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MULTIMODED REFLECTIVE DELAY LINES AND THEIR APPLICATION TO RESONANT DELAY LINE RF PULSE COMPRESSION SYSTEMS*

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Abstract

Pulse compression systems for future linear colliders, such as NLC and JLC, involve hundreds of kilometers of waveguide runs. These waveguides are highly overmoded to reduce the rf losses. In this paper we present a novel idea for utilizing these waveguides several times by using different modes. This idea is suitble for reflective delay lines. All the modes being used have low-loss characteristics. We describe mechanically simple mode transducers that switch the propagation mode from one configuration to another with no observable dispersion. We apply this technique to a resonant delay line pulse compression system. We also present experimental results that verfy these theortical developmnts.

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1 INTRODUCTION

RF pulse compression systems for future X-band linear colliders [1] contain very long runs of overmoded waveguides. A typical system might contain a few 100 km of circular waveguide under vacuum. Reducing the length of these waveguides by loading them with disks and irises increase the losses of the system. Loading, also, makes the waveguide dispersive, and distorts rf pulse shape. To reduce the lengths of these runs, multimoded RF structures and transmission lines have been suggested [2]. In these multimoded systems, the transmission lines transmit the rf power in several modes utilizing the transmission line several times.

Here we suggest a variation which is suitable for reflective delay lines. These are used in systems such as the resonant delay line pulse compression system, also known as SLED-II [3,4]. The scheme suggested here can reduce the required delay line by factor n where n is the number of modes used simultaneously. We present a design for two modes and show experimental data for

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that scheme. Then we consider the problem of using this technique for a high efficiency high power resonant delay line pulse compression system. We present component designs for an X-band system with a compression ratio of 4 and an output pulse width of 400 ns; and present experimental results for such a system

2 MULTIMODED REFLECTIVE DELAY LINES

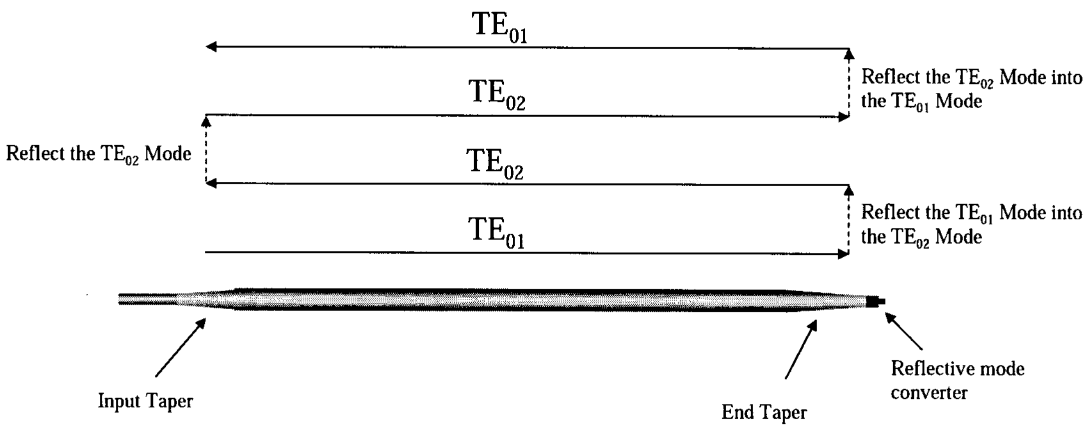


Figure 1: Dual-Moded Delay Line

Consider the delay line shown in Figure 1. The rf signal is injected from the left into the delay line waveguide in the TE_{01} mode. We choose the dimensions of the input port such that only this azimuthally symmetric TE mode can propagate. The waveguide is then tapered up to a diameter that supports several TE_{0n} modes. The TE_{01} mode travels all the way to the end of the delay line and then gets reflected. However, the end reflector also converts the reflected mode into the TE_{02} mode. The TE_{02} mode travels back to the beginning of this line and, since the input of the line cuts off this mode, it is reflected. If the input taper is designed carefully, the TE_{02} mode can be completely reflected without scattering into other modes. Then, because of reciprocity, the TE_{02} wave is converted back to TE_{01} at the other end of the line. This mode then travels back and exits the line.

The total delay in the delay line is twice that seen by a single-moded line. Hence, one can cut the delay-line length by a factor of two.

This scheme can be repeated for more than two modes. For example one can use TE_{01}, TE_{02}, TE_{03}, and TE_{04} for a factor-of-four reduction in length. In this case the end-taper has to both transform the TE_{01} mode, on reflection, to the TE_{02} mode and transform the TE_{03} mode, on reflection, to the TE_{04} mode. The input taper has to transmit the TE_{01}, reflect the TE_{02} into the TE_{03} mode and reflect the TE_{04} mode into itself. The design techniques presented below allow us to produce tapers with these properties. Designs for four-mode systems have been presented by the author[5]. Designs for three-mode systems using adiabatic tapers have been presented by S. Kazakov [6]. However, these designs are still in its initial stages and have no supporting experimental data.

