Ultrasonic Continuous Wave Spirometer

Todd A. Ell

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ULTRASONIC CONTINUOUS WAVE
SPIROMETER

by

Todd A. Ell

Bachelor of Science, University of North Dakota, 1982

A Thesis
Submitted to the Graduate Faculty
of the
University of North Dakota
in partial fulfillment of the requirements
for the Degree of
Master of Science

Grand Forks, North Dakota

May

1983
This thesis submitted by Todd A. Ell in partial fulfillment of the requirements for the Degree of Master of Science from the University of North Dakota is hereby approved by the Faculty Advisory Committee under whom the work has been done.

J. Hootman  
(Chairperson)

Nagy N. Bemjian

Edward Nelson

This Thesis meets the standards for appearance and conforms to the style and format requirements of the Graduate School of the University of North Dakota, and is hereby approved.

A. William Johnson  
Dean of the Graduate School
Title: Ultrasonic Continuous Wave Spirometer

Department: Electrical Engineering

Degree: Master of Science

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# TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>SECTION</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>LIST OF ILLUSTRATIONS</td>
<td>vi</td>
</tr>
<tr>
<td>LIST OF TABLES</td>
<td>vii</td>
</tr>
<tr>
<td>ACKNOWLEDGEMENTS</td>
<td>viii</td>
</tr>
<tr>
<td>ABSTRACT</td>
<td>ix</td>
</tr>
<tr>
<td>CHAPTER 1 INTRODUCTION</td>
<td>1</td>
</tr>
<tr>
<td>CHAPTER 2 BASIC INSTRUMENT OPERATION</td>
<td>3</td>
</tr>
<tr>
<td>CHAPTER 3 SYSTEM DESCRIPTION AND MODELING</td>
<td>10</td>
</tr>
<tr>
<td>CHAPTER 4 STEADY-STATE RESPONSE</td>
<td>20</td>
</tr>
<tr>
<td>CHAPTER 5 DYNAMIC RESPONSE</td>
<td>26</td>
</tr>
<tr>
<td>CHAPTER 6 SUMMARY</td>
<td>33</td>
</tr>
<tr>
<td>APPENDIX A DERIVATION OF RELATIVE FREQUENCY SHIFTS</td>
<td>36</td>
</tr>
<tr>
<td>APPENDIX B DERIVATION OF MODIFIED SYSTEM TRANSFER FUNCTION</td>
<td>37</td>
</tr>
<tr>
<td>APPENDIX C S-PLANE ROOT LOCUS PLOTS</td>
<td>38</td>
</tr>
<tr>
<td>APPENDIX D FORTRAN PROGRAM LISTING</td>
<td>48</td>
</tr>
<tr>
<td>APPENDIX E PLL MODEL PARAMETER DEFINITIONS</td>
<td>56</td>
</tr>
<tr>
<td>REFERENCES</td>
<td>58</td>
</tr>
</tbody>
</table>
List of Illustrations

Figure 1. Simplified diagram of one-half of system........6
Figure 2. Complete system schematic.........................12
Figure 3. Schematic of digital Phase-locked loop..........13
Figure 4. Linear controls system model for second
order digital Phase-locked loop..........................14
Figure 5. Ideal transmitting and receiving responses
for a piezoelectric transducer...........................18
Figure 6. One-half the system in signal flow graph
representation.............................................19
Figure 7. One-half of modified system in signal flow graph
representation............................................23
Figure 8. Modified system block diagram....................37
Figure 9-17. S-plane Root Locus Plots.......................39-48
# List of Tables

<table>
<thead>
<tr>
<th>Table</th>
<th>page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1) Spirometric volume flow rates</td>
<td>23</td>
</tr>
<tr>
<td>2) Spirometric air velocities</td>
<td>23</td>
</tr>
<tr>
<td>3) System stability requirements</td>
<td>23</td>
</tr>
</tbody>
</table>
Acknowledgements

The author wishes to thank Professors Joe Hootman and Nagy Bengiamin for their assistance in the preparation of the ideas presented in this thesis.
ABSTRACT

There exists a problem of accurately performing spirographic measurements under physical stress situations. Existing systems, which use mechanical structures in the measurement process, have response times that are too slow, or are too bulky to be considered portable.

The proposed system solves these problems and has a number of attractive characteristics. The system uses relatively inexpensive solid state electronic components which implies a minimal of mechanical parts; portability; and a linear, fast response time.

The system presented in this thesis determines the velocity and temperature fluctuations of the human breath by measuring the difference and sum of the transit times for two continuous sound waves travelling in opposite directions along the air path. The information about the transit times is contained in the phase differences of the two sound waves across the path. A phase-locked loop is used to keep the differences across the path constant, irrespective of air - and sound - velocity variations. Therefore, the phase information is converted to frequency variations in the phase-locked loop.
CHAPTER 1

INTRODUCTION

The purpose of this thesis is to develop the design procedure for a portable, accurate spirometer - an instrument for measuring the breathing capacity of the human lungs - for use under varying physical stress.

Present systems, which involve some mechanical structure in the measurement process, have response times that are too slow, or are too bulky to be considered as portable.

The system designed solves these problems because of a number of attractive characteristics: relatively inexpensive solid state electronic components are used which implies a lack of moving parts and portability; and linear, fast response time and reliability.

The system determines the velocity and temperature fluctuations of the human breath by measuring the difference and sum transit times for continuous sound waves travelling in opposite directions along the air path. The information about the transit times is contained in the phase difference of the sound waves across the path.

A phase-locked loop (PLL) is used to keep the phase difference across the path constant, irrespective of air - and sound - velocity variations by changing the sound frequency. Therefore the phase information is converted to frequency variations.
Although the design is realizable with either an analogue or digital phase-locked loop, the digital phase-locked loop was chosen so that the frequency variation information is directly accessible in digital format. Thus, eliminating the need for analogue to digital converters. Digital information allows the use of a microprocessor for information processing/storage and, more importantly, digital control of the system.
In this chapter we will show how the information contained in the phase shifts across the acoustic paths are converted into frequency variations of useful form.

