INSTRUMENTATION FOR AN ELECTRON BEAM PLASMA SYSTEM

by

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ABSTRACT

Two systems, for plasma studies, have been constructed and shown to work. The first system deals with determining the effect of a plasma on a pulse propagating through it. The second concerns the determination of the heating of the plasma by that pulse.

For the first system it was necessary to construct a pulse generator and an impedance transition from 50 ohm coaxial cable to 109 ohm shielded twin-lead to couple the pulse into the plasma system. The pulse generator was a charged-line discharged by a mercury wetted reed switch. The risetime of the pulse so generated was 160 pico­seconds. The amplitude was variable from 0-2000 volts. Generators with falltimes of 1-3 nanoseconds have been constructed but by increasing the length of the charged-line the falltime may be increased. The impedance transitions increased the pulse voltage coupled into the plasma with some degradation on the pulse shape.

To determine the heating of the plasma by the pulse, a gridded electrostatic energy analyzer was constructed. Although it has not been tested extensively with a plasma present, preliminary studies indicate that it is functioning properly.
CHAPTER 1

INTRODUCTION

This paper deals with the development of several systems that will be used to help determine the validity of a theory proposed by Capt. William Seidler (1975) of The United States Air Force. This theory deals with the characteristics of propagation of a high amplitude electromagnetic pulse through the D-region of the ionosphere.

To determine the validity of this theory it is necessary to; (1) simulate the D-region in the laboratory, (2) generate high amplitude pulses in the simulated D-region, (3) measure the effect of the simulated D-region on the high amplitude pulses and (4) measure the effect of the pulses on the simulated D-region.

At The University of Arizona, under the direction of Dr. Robert Carlile, an electron beam plasma system has been built which can simulate the D-region under scaled conditions. Due to these scaled conditions it has been determined that the pulses, needed to simulate the electromagnetic pulse in the system, have risetimes of approximately 100 picoseconds, amplitudes of 1000-2000 volts and falltimes of from 1-5 nanoseconds.

The above plasma system consists of a stainless steel vacuum vessel which encloses a 109 ohm parallel plate waveguide used to propagate the electromagnetic pulse in the plasma region of the
vacuum vessel. The waveguide is connected through the vacuum vessel by means of two 109 ohm shielded twin-lead feedthroughs. This system is shown in figure 1.

The systems to be considered in this paper are; (1) the generation of high amplitude pulses in the waveguide, (2) the design of a system to measure the pulses before and after they have traveled through the waveguide and (3) the design of an electrostatic energy analyzer system.

The generation of pulses, with amplitudes and risetimes described previously, poses a significant problem. Under the direction of Dr. Donald Dudley, at The University of Arizona, a pulse generator, utilizing a mercury wetted reed switch, has been constructed with measured risetimes of 400 picoseconds and amplitudes of up to 2000 volts. By using this mercury switch in conjunction with a charged-line (Ramo, Whinnery and Van Duzer, 1965), it was hoped that the required specifications will be realized. The pulse generator will be based on a 50 ohm coaxial system.

Once generated the pulses must be coupled into the plasma region. This requires that the pulses travel from a 50 ohm coaxial system to a 109 ohm shielded twin-lead system. Some loss of the signal transmitted through to the plasma will result due to reflections at the feedthroughs. To minimize these reflections an impedance transition will be used to match the 109 ohm system to the 50 ohm system.

As the pulse travels through the plasma, the plasma will absorb some of the pulse energy. Different frequency components of the pulse
Figure 1. Vacuum Vessel And Waveguide System
will be absorbed more than others. In this way the plasma will act like a complex filter. To determine the effect that the plasma has on the pulse, one need only obtain a display of the pulse before and after it has traveled through the plasma. Any differences will be due to the plasma absorbing the pulse energy.

To view pulses with risetimes in the picoseconds, it is necessary to employ a sampling oscilloscope. A sampling oscilloscope requires that either the system to be observed be pretriggered or the vertical signal to the oscilloscope be delayed by a certain amount after the oscilloscope is triggered. Because of the difficulties in pretriggering the charged-line pulse generator, the latter method will be employed. This will be accomplished by means of a delay line.

A block diagram of the above system is shown in figure 2. The problem with this system is that only one half of the input pulse travels through the waveguide. An alternate system is shown in figure 3. This system allows the full amplitude of the input pulse to travel in the waveguide, but it also necessitates obtaining a display of the input pulse either before the experiment or after it.

The energy absorbed by the plasma will cause a heating of the plasma. This heating may be observed by measuring the electron energy. A method of measuring the electron energy which has been successfully employed is the gridded electrostatic energy analyzer (Porkolab, Arunasalam and Ellis, 1972; Mau, 1974; Mix, Swain, and Chang, 1973). This device will allow the effect of the pulse on the plasma to be displayed.
Figure 2. Transfer Function Measurement System
Figure 3. Alternate Transfer Function Measurement System
In Chapter 2 the basic theory concerning a charged-line pulse generator, impedance transitions and an electrostatic probe is discussed. Chapter 3 contains a description along with constructions details for the charged-line pulse generator, impedance transition, electrostatic probe, and the systems for the measurement of the electron energy distribution and plasma transfer characteristics. Chapter 4 contains the experimental data obtained from the systems discussed in Chapter 3 and compares it with theoretical predictions. Chapter 5 contains suggestions for improvements to the systems.
THEORETICAL DISCUSSIONS

Presented in this chapter are theoretical discussions of charged-line pulse generators, impedance transitions, and gridded electrostatic probes. Since most of the work dealt with here concerns pulses, a discussion of the characterization of the type pulses used in this work will also be presented.

It is hoped that from this chapter an understanding of the design criteria for the various components of the systems will be obtained.

2.1 Transmission Line Equations

In the transmission of electromagnetic energy, when the wavelength becomes comparable with the dimensions of the circuit used to contain that waveform, it is necessary to result to the distributed transmission line equations in describing the propagation of that waveform.