3 THE END MODE CONVERTER

The mode converter at the end of the delay line is shown in Figure 2. It is basically a step in the circular waveguide. If the big waveguide supports only the TE_{01} and the TE_{02} modes among all TE_{0n} modes and the small waveguide supports only the TE_{01} mode, one can choose the diameter of the small guide so that the power transmitted from the large guide to the small guide at the step is independent of which mode is incident. That is, the coupling between the TE_{01} mode in the large waveguide and the TE_{01} in the small waveguide is *equal* in magnitude to the coupling between the TE_{02} in the large waveguide and the TE_{01} in the small waveguide. Then the junction can be viewed as a *symmetrical*, *loss-less*, and *reciprocal* three-port device with two similar ports, namely, port 1 (representing the TE_{01} in the large guide) and port 2 (representing the TE_{02} in the large guide)

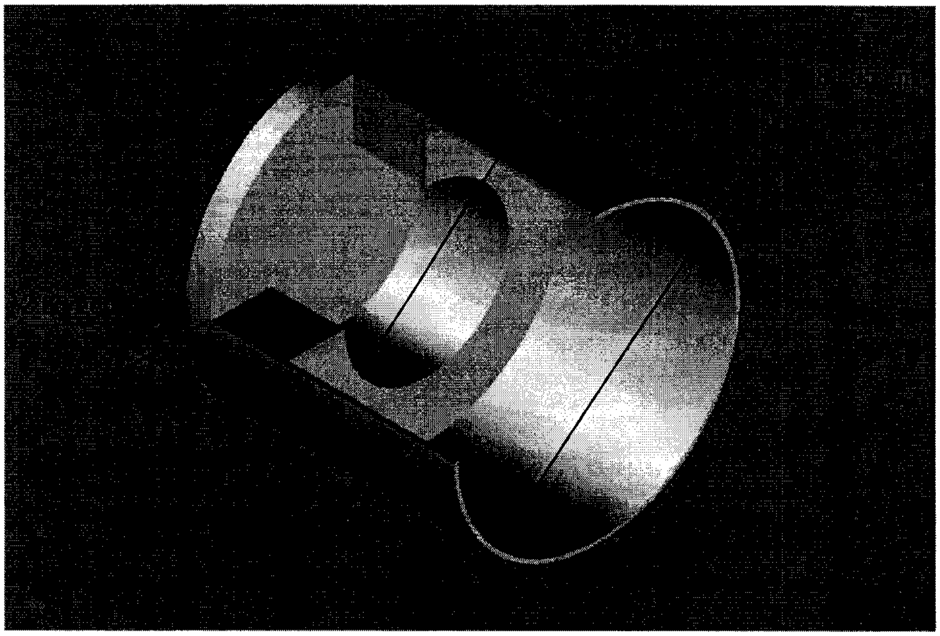


Figure 2: TE_{01}-TE_{02} reflective mode converter. The dimensions shown are for an operating frequency of 11.424 GHz.

The scattering matrix for such a network is unitary and symmetric. By imposing these two constraints on the scattering matrix $\underline{\underline{S}}$ of our device and at the same time taking into account the symmetry between port 1 and port 2; at some reference planes, one can write:

$$\underline{\underline{S}} = \begin{pmatrix} \frac{e^{j\phi} - \cos\theta}{2} & \frac{-e^{j\phi} - \cos\theta}{2} & \frac{\sin\theta}{\sqrt{2}} \\ \frac{-e^{j\phi} - \cos\theta}{2} & \frac{e^{j\phi} - \cos\theta}{2} & \frac{\sin\theta}{\sqrt{2}} \\ \frac{\sin\theta}{\sqrt{2}} & \frac{\sin\theta}{\sqrt{2}} & \cos\theta \end{pmatrix}. \qquad (1)$$

With the proper choice of the reference planes, this expression is quite general for any symmetric three-port network. The scattering matrix properties are determined completely with only two parameters: θ and ϕ. The scattered rf signals $\underline{V}^-$ are related to the incident rf signals $\underline{V}^+$ by

$$\underline{V}^- = \underline{\underline{S}}\,\underline{V}^+; \qquad (2)$$

where $V_i^{\pm}$ represents the incident/reflected rf signal from the i

$$V_3^+ = V_3^- e^{j\psi}. \qquad (3)$$

The resultant, symmetric, two-port network, then, has the following form:

$$\underline{\underline{S}}_{two-port} = \begin{pmatrix} \cos\left(\dfrac{\zeta-\phi}{2}\right)e^{j\left(\frac{\phi}{2}+\frac{\zeta}{2}+\alpha\right)} & j\sin\left(\dfrac{\zeta-\phi}{2}\right)e^{j\left(\frac{\phi}{2}+\frac{\zeta}{2}+\alpha\right)} \\ j\sin\left(\dfrac{\zeta-\phi}{2}\right)e^{j\left(\frac{\phi}{2}+\frac{\zeta}{2}+\alpha\right)} & \cos\left(\dfrac{\zeta-\phi}{2}\right)e^{j\left(\frac{\phi}{2}+\frac{\zeta}{2}+\alpha\right)} \end{pmatrix}, \qquad (4)$$

where the angle ζ is given by

$$e^{j\zeta} = \frac{\cos\theta - e^{j\psi}}{\cos\theta e^{j\psi} - 1}, \qquad (5)$$

and α is an arbitrary angle added to (4) so that the reference planes can be chosen at will.