The total system can be divided into two identical parts, each half determines the transit time of the sound waves traveling in one direction. Because there is no interaction between halves, this and all preceding chapters will study the various responses of only one half of the system.

Figure 1 shows a simplified diagram of one-half of the measurement system. T is the transmitter, R and R' are the two receivers. The phase detector (PD), sequential loop filter (F) and digital controlled oscillator (DCO) form the phase-locked loop.

In general, a DPLL system consists of two main functional blocks, a phase comparator or detector, and a digital controlled oscillator. The DCO is set to operate at an angular frequency of \( \omega_c \) in the absence of a digital control signal. When a control signal is present, then the instantaneous frequency deviation of the DCO is proportional to the control signal. The control signal comes from the PD whose output is proportional to the phase difference between two input signals, one of which is the output of the DCO.
To illustrate how a DPLL operates, assume that the loop is in lock at t=0 (i.e. the input freq./phase and output freq./phase of the DCO are equal), and that at t=0+ the input frequency changes by \( \omega \). At this time the phase detector output will give a positive signal to the DCO which in turn increases the frequency of the DCO. A new equilibrium point will be reached when the frequency of the DCO is equal to the frequency of the input signal. A filter is usually included in the control signal path to smooth the control signal.

The time it takes for a given phase plane of the transmitted sound wave to traverse the distances to the receivers are given by

\[
T_1 = \frac{d}{(c + v_a \cos \theta)} \quad (1)
\]

and

\[
T_2 = \frac{d}{(c + v_a \cos \theta)} \quad (2)
\]

where \( d \) is the separation between the transmitter and the receiver, \( c \) is the velocity of propagation of sound in still air, and \( v_a \cos \theta \) is the velocity component of the air in the direction of the propagation of the sound.

The transmitted sound waves can be expressed as

\[
T(t) = Acos(\omega t + \phi')
\]

where \( \omega \) is the frequency in rad/sec and \( \phi' \) is the initial phase output.
Knowing that each acoustic path introduces a time delay $T_i$, the received signals can be written as

$$P(t) = Acos(\omega(t - T_i) + \phi').$$

Signal magnitude attenuation is neglected because the magnitude variations are eliminated from the received signals upon entering the phase detectors—provided the magnitude is sufficient to trigger them. Notice that this implies that the transmitted signal need not be sinusoidal but only periodic.

The input phase to the phase-locked loop is given by

$$\phi_i = K_{d3} \cdot \omega(T_2 - T_1) =$$

$$K_{d3} \cdot \frac{\omega(d - d_1)}{(c + v a cos)}$$

where $K_{d3}$ is the gain constant of the phase detector.
Figure 1. Schematic of one-half of system

Acoustic path and Transducers

Phase-locked loop

Acoustic path and Transducers
The phase error $\phi_e$, assuming the reference phase is zero (without loss of generality), is then given by

$$\phi_e = K_d \omega (d_2 - d_1)/(c + v \cos \alpha). \tag{4}$$

The phase-locked loop phase detector output is zero for a given phase difference $\overline{\phi}$ (for this case $\overline{\phi} = (\frac{1}{2} \pm n)\pi$). If the differences between distances $d_1$ and $d_2$ are adjusted, under zero air velocity conditions, to give zero output from the phase detector, the output frequency will be the free running or center frequency $\omega_c$. Under these conditions we obtain

$$\overline{\phi} = K_d \omega (d_2 - d_1)/c' \tag{5}$$

where $c'$ is the velocity of propagation of sound at the time of adjustment.

Now, if $\phi_e \neq \overline{\phi}$ the phase detector output will adjust the output frequency such that $\phi_e$ tends toward $\overline{\phi}$. Assuming perfect phase-lock, (i.e. $\phi_e = \overline{\phi}$) for all $T_2 - T_1$, and equating equations 4 and 5 we obtain

$$\frac{\omega - \omega_c}{\omega_c} = \frac{v \cos \alpha}{c'} + \frac{c}{c' - 1}. \tag{6}$$

The same procedure can be used to derive the governing equation for the second half of the system.

$$\frac{\omega_1 - \omega_c}{\omega_c} = \frac{v \cos \alpha}{c'} + \frac{c}{c' - 1}. \tag{7}$$
Summing and differencing (6) and (7), we obtain

\[
\frac{\omega_1 - \omega}{c_1} \frac{\omega_1 - \omega}{c_1} = 2v_d \cos \theta / c' \quad (8)
\]

\[
\frac{\omega_1 - \omega}{c_2} + \frac{\omega_1 - \omega}{c_2} = 2(c/c' - 1) \quad (9)
\]

where \( \omega_1 \) and \( \omega_2 \) denote the two halves of the system.

The results of (6) and (7) show that the distances drop out of the equations and only relative frequency variations exist and need be measured.

Temperature variations are obtained from equation 9 using equation 10, given in [3] as

\[
c = 331.5 + .607 T_0 \text{ m/sec} \quad (10)
\]

where \( T_0 \) is the temperature in degrees centigrade.

A closer inspection of the basic system shown in figure 1 reveals that the system would operate in the same way if PD, were removed and only one acoustic path was incorporated into the system. This is true, and this method is exactly how an anemometer was built as given in [2].

The major reasons for not using the Single Path Anemometer method is to reduce cost and size. The derivation given earlier in this chapter assume that the transducers and receivers introduce no time delay. This is never the case, and the added delay introduces an
error in the desired response of the system.

There are two methods of reducing this error. One method is to use expensive condenser microphones as transducers with very small delays. This is what was done in [3], at a very high cost (75% of the total system cost). The second method, which is used in this system, is to use relatively inexpensive matched Piezoceramic air transducers, which have a larger delay, and to reduce the effect of this delay by introducing a second acoustic path as a reference, within the PLL feedback path. Exactly how this is done is considered in chapters 4 and 5.

