propagation of the Gaussian-like pulse along a conventional CPW and a suspended one (SCPW), respectively. The input pulse in the experiment has a rise time of 58 ps and a FWHM of 52 ps. The CPW structure fabricated has the following dimensions: w = 1.2 mm, s = 2.0 mm, a = 20 mm, h_1 = 0 and h_2 = 5.4 mm. The dielectric substrate is the DI-CLAD 810 produced by Arlon, which is 1.27 mm thick and has a dielectric constant of 10.5. Fig. 7 shows the pulse waveform after traveling 30 mm and 60 mm along the transmission line. The time delay and amplitude of the signals have been normalized for easy comparison of the waveforms. Since the CPW structure is strongly dispersive in the frequency bandwidth of the input signal, a considerable distortion in the pulse waveform is observed as it travels along the line. At a distance of 60 mm, the pulse has a rise time of 113 ps and a FWHM of 96 ps, both nearly doubled compared to those of the input pulse. For a comparison of the experimental results and theoretical predictions, we have plotted the simulated pulse waveform after 60 mm propagation on the same figure. The input signal for computer simulation is identical with the experimental pulse, which was digitized and transferred from the sampling oscilloscope through the GPIB interface. It can be seen from the figure that the rising edges of the simulated and measured pulses are in good agreement, while some discrepancies were found in the falling tails of the two signals. These discrepancies are mainly due to signal reflections at the transient parts between the stripline and coaxial cable, since the characteristic impedance of the fabricated transmission lines is not 50 Ω in exact.

Fig. 8 shows the measured as well as simulated evolution of the same input pulse along a suspended coplanar waveguide. The dimensions and parameters of the SCPW fabricated are as follows: w = 1.2 mm, s = 0.5 mm, a = 20 mm, h = 1.27 mm, $h_1 = h_2$ = 2.7 mm and ϵ_r = 10.5. To properly fix the dielectric substrate and to ensure good electrical contact of the two grounding planes, grooves of 1 mm deep were milled in the inner walls of the enclosure housing. Again all the signals have been

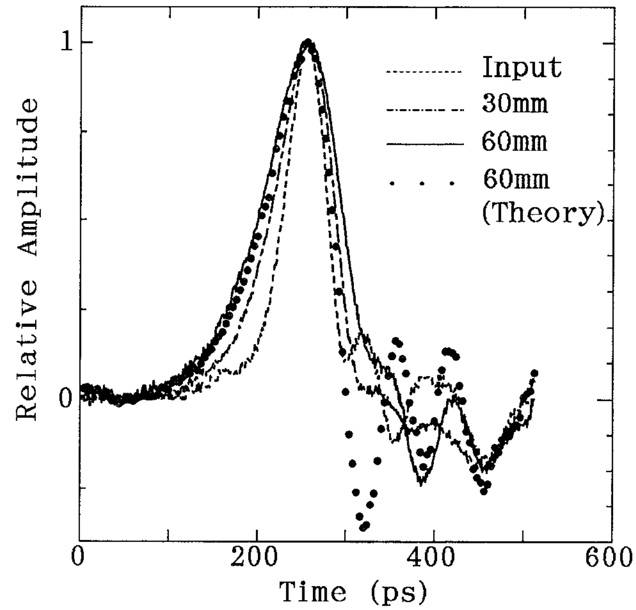

Fig. 7. Experimental (lines) and simulation (dots) results of pulse propagation along a conventional CPW (w = 1.2 mm, s = 2 mm, a = 20 mm, h_1 = 0, h = 1.27 mm, h_2 = 5.4 mm, ϵ_r = 10.5)

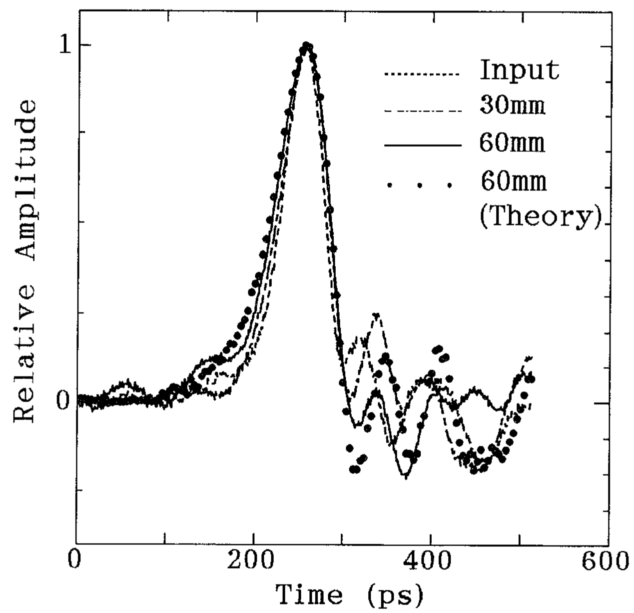

Fig. 8. Experimental (lines) and simulation (dots) results of pulse propagation along a SCPW (w = 1.2 mm, s = 0.5 mm, a = 20 mm, h = 1.27 mm, h_1 = h_2 = 2.7 mm, ϵ_r = 10.5)

normalized so that they lie atop each other. As can be seen in the figure, the rising edge of the pulse changes only slightly throughout the distance of propagation, and the FWHM of the pulse after traveling 60 mm is 65 ps, compared to the 96 pulsewidth for the case of the CPW as shown above. One feature of the measured pulses in Fig. 8 is that there exist some small oscillation rings in front of the main peak, which is not found in the computer simulation results. This is probably due to signal reflections from the bottom side of the SCPW structure, since the signals in the air layer beneath the dielectric substrate travel at a greater velocity than those within the substrate. The signal reflection can be eliminated by placing a microwave-absorbing material on the bottom side of the SCPW, as has been proposed by Paulter *et al.* for a similar problem [23]. As a whole, the measured pulses in Fig. 8 are in good agreement with theoretical predictions, and an appreciable suppression effect of pulse distortion with the SCPW has been confirmed.

V. Conclusions

In this paper we have studied the modal dispersion of coplanar waveguides with various geometries and dimensions, and proposed a method of structurally controlling the dispersion characteristics of these transmission lines. In particular, we show that the suspended coplanar waveguide (SCPW), which is far less dispersive than conventional CPWs on similar dielectric substrates, is a very promising transmission structure for ultrashort pulses with picosecond durations. We have investigated the propagation of picosecond pulses along the SCPW as well as conventional CPWs by both computer simulations and experimental measurements, both revealing a substantial improvement in pulse propagation performance with the proposed SCPW structure.

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