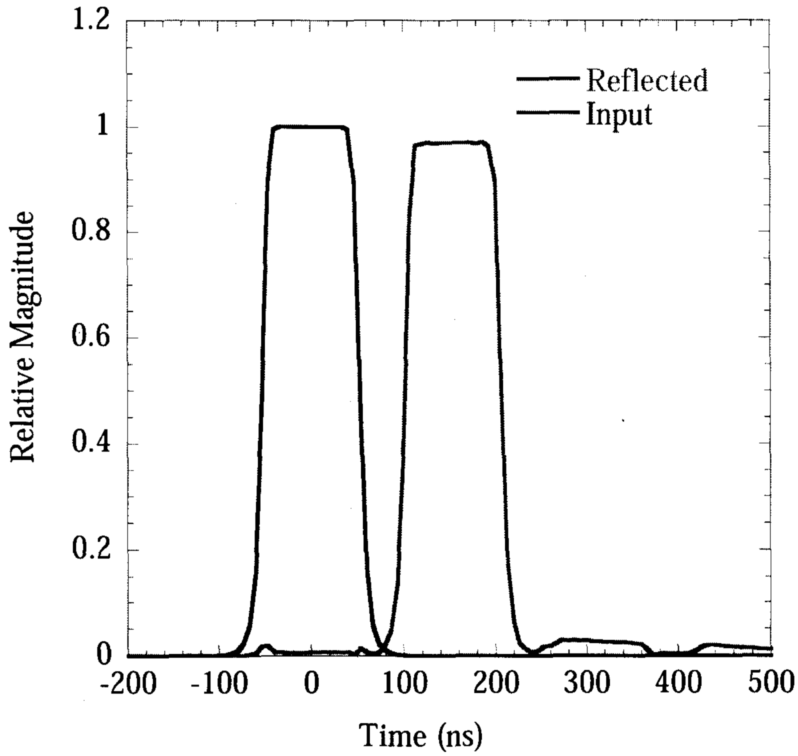
From Eq. (5), changing the angle ψ of the third port terminator can set the angle ζ to any desired value. Hence, from Eq. (4) the coupling between the first and the seconded ports can be varied from 0 to 1.

The design procedure for this reflective mode converter can be summarized as follows: one chooses the relative diameters between the small and large waveguides such that the coupling of both TE_{01} and TE_{02} modes in the large waveguide are equal. Then, according to the above theory, there exists a position for a short circuit termination such that the coupling between the TE_{01} mode and the TE_{02} mode in the large waveguide is equal to 1. Choosing the diameters and then finding the location of the short circuit was done using numerical mode matching codes and verified with finite element code, HFSS [7]

The only step left in the design of this end mode converter is a careful taper design that transforms the diameter of the delay line into the diameter of a waveguide that can support only TE_{01} and TE_{02} modes. The taper needs to transfer both modes perfectly. We used adiabatic tapers for the initial proof of principle experiment. We then used more sophisticated design techniques for the pulse compression design and experiment, see section 4 below.

3 PROOF OF PRINCPLE EXPERIMENT

To prove the concept we used a 12.065 cm diameter waveguide for the delay line. At our operating frequency of 11.424 GHz, this line supports 4 TE_{0n} modes. We used adiabatic tapers for both ends of the line. The design techniques used for these tapers are similar to those used in [8]. With a line length of 35.35 m the delay through the line is about 75 ns. With a short circuit at the end we get a round trip delay of about 150-ns, shown in Fig 3a. The measurement techniques are reported in [3] and summarized in section 4C. Figure 3b shows the delay after placing the mode converter at the end of this line. The delay was doubled at the expense of increased loss. Using larger diameter waveguide for the delay line can reduce the loss.