Notice that the assumption of a perfect phase-lock is made in the derivation of equations 8 and 9. This assumption is guaranteed by using a second order phase-locked loop which is perfect phase-locked as long as the frequency variations stay within the lock range of the PLL. This requirement introduces the possibility that the system may not be stable. Chapter 5 will deal with the stability of the proposed system.
In this chapter is a detailed description of the system and how each section is modeled. Figures 2 and 3 show the complete system schematic and a detailed schematic of the digital phase-locked loop, respectively.

**Digital Phase-locked Loop Model**

The model used for the digital phase-locked loop (DPLL) was taken directly from [4]. Although the DPLL and analogue phase-locked loop perform the phase-locking function by entirely different methods, linear control systems models for the loops are analogous, enabling the system to be constructed in either the digital or analogue world. This model is shown in figure 4. The parameters shown in figure 4 are defined in appendix E.

**Acoustic Path Model**

It is known that the acoustic paths introduce a time delay between the outputs and inputs of the transducers. Therefore, the input frequency and initial phase to the receivers can be written as

\[ f_i(t) = f_0(t - T_i) \]

and

\[ \phi_i'(t) = -f_0(t - T_i) \cdot T_i \]  

Because the PLL model operator is the total phase angle, as it differs from the rate caused by the loop center frequency \( f_c \), these equations need to be modified for incorporation into the system model.
as follows.

The input phase \( \phi_i \), due to changes in frequency input and initial phase angle is for each path
\[
\phi_i = \int_0^t (f_i - f_c) dt + \phi_i' 
\]
and the output frequency due to changes in rate of output phase \( \phi_o \), again, for each path is
\[
f_o = f_c + d\phi_o/dt. 
\]
Figure 2. Complete system schematic.
Figure 3. Schematic of Digital Phase-Locked Loop.
Figure 4. Linear controls systems model for second order Digital Phase-locked loop.
Combining equations 10, 11, 13 into 12 we obtain for

\[ \tau_i = \int_0^t \left( f_c(t-T_i) + d\phi_0(t-T_i)/dt - f_c(t) \right) dt - \left( f_c(t-T_i) + d\phi_0(t-T_i)/dt \right) T_i \]

where \( i \) represents the path taken. Taking the Laplace transform of \( \tau_i \), assuming \( f_c(t-T_i) = 0 \) for \( T_i > t \), results in

\[ \tau_i(s) = \phi_0(s) \left( 1 - sT_i \right) \cdot \exp(-sT_i) - f_c(s) \left( (sT_i - 1) \cdot \exp(-sT_i) + 1 \right)/s. \quad (15) \]

**Transducer - filter model**

If the gain transfer function of the transducers and any filter inserted into the feedback paths of the phase detectors to reduce noise is expressed in factored form as

\[ T(s) = \prod_{i=1}^{n} (b_{2i}s^2 + b_{3i}s + b_{5i})/(b_{4i}s^2 + b_{6i}s + b_{1i}). \quad (16) \]

then the delay, using the definition given in [5], as

\[ T_k = d(\tau(\omega))/d\omega \]

can be written as

\[ T_k(\omega) = \sum_{i=1}^{n} \left( (-b_{3i} + b_{5i}\omega^2)/((b_{3i} - (b_{4i}\omega)^2) + (b_{5i}\omega)^2) \right) \]

\[ + \left( -b_{5i} + b_{6i}\omega^2)/((b_{6i} - (b_{4i}\omega)^2) + (b_{5i}^2) \right) \quad (17) \]

If the changes in the output frequency \( \omega_o \) from the center frequency \( \omega_c \) are very small \( T_k \) can be approximated as a constant whose value is given by equation 17 with \( \tau \) being replaced by \( \omega_c \). Therefore, the
model for the transducers and filters will be simply a constant delay represented as \(\exp(-T_K s)\).

The ideal transmitting response of a piezoceramic transducer is defined by \(T_T(s)\) in [6] as

\[
T_T(s) = K s^2 / \left(s^2 + \left(\frac{\omega_n T}{Q_T}\right) + \omega_n^2\right)
\]

The ideal receiving response of a piezoceramic transducer \(T_R(s)\) is given by [7] as

\[
T_R(s) = K / \left(s^2 + \left(\frac{\omega_n R}{Q_R}\right) + \omega_n^2\right)
\]

where, in the previous equations, \(\omega_n\) is the resonance frequency and \(Q\) is the quality of the transducers.

If the crystals are operated at their resonance frequency the corresponding delay constant is

\[
T_K = 2(Q_T / \omega_n T + Q_R / \omega_n R).
\]

Because the crystals have gain characteristics of a sharp bandpass filter, as shown in figure 5, further filtering is unnecessary.

Figure 6 shows the complete signal flow graph, for one-half of the system.

The relationship of the DPLL to the analog PLL and why the digital was chosen over the analog can be explained by referring to figure 3. We see that if a divide-by-2IN counter with parallel outputs is
incorporated into the phase-locked loop, the frequency offset information can be latched into a bank of registers by $f_0$, this eliminates the need for converting the output frequency into digital form for processing by a microprocessor, thus reducing the components needed for conversion into a bank of latches. With this technique a higher degree of accuracy may be attained by adding more stages to the divide-by-$2^LN$ counter and latch.
Figure 5. Ideal transmitting and receiving response for a piezoelectric transducer.
Figure 6. One-half the system in signal flow graph representation.
CHAPTER 4

STEADY-STATE RESPONSE

In this chapter, we will show how equations 8 and 9 are realized by the system under steady-state conditions (i.e., $\phi_e = 0$, $v_a \neq 0$). We will first neglect transducer-filter delay and later address this problem in more detail. Under steady-state conditions $\phi_e$ is given as [8]

$$\phi_e = (4K)N(f_0 - f_c)/Kd_2$$

Inserting equation 6 we obtain

$$\phi_e = (4K)N(v_a \cos \theta/c'/c + c/c' - 1)/Kd_2$$

and from the other half of the system

$$\phi_{eB} = (4K)N(-v_a \cos \theta)/c'/c - 1)/Kd_2.$$

The center frequencies must be different for each acoustic signal so that the signals do not interfere. The summing and differencing of the above equations yield

$$\phi_e - \phi_{eB} = (8K)N(v_a \cos \theta/c'/c)/Kd_2$$

and

$$\phi_e + \phi_{eB} = (8K)N(c/c' - 1)/Kd_2.$$

These equations assume perfect phase-lock. The second-order DPLL will track its incoming signal with zero phase error within its lock range.
The second-order DPLL lock range is given in [8] as

\[ \Delta f / f = (f_{max} - f_c) / f_c = M/8K N(1 + 1/2K^2) \text{ Hz.} \]  

(20)

To determine proper values for the parameters, in the above equation, we must determine what range of values that \( v_a \) can obtain spirometry. This is done in the next section.

