1. OBJECTIVES

The primary objective of this work has been to demonstrate (proof-of-concept) the potential utility of Frequency-Modulated Continuous Wave (FM-CW) radars for evaluating refractory furnace wall thickness and regression. Two FM-CW radar systems were assembled (primarily using in-house equipment) for this purpose; namely an S-band (2.6-3.95 GHz) and an X-band (8.2-12.4 GHz) systems. Additionally, during the course of this investigation, it became apparent that other microwave methodologies, such as the time-domain reflectometry approach, may also be successfully used for this purpose. Consequently, a reasonably extensive investigation, using this method was also conducted. An additional goal of this endeavor was to demonstrate the utility of these microwave methods as a function temperature as well. Using a furnace available at UMR campus high temperature measurements were also conducted using these microwave radar and time-domain systems. This report describes the foundation of these techniques as well as the results obtained at room temperature and those at higher temperatures using this furnace. A discussion of the results and the limitations of these techniques along with several recommendations for future enhancement of these methods is also provided.

2. INTRODUCTION

Furnaces are the most crucial components in the glass and metallurgical industry. Like any other components in an industry, furnaces require periodic maintenance and repair. Today, furnaces are being operated at higher temperatures and for longer periods of time thus increasing the rate of wear and tear on the furnace refractory lining. As a result of the competitive market facing these industries, longer furnace lifetime with shorter maintenance downtime are increasingly required. Higher fuel consumption, low production and safety are issues that accompany delayed maintenance. Consequently, there is a need to know the state of a refractory wall to prevent premature or unnecessary maintenance shutdowns.

For many years the observation skills of an experienced operator has been the primary source of evaluating the wear associated with a refractory wall. The rate of regression of a refractory lining depends on the type of the refractory lining, the materials
being melted, seepage, mechanical stresses, and temperature [1]. Moreover, the regression of a refractory lining is also not uniform throughout a furnace and it is more prominent at the metal line along the sidewalls as this region is exposed to hot gaseous byproducts and flowing molten material. Hence, more accurate measurement techniques are required to determine the local residual thickness of a refractory lining so as to utilize the refractory lining to the maximum extent possible. The use of isotope radiators, thermocouples and endoscopes has also been investigated for monitoring regression. These techniques are capable of providing scanned thermal images showing the profile of the refractory wall. However, these techniques can only provide relative profile information and cannot provide absolute thickness measurements [2-5]. A novel laser technique was also studied for monitoring refractory regression and provided accurate thickness measurements [4]. However, the technique requires the use of bulky equipment and is relatively expensive. Another technique that has been used involves the drilling of holes in a refractory wall at specific locations, known to be prone to high rate of erosion, and inserting a rod to measure the refractory thickness. This technique is destructive and expensive [2].

A nondestructive, real-time, reliable, and inexpensive technique which allows for accurate evaluation of refractory lining thickness of a refractory lining is highly desirable. Such a technique will provide for proper scheduling of maintenance. Furthermore, the ability to determine local regression allows for run time maintenance techniques to be administered more effectively. This will significantly help to reduce the cost of furnace operation and provide increased safety [6].

Microwave nondestructive evaluation techniques have evolved over the past few decades [7]. With the availability of less expensive and compact microwave hardware, it is possible to design custom systems for a particular application. The microwave hardware can be designed and built to be relatively inexpensive, rugged, and portable and can be used for real-time and in-situ inspection. Microwave signals are able to penetrate inside of dielectric materials such as plastics and ceramics [7]. The relative (to free-space permittivity, \(\varepsilon_0\)) dielectric constant of a material is a complex parameter. The real part,
\( \varepsilon_r' \), is the relative permittivity which represents the ability of the material to store microwave energy. The imaginary part, \( \varepsilon_r'' \), is the loss factor which represents the ability of the material to absorb microwave energy [8]. Ceramics and refractory bricks (at room temperature) are in the family of low loss dielectric materials. The type of refractory used in a furnace depends on the composition of the material being melted in that furnace. Refractory walls are expected to be low loss materials, and hence low-power microwave signals can be used to monitor their thickness. The use of low-powered microwave energy does not pose any hazard to the operators and does not produce any undesirable byproducts.

As stated in the Objectives Section, it was proposed to investigate the potential utility of Frequency-Modulated Continuous-Wave (FM-CW) radars for refractory wall thickness and recession measurement. This radar technique is a well-established microwave method for distance measurement [9]. With the advancement in RF technology and the availability of less expensive and wider band microwave components, radars have become increasingly utilized for numerous applications.

This investigation was carried out to determine the viability of the microwave FM-CW radar technique to measure refractory wall thickness. Experiments were conducted on several types of refractory bricks, commonly used in furnaces, at room temperature and high temperatures to evaluate the effect, if any, of heat on the microwave measurements. During the course of this investigation, another microwave technique, Time Domain Reflectometry (TDR), was also considered for the purpose of refractory wall thickness measurement. TDR is another well established microwave technique which has been used for many applications including locating faults in transmission lines (i.e., providing relative distance information) [9-11].
2.0 FREQUENCY-MODULATED CONTINUOUS-WAVE RADAR

2.1. Background

A radar is a system which is primarily used for locating and detecting objects by means of detecting reflected radio waves from an object. The first practical radar was developed during World War II to assist in navigation and to detect enemy crafts [9]. Since then, radars have become crucial component not only for navigation and ranging but also other applications such as weather monitoring, remote sensing, ocean wave speed and direction monitoring, monitoring of soil moisture, and geological surveying [9]. Radars operate by transmitting a particular type of electromagnetic wave, and detecting the reflected signal from the target and processing this signal for information such as range, velocity, size and shape [12]. Nowadays, a wide variety of radars are available. Although each type of radar is specifically designed for a particular application, the basic principle remains the same.

Pulsed radar was the first system designed to provide range information [9]. A pulsed radar transmits a stream of pulses and measures the range of a target by monitoring the time of return of reflected pulses from a target. Continuous-Wave (CW) radars on the other hand continuously transmit a low-power microwave signal and require only a few milliwatts of power. The design of a CW radar is comparatively simpler than that of a pulsed radar. However a CW radar is only capable of measuring relative velocity of a moving target and cannot provide any range information without some sort of time or signal triggering (i.e. modulation). To determine range, a timing mark must be applied to the CW signal. This can be achieved either by employing amplitude-modulated (AM) or frequency-modulated (FM).

FM-CW radars are commonly used for remote sensing and range detection. The relatively simple microwave hardware design and the low power levels required for operation, makes the FM-CW technique an attractive method for determining refractory wall thickness. Another important feature of an FM-CW radar is that the radar output is down converted (i.e. mixed) to low frequencies. Hence, a low-cost, low-frequency methods.
spectrum analyzer (or spectrum analysis system, i.e. filter banks) can be used to display the output signal.

For refractory wall thickness measurement, the cold face of the wall and the hot face of the wall are the targets of interest. An FM-CW radar can be designed to be portable and its output can be displayed on a laptop PC using a data acquisition tool such as LabVIEW.

2.2. FM-CW Radar Range Detection

CW radar is not capable of determining range as there is no timing mark present to provide for a time reference [9]. The timing mark helps to determine the time that it takes a transmitted signal to travel to the target and be reflected by the target and return to the radar. This two-way travel time is then used to determine the range of a target (or thickness of a refractory wall). The more distinct the timing mark, the more accurately the two-way travel time can be measured. To have a sharp timing mark, the frequency spectrum of the transmitted signal has to be wider [9]. Frequency modulation can broaden the spectrum of a CW signal. Frequency modulation of the transmitted microwave signal can be achieved by modulating a voltage controlled oscillator (VCO) with a triangular modulating waveform. The minimum and maximum voltage value of the triangular waveform sets the minimum and maximum output frequency, $f_{\text{min}}$ and $f_{\text{max}}$, of the VCO. During half of the cycle of the triangular wave, the VCO output is swept from $f_{\text{min}}$ to $f_{\text{max}}$ and then from $f_{\text{max}}$ to $f_{\text{min}}$ during the next half of the cycle. The bandwidth, $B$, of the resulting FM-CW signal is then given by $f_{\text{max}} - f_{\text{min}}$. When this signal is transmitted, part of it gets reflected back by the target and is received by the radar. The two-way travel time is proportional to the difference in frequency, $\Delta F$, between the transmitted and received signals [13].