(a)

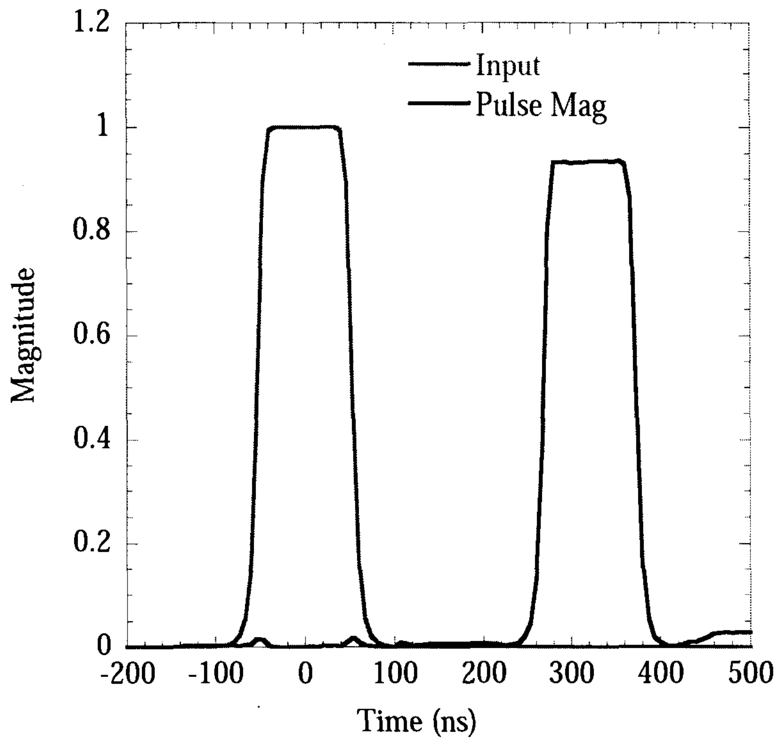


Figure 3: (a) Measured delay through 75 feet of WC475 waveguide terminated with a flat plate. The round trip delay time is 154 ns. (b) Measured delay through 75 feet of WC475 waveguide terminated with the TE_{01}-TE_{02} mode converter. The round trip delay time is 320 ns. The operating frequency is 11.424 GHz.

4 DUAL-MODE RESONANT DELAY LINE PULSE COMPRESSION SYSTEM

A. Resonant Delay Line design.

At the NLC required pulse length of 400 ns [1], the dual moded delay line needs to have an approximate length of 30 m. At this length, using both the TE_{01} and the TE_{02} modes in a single delay line forced us to increase the line diameter to reduce the losses. The diameter of choice for this line is 17.082 cm. In a continuous interval of diameters that have acceptable losses this particular choice makes the waveguide delay line operate as far as possible away from any TE or TM modes that have a cutoff frequency near our operating frequency of 11.424 GHz. This helps reduce losses due to spurious mode excitation.

At this diameter, there are 6 possible TE_{0n} modes that can propagate. Designing adiabatic tapers at this level of overmoding results in tapers that are approximately 1 m long. To make the manufacturing of these taper manageable, we introduced in the next section a new design technique for these tapers. Also the delay line needs to be adjusted in length to bring it into resonance [3]. This means that the mode converter at the end needs to be movable. We had two choices, either to combine the end-taper with mode converter and move the assembly as a whole or to split the reflective mode converter at the plane separating its large waveguide from its small waveguide and only move the later [9]. We choose the later solution. This is shown in Fig. 4. We choose a gap between the movable cup and the waveguide wall of about 0.76 mm. This is possible because TE_{0n} modes do not have axial currents.

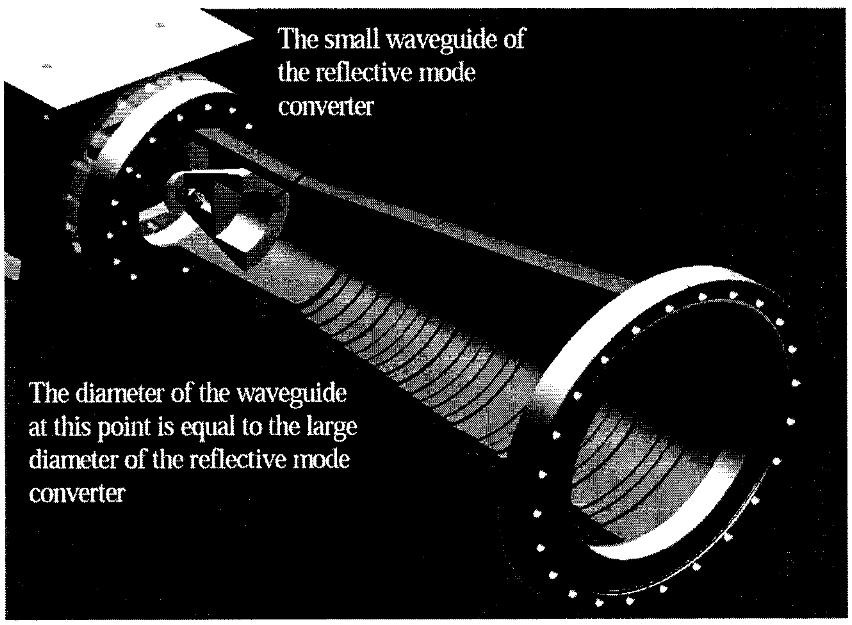


FIGURE 4 End taper and mode converter. The movable cup at the end of the taper reflects the TE_{01} mode into the TE_{02} mode, and vice versa.