The distributed transmission line equations are:

\[ \frac{\partial V}{\partial z} + RI + \frac{L}{C} \frac{\partial I}{\partial t} = 0 \]  

\[ \frac{\partial I}{\partial z} + GV + \frac{C}{L} \frac{\partial V}{\partial t} = 0 \]

Using the phasor representation of the voltage and current results in a simplification of the above equations since:
where $V'$ is the phasor representation of $V$ and $I'$ is the phasor representation of $I$.

Rewriting the transmission line equations using 2.1-3 and 2.1-4 results in:

\[
\frac{3V'}{\partial t} = j\omega V' \quad (2.1-3)
\]
\[
\frac{3I'}{\partial t} = j\omega I' \quad (2.1-4)
\]

where $V'$ is the phasor representation of $V$ and $I'$ is the phasor representation of $I$.

Rewriting the transmission line equations using 2.1-3 and 2.1-4 results in:

\[
\frac{3V'}{\partial z} + ZI' = 0 \quad (2.1-la)
\]
\[
\frac{3I'}{\partial z} + YV' = 0 \quad (2.1-2a)
\]

where $Z$ is the series impedance and $Y$ is the shunt admittance per unit length of the transmission line.

Equations 2.1-la and 2.1-2a provide a means of obtaining the steady state solution to transmission line problems but their application to transient problems would be very difficult. For this purpose the Laplace transform provides a convenient tool.

Taking the Laplace transform of equations 2.1-1 and 2.1-2 results in:

\[
\frac{3\bar{V}}{\partial z} + (R + Ls)\bar{I} - LI_0 = 0 \quad (2.1-5)
\]
\[
\frac{3\bar{I}}{\partial z} + (G + Cs)\bar{V} - CV_0 = 0 \quad (2.1-6)
\]

where $\bar{V}$ and $\bar{I}$ are the Laplace transform of the voltage and current respectively.
Carl Durney and Curtis Johnson (1969, pp. 373-374) show that after combining the above two equations an intermediate solution may be obtained such that:

$$R^\pm(z,s) = R_0^\pm(s)e^{\mp\gamma(s)z}$$

$$+ CZ_0(s)e^{\mp\gamma(s)}\int_0^z e^{\pm\gamma(s)z}A^\pm(z,s)dz \quad (2.1-7)$$

where

$$Z_0(s) = \sqrt{(R + Ls)/(G + Cs)} \quad (2.1-7a)$$
$$R^+(z,s) = \bar{V} + Z_0(s)\bar{I} \quad (2.1-7b)$$
$$R^-(z,s) = \bar{V} + Z_0(s)\bar{I} \quad (2.1-7c)$$
$$A_0^+(z,s) = V_0 + \frac{L}{CZ_0(s)}I_0 \quad (2.1-7d)$$
$$A_0^-(z,s) = V_0 + \frac{L}{CZ_0(s)}I_0 \quad (2.1-7e)$$
$$\Gamma(s) = \sqrt{(R + Ls)(G + Cs)} \quad (2.1-7f)$$

The term on the right side of the equality in equation 2.1-7 is composed of two terms. The first is the homogeneous solution while the second is the particular solution to the transmission line equations. In some cases, as stated by Durney and Johnson (1969, p. 374), the particular solution might be easier found directly than by the application of the second term on the right side of equation 2.1-7.

From 2.1-7b and c, the Laplace transform of the voltage and current may be found.

$$\bar{V} = \frac{R^+ + R^-}{2} \quad (2.1-8)$$
Since 2.1-8 and 2.1-9 are the Laplace transform of the voltage and current, the voltage and current may be found by taking the inverse Laplace transform of 2.1-8 and 2.1-9 once \( R^+ \) and \( R^- \) are known.

2.2 Charged-Line Pulse Generators

Shown in figure 4 is the basic charged-line pulse generator that will be used to produce the pulses for this paper. The line from \( z = 0 \) to \( z = \ell \) is charged up to some voltage, \( V \). At some time, \( t \), the switch closes causing the voltage, \( V \), to be discharged into the load. Once discharged, the switch is opened and the line is again charged up to the voltage, \( V \), waiting for the cycle to begin again. The exact shape of the waveform produced by the charged-line described above, may be found from the previous transient analysis of the transmission line equations.

The charged-line is assumed to be lossless and to have some characteristic impedance, \( Z_0 \). The load, positioned at \( z = \ell \), has some characteristic impedance, \( Z_1 \), which may or may not be different that \( Z_0 \).

The initial conditions on the charged-line are:

\[
V_0(z) = V \quad \text{(2.2-1)}
\]
\[
I_0(z) = 0 \quad \text{(2.2-2)}
\]
Switch

Load

Charged-Line Of Length $\ell$

And Characteristic Impedance $Z_0$

Load

Impedance

$Z_1$

$z = 0$

$z = \ell$

Figure 4. Model Of A Charged-Line Pulse Generator
This implies that 2.1-7d and e reduce to:

\[ A^+_0 = V \]  \hspace{1cm} (2.2-3a)
\[ A^-_0 = V \]  \hspace{1cm} (2.2-3b)

Since the line is assumed to be lossless, equations 2.1-7a and f reduce to:

\[ Z_0 = \sqrt{L/C} \]  \hspace{1cm} (2.2-4)
\[ \Gamma = \sqrt{LC} \]  \hspace{1cm} (2.2-5)

Substituting the above into equation 2.1-7 results in the intermediate solution for the voltage and current.

\[ \bar{R}^+(z,s) = \frac{\bar{R}_0^+ e^{-\Gamma s}}{s} + \frac{V}{s} \]  \hspace{1cm} (2.2-6)
\[ \bar{R}^-(z,s) = \frac{\bar{R}_0^- e^{\Gamma s}}{s} + \frac{V}{s} \]  \hspace{1cm} (2.2-7)

Using 2.2-6 and 2.2-7 in equation 2.1-8 and 2.1-9 gives the Laplace transform of the voltage and current at the position \( z \).