**Spirometric Parameter Ranges**

Tables 1 and 2 list the average volume flow rates with the corresponding air velocities and average lung volumes respectively [9],[10],[11]. The following definitions will clarify the terms used in the tables.

- **Maximum expiratory/inspiratory volume flow rate (MEV/MJV)** - the maximum volume flow rate obtained after maximum inspiration/expiration.

- **Maximum breathing capacity (MBC)** - the maximum sustained volume flow rate under physical stress.

- **Spontaneous breathing capacity (SBC)** - the volume flow rate under quiet rest conditions.

- **Total lung capacity (TLC)** - the total volume of lungs upon maximum inhalation.
Vital capacity (VC) - the largest volume of air that can be expired after a maximum inspiration.

Inspiratory reserve volume (IRV) - the volume capable of being inspired after quiet expiration.

Expiratory reserve volume (ERV) - the volume capable of being expired after quiet inspiration.

The air velocities given in table 1 were determined by the following equation

\[ v = \frac{\text{volume flow rate}}{\text{cross sectional area}} = \frac{4v}{\pi d^2} \]  

(21)

where \( v \) is the volume flow rate and \( d_0 \) is the diameter of the circular breathing tube. These air velocities were obtained if breathing is done through a 1.5 inch diameter tubing, which is assumed not to affect the normal breath rates.

The lung volumes are determined by the system by integrating the air velocity over the cross sectional area used in equation 21. Table 2 is included to give the range of volumes that will be encountered.
Table 1. - Spirometer Volume Flow Rates

<table>
<thead>
<tr>
<th>Type</th>
<th>Volume Flow Rate (liters/sec)</th>
<th>Volume Velocity (meters/sec)</th>
</tr>
</thead>
<tbody>
<tr>
<td>MBV</td>
<td>12.0</td>
<td>10.7</td>
</tr>
<tr>
<td>MIV</td>
<td>9.0</td>
<td>8.0</td>
</tr>
<tr>
<td>MBC</td>
<td>1.67</td>
<td>1.48</td>
</tr>
<tr>
<td>SRC</td>
<td>0.17</td>
<td>0.150</td>
</tr>
</tbody>
</table>

Table 2. - Spirometric Air Velocities

<table>
<thead>
<tr>
<th>Type</th>
<th>Volume (liters)</th>
</tr>
</thead>
<tbody>
<tr>
<td>TLC</td>
<td>6.00</td>
</tr>
<tr>
<td>VC</td>
<td>5.0</td>
</tr>
<tr>
<td>IRV</td>
<td>2.5</td>
</tr>
<tr>
<td>ERV</td>
<td>2.0</td>
</tr>
</tbody>
</table>

Table 3. - Stability Requirements

\[ T_2 > T_1 \]
\[ \omega_n / f > 6(T_2^2 - T_1^2) / (T_3^3 - T_1^3) \]
\[ \omega_n < \sqrt{2/3(T_2^2 - T_1^2)} \]
Lock Range Requirements

In this section we will check to see if the lock range of the DPLL is sufficient for our purposes.

Neglecting changes in $c$ and setting $\cos \theta = 1$ equations 6 and 20 become

$$M/8K N(1 + \frac{1}{2}K^2) > v/c'. $$

Using $c = 343.57$ m/s, $K_2 = 8$ (which is the minimum value of $K$ for a $K$-counter using the SN74LS297 digital PLL filter), and setting the system clocks equal (i.e., $M = 4N$), $v$ satisfies

$$v < 20.21 \text{ m/sec}$$

which is true for the maximum value obtained in spirometry ($v_a(\text{MEV}) = 10.7$ m/s).

Transducer-filter Delay Affects

Now we will look more closely at the effect transducer-filter delays ($T_k$) have on the static response of the system. Starting at equations 1 and 2 we must include the transducer delays $T_k$ as

$$T_1 = d_1/(c + v_1 \cos \phi) + T_{k1}$$

and

$$T_2 = d_2/(c + v_2 \cos \phi) + T_{k2}. $$
Using the same argument as before, the phase error is given by

$$\phi_{e1} = K_d d_3 \frac{d - d}{(c + v_a \cos \theta) + (T_{k2} - T_{k1})}. \quad (22)$$

Substituting $T_e = T_{k2} - T_{k1}$, and following the same steps which led to equations 6 and 7 of chapter 2, we obtain for perfect phase-lock (derivation given in Appendix A)

$$(\omega_a - \omega_c)/\omega_c = (c/c' + v_a \cos \theta/c' - 1) \cdot (1 - (c + v_a \cos \theta)(T_e)/(d_2 - d_1))^2. \quad (23)$$

By comparing this with equation 6 and knowing $10v_a < c$, it is seen that the conditions for ideal response is

$$| (c T_e)/(d_2 - d_1) | << 1. \quad (24)$$

Thus is necessary to maximize the separation of the receivers for any given transducer-filter delay. For $c = 343.57 \text{ m/s (20 °C 0.0% humidity)} \), $(d_2 - d_1) = 0.01 \text{ m}$ and $T_e = 2.5 \text{ microseconds}$ the resulting error from the ideal of equation 21 is less than 1%.