We used same design technique for the end taper and input taper. Then, we added an iris at the small diameter of the input taper where the TE_{01} mode is the only propagating mode among all TE_{0n} modes. This iris has a reflection coefficient, for the TE_{01} mode, of 0.607, the optimum iris for a compression ratio of 4 [3].

B. Input and End-Taper Designs

We note that the TE_{0n} modes have neither axial currents nor normal electric fields on the waveguide walls. This allows the design of these tapers in steps using abrupt transitions. We give up completely the idea of adiabaticity. We divided the tapers into several segments and used a mode matching code to simulate their response. For a detailed description of mode matching algorithms the reader is referred to Ref [10]. In circularly symmetric structures TE_{0n} modes do not couple with TM_{0n} modes. This allows us to optimize the speed of our mode matching code. Because of this optimized speed we can vary the parameters of each section and optimize its performance based on a goal function. The goal function used in this code was the multiplication of all the magnitudes of the scattering parameters of interest. The goal function for the input taper is the magnitude of the coupling coefficient between the TE_{01} mode at the input and output of the taper times the magnitude of the coupling coefficient between incident TE_{02} mode at the output port of the taper and the reflected TE_{02} mode at the same port. The resultant shape of the input taper and its computed frequency response are shown in Figs. 5 and 6. The resultant shape of the end taper and its computed frequency response are shown in Fig 7 and 8.

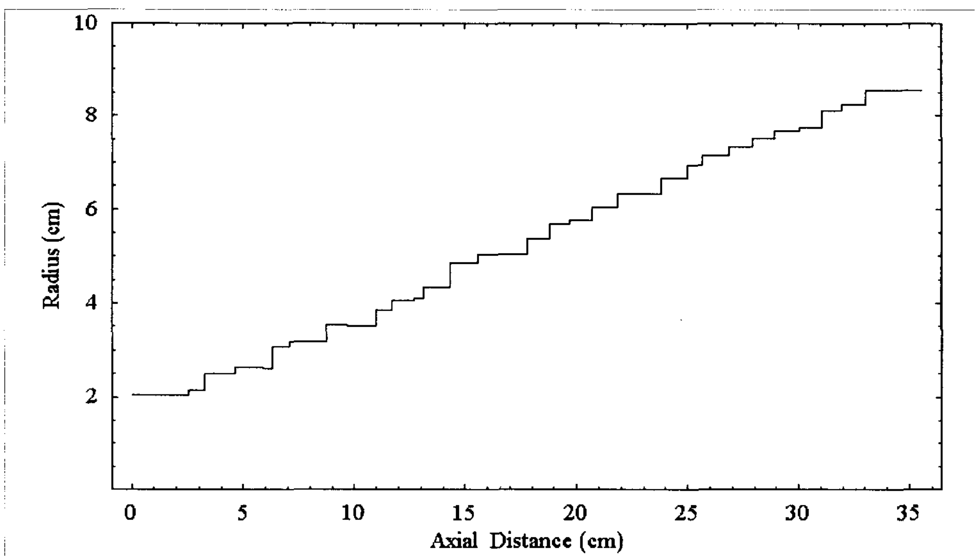


Figure 5 Input taper designs produced by the mode matching optimization code.