\[ \bar{V} = \frac{\bar{R}_0^+ e^{-\Gamma z} + \bar{R}_0^- e^{\Gamma z}}{2} + \frac{V}{s} \]  \hspace{1cm} (2.2-8)
\[ \bar{I} = \frac{\bar{R}_0^+ e^{-z} + \bar{R}_0^- e^{z}}{2Z_0} \]  \hspace{1cm} (2.2-9)

At \( z = 0 \), \( I = 0 \), since the line is an open circuit at this point. This first boundary condition may be used to obtain part of the intermediate solution to the voltage and current waveforms.

\[ \bar{R}^+_0 = \bar{R}^-_0 \]  \hspace{1cm} (2.2-10)
At $z = \xi$, $V = IZ_1$. Applying this boundary condition with equations 2.2-9 and 2.2-10 to equation 2.2-8 finalizes the intermediate solution.

$$R^*_0 = \frac{2V}{s} \frac{1}{e^{-\Gamma \xi} \left(1 - \frac{Z_1}{Z_0}\right) + e^{\Gamma \xi} \left(1 + \frac{Z_1}{Z_0}\right)}$$ \hspace{1cm} (2.2-11)

The Laplace transform of the voltage at the load is obtained by substituting the intermediate results into equation 2.2-8.

$$\bar{V} = \frac{1}{s} \left[ 1 - \frac{e^{\Gamma \xi} + e^{-\Gamma \xi}}{e^{-\Gamma \xi} \left(1 - \frac{Z_1}{Z_0}\right) + e^{\Gamma \xi} \left(1 + \frac{Z_1}{Z_0}\right)} \right]$$ \hspace{1cm} (2.2-12)

Expanding this by polynomial long division results in an infinite summation which is easier handled than the original expression.

$$\bar{V}_{z=\xi} = \frac{V}{s} \left[ 1 - \frac{Z_0}{Z_0 + Z_1} - \sum_{n=1}^{\infty} e^{2nk \xi} \frac{Z_0}{Z_0 + Z_1} \left(\frac{1 - \frac{Z_1}{Z_0}}{1 + \frac{Z_1}{Z_0}}\right)^n \left(\frac{1 - \frac{Z_1}{Z_0}}{1 + \frac{Z_1}{Z_0}}\right)^{n-1} \right]$$ \hspace{1cm} (2.2-13)

where

$$k = \sqrt{LC} \text{ (the reciprocal of the phase velocity of a wave in the charged-line)}$$
By taking the inverse Laplace transform of equation 2.2-13, the voltage waveform, generated by the charged-line, at the load is obtained.

\[
V(t) = \left[ 1 - \frac{Z_0}{Z_0 + Z_1} \right] U(t)
\]

\[
- \sum_{n=1}^{\infty} \left( \frac{Z_0}{Z_0 + Z_1} \right)^n \left( \frac{1 - \frac{Z_1}{Z_0}}{1 + \frac{Z_1}{Z_0}} \right) \left( \frac{1 - \frac{Z_1}{Z_0}}{1 + \frac{Z_1}{Z_0}} \right)^{n-1} U(t - 2n/\sqrt{LC})
\]

(2.2-14)

\(U(f)\) is a unit step function beginning at the time when \(f = 0\). The term \(2n/\sqrt{LC}\) in the above is the time it takes for a wave to travel the length of the charged-line \(2n\) times.

It may be concluded from 2.2-14 that the initial value of the wave is \(Z_1/(Z_1 + Z_0)\) which is just what one would expect, the characteristic impedances of the charged-line and load are acting as a voltage divider.

When the switch closes a wave travels to the load and a wave is reflected back toward \(z = 0\). When the wave reaches \(z = 0\) it is reflected back toward the load by the open circuit. Upon reaching the load, part of the wave is transmitted through, subtracting from the voltage at the load, and part is again reflected back toward \(z = 0\). In this way the waveform at the load is stepped down. The rate at which the voltage decreases is therefore determined by the
length of the charged-line and the reflection coefficient, which is
determined by the ratio of the impedances.

Shown in figure 5 are the plots of the voltage for $Z_1 > Z_0$, $Z_1 = Z_0$ and $Z_1 < Z_0$.

It will not be proven here but for $Z_1 >> Z_0$, the above charged-
line pulse generator may be characterized by a discrete capacitor,
charged to a value equal to the voltage on the charged-line, and with
an internal impedance equal to the characteristic impedance of the
pulse generator. The pulse shape may then be characterized by this
capacitor discharging into a resistor whose value is the character-
istic impedance of the load.

2.3 Frequency Spectrum Of The Pulse

The frequency spectrum of the pulse generated by the pulse
generator would be of considerable help in determining the necessary
bandwidth of the equipment. For this purpose it is necessary to
determine a functional relationship for the pulse. After having done
this the spectrum of the pulse may be found by taking the Fourier
transform of the functional relationship.

The pulse generated by the pulse generator discussed pre-
viously, may be characterized by its risetime, falltime and amplitude.
The falltime is obviously exponential since it may be represented by
a capacitor decay. The risetime will be assumed exponential. In
Fourier transforms, the amplitude makes no difference to the frequency
spectrum so it will be assumed unity.
Figure 5. Pulse Generated By A Charged-Line Generator

If the falltime is much greater than the risetime, the pulse may be simulated by the sum of a rising and a falling exponential such that:

\[ f(t) = (e^{-t/t_f} - e^{-t/t_f})u(t) \]  

(2.3-1)

where

- \( t_r \) = the risetime
- \( t_f \) = the falltime

The Fourier transform of some function, \( f(t) \), is given by:

\[ F(\omega) = \int_{-\infty}^{\infty} f(t)e^{-j\omega t}dt \]  

(3.3-2)

The Fourier transform of 2.3-1 results in the spectrum of the pulse generated by the pulse generator.