Each $T_k$ is composed of two parts; the receiver delay and the transmitter delay. For each half of the system, as shown in figure 1, the same transmitter is used for each path. Therefore, $T_e$ is composed of only the difference between receiver delays. Because of the fine tolerances required in manufacturing piezoceramic air transducers closely matched transducers are not uncommon and any slight difference can be tuned to a very small difference, using common crystal tuning techniques.
In this chapter we will study the stability requirements, how the system obtains perfect phase-lock, and the maximum allowable step change in air velocity.

**Stability Requirements**

Figure 6 shows one-half the system in signal flow graph representation. Using Masons Loop Rule [12], we obtain the transfer function

\[
\frac{\Psi}{f} = sK \frac{P(s)}{d_3}
\]

\[
(2^2 + 2\zeta \omega_n s + \omega_n^2) - K d_3 (2\zeta \omega_n s + \omega_n^2) P(s) \tag{25}
\]

where

\[
P(s) = \exp(-sT) \left(1 - sT \right) - \exp(-sT) \cdot (1 - sT),
\]

\[
\zeta = \frac{1}{2} \left(\omega_1 - \omega_2\right)^{\frac{1}{2}},
\]

\[
\omega_n = \left(\omega_1 \cdot \omega_2\right)^{\frac{1}{2}},
\]

\[
\omega_1 = \left(\frac{K}{d_1} \frac{Mf}{2K \cdot N}\right),
\]

and

\[
\omega_2 = \left(\frac{K}{d_2} \frac{Mf}{4K \cdot N L}\right).
\]

First, we will look at the stability requirements. The Routh-Hurwitz Stability Criterion [13] can be applied to this system only if the delay terms are approximated by a few terms of the power series

\[
\exp(-sT) = 1 - sT + \frac{(sT)^2}{2!} - \frac{(sT)^3}{3!} + \ldots.
\]
We will use the first three terms of the series. Therefore, the Routh-Hurwitz criterion will yield only approximate stability information. Substituting into $P(s)$ we obtain

$$P(s) = \frac{1}{2}(4s(T - T_1) - 3s(T_2^2 - T_1^2) + s^3(T_3 - T_1)). \quad (26)$$

At this point we come to the first major problem. Substituting equation 26 into 25, we find that the system, as it is configured, will always be unstable. This arises because of the minus sign in the denominator of equation 25. This problem is easily corrected, and in doing so we reduce the number of parts used in the system. What is done is shown in figure 7. We eliminate the third phase detector and place the second acoustic path inside the phase-lock loop feedback path. The following transfer function results from these changes

$$\phi_{e_1}/f_c = -sP(s)/(s^2 + (2\zeta_\omega s + \omega_n^2)P(s)). \quad (27)$$

The derivation of equation 27 is given in Appendix B.

Notice that this modification has two effects; the minus sign is eliminated, and the order of the transfer function is reduced. Substituting equation 26 into 27 we obtain

$$\phi_{e_1}/f_c = -s((T_3^3 - T_1^3)s^2 - 3(T_2^2 - T_1^2)s + 4(T_3^2 - T_1^2)) (A_3^3 + A_2^2 s + A_1 s + A_0) \quad (28)$$
where

\[ A_3 = 2 \zeta \omega_n \left( T_2^3 - T_1^3 \right), \]

\[ A_2 = \omega_n \left( T_2^3 - T_1^3 \right) - 6 \zeta \omega_n (T_2 - T_1), \]

\[ A_1 = 2 + 8 \zeta \omega_n (T_2 - T_1) - 3 \omega_n T_2^2 T_1^2, \]

and

\[ A_0 = 4 \omega_n T_2 T_1. \]
Figure 7. One-half of modified system in signal flow graph representation.
The Routh-Hurwitz stability requirements are
\[ A_i > 0 \quad i = 0 \text{ to } n, \]
and
\[ A_{21} A_3 - A_{20} A_0 > 0. \]
The first four requirements are fulfilled if
\[ T > T \]
\[
\omega_n / \zeta > 6(T_{22}^3 - T_{21}^3) / (T_{22}^3 - T_{11}^3)
\]
\[
\omega_n < (2/(3(T_{22}^3 - T_{11}^3)))^{1/2}.
\]
Whether all five requirements can be fulfilled is dependent on the values of \( T_1 \) and \( T_2 \). These requirements are listed in Table 3.

It is important to recognize that the steady-state response of the modified system does not differ from the response of the original system.

Appendix C contains plots of s-plane root locations of the modified system under varying conditions. As can be seen from these plots, the system can usually be stabilized for typical system values and the stability range is highly dependant upon transducer delay and acoustic path distances. Two major points can be drawn from these s-plane plots. First, decreasing the transducer delays or decreasing
acoustic path distances has a marked improvement on the system response at the cost of having to raise the DPLL resonance frequency which is determined primarily by the clock frequency of the DPLL-filter integrated circuit. Another method of raising the resonance frequency is to use different clock frequencies for the various DPLL components (i.e. $M=4N$). Both of these methods would reduce the resolution of the output (refer to equations 25 and 18).

Second, reducing the separation distance between the two receivers also improves the system response.

Appendix D contains the Fortran program and parameter values used to generate the data points of the s-plane plots.

**Step Input Response**

The system response to a step change in air velocity is controlled by a highly non-linear equation and evades simple analysis. The transfer function given by equation 28 does not give the response required but is only used to determine the system's stability. What we would be looking for is the output's ($\phi_2$) response to changes in air velocity. The steady-state response shows that this output is 100% sensitive to changes in air velocity but tells us nothing on how this steady-state value is reached.