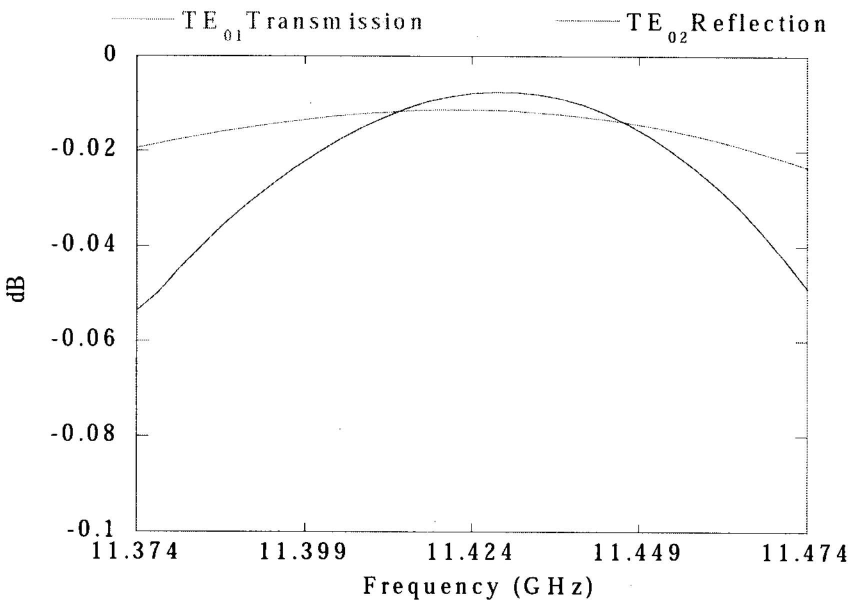


Figure 6. The Simulated frequency response of the input taper

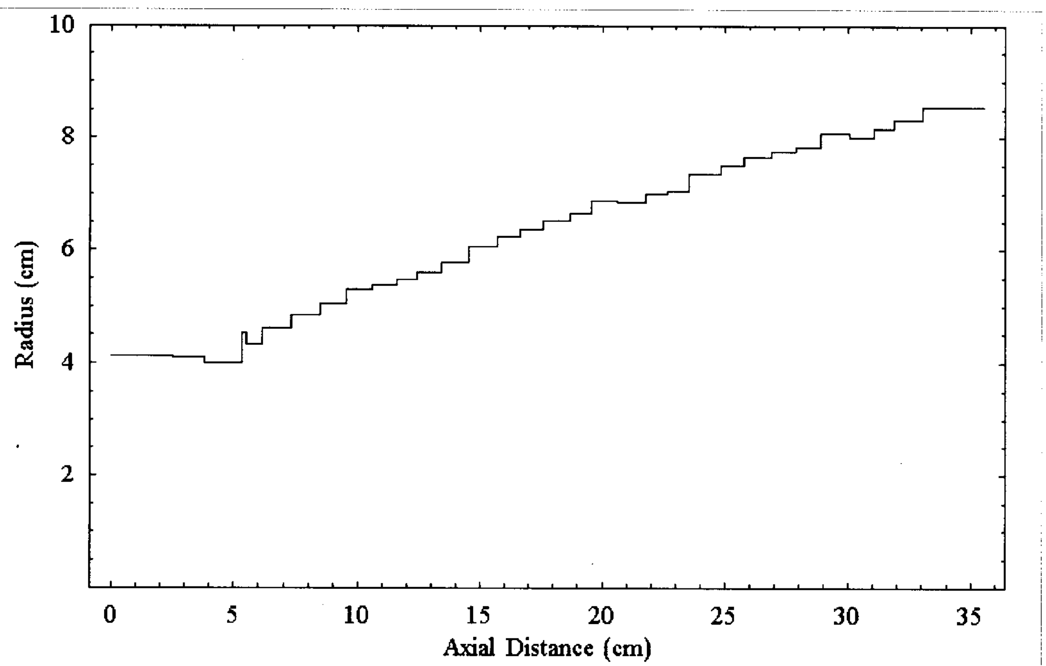


Figure 7 End taper designs produced by the mode matching optimization code.