\[ F(\omega) = \frac{t_f - t_r}{1 + j\omega(t_f + t_r) - \omega^2t_ft_r} \]  

(2.3-3)

The normalized magnitude of \( F(\omega) \) is given by:

\[ \left| \frac{F(\omega)}{t_f - t_r} \right| = \frac{1}{(\omega^2t_f^2 + 1)(\omega^2t_r^2 + 1)} \]  

(2.3-4)

This normalized magnitude of \( F(\omega) \) is shown plotted in figure 6 for \( t_f = 1 \) nanosecond and \( t_r = 100 \) picoseconds. It can be seen from this figure that the pulses are composed of all frequencies from D.C. on up. This implies that all equipment used to propagate the pulses must
Figure 6. Normalized Magnitude Log Plot Of The Frequency Spectrum Of An Exponential Pulse
be very broadband if the pulse shape is not to be altered. The fall-time affects the "frequency breakpoint" A in the diagram while the risetime controls the "frequency breakpoint" B. Increasing the risetime lowers the breakpoint B and increasing the falltime lowers the breakpoint A. Similarly decreasing the risetime or falltime raises the corresponding breakpoint.

2.4 Impedance Transitions

In many applications it is necessary to change from one impedance level to another. If this change is done in only one step, the reflection coefficient will be:

\[
\rho = \frac{Z_1 - Z_0}{Z_1 + Z_0}
\]  

This may become quite large depending upon \(Z_1\) and \(Z_0\). A way to improve this would be to have the impedance change smoothly along some given length. An analysis of any impedance taper is very lengthy and involved. Durney and Johnson (1969, p. 369-371) have analyzed exponentially varying impedance tapers which maintain a constant and an exponentially varying propagation constant, \(\beta\), with \(z\). They found for the constant \(\beta\) case that the reflection coefficient was given by

\[
\rho = \frac{1}{2} \ln \frac{Z_1}{Z_0} \exp(-j\beta_1 l) \frac{\sin \beta_1 l}{\beta_1 l} 
\]  

(2.4-2)
and for the exponential case:

\[
\rho = \frac{j}{4} \ln \left( \frac{Z_1}{Z_0} \exp\left[- \left(2j\beta_1/p\right)(e^{p\ell} - 1) - p\ell\right] - 1 \right) \tag{2.4-3}
\]

where \( p \) is defined by the relationship

\[
\beta = \beta_1 e^{p\ell} \tag{2.4-4}
\]

Equation 2.4-2 is plotted in figure 7. As may be seen, the reflection coefficient is slightly worse at D.C. then the step impedance change but as the frequency increases the tapered line's reflection coefficient decreases greatly. Also it may be noted that the longer the tapered section is the lower the reflection coefficient will be for a given frequency. Equation 2.4-3 resembles equation 2.4-2 and this might lead one to wonder if the reflection coefficient behaves similarly to figure 7. As stated by Durney and Johnson (1969, p. 371) the plot of \( \rho \) versus \( \beta_1 \ell \) for equation 2.4-3 looks very similar to that of figure 7.

For the above reason it appears that a gradual impedance would perform better in the broadband case then would a step impedance change. This is noted by Brown, Sharpe and Hughes (1961, p. 66) when they state that "If this transition takes place over an electrical distance of one-quarter wavelength or greater, then the device (transition) behaves in the fashion of a transformer."
Figure 7. Reflection Coefficient For An Exponential Taper

2.5 Gridded Electrostatic Energy Analyzers

A popular instrument for the measurement of the electron energy distribution function is the gridded electrostatic energy analyzer. One such device is shown in figure 8. For this discussion grids 1, 3 and 5 will be assumed biased at ground potential. Grid 2, which is called the repeller, will be maintained at a positive potential large enough to exclude ions from the analyzer. Grid 4, known as the discriminator, is biased at some variable negative voltage.

Electrons with energy greater than the voltage on the discriminator will reach the collector. Only those electrons with velocity parallel to the axis of the analyzer will reach the collector. By varying the discriminator voltage from 0 to some large value where no electrons are collected, a plot of the relative number of electrons with energy greater than the discriminator voltage versus the discriminator voltage is obtained. Differentiating the collector current with respect to the discriminator voltage, the electron parallel energy distribution function, \( F(\varepsilon) \), is obtained. For the following analysis the axis of the analyzer will be assumed to lie in the \( z \) direction.

Given some electron energy distribution function, \( F(\varepsilon) \), the current collected by the collector is given by

\[
I = Aen\bar{\varepsilon}_z
\]  \hspace{1cm} (2.5-1)

where

\( A = \) the crosssectional area of the analyzer
Figure 8. Gridded Electrostatic Energy Analyzer
\( \bar{v}_z \) = the average electron velocity in the z direction

The average electron velocity in the z direction may be obtained from the electron velocity distribution function.

\[
\bar{v}_z = n^{-1} \int_0^{\infty} F(v_z) v_z \, dv_z
\]  \hspace{1cm} (2.5-2)

By combining equations 2.5-1 and 2.5-2, the current collected by the analyzer in terms of the distribution function may be obtained.

\[
I = Ae \int_0^{\infty} F(v_z) v_z \, dv_z
\]  \hspace{1cm} (2.5-3)

Differentiating the analyzer current with respect to the electron velocity in the z direction gives

\[
\frac{dI}{dv_z} = AeF(v_z) v_z \, dv_z
\]  \hspace{1cm} (2.5-4)

Since the electron energy may be expressed in terms of its mass and velocity by the relationship

\[
eV = \frac{1}{2} m v_z^2
\]  \hspace{1cm} (2.5-5)

the final result is

\[
\frac{dI}{deV} = \frac{Ae}{m} F(v_z)
\]  \hspace{1cm} (2.5-6)
The above result is expressed in terms of the parallel electron velocity distribution function but since the electron's velocity and energy are related by expression 2.5-5 and since we are only concerned with one direction, it may also be expressed in terms of its parallel energy distribution function. Therefore by differentiating the collector current with respect to the discriminator voltage, the electron energy distribution function in the direction of the axis of the analyzer is obtained with some multiplication factor.