Figure 2.1 shows the instantaneous frequencies of the transmitted and received signal for a target at a distance of $R$ in free-space. The received signal is a time-shifted version of the transmitted signal. The time shift is equal to the two-way travel time, $T$. $T$ can be determined from the frequency shift, $\Delta F$. A microwave mixer is a device that can be used
to determine this frequency shift. Figure 2.2 shows an expanded view of a segment of the frequency sweep. When the received signal is mixed with part of the transmitted signal, a signal proportional to the two-way travel time, $\Delta F$, is obtained and Figure 2.3 shows the instantaneous output, $\Delta F$, from the radar.

![Figure 2.1. Instantaneous transmitted and received signals for a target at a distance R.](image)

$\Delta F$ is proportional to the two-way travel time which in turn is proportional to the range of the target [13]. From Figure 2.2 it can be noted that $T$ and $\Delta F$ are related by:

$$\Delta F = \frac{dF}{dt} T$$  \hspace{1cm} (1)

From Figure 2.1:

$$\frac{dF}{dt} = \frac{2B}{T_R}$$  \hspace{1cm} (2)
which can be simplified using Equation (1) to:

$$\Delta F = \frac{2BT}{T_R} \quad (3)$$

Also,

$$T = \frac{R}{v_p} \quad (s) \quad (4)$$

where, $d$ is distance and $v_p$ is the velocity of propagation. The two-way travel time, $T$, is proportional to twice the range, hence to get the range of the target when the wave is propagating in free-space:

$$T = \frac{2R}{c} \quad (s) \quad (5)$$

where, $c$ is the wave (phase) velocity in free-space. As a result:

Figure 2.2. Expanded view of a segment of the frequency sweep.
\[
\Delta F = \frac{4BR}{cT_R}
\]  

(6)

Rearranging this equation to get the range result is:

\[
R = \frac{c\Delta F T_R}{4B}
\]  

(7)

Figure 2.3. Instantaneous difference between the transmitted and received signals (i.e. the mixer output signal from radar).

Also, \(T_R\), the time period of the triangular modulating waveform, can be replaced by \(1/f_m\), where \(f_m\) is the frequency of the modulating triangular waveform. Hence, we finally get:

\[
R = \frac{c\Delta F}{4Bf_m} \quad (m)
\]  

(8)

\(\Delta F, B, c\) and \(f_m\) are all known parameters, and hence the range of the target, \(R\), can be determined. For the purposes of making refractory wall thickness measurements, the
transmitted wave propagates through the wall having a relative dielectric constant of $\varepsilon_r$. The velocity of propagation is dependent on the relative dielectric constant of the material. For low-loss materials, the relative permittivity can be used to determine approximate velocity of propagation, as will be discussed later. Considering propagation in a dielectric medium, the final range or thickness measurement equation become:

$$R = \frac{v_p \Delta F}{4Bf_m} \quad (m) \quad (9)$$

where,

$$v_p = \frac{3 \times 10^8}{\sqrt{\varepsilon_r}} \quad (m/s) \quad (10)$$

For a radar to be capable of detecting or distinguishing between adjacent targets, the system must have a relatively fine range resolution. Range resolution, $RR$, is (theoretically) set by $B$ as:

$$RR = \frac{v_p}{2B} \quad (11)$$

As a result, finer range resolution can be achieved through wider signal bandwidth [13]. It must be noted here that this parameter will be of concern only when one is interested in measuring thin refractory wall thickness (i.e., when a wall has experienced a significant amount of recession). In this case it will be imperative to be able to resolve a reflection from the cold surface (outside) of a wall and the hot surface. The operating bandwidth of a system is highly dependent on the performance characteristics of the microwave components. Also, wideband microwave devices are increasingly available from commercial vendors. To obtain larger operating bandwidth one may also operate at higher frequencies. In this case the percent available bandwidth will be larger resulting in finer range resolution.
2.3. FM-CW Radar Operation

A simple block diagram for an FM-CW radar is shown in Figure 2.4. A triangular wave with a frequency of $f_m$ is used to modulate a microwave oscillator. The output of the microwave oscillator is a swept frequency signal with a bandwidth of $B$. A part of the transmitted signal must be available as the reference signal. The directional coupler is a four-port device which allows a major portion of the FM-CW signal through while a portion of the signal is made available as the reference. The primary signal from the directional coupler is passed through a circulator to a horn antenna while the reference signal is sent to the local oscillator (LO) port of the mixer. A circulator is a three-port device which separates the forward and backward traveling waves thus allowing the use of one antenna for transmitting and receiving a signal. The horn antenna is used to transmit the FM-CW signal and receive the reflected signal from the target. The received signal from the antenna is sent to the mixer through the circulator. The mixer is a three-port device which utilizes a nonlinear solid-state device for detection. The reference signal is applied to the local oscillator (LO) port and the received signal from the circulator is applied to the radio frequency (RF) port of the mixer. When the mixer is fed with two signals that are different in frequency, the output of the mixer, known as the intermediate frequency (IF), comprises of the sum and difference frequencies of the LO and RF signals as well as other intermodulation products [14]. The desired difference frequency component can be obtained by proper filtering. A filter is usually incorporated in the mixer package. The low frequency IF signal is then fed to a spectrum analyzer and the range information is extracted using Equation (9).

A typical output of an FM-CW radar is shown in Figure 2.5, and the measurement setup for determining the thickness of a refractory lining in a glass furnace is shown in Figure 2.6. Two frequency peaks can be noted in Figure 2.5, the first peak is due to the reflection at the interface between the antenna and the cold face of the wall, and the second peak is due to the reflection from the interface between the hot face of the wall and the molten glass.
A reflection occurs whenever there is an impedance (or dielectric) discontinuity in the medium in which the signal travels. A discontinuity is characterized by a change in the dielectric or material properties of the medium of propagation. The greater the change in the dielectric properties the larger the reflected signal will be. When measuring refractory wall thickness, a reflection occurs at the antenna-outer wall interface and also at the interface between the inner wall and the contents of the furnace. For determining the thickness of the wall, the difference in frequency between the antenna-outer wall interface, $F_1$, and the inner wall-furnace contents interface, $F_2$, must be considered. This difference in frequency is given by $\Delta F$ and by using Equation (9) the thickness of the wall can be determined.

Figure 2.4. Basic components of an FM-CW radar.
Figure 2.5. A typical FM-CW radar output spectrum.

Figure 2.6. Measurement setup to monitor refractory lining thickness for a furnace.
3. FM-CW RADAR SYSTEM DESIGN

3.1. Approach
Two FM-CW radars were considered for refractory wall thickness measurement; an X-band (8.2-12.4 GHz) and an S-band (2.6-3.95 GHz) versions. Each has its own set of advantages. In general, the X-band radar has a higher bandwidth and hence higher range resolution. Greater range resolution allows for measurement of thinner refractory walls. Also, the X-band radar system is more compact in size and hence may be more portable. However, if it is required to monitor refractory walls with relatively higher loss tangents, signals at S-band frequencies are expected to experience less attenuation, and are therefore capable of greater penetration depth (i.e. thicker wall inspection).

3.2. S-Band Triangular Wave Modulator
The triangular wave generator is used to modulate the RF source so that microwave oscillator can be swept in the desired frequency band to generate the required FM signal. The frequency of the triangular wave generator is set at the modulation rate of $f_m$. The DC bias and the peak-to-peak voltage of the triangular waveform determine the bandwidth and the maximum and minimum frequencies in the band. The modulation rate, $f_m$, is the rate at which the RF source sweeps through the band.

The triangular wave generator was designed using a Maxim max038CPP high frequency wave generator IC. The IC is capable of generating triangular, sawtooth, sine, square, and pulse waveforms. An external resistor and capacitor can be used to control the output frequency range between 0.1 Hz and 20 MHz [15]. A variable resistor and a fixed capacitor were used as external control components. By adjusting the resistance the modulation rate can be controlled. The output from the wave generator is fed into an amplifier and an offset voltage section. Both sections are designed using the Maxim OP37FP which is a precision operational amplifier. The amplifier controls the peak-to-peak voltages of the triangular wave to generate the required bandwidth while the offset controls the center frequency. Both sections are controlled by variable resistance. The circuit schematic for the triangular wave generator is shown in Figure 3.1.
3.3. S-Band Oscillator

The triangular waveform is fed to a microwave oscillator. The parameters of the modulating waveform set the bandwidth, center frequency, and sweep rate of the FM signal. The microwave oscillator was designed using an in-house Micro Lambda MLMB-0208 YIG (Yttrium Iron and Garnet) tuned oscillator. The YIG oscillator is capable of generating frequencies between 2-8 GHz [16]. By feeding the voltage triangle waveform through a Darlington buffer amplifier, the required current levels were obtained. The output from the buffer amplifier was fed to the TUNE pins of the YIG oscillator.