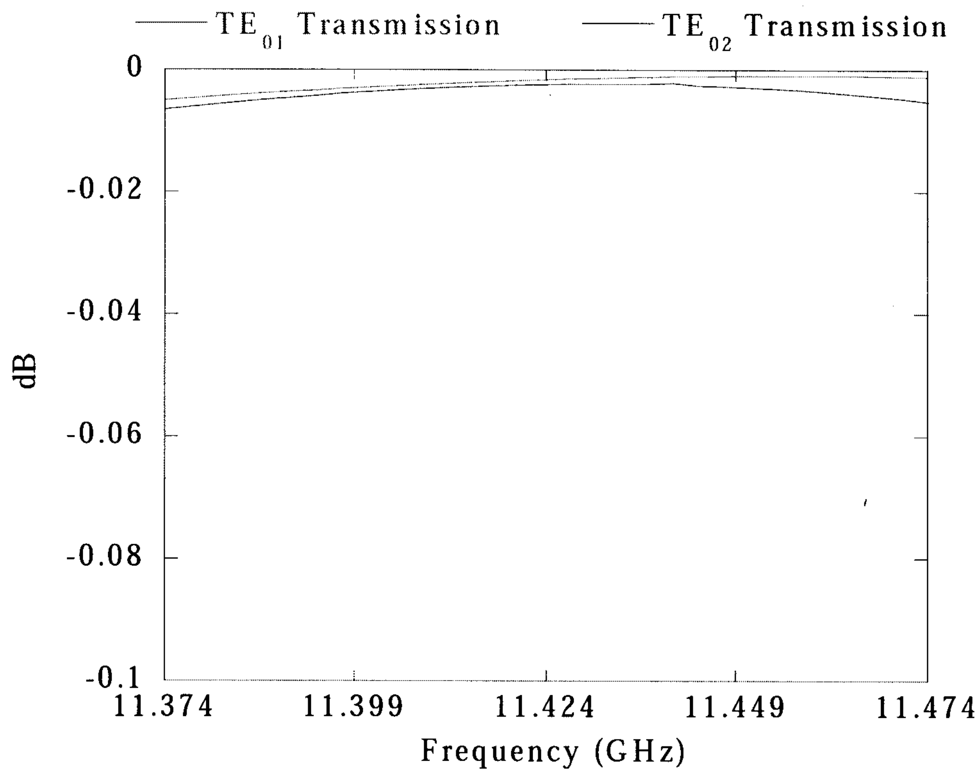


Figure 8. The Simulated frequency response of the end taper

C. Pulse Compression System Response

All measurements were performed using an HP8510C network analyzer with the results examined in the time domain using a PC. The system is shown in Fig. 9. The frequency domain measurements, shown in Fig. 10, were transferred to the PC via a GPIB link and multiplied by the FFT of a maximally flat pulse modulating an 11.424 GHz signal. If we define the compressed pulse width as τ (equal to *two* times the round trip time of the rf through the delay line), the input pulse width should be of the form $n\tau$, where n is an integer equal to the compression ratio. For a given compression ratio, the phase of the input pulse should be reversed 180° at the time $\tau(n\text{-}1)$. In our case the value of τ is chosen to be 400 ns. The test pulse has the following form:

$$V_{in}(t) = \frac{1}{\sqrt{1+\left(\frac{2t}{\tau n}\right)^{2l}}} - \frac{2}{\sqrt{1+\left(2\frac{(t-\frac{\tau(n-1)}{2})}{\tau}\right)^{2k}}} \qquad (6)$$

where l controls the pulse rise time and k controls the phase reversal rise time. The time domain output is produced by taking the IFFT of this frequency domain product. Note that once we obtain the frequency characteristics of the system from the network analyzer, we can calculate the time domain response for any arbitrary input pulse.

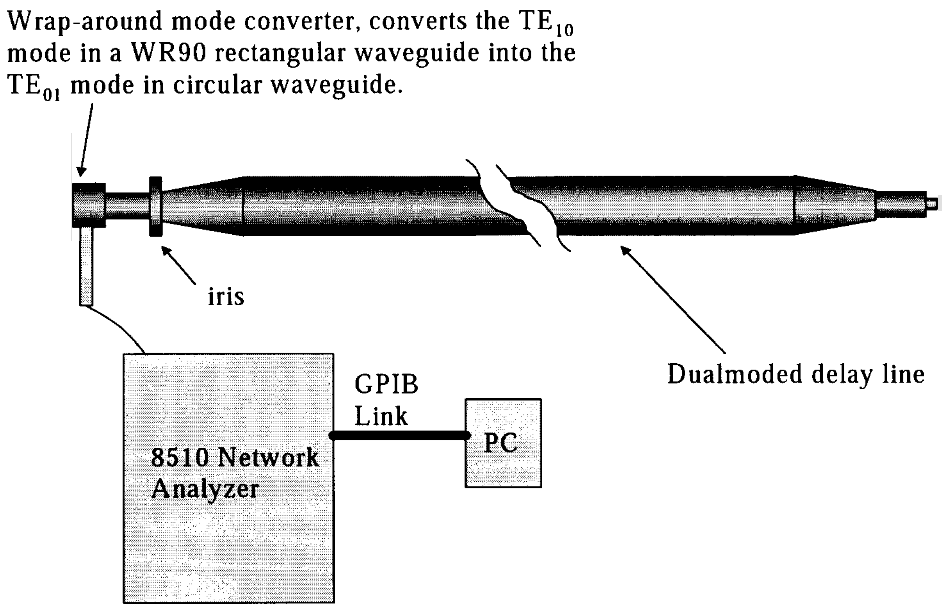


Figure 9 Experimental Setup for Characterizing the Resonant Delay Line Pulse Compression System.