With the gridded electrostatic analyzer the ion velocity distribution function is also obtainable. To obtain this, the repeller must be biased at a negative potential to repel electrons while the discriminator is biased at some variable positive potential.
CHAPTER 3

THE EXPERIMENTAL SYSTEMS

This chapter deals with the construction details of a charged-line pulse generator, an impedance transition from 50 ohm coaxial cable to 109 ohm shielded twin-lead and an electrostatic probe. System setups for measuring the transfer characteristics of the plasma and the electron distribution function are also discussed.

3.1 Pulse Generator

It is desired that the pulse generator produce pulses whose shape is as shown in figure 9a. The desired risetime and falltimes are respectively 100 picoseconds and 1-5 nanoseconds. The coaxial charged-line pulse generator shown in figure 9b is capable of meeting these requirements provided that the characteristic impedance of the charged-line is less than the characteristic impedance of the load.

3.1.1 Design Criteria

In the design of the coaxial charged-line pulse generator, shown in figure 9b, there are three factors which must be considered; (1) the risetime, (2) the falltime and (3) the maximum peak amplitude of the pulse.

The risetime is limited almost exclusively by the risetime of the switch and any series inductances associated with that switch. The falltime is determined by the length and phase velocity of the
Figure 9. Pulse Shape Desired And A Proposed Charged-Line Pulse Generator To Produce This Pulse
charged-line, which is related to its characteristic impedance. The maximum peak amplitude is limited by the voltage capabilities of the switch and the ratio of the charged-line and load impedance.

The switch used in the pulse generator will be a mercury wetted reed switch because of its small physical size, no contact bounce and high voltage handling ability. It is dimensioned as shown in figure 10. Switching is accomplished by a solenoid aligned with the axis of the switch. The solenoid is then modulated by a multivibrator circuit oscillating at the required frequency, no higher than the maximum switching frequency of the mercury switch, 200 Hz.

Since the leads of the switch are smaller than the center conductor of the coaxial cable, there will be some series inductance present. The risetime may be minimized by minimizing this series inductance.

\[ L = \frac{\mu}{2\pi} \ln \frac{r_0}{r_i} \]  

(3.1-1)

where

\[ \mu = \text{permeability} \]
\[ r_0 = \text{inner radius of the outer conductor} \]
\[ r_i = \text{outer radius of the inner conductor} \]

Therefore to minimize the series inductance, \( r_0 \) must approach \( r_i \), but their ratio should be such that the characteristic impedance of the switch does not exceed 50 ohms, in order to avoid reflections. A method of accomplishing this is to imbed the switch in the center
Minimum Trim Length = 1.08"

Maximum Lead Diameter = 0.023"

Maximum Diameter = 0.0690"

Figure 10. Dimensions Of The Mercury Wetted Reed Switch
conductor of the charged-line as shown in figure 11. This forms the basis for the charged-line pulse generator used in this experiment. With this construction, the minimum length of the charged-line is essentially fixed at the length of the switch, approximately one inch.

Because the switch can only handle a certain voltage it is desirable to maximize the pulse voltage for a given voltage on the charged-line. Since, as was shown in Chapter 2,

\[ V_p = V_0 \times \frac{Z_1}{Z_1 + Z_0} \]  \hspace{1cm} (3.1-2)

where

\[ V_p = \text{the peak pulse voltage} \]
\[ V_0 = \text{the voltage on the charged-line} \]

and

\[ Z_0 = \sqrt{\frac{\mu}{\varepsilon}} \frac{\ln(r_0/r_1)}{2\pi} \]  \hspace{1cm} (3.1-3)

this requires that the characteristic impedance of the charged-line be minimized.

The falltime is directly related to the length and diameter of the inner conductor.

\[ t_f = 2.2RC \]  \hspace{1cm} (3.1-4)
\[ C = \frac{2\pi \varepsilon l}{\ln(r_0/r_1)} \]  \hspace{1cm} (3.1-5)

where

\[ R = Z_1 + Z_0 \]
Figure 11. Modified Charged-Line Pulse Generator
\[ l = \text{the length of the charged-line} \]

The resistor shown in figure 10 serves to make the source of the high voltage supply to the switch appear as an open circuit. To accomplish this, it must be much larger than \( Z_1 \) or \( Z_0 \). It must also be chosen such that the charged-line has a chance to charge up to its peak value while the switch is open.

With 3.1-2, 3.1-4 and the minimum length of the charged-line in mind, a pulse generator may be built to the required specifications.

3.1.2 Construction

The tube used for the outer conductor shown in figure 11 is made of brass because of its availability and ease of machining. The inner conductor is also of brass for the same reasons in addition to which the resistor and switch are to be soldered to it. The outer radius of the outer conductor is 7/8 inch while its inner radius is 0.745 inch. Type N connections are readily available for this size tubing and it is large enough to allow for reasonable ease of construction. Teflon is chosen for the dielectric for its dielectric strength, low loss tangent, availability and machining ease. The resistor was selected to be 20 megohms. This allows the charging time of the charged-line to be about 20 times less than the frequency of the switch and it, for all practical purposes, allows the source end of the charged-line to appear as an open circuit. The high voltage for the pulse generator is obtained from a type 812 klystron power supply.
Three charged-line pulse generators have been constructed as of the writing of this paper. The first was of the design of figure 9b, the other two were constructed as shown in figure 11. Their construction details, theoretical peak pulse values and falltimes are shown in table 1.

### 3.2 Impedance Transitions

Once the pulses have been generated they must be coupled to the 109 ohm shielded twin-lead feedthroughs. For this purpose it is necessary to construct impedance transitions from 50 ohm coaxial cable to 109 ohm shielded twin-lead.