The schematic of the YIG-tuned oscillator is shown in Figure 3.2. A single ± 15 Volts DC power supply was used to power both the triangular wave generator as well as the YIG oscillator. DC power supply to the YIG was provided through reverse polarity detector circuits to prevent reverse polarity biasing which may damage the YIG. The microwave oscillator section was mounted on a one layer printed circuit board.
The output of the YIG-tuned oscillator or a frequency sweep from 2.78 GHz to 3.6 GHz is shown in Figure 3.3. The oscillator provides a stable output power level of approximately 12 dBm throughout the band, however a significant level of harmonics are also generated.

The harmonics should not be transmitted as they will cause unwanted intermodulation products in the mixer, which may be within the frequency spectrum of interest, and hence cannot be easily removed after mixing. To avoid the propagation of the harmonics of the oscillator through the radar system, a stepped-impedance low-pass filter was designed [14]. The filter was designed to provide maximum attenuation after 5 GHz and a 3-dB cutoff at 4 GHz. To provide a flat response through the passband, a low-pass filter prototype with 0.5-dB passband ripple was chosen to implement the desired filter. The maximum roll-off rate was provided by a 10th order filter [18]. Also, to provide a 3-dB cutoff at 4 GHz it was noted that the filter should be designed for a higher cutoff frequency. By choosing a cutoff at 5.5 GHz, the filter was expected to provide at least 20-dB attenuation at 5 GHz and a 3-dB cutoff point at 4 GHz. The filter was designed using
the element values for a 10\textsuperscript{th} order equal-ripple low pass filter with 0.5-dB passband ripple [19]. Considering the output frequency spectrum of the microwave oscillator, an attenuation greater than 20-dB in the stopband was desired. To provide such a response, two filters were cascaded.

The detailed design of the filter is discussed in Appendix A. The designed low pass filter is shown in Figure 3.4. The frequency response of the low pass filter is shown in Figure 3.5 and the output frequency spectrum of the microwave oscillator section after filtering is shown in Figure 3.6. The filter has an insertion loss of 2-dB in the passband. To avoid undesired reflections, an isolator is placed between the oscillator and the filter. The triangle wave generator, YIG oscillator section, low-pass filter, and an isolator were placed in an enclosure, shown in Figure 3.7 to protect the circuits as well as to provide for easy connections and portability.

Figure 3.3. YIG oscillator output frequency spectrum.
Figure 3.4. $10^{th}$ order low pass filter.

Figure 3.5. Frequency response of two cascaded $10^{th}$ order low-pass filters.
3.4. X-Band Triangular Wave Modulator

For the X-band radar system an in-house HP8350B oscillator was used as the source. To sweep the oscillator to produce the desired FM signal, a modulating waveform must be applied to the sweep in/out port of the HP8350B oscillator. A triangular wave
A modulator similar to that outlined in Section 3.2 can be used to provide the modulating triangular waveform. To reduce the number of hardware components used in the radar, a software-based triangular wave generator was designed in LabVIEW. Apart from reducing the number of hardware components, the software-based wave generator is highly versatile. The amplitude, offset and frequency of the triangular waveform can be adjusted with a high degree of precision. Since the spectrum analyzer was also designed using the same software package, the same interface card can be used to connect to the oscillator. The front panel of the triangular wave generator is shown in Figure 3.8. The (software) design of the triangular wave generator is provided in Appendix B. in Figure B.1.

![Figure 3.8. Triangle wave generator front panel.](image)

### 3.5. Spectrum Analyzer Design in LabVIEW

The output IF signal from the mixer has a low-frequency spectrum which comprises of the difference frequency component between the LO and RF signal. A spectrum analyzer has to be used to display the spectrum of the output IF signal. A regular spectrum analyzer can be used to display the output spectrum; however, to increase mobility, ruggedness and to considerably reduce the cost of the system, a software spectrum analyzer was designed in LabVIEW. The front panel of the Spectrum Analyzer, using LabVIEW, is shown in Figure 3.9. The spectrum analyzer was designed with user-
controllable windowing functions, averaging parameters and filtering function, to enable processing of the data. The number of acquired points per scan can also be controlled by the operator to provide suitable resolution bandwidth for the analyzer display. The spectrum analyzer was also designed to save the acquired data when prompted. Moveable cursors are provided in the display screen to enable the operator to select the frequency points of interest. The spectrum analyzer has a built-in “thickness calculator” which displays the measured thickness in inches and also as a bar graph, by entering the operating parameters of the radar and the frequency points of interest into the calculator. The software designs of the spectrum analyzer and the thickness calculator are provided in Appendix B in Figure B.2 and Figure B.3, respectively.

An X-band oscillator section, similar to the S-band oscillator, can be designed to replace the HP8350B oscillator. This will considerably reduce the cost and also increase the portability of the X-band radar system. However, in this work, as much as possible, available laboratory equipment was used since this was a proof-of-concept investigation and the potential of the methodology was the primary objective of the work. Furthermore,
by using a laptop PC for data acquisition, the size of the overall system can be further reduced. The complete S-band and X-band radar system or shown in Figure 3.10 and Figure 3.11, respectively.

Figure 3.10. S-band radar system for refractory wall thickness measurement.

Figure 3.11. X-band radar system for refractory wall thickness measurement.
4. DIELECTRIC PROPERTY CONSIDERATIONS

4.1. Background

To accurately determine the thickness of the refractory wall, the FM-CW radar technique required an accurate relative permittivity for the refractory bricks. This chapter introduces a simple and approximate technique used in this investigation to determine the relative permittivity of the refractory bricks used. A validation for using the relative permittivity alone and not the loss factor to determine the velocity of propagation mentioned in Chapter 2 is justified and the error in thickness measurements in case of a relative permittivity profile when the refractory bricks are subjected to high temperatures are discussed.

4.2. Measurement of Relative Permittivity

From Equation (10) it is clear that the relative permittivity, $\varepsilon'$, of the material is embedded in the velocity of propagation. Hence, it is necessary to know $\varepsilon'$ of the refractory walls of interest. In this investigation, measurements were conducted on Alumina-Zirconia-Silica (AZS) bricks as they are one of the most widely used refractory material for furnace side walls [20]. Two different types of AZS refractory bricks were used in this study, namely cast AZS refractories and fused-cast AZS refractories. Three different grades of cast (Emhart 315 and Emhart 333 and bonded) AZS bricks were used in this study. The fused-cast AZS refractory bricks have a similar chemical composition as the cast bricks, however they have about 1% porosity. The chemical composition of each of the cast samples are given in Table 1 [21-22]:

The dielectric properties of a material can be accurately measured using several methods including the completely-filled waveguide method [7]. However, this method requires that a the material be precisely machined to fit inside of rectangular waveguide sample-holder. This method was attempted in this investigation. However, precise machining of refractory bricks turned out to be a more challenging task that expected, and the resulting sample did not meet the strict requirements needed for this dielectric measurement method. However, since the refractory brick are expected to be low loss at the frequency bands used in this investigation a different (not exact but one that produces
reasonably accurate results) was implemented. Subsequently, to determine $\varepsilon'$, for the X-band frequencies, the X-band FM-CW radar was used to measure a 15 cm-thick 315 AZS brick. The output spectrum from the radar is shown in Figure 4.1.

Table 1. Chemical composition for Emhart Glass 315 and 333 AZS bricks.

<table>
<thead>
<tr>
<th></th>
<th>$\text{Al}_2\text{O}_3$</th>
<th>$\text{SiO}_2$</th>
<th>$\text{ZrO}_2$</th>
<th>Other</th>
<th>Porosity</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Wt %</td>
<td>Wt %</td>
<td>Wt %</td>
<td>Wt %</td>
<td></td>
</tr>
<tr>
<td>315</td>
<td>69</td>
<td>10.7</td>
<td>20</td>
<td>0.3</td>
<td>17-20%</td>
</tr>
<tr>
<td>333</td>
<td>73.2</td>
<td>15.1</td>
<td>11.1</td>
<td>0.4</td>
<td>17-19%</td>
</tr>
</tbody>
</table>

Figure 4.1. X-band radar output spectrum for 15 cm-thick 315 AZS brick ($B = 1.5$ GHz, $f_m = 100$ Hz).

The parameters of the radar were set as follows:

$$f_m = 100 \quad (Hz)$$
$$B = 1.5 \quad (GHz)$$

From Figure 4.1,
\[ \Delta F = 1266 - 600 \]
\[ \Delta F = 666 \text{ (Hz)} \] (13)

Rearranging Equation (9) and Equation (10) and assuming that the bricks are relatively low loss (at room temperature and at this frequency band) we get:

\[ \varepsilon_r = 4.77 \] (14)

A few similar measurements were carried out and the average relative permittivity was determined to be about 4.35.