**Maximum Allowable Step Input**

The next question we must answer is what is the maximum step change
in air velocity \( v_a \) that is allowed before an ambiguity occurs in the phase difference. This occurs when \( |\phi_{e_1}| > \pi \). Starting with \( c = c' \) neglecting \( \cos \theta \) and changes in \( \omega_0 \) (i.e. \( \omega_0 = \omega_c \)) we obtain

\[
\pi < |\omega_c (d_2 - d_1)/c' - \omega_0 (d_2 - d_1)/c + v_a \cos \theta| = \\
\left| -v_a c (d_2 - d_1)/(c') \right|^2.
\]

For \( c' = 343.0 \text{ m/s}, \ \omega_c = 2\pi \ 40K \text{ rad/s}, \text{ and } d_2 - c = 0.01 \text{ m},

\[
v_a > \pi (c')^2 / (\omega_c (d_2 - d_1)) \approx 147.0 \text{ m/s}.
\]

This is enormous, and for smaller separation becomes even greater. Therefore, this constraint presents no problems for most applications. This ambiguity occurs because of the saw-tooth shape of the phase detector transfer function.
In this thesis, we have been concerned with quantitative mathematical modeling of the various components of the system. The differential equations describing the dynamic and static performance of the system was utilized to construct this mathematical model.

Various design considerations are given in the text to aid in construction of the system with the desired characteristics. Major advantages resulting from the designs used include:

1) Only the air temperature at the time of system calibration need be known to completely determine air velocities and temperatures measured.

2) The cost of construction, as compared to other similar devices, is significantly reduced by configuring the system so that inexpensive piezoelectric transducers can be used.

3) Long term reliability and durability results from the use of solid state electronic components and the absence of mechanical parts with the exception of the piezoelectric transducers.

Further considerations would be to determine energy balances by either
estimating or measuring mass flow rates and applying the laws of thermodynamics thus enabling the user to determine breathing efficiencies.
APPENDICES
APPENDIX A

DERIVATION OF RELATIVE FREQUENCY SHIFTS

The phase error under zero air velocity conditions is

\[ \phi_e = \omega_c ((d_2 - d_1)/c' + T_e) \]  

and the phase error under non-zero air velocity conditions is

\[ \phi_e = \omega_o ((d_2 - d_1)/(c + v_a \cos \phi) + T_e). \]  

Under perfect phase-lock conditions these two are equal, and can be equated. Doing this we get

\[ \omega_c ((d_2 - d_1)/c' + T_e) = \omega_o ((d_2 - d_1)/(c + v_a \cos \phi) + T_e). \]  

Rearranging (3) in the following steps;

\[ \omega_c ((d_2 - d_1)/c' - \omega_o ((d_2 - d_1)/(c + v_a \cos \phi)) = T_e (\omega_o - \omega_c), \]

\[ c + v_a \cos \phi - \omega_o c'/\omega_c = c' \cdot T_e (\omega_o - \omega_c)/((d_2 - d_1)\omega_c), \]

\[ \omega_o/\omega_c = (c + v_a \cos \phi)/c' + T_e (c + v_a \cos \phi)(\omega_c - \omega_o)/((d_2 - d_1)\omega_c), \]

\[ (\omega_o - \omega_c)/\omega_c = c/c' + v_a \cos \phi/c' - 1 + T_e (c + v_a \cos \phi)(\omega_c - \omega_o)/((d_2 - d_1)\omega_c), \]

\[ ((\omega_o - \omega_c)/\omega_c)(1 - (c + v_a \cos \phi)(T_e)/(d - d_1)) = c/c' + v_a \cos \phi/c' - 1, \]

and finally,

\[ (\omega_o - \omega_c)/\omega_c = (c/c' + v_a \cos \phi/c' - 1)(1 - (c + v_a \cos \phi)(T_e)/(d - d_1))^{-1}. \]
APPENDIX B

DERIVATION OF MODIFIED SYSTEM TRANSFER FUNCTION

Careful inspection of the two systems in figure 8 show that they are identical. The DPLL Transfer function is

\[
\frac{\phi_o}{\phi_i'} = \frac{N(s)}{D(s)} = \frac{(2\zeta\omega_n s + \omega_n^2)}{(s^2 + 2\zeta\omega_n s + \omega_n^2)}. \quad (1)
\]

Also, \( \phi_e / \phi_i' = s / D(s) \). \( \quad (2) \)

From figure 8 we obtain; \( \phi_i' = -\phi_i + \phi_o \), \( \quad (3) \)

and from the text the acoustic path function is

\[
\phi_i = P(s)\phi_o + P(s)f_c/s. \quad (4)
\]

Combining (1) and (3): \( \phi_i = \phi_o (1 - D(s)/N(s)) \), \( \quad (5) \)

(1) and (2) \( \phi_o = \phi_e N(s)/s^2 \), \( \quad (6) \)

and (4) and (5) \( P(s)f_c/s = \phi_o (1 - D(s)/N(s)) - P(s) \). \( \quad (7) \)

Finally, combining (1), (6), and (7) we obtain

\[
\frac{\phi_e}{f_c} = -sP(s)/(s^2 + P(s)N(s)).
\]

--- 37 ---
APPENDIX C

S-PLANE ROOT LOCUS PLOTS
Figure 9. S-plane Root locus Plot.

\[ \zeta = 0.707, \omega_n = 76 \text{ to } 1000, T_k = 30 \text{ ms}, d_{2} - d_{1} = 1.0 \text{ cm}. \]
\( \zeta = 0.707, \ \omega_n = 46 \text{ to } 1000, \ T_k = 30 \text{ ms}, \ d_2 - d_1 = 2.0 \text{ cm}. \)
Figure 11. S-plane Root Locus Plot.

\[ \zeta = 0.707, \quad \omega_n = 31 \text{ to } 1000, \quad T_k = 30 \text{ ms}, \quad d_2 - d_1 = 3.0 \text{ cm}. \]
Figure 12. S-plane Root locus Plot.

$\zeta = 0.707, \omega_n = 480$ to 1000, $T_k = 15$ ms, $d_2 - d_1 = 0.5$ cm.
Figure 13. S-plane Root locus Plot.

\( \zeta = 0.707, \omega_n = 300 \text{ to } 1000, T_k = 20 \text{ ms}, d - d_1 = 0.5 \text{ cm}. \)
Figure 14. S-plane Root locus Plot.