Shown in figure 12 are the end views of both a coaxial cable and a shielded twin-lead. The characteristic impedance of the coaxial cable is given by 3.1-3. Ramo, Whinnery and Van Duzer (1965, table 8.09) give the characteristic impedance of the shielded twin-lead as:

\[ Z_0 = \sqrt{\frac{\mu}{\epsilon}} \frac{1}{\pi} \left[ \ln \left( 2p \left[ \frac{1 - q^2}{1 + q^2} \right] \right) - \frac{1 + 4p^2}{16p^4} (1 - 4q^2) \right] \]  

(3.2-1)

where

\[ p = \frac{s}{d} \]

\[ q = \frac{s}{D} \]

By the use of equation 3.1-3 and 3.2-1, it is possible to design an impedance transition with a very simple geometry, meeting the desired impedance taper. The design is shown in figure 13.
### Table 1. Summary Of The Constructed Pulse Generator Parameters

<table>
<thead>
<tr>
<th>Generator Number</th>
<th>Design Figure</th>
<th>Measured L (Inches)</th>
<th>Measured D (Inches)</th>
<th>Theoretical Capacitance (pf)</th>
<th>Theoretical Impedance (ohms)</th>
<th>Theoretical Falltime (ns)</th>
<th>Theoretical Amplitude (%V&lt;sub&gt;in&lt;/sub&gt;)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>9(b)</td>
<td>1.375</td>
<td>0.625</td>
<td>22.125</td>
<td>5.265</td>
<td>2.690</td>
<td>90</td>
</tr>
<tr>
<td>2</td>
<td>11</td>
<td>1.187</td>
<td>0.625</td>
<td>19.117</td>
<td>5.265</td>
<td>2.323</td>
<td>90</td>
</tr>
<tr>
<td>3</td>
<td>11</td>
<td>1.062</td>
<td>0.530</td>
<td>8.818</td>
<td>10.21</td>
<td>1.168</td>
<td>83</td>
</tr>
</tbody>
</table>
Figure 12. Cross Sections Of A Shielded Twin-Lead And A Coaxial Cable
Figure 13. Impedance Transition From 50 Ohm Coaxial Cable To 109 Ohm Shielded Twin-Lead
Dimensioned as shown and with a dielectric of dielectric constant 8.0, the impedances at the coaxial end and the shielded twin-lead end are 52 ohms and 102 ohms respectively.

The exact impedance of this taper as a function of its length is a horrendous problem and has not been analyzed. The sole reason for the choice of this particular taper is its simplicity of construction. The length of this taper, as noted in section 2.4, should be as long as possible. As dictated by available funds, this length was 8 inches.

3.3 System For The Determination Of The Plasma Transfer Characteristics

Once the pulses have been generated and coupled into and out of the plasma it is possible to determine the effect of the plasma on the pulse by displaying the input and output pulse simultaneously. Any difference between the pulse shape with and without a plasma present may then be attributed to the plasma. Two systems to obtain the above measurements were shown in figures 2 and 3.

The sampling oscilloscope is a Textronics type 661 with 4S2 and 5T1 plug-ins. This combination require the delay lines to have a minimum delay of 40 nanoseconds. The risetime of the combination is less than or equal to 100 picoseconds. In order for the delay lines to have little or no effect on the displayed pulse, they must have risetimes much less than the risetime of the oscilloscope.

As pointed out by I.A.D. Lewis and F.H. Wells, (1959, p. 45), the risetime of a coaxial cable, given by
\[ t_r = 3.74 \times \frac{\alpha_0^2 \lambda^2}{f_0} \]  

(3.3-1)

where

\[ \lambda = \text{the length of the cable in meters} \]
\[ \alpha_0 = \text{the attenuation constant at } f_0 \text{ in nepers/meter} \]
\[ f_0 = \text{some frequency at the upper end of the frequency band being considered} \]

is "adequately justified as a rough practical measure" of the risetime of the cable.

From figure 6 it may be noted that the frequency spectrum, for the pulse under consideration here, is infinite. At 6 Ghz the frequency content is down to about one percent. For this reason and that coupling into any mode other than the TEM mode is undesirable, it was decided that the delay cable used would have an upper mode lower cutoff frequency above this frequency. This dictated a one half inch semiflexible cable with a foamed dielectric. The upper mode lower cutoff frequency of this cable is 9 Ghz with a published attenuation constant, at this frequency, of about 0.01612 nepers/foot. Since this cable has a velocity propagation constant of 0.81, the minimum total length of cable necessary to provide 40 nanoseconds of delay is about 32 feet. It being better to have too much than too little cable, the length was chosen at 35 feet. The total attenuation of the delay cable is therefore 0.5642 nepers.

The risetime of the delay line, calculated from the above information and equation 3.3-1 is 132 picoseconds. Since the
attenuation quoted above is the worst case attenuation it is hoped that the risetime of the delay line will be better than 132 picoseconds.

It is well known that if the risetime of the measuring instrument is close to the risetime of the waveform being measured, the measured risetime will be longer than the actual risetime. If the risetime of the measuring equipment is known then the actual risetime may be calculated from

\[ t_m = \sqrt{t_a^2 - \frac{1}{2} t_e^2} \]  

(3.3-2)

where

- \( t_m \) = the measured risetime
- \( t_a \) = the actual risetime
- \( t_e \) = the equipment risetime

For the systems shown in figure 2 and 3, the equipment risetimes would consist of the delay cable, impedance generator, pulse generator, power divider, attenuator, waveguide and oscilloscope. It will be assumed that the dominate effects are due to the pulse generator, delay lines, impedance tapers and the oscilloscope. Therefore the risetime of the input pulse is given by:

\[ t_a = \sqrt{t_m^2 - \frac{t_2^2}{2} - \frac{t_3^2}{3}} \]  

(3.3-3)

\[ t_a = \sqrt{t_m^2 - \frac{t_2^2}{2} - \frac{t_3^2}{3} - \frac{2t_4^2}{4}} \]  

(3.3-4)

where

- \( t_2 \) = the risetime of the oscilloscope
\[ t_3 \] = the risetime of the delay lines
\[ t_4 \] = the risetime of the tapers

Using 3.3-3, the actual risetime of the pulse may be calculated. Once this is known, 3.3-4 may be used to calculate the risetime of the tapers.