This is reasonable method of measuring permittivity for this investigation. However, more elaborate techniques can provide more accurate measurements of relative permittivity and also the loss factor. These techniques include filling a waveguide with a sample of the refractory brick and measuring the reflection and transmission properties [7].

Similar measurements were conducted on the bonded and fused-cast AZS bricks at both S-band and X-band frequencies. Table 2 summarizes the approximate relative dielectric permittivity for the different AZS bricks used in this study.

<table>
<thead>
<tr>
<th></th>
<th>315 AZS</th>
<th>333 AZS</th>
<th>Bonded AZS</th>
<th>Fused-cast AZS</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \varepsilon_r )</td>
<td>S-band</td>
<td>X-band</td>
<td></td>
<td></td>
</tr>
<tr>
<td>( \varepsilon_r )</td>
<td>6.28</td>
<td>4.35</td>
<td>6.6</td>
<td>7.27</td>
</tr>
<tr>
<td></td>
<td>6.98</td>
<td>5.04</td>
<td>4.86</td>
<td>4.95</td>
</tr>
</tbody>
</table>

4.3. Influence of Loss Tangent on the Velocity of Propagation

As mentioned earlier, the dielectric constant is generally a complex parameter. Although ceramics are considered to be low loss materials, the loss tangent affects the velocity of propagation. Loss tangent is the ratio of the loss factor to the permittivity [14]. Loss factor represents the attenuation the material offers to microwave signals. In this
case study, the relative permittivity is only used to determine $v_p$. However, the actual velocity of propagation in a lossy medium is given by [8]:

$$v_p = \frac{\omega}{\beta}$$

(15)

where,

$$\beta = \omega \sqrt{\mu \varepsilon} \left\{ \frac{1}{2} \left[ \sqrt{1 + \left( \frac{\varepsilon'}{\varepsilon_r} \right)^2} + 1 \right] \right\}^2$$

(16)

Equation (15) as a function of loss tangent is shown in Figure 4.2.

From Figure 4.2 it can be concluded that the velocity of propagation is little affected by a low loss tangent. This justifies the use of only the relative permittivity for determining velocity of propagation.
4.4. Error in Thickness Measurement Evaluation for Walls with a Dielectric Profile

In the previous section, an approximate relative permittivity for the AZS bricks at room temperature was determined. However, for certain ceramic or refractory materials, the dielectric properties may be temperature dependent [23-24]. Since the furnace refractory walls are subjected to high temperatures on one side and near room temperatures on the other side, a dielectric profile can be considered to exist along the refractory wall. This section introduces the effect of a relative permittivity profile on refractory wall thickness measurements.

The following general example is used to illustrate the influence of a dielectric profile existing in a refractory wall. Considering a 100 cm-thick refractory wall with a relative permittivity of 4 at room temperature, let’s assume that the wall has a permittivity profile shown in Figure 4.3 when one end is subjected to high temperature. In this case, the velocity of propagation of the signal is dependent on the integral of the dielectric profile. Considering a X-band radar output with $B = 1 \text{ GHz}$ and $f_m = 100 \text{ Hz}$ and rearranging Equation (9):

$$\Delta F = \frac{4RBf_m}{v_p} \quad (\text{Hz})$$

The integral of the dielectric profile is given by:

$$\varepsilon_{r\text{(profile)}} = 4.718$$

Hence,

$$\Delta F \approx 2987 \quad (\text{Hz})$$

When using the room temperature relative permittivity to calculate the thickness with this $\Delta F$, then the total thickness is measured to be 112 cm. This represents an error of 12%. Thus, the results of the measurement are influenced by the dielectric profile and careful consideration of the profile, if any, must be taken to provide accurate refractory thickness measurement.
Figure 4.3. Assumed dielectric profile along the length of the refractory wall due to temperature difference.
5. EXPERIMENTATION AND MEASUREMENT RESULTS

5.1. Approach

The previous chapters introduced the basic concepts of FM-CW radar technique for refractory wall thickness measurements. The importance of dielectric constant and the effects of a dielectric profile on the measurement techniques were briefly discussed. Subsequently, experiments were conducted on a few AZS samples to determine the viability of these microwave techniques for refractory wall thickness measurement. Measurements were conducted at both room temperature and at high temperatures to study the potential effect of temperature on the materials as well as on the measurement techniques. Due to the limited dimensions of the available of refractory bricks, S-band radar measurements could only be conducted on one type of brick, however X-band radar measurements were conducted for a number of bricks and different configurations.

5.2. Room Temperature Measurements

Initial measurements were conducted at room temperature. Measurements were carried out on a 105 cm-thick type 333 AZS brick using the X-band radar system. The output frequency spectrum for \( B = 900 \text{ MHz} \) and \( f_m = 100 \text{ Hz} \) is shown in Figure 5.1.

From Figure 5.1:

\[
\Delta F = 3420 - 560 \quad (\text{Hz})
\]
\[
\Delta F = 2860 \quad (\text{Hz})
\]  

(20)

also,

\[
\nu_p = \frac{3 \times 10^8}{\sqrt{5.04}} \quad (\text{m/s})
\]  

(21)

Now using Equation (9):

\[
R = 106.2 \quad (\text{cm})
\]  

(22)
Figure 5.1. X-band radar output spectrum for the 105 cm-thick 333 AZS brick ($B = 900$ MHz, $f_m = 100$ Hz).

A similar measurement was conducted on a 45 cm-thick type 315 AZS brick and the output frequency spectrum is shown in Figure 5.2. The measured thickness from the data was 46 cm. The X-band radar is also capable of measuring regression. A measurement was conducted with several bonded AZS bricks, each ~7.5 cm-thick, placed one after the other. Measurements were conducted by subsequently reducing the number of bricks. The results of the measurement are shown in Figures 5.3-5.6. From the measured spectra the measured results were 37.8 cm, 22.7 cm, 15.8 cm, and 7.4 cm where the actual thicknesses were 37.5 cm, 22.5 cm, 15 cm, and 7.5 cm respectively. The measured thickness values are very close to the actual values since the permittivity used to calculate the thickness was estimated using one of the bricks used in these experiments. Moreover, these measurements were conducted in the laboratory under relatively ideal conditions. In practice, these measured thickness values are still expected to be close to the actual values. However, in practice multiple measurements need be conducted to obtain an average value for the thickness along with a standard deviation of the measurements indicating the deviation from actual thickness values. From these results the X-band radar technique proves to be a viable technique for refractory wall thickness measurement.
Figure 5.2. X-band radar output spectrum for the 45 cm-thick 315 AZS brick ($B = 900$ MHz, $f_m = 100$ Hz).

Figure 5.3. X-band radar output spectrum for 5 bonded AZS bricks ($B = 900$ MHz, $f_m = 100$ Hz).
Figure 5.4. X-band radar output spectrum for 3 bonded AZS bricks ($B = 900$ MHz, $f_m = 100$ Hz).

Figure 5.5. X-band radar output spectrum for 2 bonded AZS bricks ($B = 900$ MHz, $f_m = 100$ Hz).
5.3. High Temperature Measurements

The FM-CW radar measurement technique provided reasonably accurate thickness measurements at room temperature. Additional experiments were conducted by heating the AZS bricks and measuring the thickness of the heated bricks. Three high temperature experiments were conducted. The first experiment was conducted on a 15.2 cm-thick type 315 AZS brick. One end was placed at the opening of a box furnace. Thermocouples were placed on the hot and cold faces of the AZS bricks to monitor the temperature on these surfaces. The temperature was gradually increased in steps from room temperature to 1130°C so that a thermal equilibrium existed between the hot and cold face of the brick. Measurements were conducted using the X-band radar system. The experimental setup is shown in Figure 5.7. The output spectrum for $B = 1.5 \text{ GHz}$ and $f_m = 100 \text{ Hz}$, over the entire temperature range is shown in Figure 5.8.

From Figure 5.8, the measured thickness was 15.9 cm. Although slight variations are observed from each measurement, the overall thickness remained the same over the entire temperature range. It is important to mention that a fuse in the furnace blew when the temperature reached around 1130°C. Thus, for this particular experiment no more...
measured data was obtained beyond this temperature. From the results shown in Figure 5.8 it is apparent that the signal from the back of the brick (i.e., F2) possesses a signal-to-noise (clutter) ratio of ~5dB. It is also apparent that the back of the brick would have been detected at higher than 1130\(^\circ\)C in this case (the expected internal temperature for furnaces is between 1400\(^\circ\)C-1600\(^\circ\)C).