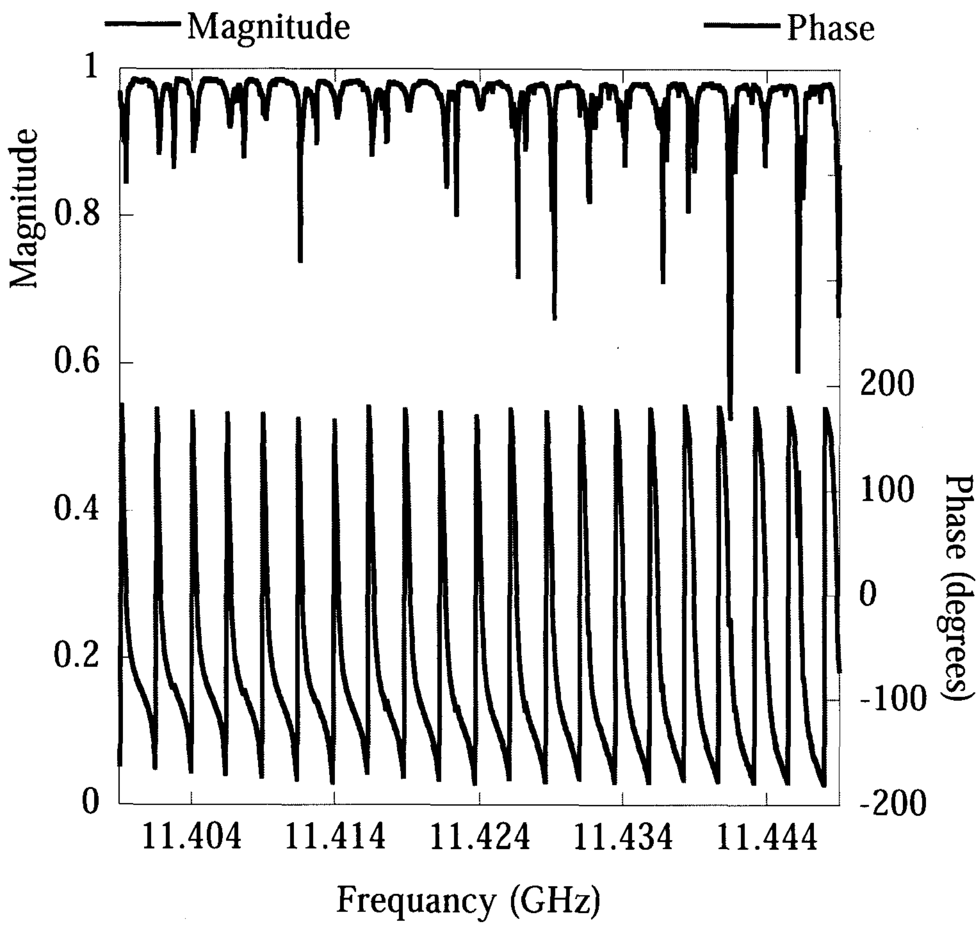


Figure 10 Frequency response of the dualmoded delay line with an iris inserted between the mode launcher and the delay line. The iris has a reflection coefficient of 0.607.

The results of these measurements are shown in Fig. 11. Taking into account the computed response of the tapers, the theoretical losses of the delay line, and the losses due to external mode launchers, the system pulse compression gain of 3.2 is very close to the theoretical estimates [3]. Note that the width of the time bins is governed only by the delay though the line. The 400 ns phase flip happens only at the last time bin. The first three time bins are free to take any value. There width is 400 ns indicating two round trips, one for TE_{01} mode and another for the TE_{02} mode.

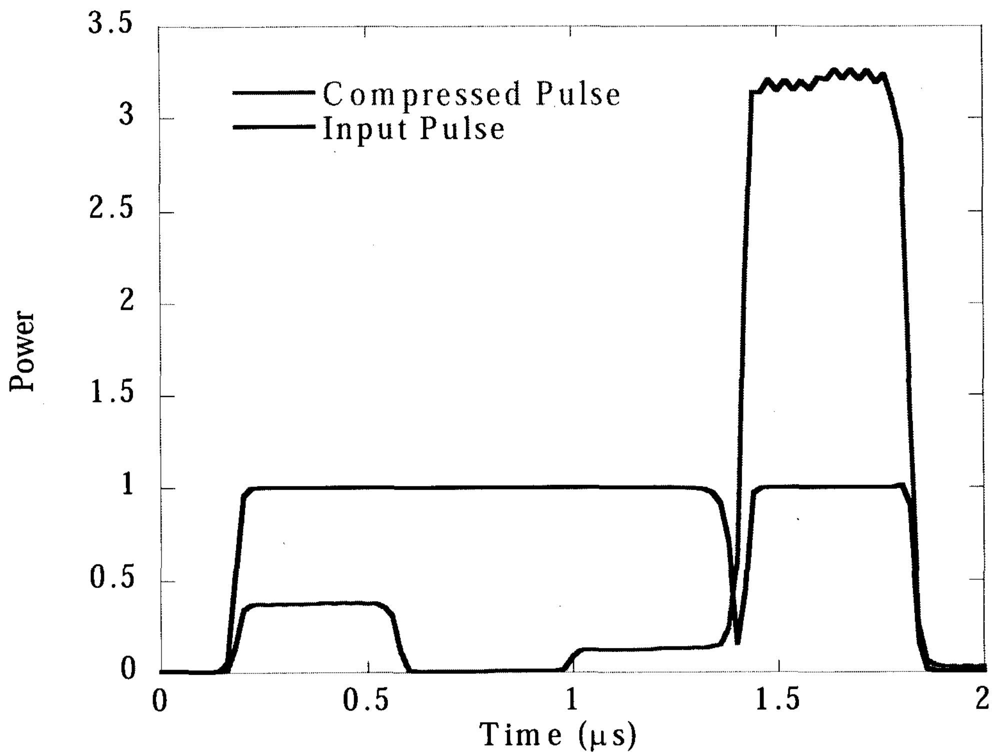


Figure 11. Pulse compression system response.

4 SUMMARY

We have demonstrated a new idea for multimoded reflective delay lines. We designed and built the rf components needed for a dual moded system. Our experimental results agreed well with theory. We also presented designs for components to build a pulse compression system that can operate with this delay lines. The idea of multiple modes need not stop at 2 modes. The difficulty of designing multiple function tapers can be alleviated with the use of modern optimization tools and fast simulation codes.

5 ACKNOWEDGMENT

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