### 3.4 Electrostatic Probe

A detailed drawing of the electrostatic probe used in this experiment is shown in figure 14. The case and collector are stainless steel since the probe is to be used in a vacuum system. The grids are a commercially available copper hex grid, used in klystrons, with an outer diameter of 0.3125 inch and hole diameters of approximately 0.020 inch. The entrance grid is recessed into the stainless steel case and hard soldered in place. Boron nitride is used to separate the grids for its heat resistance and ease of machining. Grid wires, insulated with teflon, are run through holes in the spacers and out of the vacuum system through a 0.375 inch stainless steel tube. To provide a vacuum tight feedthrough for the wires, Torr Seal was applied to the vacuum side of the tubing. For the collector wire, a small semiflexible copper coaxial cable is used to help prevent stray signal pick-up.

To provide a uniform electric field between the grids, it is required that the spacing be at least twice the grid hole diameter. Also, as pointed out by T.K. Mau (1974, p. 152), the entrance grid hole diameter must be less than twice the Debye length. Debye lengths
Figure 14. Electrostatic Analyzer
encountered in this experiment are roughly of the order of centimeters. Therefore both above conditions are satisfied.

The grid biasing scheme is shown in figure 15. The entrance grid is biased at the local plasma space potential, typically 0-20 volts, in order to eliminate the plasma sheath at the aperture. Secondary emission at the collector is minimized by biasing the collector at a positive voltage (for electron collection). The discriminator and repeller are biased as explained in Chapter 2.

### 3.5 Electron Distribution Function

The biasing scheme shown in figure 15 is adequate for manually obtaining the electron distribution function. This method becomes very tedious if many sets of curves are desired. An alternate scheme, used by L.P. Mix, D.W. Swain and J. Chang (1973, pp. 1703-1708), is shown in figure 16.

Here the audio oscillator provides a small A.C. signal superimposed upon the D.C. discriminator bias. The differential output collector current associated with the A.C. signal is filtered out of the noise by the lock-in amplifier, which provides a D.C. signal proportional to the amplitude of the differential collector current. The sampling oscilloscope provides a means of only looking at those points with a certain time relationship to the trigger signal. In this way, an electron distribution function at any time after the pulse has travelled through the plasma, is obtainable.
Grid Number

1  2  3  4  5  Collector

Repeller Voltage

Discriminator Voltage

0-20 Volts

Collector Voltage

To Oscilloscope

Figure 15. Grid Biasing Scheme
Figure 16. System To Obtain The Distribution Function Directly
There is a certain drawback associated with this system, when used in conjunction with the pulse generator. The audio oscillator must operate much slower than the pulse generator and the ramp generator must operate much slower than the audio oscillator. This constrains the oscillator to frequencies around 20 Hz and the ramp generator to slopes of about 1 volt/second. To obtain a plot of the electron distribution function, with the fastest electrons with energies of 90 eV, would require about 2 minutes. This is still much better than the manual method.
CHAPTER 4

EXPERIMENTAL RESULTS

Before any meaningful data may be taken it is necessary to determine the qualitative effectiveness of the impedance tapers and the risetime of the delay line.

4.1 Impedance Transitions

To determine, qualitatively, if the impedance transitions are beneficial, a Time Domain Reflectometer, TDR, analysis was run on the waveguide system, setup as shown in figure 17a. If instead of the transitions there had been a direct mismatch of 50 ohms to 109 ohms the expected TDR plot would be as shown in figure 17b. The actual TDR plot may be predicted from the reflection coefficient analysis of tapers presented in Chapter 2. Since the tapers have a reflection coefficient that varies with frequency, being higher at low frequencies than high frequencies, the TDR plot would consist of gradual rises from the 50 ohm baseline to the 109 ohm level or lower. If the tapers were infinitely long then the TDR plot would just be that for a 50 ohm system (since the average reflection coefficient for a given frequency band decreases as the length of the taper increases as may be seen from figure 7).

The actual TDR plot for the tapers and waveguide system is shown in figure 17c. As may be seen, there are small sharp reflections with reflection coefficients of about 0.08, and large gradual rises to
Figure 17. TDR Analysis Of The Waveguide System And Impedance Transitions
the 109 ohm baseline, as predicted. This indicates that the reflection coefficient varies, qualitatively, as predicted.

The purpose behind using the tapers is to obtain a higher voltage through the waveguide with as little effect on the shape of the pulse as possible. If the tapers had not been used, only approximately 60 percent of the input pulse would propagate down the waveguide and 60 percent of that, or 36 percent of the input pulse, would propagate out of the system. This would mean that 24 percent of the input pulse would have been reflected back down the waveguide. With the tapers, as will be shown later, the actual pulse amplitude, propagating out of the system, was approximately 67 percent of the input pulse amplitude, much better that the 24 percent for no tapers. This indicates that the tapers are very beneficial.

4.2 Risetime Of The Delay Line

The delay line risetime may be measured with the TDR step output and a sampling oscilloscope. The TDR step output has a risetime of about 50 picoseconds. A sampling oscilloscope with a risetime of 25 picoseconds was obtained in order to measure the risetime of the delay line. The effective risetime of this combination is 56 picoseconds.

The measured risetime of the delay line with the system above is 136 picoseconds. After correcting for the risetimes of the measuring equipment, the actual risetime of the delay line was found to be 125 picoseconds. This compares very favorably to the calculated value of 132 picoseconds.
4.3 Pulse Characteristics Of The Transfer Function Measurement System

The input and output pulses, measured by the system shown in figure 2, for the three pulse generators, constructed with the dimensions given in table 1, are shown in figure 18 and 19 respectively for a voltage on the charged-line of 1000 volts. Table 2 gives a summary of the pertinent information obtained from these figures after correcting for the risetimes of the various equipment by 3.3-3. The output pulse for the third pulse generator has not been obtained due to equipment failure which has not been corrected at the time of the writing of this paper.