The same experiment was repeated with two 15.2 cm-thick type 315 AZS bricks placed one after the other. This was done after the fuse was replaced and the furnace temperature could be increased beyond 1130\(^\circ\)C. Due to a slight skew in the setup, a small tapered gap with a width of 0.2 cm at the base was created. The test measurement setup is shown in Figure 5.9.

Measurements were carried out at both S-band and X-band frequencies and the results for the measurements are shown in Figure 5.10 and Figure 5.11 respectively. From the S-band output spectrum, \(B = 1.2\) GHz, \(f_m = 100\) Hz the thickness is measured to be 31.1 cm.

Figure 5.7. Measurement setup for high temperature measurements.
Figure 5.8. X-band radar output spectrum at various hot face temperatures for 15.2 cm-thick type 315 AZS brick ($B = 1.5$ GHz, $f_m = 100$ Hz).

Figure 5.9. Test measurement setup for two 15.2 cm type 315 AZS brick.
Figure 5.10. S-band radar output spectrum at various hot face temperatures for 30.5 cm-thick type 315 AZS brick ($B = 1.2$ GHz, $f_m = 100$ Hz).

Figure 5.11. X-band radar output spectrum at various hot face temperatures for 30.5 cm-thick type 315 AZS brick ($B = 2.5$ GHz, $f_m = 100$ Hz).

From the X-band output spectrum, three peaks were observed, the first being the reflection from the antenna-brick interface, the second peak from the brick-brick gap, and the third from the back end of the brick. With the X-band radar set to $B = 2.5$ GHz, $f_m =$
100 Hz the gap is located at 15.7 cm and the back at 30.7 cm. From these results it was observed that temperature had no effect on the measurement for cast 315 type AZS bricks. The results shown in Figure 5.11 clearly indicate the influence of signal bandwidth on range resolution since the reflection from the gap is also shown in the measured spectra.

The setup for the third high temperature experiment consisted of three 7.6 cm fused-cast AZS bricks placed next to each other. The bricks were placed in the opening of a box furnace. The measurement setup is shown in Figure 5.12. The temperature was increased to 1200°C. The X-band output spectrum for the setup at room temperature is shown in Figure 5.13.

![Experimental setup for measuring the thickness of three 7.62 cm-thick fused-cast AZS bricks.](image)

Figure 5.12. Experimental setup for measuring the thickness of three 7.62 cm-thick fused-cast AZS bricks.
From the output spectrum, only the reflection from the antenna-brick interface is prominent. Although two very small peaks representing the interface between each brick are visible, no reflection from the back-end was observed. This peak was expected to appear at approximately 2700 Hz. The X-band radar output spectrum at various temperatures is shown in Figure 5.14 and Figure 5.15.

As the temperature was increased, a peak appears at around 3000 Hz as seen in Figure 5.14. This peak represents a thickness of 24.8 cm if the room temperature dielectric permittivity is used for calculation. However this feature disappears again as the temperature was further increased as seen in Figure 5.15. Also the peak located at 2200 Hz, clearly seen in Figure 5.13 and Figure 5.14, can be seen vanishing as the temperature was increased from 707°C to 1200°C in Figure 5.15. This is most likely an indication of the fact that the material properties of the bricks are changing as a function of increasing temperature. At high temperatures, the material seemed to significantly absorb the microwave signals, suggesting an increase in the loss tangent of the bricks.

![Figure 5.13. X-band radar output spectrum at room temperature for three fused-cast AZS bricks ($B = 2.5$ GHz, $f_m = 100$ Hz).](image-url)
Figure 5.14. X-band radar output spectrum at hot face temperatures of 395°C to 630°C for three fused-cast AZS bricks ($B = 2.5$ GHz, $f_m = 100$ Hz).

Figure 5.15. X-band radar output spectrum at hot face temperatures of 707°C to 1200°C for three fused-cast AZS bricks ($B = 2.5$ GHz, $f_m = 100$ Hz).

To further investigate this phenomenon, the bricks were heated up to 1200°C and the temperature was maintained until a steady state flow of heat was observed. This was
achieved by monitoring the temperatures at the hot and cold faces of the bricks. Once a thermal equilibrium was observed, the furnace was shutdown and the bricks were allowed to cool down and measurements were conducted while the bricks cooled down. The S-band radar output spectrums at 1200°C to 49°C are shown in Figures 5.16-5.19. From the S-band radar output, no significant information could be observed. It is important to note that the relative size of these bricks at this frequency (i.e., the horn antenna used) was small. Therefore, it is expected that this size limitation could have also contributed to the fact that not much useful information was obtained in this case.

Figure 5.16. S-band radar output spectrum at hot face temperatures of 1200°C to 810°C for three fused-cast AZS bricks ($B = 1.2$ GHz, $f_m = 100$ Hz).
Figure 5.17. S-band radar output spectrum at hot face temperatures of 729°C to 606°C for three fused-cast AZS bricks ($B = 1.2$ GHz, $f_m = 100$ Hz).

Figure 5.18. S-band radar output spectrum at hot face temperatures of 226°C and 140°C for three fused-cast AZS bricks ($B = 1.2$ GHz, $f_m = 100$ Hz).
Figure 5.19. S-band radar output spectrum at hot face temperatures of 87°C and 49°C for three fused-cast AZS bricks ($B = 1.2$ GHz, $f_m = 100$ Hz).

The most likely explanation for this behavior is that the fused-cast bricks have a glass phase while the cast AZS bricks have a crystalline phase. When a glass structure is subjected to high temperatures, the medium provides high loss to microwave energy [25-26]. The digital images of the cast and fused-cast AZS bricks are shown in Figures 5.20-5.23 respectively. The results in Figure 5.15 show that starting at a temperature around 700°C the signal from the back of the bricks is no longer detected. This is very close to the glass transition temperature.

By digital image analysis, the fused-cast AZS was found to contain 20% to 30% glass by volume whereas the cast 315 AZS has negligible amounts of glass. Although this analysis supports the stated theory, this explanation needs to be further verified.
Figure 5.20. Reflected light optical microscopy of Emhart 315 AZS brick.

Figure 5.21. Cathodoluminescence optical microscopy of Emhart 315 AZS brick.
5.4. Dielectric-Backed Room Temperature Measurements

In the measurements conducted thus far, the thickness of the AZS bricks were measured with air at the back-end. However, in a furnace this may not be the case. The hot face of the refractory wall may be in contact with the molten contents of the furnace. Since a reflection of microwave signals occur only when there is a change in dielectric properties in the medium of propagation, these microwave techniques are only effective
in measuring the thickness of the refractory wall when there is a contrast in the dielectric properties of the refractory wall and the molten content of the furnace.

Measurements were conducted with a fused cast AZS brick backed with air, conductor, rubber and wood. Two types of rubber sheet, neoprene and black (carbon loaded), were used. The dielectric constant, at S-band and X-band respectively, for the neoprene rubber sheet was measured to be 2.6 - j0.05 and 2.6-j0.02, that of the black rubber sheet was 6.1-j0.24 and 6.1-j0.18, and that of wood was estimated to be 3.5-j0.05 (this is highly moisture and wood grain direction dependent). The X-band and S-band output spectra are shown in Figure 5.24-5.25 respectively. From these output results it is clearly seen that the reflected signal depends on the contrast of the dielectric constant at the brick-dielectric interface. The conductor-backed and the air-backed measurements provided the greatest reflected signal, as expected.

![Figure 5.24. X-band output spectrum for various brick-dielectric interfaces.](image)

In Figure 5.24, the output spectra for the rubber and wood backed measurements seem to be shifted from those of the conductor and air backed measurements. This is believed to occur as the microwave penetrates the rubber and wood, which were approximately 2 to 3 cm-thick, and the reflections occur at the refractory-rubber (or wood) boundary and back-end of these samples overlapped (i.e., the effect of not being able to individually resolve the two sides of the rubber and wood backings since the range resolution is greater than their thickness of 2-3 cm).
6. DISCUSSION OF RESULTS AND CONSIDERATIONS

The FM-CW radar technique was used to determine the thickness of a number of cast and fused-cast AZS bricks. The technique showed great promise as a viable method for refractory wall thickness measurement. S-band and X-band FM-CW radars were successfully designed to demonstrate their applicability, portability and ruggedness. The technique provided good results with a high degree of accuracy. The measured thickness results using the X–band radar technique show a linear relationship with respect to frequency component $F_2$, shown in Figure 6.1. The figures illustrate that the linear relationship can be utilized to determine the thickness of a refractory wall with known dielectric properties and measured $F_2$. The figure also shows the expected linear relationship for materials with different relative dielectric permittivity.