$\zeta = 0.707$, $\omega_n = 150$ to $1000$, $T_k = 30\text{ ms}$, $d_2 - d_1 = 0.5 \text{ cm}$.
Figure 15. S-plane Root locus Plot.

$\zeta = 1.00$, $\omega_n = 100$ to 1000., $T_k = 30$ ms, $d - d_0 = 0.5$ cm.
Figure 16. S-plane Root locus Plot.

$\zeta = 2.0, \omega_n = 60$ to $1000, T_k = 30 ms, d_{21} = 0.5 \text{ cm}$. 
Figure 17. S-plane Root locus Plot.

\[ \zeta = 2.83, \omega_n = 46 \text{ to } 1000, T_k = 30 \text{ ms}, d \equiv d = 0.5 \text{ cm}. \]
APPENDIX D

FORTRAN PROGRAM LISTING
**49**

```plaintext
10 DIMENSION DELAY1(30), DELAY2(30), P1(30), P2(30), POLES(30)
20 DIMENSION PR(30), PI(30), D(22), ZEROS(30), ZR(30)
30 DIMENSION A(30), B(30), C(30), PLLNUM(30), P(30)
40 C *************************************************************
50 C
60 C
70 C THIS SECTION GENERATES THE POLYNOMIAL P(S)
80 C (SEE TEXT OF THESIS.)
90 C
100 C P(S) = COEFFICIENT POLYNOMIAL OF P
110 C IP = ORDER OF P
120 DO 100 III = 3, 16
130   D1 = .05
140   D2 = .055
150   C0 = 343.0
160   VA = 0.0
170   ID = 2
180   TFIL = 30.E-3
190   T1 = D1 / (VA + C0) + TFIL
200   T2 = D2 / (VA + C0) + TFIL
210   CALL PDELT(T1, ID, DELAY1)
220   CALL PDELT(T2, ID, DELAY2)
230   P1(1) = 1.0
240   P1(2) = -T1
250   IP1 = 1
260   P2(1) = 1.0
270   P2(2) = -T2
280   IP2 = 1
290   G = -1.
300   CALL PMUL(B, IB, DELAY1, ID, P1, IP1)
310   CALL PMUL(A, IA, DELAY2, ID, P2, IP2)
320   CALL FORM(G, A, IA, B, IB, P, IP)
330 C *************************************************************
340 C
350 C CK LOOP
360 C
370 C
380 C LOOP
390 C D LOOP
400 C OTAL LOOP
```

**49**

This section generates the transfer function polynomials.

**Omega1** = Resonance frequency of first

**Omega2** = Resonance frequency of second

**OmegaN** = Natural resonance freq. of T
FREQC = FREE RUNNING FREQ. OF PHASE-LOCK LOOP
DAMP = DAMPING FACTOR OF PHASE-LOCKED LOOP

PLLNUM(S) = NUMERATOR POLYNOMIAL OF PLL.
IPLLN = ORDER OF NUMERATOR

L = 1.0
N = 256 / (2*L)
M = 4*N
FREQC = 40000.
K1 = 2**(III)
K2 = K1
K11 = 2
KD2 = 2
OMEGA1 = (KD1 * M * FREQC) / (2 * K1 * N)
OMEGA2 = (KD2 * M * FREQC) / (4 * K2 * L * N)
OMEGAN = SQRT(OMEGA1*OMEGA2)
DAMP = SQRT(OMEGA1/OMEGA2) / 2.0
PLLNUM(1) = OMEGAN**2
PLLNUM(2) = 2*DAMP*OMEGAN
IPLLN = 1


T(S) = CHARACTERISTIC EQUATION
IT = ORDER OF T(S)
U(S) = REAL ROOT ARRAY
V(S) = IMAGINARY ROOT ARRAY
G = 1.0
D(1) = -0.0
D(2) = 0.0
D(3) = 1.0
IDD = 2
CALL PMUL(C,IC,PLLNUM,IPLLN,P,IP)
CALL FORM(G,C,IC,D,IDD,POLES,IPOLES)
CALL PROOT(IPOLES,POLES,PR,PI,1)
CALL PRINT(1, OMEGAN,DAMP
1 FORMAT(1X,'OMEGAN=',F15.5,2X,'DAMPING FACTO
R=',F15.5)
SUBROUTINE NORMP(X,IX,EPS)
C THIS SUBROUTINE ZEROS COEFFICIENTS OF A POLYNOMIAL
C THAT ARE LESS THAT A THRESHOLD VALUE, THUS REDUCING
C THE ORDER OF THE POLYNOMIAL.
DIMENSION X(10)
1 IF (IX) 4,4,2
2 IF (ABS(X(IX)) - EPS) 3,3,4
3 IX = IX - 1
GO TO 1
4 RETURN
END

SUBROUTINE PDELT(T,N,DELAY)
C THIS SUBROUTINE GENERATES THE NTH ORDER POLYNOMIAL
APPROXIMATION TO A DELAY (I.E. EXP(-TS)).

DELAY = POLYNOMIAL APPROXIMATION OF DELAY
N = ORDER OF POLYNOMIAL N [= 10
T = DELAY TIME

DIMENSION DELAY(11)
FACT = 1.
SIGN = 1.0
MINUS = -1.0
DO 10 I = 1,N
SIGN = SIGN * MINUS
FACT = FACT * I
10 DELAY(I+1) = SIGN * (T**I) / FACT
DELAY(1) = 1.
RETURN
END

SUBROUTINE FORM(G,A,N,B,M,C,IX)
THIS SUBROUTINE FORMS THE WEIGHTED SUM OF TWO POLYNOMIALS

C(S) = B(S) + G * A(S)
G = SCALER WEIGHTING FACTOR
A = POLYNOMIAL COEFFICIENT ARRAY FOR A(S), CONSTANT FIRST
N = ORDER OF A(S), N [= 10
B = POLYNOMIAL COEFFICIENT ARRAY FOR B(S), CONSTANT FIRST
M = ORDER OF B(S), M [= 10
C = POLYNOMIAL COEFFICIENT ARRAY FOR RESULTING C(S)
IX = ORDER OF C(S)