The input and output pulse amplitudes as a function of the voltage across the switch are shown in figure 20. As would be expected, the input and output risetimes and falltimes were found to be independent of the applied voltage. From figure 20, it is seen that the peak pulse voltage varies linearly with the voltage across the switch as predicted by 2.2-14. The measured input falltimes agree very closely with the theoretical, any differences may be attributed to measurement errors. The difference in the measured and theoretical $V_0$ peak may be attributed to the delay lines, which as was noted in Chapter 3, attenuate the signal slightly.

For the tapers to be performing their job as planned the shape of the output pulse would be exactly the same as the input pulse. As may be seen by comparing figures 18 and 19, this is not
Figure 18. Input Pulses To The System With A Voltage On The Charged-Line Of 1000 Volts
Figure 19. System Output Pulses For A Voltage On The Charged-Line Of 1000 Volts
Table 2. Summary Of The Pulse Generator Characteristics

<table>
<thead>
<tr>
<th>Generator Number</th>
<th>Theoretical Input Pulse</th>
<th>Measured Input Pulse</th>
<th>Measured Output Pulse</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Falltime (nsec)</td>
<td>$V_0$ Peak (% $V_{in}$)</td>
<td>Risetime (psec)</td>
</tr>
<tr>
<td>1</td>
<td>2.690</td>
<td>90</td>
<td>205</td>
</tr>
<tr>
<td>2</td>
<td>2.323</td>
<td>90</td>
<td>156</td>
</tr>
<tr>
<td>3</td>
<td>1.168</td>
<td>83</td>
<td>156</td>
</tr>
</tbody>
</table>
Figure 20. Peak Pulse Voltage Versus The Voltage Across The Switch
the case. The risetime of the output pulse for all the generators is longer and the falltime for only one of the generators has increased, the generator with the longest falltime. This is not what would have been expected from section 2.4. It would have been expected for the tapers to have increased the falltime more than the risetime. The increased risetime, though, may not be due to the tapers but rather to discontinuities in the waveguide system, possibly where the tapers connect to the shielded twin-lead feedthroughs or where the feedthroughs connect to the waveguide. The increased falltime for the one pulse generator may be due to the fact that its falltime frequency content is such that the tapers affect it more than the other generators.

The peak pulse amplitude of the output pulse is much better than if no taper had been used. From this an "average" reflection coefficient is found to be about 0.18.

If no tapers had been used the pulse traveling in the waveguide would have had an amplitude of 60 percent of the input pulse where as with the tapers the pulse in the waveguide has an approximated amplitude of 82 percent of the input pulse. Since the amplitude of the pulse was deemed to be more important than the risetime and falltime, the tapers are of beneficial value.

4.4 Electron Distribution Function

In the vacuum system, electrons are obtained by means of a hot cathode. The electrons are accelerated through a potential defined by a grid at ground potential and the cathode at some negative potential.
With no plasma present the cathode was biased at a negative 90 volts thereby producing electrons with an energy of 90 eV. If there are no collisions present then the electron energy distribution function would consist of a single vertical line at 90 eV. Due to scattering at the grid, neutral particles and interelectron collisions some spread in the distribution function is expected.

To obtain the ambient electron energy distribution function of the above system, the electrostatic probe bias scheme given in figure 15 is used. The system shown in figure 16 was not used because a lock-in amplifier was not available at the time. Since there are few if any ions with high energies present, the repeller was biased at 20 volts. When the discriminator voltage was varied between 0 and 100 volts, the collector voltage plot shown in figure 21 was obtained. The derivative of this curve is shown in figure 22. This is the electron energy distribution function for the vacuum system with no gas present. Most of the electrons have energies of around 84 eV. The difference between this and the theoretical value of 90 eV is most probably due to errors in the voltage measurement and scattering effects.

No distribution function was obtained with a plasma present due to problems in obtaining a plasma.
Figure 21. Collector Current As A Function Of Discriminator Voltage
Figure 22. Electron Distribution In A Vacuum
CHAPTER 5

CONCLUSIONS

This paper describes two systems, one for the determination of the plasma pulse transfer characteristics and the second for the determination of the electron heating caused by the high amplitude pulse propagating through the plasma.

For both of these systems it was necessary to construct a charged-line pulse generator and impedance transitions from 50 ohm coaxial cable to 109 ohm shielded twin-lead. The pulse generator behaved very close to theoretical predictions and little improvement is possible. The transitions on the other hand, did degrade the pulse shape somewhat more than expected. This was due to the fact that the reflection coefficient varied with frequency and the numerous discontinuities between the coaxial cable and the parallel plate waveguide. Improvements are possible here by eliminating the discontinuities. An improved transition is suggested below.

For the determination of the heating of the plasma it was necessary to construct an electrostatic probe. This probe gave reasonable agreement to the non-plasma electron energy. It has not yet been tested enough with a plasma present to determine its reliability. The method used thus far for obtaining the distribution function was to obtain a number of points for the integral of the distribution
function, plot them and differentiate the resulting plot by hand. An automatic system has been described which would eliminate this procedure. Once this system is employed, obtaining the electron distribution function would be very easy.

5.1 Suggestions For Improvements

The device which could stand the most improvements is the impedance transitions. By eliminating the shielded twin-lead feed-throughs and incorporating the transition in the feedthroughs themselves, it is possible to eliminate some of the discontinuities in the transitions. Shown in figure 23 is a possible substitute to the present impedance transition. As will be noted the shielded twin-lead has been eliminated and the taper takes place inside the coaxial cable by reducing the size of the inner conductor thereby reducing the impedance of the line. Another advantage of this arrangement is that an expensive dielectric is not needed.
Figure 23. Improved Impedance Transition
LIST OF REFERENCES