The FM-CW radar technique provided accurate thickness measurements for the cast AZS brick both at room temperature and at high temperatures. The thickness of the fused-cast AZS bricks, on the other hand, could not be measured at high temperatures. Further investigations have to be conducted on the fused-cast AZS bricks and also on other refractory materials used for lining furnace walls to determine their dielectric properties at high temperatures.

Measurements were also conducted with the AZS bricks subjected to high temperatures. The thickness measurement of the cast AZS bricks was successful in a wide range of temperatures. However, the thickness measurement of the fused-cast AZS bricks was not very successful at much higher temperatures. The properties of the fused-cast AZS samples were observed to change significantly with increasing temperature since it is believed that the fused-cast AZS bricks’ prominent glassy phase significantly contributes to the signal attenuation as the temperature increases. However, even at higher temperatures as the wall gets thinner the signal will be able to sufficiently penetrate inside the brick and provide thickness information. Higher incident power levels may also accomplish the same for thicker walls. It is important to have an idea about the complex dielectric properties of these refractory walls as a function of increasing temperature. This information can help the radar designers to overcome the
problem of the apparent increase in the dielectric losses. However, it is important to note that the evaluation of wall thickness is more important for thinner walls. This method clearly showed its capability for this purpose even for fused-cast AZS bricks. How thin of a wall one can evaluate the thickness of, depends primarily on the signal bandwidth. Operating at X-band or higher frequencies can clearly enable an operator to evaluate wall thickness of a few centimeters very effectively.

Figure 6.1. Relationship between the thickness of the refractory wall and $F_2$ in the output spectrum of the X-band radar when $F_1$ is at 410 Hz ($B = 900$ MHz, $f_m = 100$ Hz).

The knowledge of the dielectric properties of refractory bricks for accurate thickness measurement is important. Additionally, this technique relies on the detection of a reflected signal from the back-end of a refractory wall. Thus, a good contrast in the dielectric properties at this boundary results in a stronger reflection. Subsequently, experiments were conducted to examine the influence of change in the dielectric contrast at this boundary. These results also showed the need for knowing the dielectric properties of molten glass (as a function of temperature). Overall, this radar technique showed great promise for the evaluating the thickness of refractory walls and recession in the wall thickness.
Another issue that must be discussed is the repeatability of the results. Any method used for evaluating refractory wall thickness must provide reliable and repeatable results. This FM-CW technique certainly provided very repeatable results throughout this investigation. The most important parameters that are used for wall thickness evaluation are the frequency locations of $F_1$ and $F_2$ (and their difference). Throughout this investigation multiple measurements of a particular wall was made and in all cases the locations of $F_1$ (and $F_2$) consistently remained constant. Figures 5.8, 5.10, 5.11, 5.14 and 5.15 are very clear examples of this fact.

This FM-CW radar technique provided for many positive features that can be directly and successfully utilized for monitoring refractory wall thickness and recession. However, through the course of this investigation, it also became clear that this method is also capable of providing much more useful information about a wall than just its thickness. The following is a list of other important parameters that may be evaluated using this method.

- **Lamination in a Wall** – If there is a lamination in a refractory wall (i.e., slight separation or gap between adjacent bricks) this FM-CW method can detect the presence and the location of the lamination within the wall. If low frequencies are used, the available signal bandwidth may not be sufficient for this purpose. However, at X-band or higher frequencies, detection of such laminations could be accomplished relatively easily, as shown in Figure 5.11.

- **Planar Cracks** – Planar cracks in a brick manifest themselves as very thin (pancake-like) delamination. Similar to a lamination between two adjacent bricks, the presence and location of this type of cracks are also expected to be detected using this technique.

- **Metal and Glass Penetration** – Intrusion of materials in between adjacent bricks produces impedance mismatch (i.e., similar to a lamination) that can also be detected using this technique. Depending on the radar system properties and the
severity of the penetration, the location and extent of penetration may also be evaluated.

- **Porosity** – Once the dielectric properties of a given brick has been evaluated (using the approximate method discussed here or a more accurate method), through calibration and the knowledge of the dielectric properties of the brick as a function of temperature one may be able to evaluate the porosity (or change in porosity) of the brick on-line, through the development of an appropriate dielectric mixing model.

These additional features make the utility of this method more attractive for overall refractory wall inspection and evaluation (i.e., life-cycle inspection).

One issues that must be further studied is the dielectric properties of refractory bricks as a function of temperature. This knowledge can significantly enhance the optimum design of a radar system in term of the operating frequency, incident power level, etc. Obtaining this information is not a straightforward matter since the element of high temperature significantly limits the number of microwave methods that can be used to evaluate the high temperature dielectric properties of refractory bricks. However, it is not impossible to do so using a box furnace, an appropriate electromagnetic model, a reliable dielectric mixing model and a robust inverse model for recalculating the dielectric properties.

Another issue that may be looked into is the reflection from the antenna-refractory wall interface. As mentioned earlier, reflections occur when there is an impedance (or dielectric) mismatch in the medium of propagation. Since there is a large impedance contrast at this boundary, a significant portion of the signal is reflected. As a result, a smaller part of the signal is transmitted through the wall. Furthermore, this reflection is the closest to the radar receiver (i.e., the mixer) and therefore its magnitude is relatively large due to this fact as well. Consequently, the effective noise floor of the system also increases as shown in Figure 6.2. It must be noted that without any consideration given to this reflection (i.e., efforts to reduce it) the results both at room temperature and at high...
temperature were quite encouraging and did not greatly suffer from the presence of this reflection. Additionally, the presence of this reflection is necessary as a reference point (i.e., the cold face of the brick or $F1$) to calculate $\Delta F$. However, one may still consider reducing the reflection at this boundary. This reflection can be reduced by better matching the antenna impedance to the refractory wall. Matching can be achieved by using matching networks or by filing the horn antenna with a dielectric material [27-28].

Advantages of filling the horn antenna with a dielectric medium are three fold. First, a reasonable degree of impedance matching may be achieved; secondly, the size of the horn antenna may be significantly reduced [28], and thirdly, the spatial resolution is increased. Wavelength ($\lambda$) and frequency ($f$) have a fixed relationship through phase velocity ($v_p$) by:

\[ \lambda = \frac{v_p}{f} \]  

(23)
The relationship between phase velocity in a dielectric medium and that of free-space is given in Equation (10). Due to the fixed relationship shown in Equation (23), the wavelength for a particular frequency in a dielectric medium is shorter than its wavelength in free-space. Thus, by filling the waveguide (or horn antenna) with a dielectric medium, the wavelength of the signal is reduced, and thus a smaller horn can be used resulting in finer spatial resolution. Consequently, a K-band (18-26.5 GHz) horn antenna was filled with titanium oxide (TiO$_2$), with a measured dielectric constant of 5.7-j0.06 at X-band frequencies, in an attempt to reduce the impedance mismatch at the antenna-refractory wall boundary. The filled horn was expected to operate within the X-band frequencies. Figure 6.3 shows the theoretical cutoff frequencies for an X-band, K-band and a filled K-band waveguide. The results clearly show that filling the waveguide causes a reduction in its cutoff frequency and in this case the filled K-band waveguide is capable of operating at X-band frequency range. Subsequently, an experiment was conducted to determine the power received by an X-band horn antenna from an empty and filled K-band horn antenna placed 5 cm apart. The K-band horn was then replaced by a second X-band horn to determine the power received by the receiving X-band horn. The experimental setup is shown in Figure 6.4 and the received signal level as a function of frequency for the TiO$_2$-filled, the empty K-band horn antenna and the X-band horn is shown in Figure 6.5. The results show that the filled K-band horn antenna is capable of operating at X-band frequencies; however the power received from the TiO$_2$-filled K-band horn antenna is almost 30 dB lower than the power received from the X-band horn antenna. This indicates that there are significant “losses” that the signal apparently experienced when traveling through the filled K-band horn. The apparent “loss” could be caused by a number of factors. Although the dielectric properties of the TiO$_2$ powder were measured at X-band, the filling/compacting factor (in the horn) can modify the dielectric properties, as less compaction results in more porosity and a reduction in the permittivity of the powder. A decrease in permittivity then increases the cutoff frequency. This fact is actually some what evident in Figure 6.5, as there is marked reduction in power level below a frequency of 10 GHz.
Figure 6.3. Cutoff frequencies for X-band, K-band and filled K-band waveguides.

Figure 6.4. Experimental setup to determine operating frequency of an empty and filled K-band horn antenna.
The filled K-band horn and the X-band horn were then used to measure the thickness of a 25 cm-thick AZS brick. A measurement was also conducted by removing the antenna and replacing it with a matched load. This was done so as to calibrate the radar system. The results are shown in Figure 6.6. Filling of the horn and its waveguide to coax adapter also changes the electrical properties of the latter component. From Figure 7.5 (which is normalized to the peak of the signal from the filled K-band antenna) it is obvious that the reflection from the filled horn antenna is about 5 dB higher that that of the X-band horn antenna, indicating a higher impedance mismatch at this location. Additionally, the back of the brick is not detected with the filled K-band horn while it is detected with the X-band horn. This may be due to additional dielectric losses in the horn or other factors that must be studied in more detail if this approach is undertaken in the future. Therefore, filling a horn without additional considerations may not be a feasible approach in reducing antenna-refractory interface reflection.
By using the X-band horn the thickness was determined to be 25.3 cm. However, the measurement conducted with the filled K-band antenna was not very successful. The result showed that the reflection at the boundary increased and hence a good impedance matching was not achieved. It is important to note that for the room and high temperature measurements conducted here, this issue (i.e., this reflection) was not a limiting factor since very good thickness results were obtained without any attempt to match the antenna impedance to that of the refractory walls.
7. TIME DOMAIN REFLECTOMETRY

7.1. Background

Although not part of the original objectives of this proof-of-concept project, Time Domain Reflectometry (TDR) which is a well established microwave technique was also investigated for the purpose of evaluating the thickness of refractory walls. This method has been used to detect flaws in underground and overhead cables, power lines and fiber optic lines. The TDR technique determines the location of a defect or an impedance mismatch in a transmission line by monitoring the time it takes for a pulse to travel along the line, reach the defect and return to the end where the pulse was initiated. Reflection of the incident signal occurs whenever there is a discontinuity or change in dielectric properties (or impedance) of the medium of propagation. The TDR technique monitors these reflections in time and with this information the location of the defect can be determined. If the refractory wall is considered as the medium of propagation and the wall faces are considered as the discontinuities, reflections will occur at these boundaries. By monitoring the time at which these reflections occur, the thickness of the wall may be determined. This section presents the TDR technique and the operation of the Agilent 8753E network analyzer, with Time Domain options for conducting TDR measurements, and the potential application of the TDR technique to refractory wall thickness measurement.

7.2. Basics of Time Domain Reflectometry

The TDR technique employs the basic principles of microwave and transmission line theory. The technique depends on the time delay associated with wave propagation along a transmission line, and the reflections of microwave signals due to impedance contrast or mismatch [11]. In this technique a pulse or step function is launched into a cable or a transmission line under consideration. If there is a discontinuity or a change in impedance in the transmission line, a portion of the transmitted signal is reflected back to the source while the remaining portion propagates through or is absorbed by a load. In our case, the change in impedance is characterized by a change in dielectric properties. The magnitude of the reflected signal depends on the mismatch in impedance at the discontinuity. Thus, the larger the impedance mismatch is, the larger the reflected signal will be and vice
versa. The ratio of the reflected wave to the incident wave is known as the reflection coefficient \( \Gamma \) and the ratio of the transmitted wave to the incident wave is known as the transmission coefficient \( T \). The reflection coefficient and the transmission coefficients are given by [11,29]:

\[
| \Gamma | = \frac{Z_1 - Z_0}{Z_1 + Z_0} \quad (24)
\]

\[
| T | = 1 + | \Gamma | \quad (25)
\]

where \( Z_0 \) the impedance of the medium of propagation and \( Z_1 \) is the impedance at the interface.

The TDR monitors the time taken for the reflections, if any, to return to the source. Knowing the propagation velocity of the pulse in the line and the elapsed time between the transmitted and the received signals, the location of the discontinuity can be determined. The time at which the reflected wave is detected is the total time it takes for the wave to travel one way down the line and for its reflection to travel back down the line to the source. The time taken for the wave to travel one-way is called time-of-flight (\( TOF \)). A pulse bounce diagram can be used to analyze the reflected waveforms. The bounce diagram tracks the amplitude of a reflected wave as it reflects off of discontinuities. Consider an ideal lossless transmission line section (i.e. no dispersion and no distortion) of length \( l \) fed with a step input shown in Figure 7.1 with the source impedance \( Z_S \), load impedance of \( Z_L \), line characteristic impedance of \( Z_0 \), and velocity of propagation of \( v_p \). \( TOF \) is given by

\[
TOF = \frac{l}{v_p} \quad (s) \quad (26)
\]

When a step voltage input is applied to the line, the voltage is divided between the source impedance and the lines characteristic impedance and an input \( V_i \) is launched onto
the line. One $TOF$ later, the step input reaches the load impedance and a part of the voltage $V_1T_L$ is transmitted to the load while $V_1\Gamma_L$ is reflected, where $T_L$ and $\Gamma_L$ are the transmission and reflection coefficients at the load end. The reflected voltage arrives back at the source end after another $TOF$ and part of the signal $V_1\Gamma_LT_S$ get transmitted to the source while $V_1\Gamma_L\Gamma_S$ is reflected back into the transmission line. This process continues until the steady state condition is reached at the source and load ends. The entire process can be monitored with a pulse bounce diagram shown in Figure 7.2. The total voltage at the load or the source end will be the sum of the incident wave, reflected wave and the voltage that was already there.

Figure 7.1. Transmission line section fed with a step input.
7.3. Agilent 8753E Vector Network Analyzer

The TDR technique can be applied to refractory wall thickness measurement if the wall itself is considered as the medium of propagation and the inner and outer faces of the wall are thought to be the discontinuities. The Agilent 8753E Vector Network Analyzer (VNA), with option 010, is equipped with time domain capabilities [30]. It is capable of transforming frequency domain data to time domain data. The VNA measures the characteristics of the device under test as a function of frequency and by using an inverse Fourier transform algorithm, the analyzer transforms the frequency domain measurements into the time domain [30]. The Agilent 8753E VNA has three frequency-to-time transform modes, namely band-pass mode, low-pass step mode and low-pass impulse mode. Though the latter two modes provide more accurate measurements, the band-pass mode is selected since waveguides are used in this case to launch the wave into the medium [30].

In the band-pass mode, the network analyzer transmits a signal which is swept in the specified frequency band. This in effect can be considered as transmitting a modulated
Gaussian pulse in the time domain as shown in Figure 7.3. The shape and width of the pulse is dependent on the bandwidth and the specific frequencies of the specified frequency sweep. The pulse generated when operating the network analyzer at S-band frequencies is shown in Figure 7.4. The peak of the modulated Gaussian pulse is at 0 seconds as it is considered as the measurement reference point.

Figure 7.3. Modulated Gaussian pulse stimulus.

Figure 7.4. S-band modulated Gaussian pulse from network analyzer.
Once such a pulse is transmitted into the refractory wall, it in turn is reflected at the boundaries (i.e. the front and back end of the wall) and is finally received by the network analyzer. The measurement setup for conducting such a TDR measurement is shown in Figure 7.5, and a typical TDR output is shown in Figure 7.6.

Figure 7.5. TDR test measurement setup for thickness measurement.

Figure 7.6. Typical TDR output signal.
Considering Figure 7.6, a major reflection is seen at the waveguide-wall interface and another reflection occurs representing the reflection from the back-end of the wall. For refractory wall thickness measurement, Equation (26) is rearranged as follows:

\[ l = TOF \times v_p \quad (m) \]  

However, as mentioned earlier, \( TOF \) represents the two-way travel time of the pulse through the wall, hence, to obtain the actual thickness, Equation (27) becomes:

\[ l = \frac{TOF \times v_p}{2} \quad (m) \]  

To calculate the thickness of a refractory wall, the \( TOF \) considered is the time of the peak of the second pulse, which in this case is 3.7 ns. Assuming the relative permittivity of the wall to be 5, and from Equation (28) we can get:

\[ l = \frac{3.7 \times 10^{-9} \times 3 \times 10^8}{2 \sqrt{5}} \quad (m) \]

\[ l = 0.24 \quad (m) \]  

Thus, the thickness of the refractory wall can be calculated by knowing it relative permittivity and the \( TOF \) obtained from the TDR output. Although the network analyzer introduced in this section is expensive and not suitable for industrial environments, a portable and rugged TDR system can conceivably be designed and built. Similar to the FM-CW radar output spectrum, distinct reflections are observed from the cold and hot face of the refractory wall, hence contact or non-contact measurements can be carried out.
8. TDR EXPERIMENTATION AND MEASUREMENT RESULTS

8.1. Background

The previous sections introduced the basic concepts of the TDR measurement technique for refractory wall thickness measurements. Subsequently, experiments were conducted on a few AZS samples to determine the viability (proof-of-concept) of the microwave TDR technique for refractory wall thickness measurement. Measurements were conducted at both room temperature and at high temperatures to study the potential effect of temperature on the materials as well as on the measurement techniques.

8.2. TDR ROOM TEMPERATURE MEASUREMENTS

Initial measurements were conducted at room temperature. Measurements were carried on a 105 cm-thick type 333 AZS brick using the TDR. The S-band TDR output for a measurement conducted on a 105 cm-thick type 333 AZS brick is shown in Figure 8.1.

From Figure 8.1:

\[ TOF = 18.7 \text{ (ns)} \]
\[ \varepsilon_r = 6.28 \]
\[ v_p = \frac{3 \times 10^8}{\sqrt{6.98}} \]
Using Equation (28):

\[ l = 106.2 \text{ (cm)} \]  

(32)

A similar measurement was conducted on a 45 cm-thick type 315 AZS brick. The TDR output signal is shown in Figure 8.2. From Figure 8.2, the \textit{TOF} was determined to be 7.7 ns and the thickness was calculated to be 46 cm.

The TDR technique is also capable of detecting a reduction in thickness. The same experiment mentioned earlier for the X-band FM-CW radar, using several bricks, was repeated for the S-band TDR. The measurement results are shown in Figure 8.3. From the graphs the measured results were 21.6 cm, 30.9 cm, 37.9 cm, and 53.7 cm whereas the actually thicknesses were 22.5 cm, 30 cm, 37.5 cm, and 52.5 cm, respectively.

Figure 8.1. S-band TDR output signal for a 105 cm-thick type 315 AZS brick.
From the TDR outputs it was observed that the reflected pulse had broadened. This is thought to be due to the reflections from the sides of the bricks adding to the reflection from the back-end. Also, in the first two plots in Figure 8.3, multiple reflections can be
clearly noted. The third plot also shows a similar trend however, the pulse goes out of range. The results from the TDR measurements at room temperature proved to be promising for determining refractory wall thickness measurement.

8.3. High Temperature TDR Measurements

Additional experiments were conducted by heating the AZS bricks and measuring the thickness of the heated bricks. Due to certain technical difficulties, high temperature TDR measurements were only conducted for the fused-cast AZS bricks.

The high temperature experiment consisted of three ~7.5 cm fused-cast AZS bricks placed next to each other in the opening of a box furnace. The temperature was increased to 1200°C. The TDR output for this setup at room temperature is shown in Figure 8.4. From the TDR output a thickness of 23.4 cm was measured whereas the total thickness was 22.8 cm. It was observed that the signal level in the TDR output was comparatively lower than when measuring the cast AZS samples. This suggests that the fused cast AZS bricks were lossier than the cast AZS bricks.

Figure 8.4. TDR output signal at room temperature for three fused-cast AZS bricks.
The S-band TDR measurement outputs for various temperatures are shown in Figure 8.5-9.8. As temperature increased, the TDR output signal deteriorated. A peak was noted at 3 ns, which signifies a thickness of 16.7 cm the room temperature dielectric permittivity is used. This peak may represent the reflection at the boundary between the second and third fused-cast AZS brick which is located 15.2 cm from the front end. Further, the signal level reduces to such an extent, that the material seemed to significantly absorb the microwave signals, suggesting an increase in the loss tangent of the material.

![Figure 8.5. TDR output signal at hot face temperature of 395°C for three fused-cast AZS bricks.](image)

Figure 8.5. TDR output signal at hot face temperature of 395°C for three fused-cast AZS bricks.
Figure 8.6. TDR output signal at hot face temperature of 971°C for three fused-cast AZS bricks.

Figure 8.7. TDR output signal at hot face temperature of 1090°C for fused-cast AZS bricks.
To further investigate this phenomenon, the bricks were heated up to 1200°C and the temperature was maintained until a thermal equilibrium was attained. This was achieved by monitoring the temperatures at the hot and cold faces of the bricks. Once a thermal equilibrium was observed, the furnace was shutdown and the bricks were allowed to cool down and measurements were conducted while the bricks cooled down. The temperatures at the hot and cold faces, during each measurement, were also recorded, as shown in Table 3.

Table 3. Temperatures at the hot and cold faces of the AZS bricks.

<table>
<thead>
<tr>
<th></th>
<th>Temperature (°C)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Hot face</td>
<td>1200 1068 968 810 729 652 606 226 140 87 47</td>
</tr>
<tr>
<td>Cold face</td>
<td>238 238 238 237 236 234 210 100 86 51 44</td>
</tr>
</tbody>
</table>

The TDR outputs at 1200°C to 49°C are shown in Figure 8.9-9.15. From these output graphs, it can be observed that no significant signal was measured until the temperature was 226°C at the hot face shown in Figure 8.12. However, no relevant information could be extracted from the signal. Only when the hot face temperature dropped to 140°C was
the back end of the brick detected and using the room temperature dielectric permittivity value, the thickness was determined to be 23.4 cm. As mentioned earlier, the significant loss of the microwave signal is attributed to the glassy phase in the fused-cast AZS bricks.

Figure 8.9. TDR output signal at hot face temperature of 1200°C for three fused-cast AZS bricks.
Figure 8.10. TDR output signal at hot face temperatures of 810°C for three fused cast AZS bricks.

Figure 8.11. TDR output signal at hot face temperatures of 606°C for three fused-cast AZS bricks.
Figure 8.12. TDR output signal at hot face temperature of 226°C for three fused-cast AZS bricks.

Figure 8.13. TDR output signal at hot face temperature of 144°C for three fused-cast AZS bricks.
Figure 8.14. TDR output signal at hot face temperatures of 87°C for three fused-cast AZS bricks.

Figure 8.15. TDR output signal at hot face temperature of 49°C for three fused-cast AZS bricks.
8.4. Summary

The TDR technique was used to demonstrate its potential utility for measuring the thickness of a number of cast and fused-cast AZS bricks. The technique showed great promise as a viable method for refractory wall thickness measurement. The technique provided good results with a high degree of accuracy at room temperature. The measured thickness results using the TDR technique show a linear relationship with respect to TOF and is shown in Figure 8.16. The figures illustrate that the linear relationship can be utilized to determine the thickness of a refractory wall with known dielectric properties and measured \( TOF \). The figure also shows the expected linear relationship for materials with different relative dielectric permittivity.

![Figure 8.16. Relationship between the thickness of the refractory wall and the TOF of the reflected pulse.](image)

The thickness of the fused-cast AZS bricks could not be measured at high temperatures (for the same reason as discussed in the FM-CW radar section). Further investigations have to be conducted on the fused-cast AZS bricks and also on other refractory materials used for lining furnace walls to determine their dielectric properties at high temperatures.
The network analyzer currently used in this investigation is expensive, sensitive and not easily portable and not suitable for the industrial environment. A TDR system, similar to the radar system designed in this investigation, can be built to reduce cost, increase ruggedness and mobility.

The current measurements were conducted with a waveguide adapter which may be inefficient. Better signal coupling into the wall may be achieved by using a horn antenna (i.e., a more efficient signal coupler). Also, a tracking filter can be designed to automatically determine the thickness of the wall. With this feature, the TDR system may be mounted on an X-Y scanner and a 2-Dimensional image with thickness information over the scan region can be obtained.

To summarize, the TDR technique is capable of providing nondestructive, non-intrusive, real-time, in-situ, and online refractory wall thickness information. However, further investigations and improvements are required before these methods can be successfully implemented in the industry.
APPENDIX A.
LOW PASS FILTER DESIGN
Figure A.1. Design parameters of two cascaded stepped-impedance filters on a board with $\varepsilon_r = 2.2$. All dimensions are in mm.
APPENDIX B.
LabVIEW DESIGN OF TRIANGULAR WAVE GENERATOR, SPECTRUM ANALYZER AND “THICKNESS CALCULATOR”
Figure B.1. LabVIEW design of Triangular wave generator.
Figure B.2. LabVIEW low frequency spectrum analyzer design.

Figure B.3. LabVIEW “Thickness Calculator” design.
BIBLIOGRAPHY


18. D. M. Pozar, “Figure 8.27 (a). Attenuation versus normalized frequency for equal-ripple filter prototypes, 0.5-dB ripple level,” in *Microwave Engineering*. New York: John Wiley and Sons, 1998.