DIMENSION A(11),B(11),C(11)
IF(N-M) 1,2,2
1 IX=M+1
GO TO 3
2 IX=N+1
3 DO 4 I=1,IX
4 C(I)=B(I)+G*A(I)
IX=IX-1
RETURN
END

SUBROUTINE PEXCG(A,IA,B,IB)
SUBROUTINE PMUL(Z,IZ,X,IXA,Y,IYA)

THIS SUBROUTINE FORMS PRODUCT OF TWO POLYNOMIALS

Z(S) = X(S) * Y(S)

Z = RESULTING COEFFICIENT ARRAY, CONSTANT FIRST

A = ARRAY OF A(S)
IA = ORDER OF A
B = ARRAY OF B(S)
IB = ORDER OF B

DIMENSION A(1),B(1)
JJ=IB+1
DO 1 I=1,JJ
1 A(I)= B(I)
10 RETURN
END

DIMENSION Z(1),Y(1),Z(1)
IX=IXA+1
IY=IYA+1
IP (IX*IY) 10,10,20
10 IZ = 0
20 GO TO 50
20 20 IZ = IX + IY
20 30 DO 30 I = 1, IZ
20 40 Z(I) = X(I) * Y(J) + Z(K)
20 30 CONTINUE
20 50 IZ = IZ - 2
SUBROUTINE PROOT(N,A,U,V,IR)

C THIS SUBROUTINE USES A MODIFIED BARSTOW METHOD TO FIND
C THE ROOTS OF A POLYNOMIAL.

C
C N = DEGREE OF POLYNOMIAL, N [=19
C A = POLYNOMIAL COEFFICIENT ARRAY.
C U = REAL ROOT ARRAY
C V = IMAGINARY ROOT ARRAY
C IR = +1 IF POLYNOMIAL WRITTEN AS: A(1)+A(2)
S**2+A(3)S**2+...
C = -1 IF POLYNOMIAL WRITTEN AS; A(1)S**N
C
C DIMENSION A(20),U(20),V(20),H(21),B(21),C(21)

NDERAY = IR
NC = N + 1
DO 1 I = 1, NC
H(I) = A(I)
P = 0.
Q = 0.
R = 0.
3 IF (H(1)) 4,2,4
2 NC = NC - 1
V(NC) = 0.
U(NC) = 0.
DO 1002 I = 1, NC
H(I) = H(I+1)
1002 GO TO 3
4 IF (NC - 1) 5,100,5
5 IF (NC - 2) 7,6,7
6 R = -H(1)/H(2)
50 GO TO 50
7 IF (NC - 3) 9,8,9
8 P = H(2)/H(3)
9 Q = H(1)/H(3)
70 GO TO 70
9 IF (ABS (H(NC-1)/H(NC))) -ABS (H(2)/H(1))) 10, 19,19
10 IREV = -IREV
M = NC / 2
DO 11 I = 1, M
11 NL = NC + 1 - I
F = H(NL)
H(NL) = H(I)
H(I) = F

IF (Q) 13,12,13
P = 0.
GO TO 15
P = P/Q
Q = 1. / Q
IF (R) 16,19,16
R = 1. / R
E = 5.E-10
B(NC) = H(NC)
C(NC) = H(NC)
B(NC+1) = 0.
C(NC+1) = 0.
NP = NC - 1
DO 49 J = 1, 1000
I = NC - I1
B(I) = H(I) + R*B(I+1)
C(I) = B(I) + R*C(I+1)
IF (ABS(B(1)/H(1)) - E) 50,50,24
IF (C(2)) 23,22,23
R = R + 1
GO TO 30
R = R - B(1)/C(2)
DO 37 I1 = 1, NP
I = NC - I1
B(I) = H(I) - P*B(I+1) - Q*B(I+2)
C(I) = B(I) - P*C(I+1) - Q*C(I+2)
IF (H(2)) 32,31,32
IF (ABS(B(2)/H(1)) - E) 33,33,34
IF (ABS(B(2)/H(2)) - E) 33,33,34
IF (ABS(B(1)/H(1)) - E) 70,70,34
CBAR = C(2) - B(2)
D = C(3)**2 - CBAR*C(4)
IF (D) 36,35,36
P = P - 2
Q = Q * (Q+1)
GO TO 49
P = P + (B(2)*C(3) - B(1)*C(4)) / D
Q = Q + (-B(2)*CBAR + B(1)*C(3)) / D
49 CONTINUE
E = E*10.
GO TO 20
NC = NC - 1
V(NC) = 0.
IF (IREV) 51,52,52
U(NC) = 1. / R
GO TO 53
U(NC) = R
DO 54 I = 1, NC
54 H(I) = B(I+1)
3080 GO TO 4
3090 70 NC = NC - 2
3100 IF (IREV) 71,72,72
3110 71 QP = 1. / Q
3120 PP = P / (Q * 2.0)
3130 GO TO 73
3140 72 QP = Q
3150 PP = P / 2.0
3160 73 F = (PP)**2 - QP
3170 IF (F) 74,75,73
3180 74 U(NC+1) = -PP
3190 U(NC) = -PP
3200 V(NC+1) = SQRT(-F)
3210 V(NC) = -V(NC+1)
3220 GO TO 76
3230 75 IF (PP) 81,80,81
3240 80 U(NC+1) = -SQRT(F)
3250 GO TO 82
3260 81 U(NC+1) = -(PP / ABS(PP)) * (ABS(PP) + SQRT(F))
3270 82 CONTINUE
3280 V(NC+1) = 0.
3290 U(NC) = QP / U(NC+1)
3300 V(NC) = 0.
3310 76 DO 77 I = 1, NC
3320 77 H(I) = B(I+2)
3330 GO TO 4
3340 100 RETURN
3350 END
### APPENDIX E

**PLL MODEL PARAMETER DEFINITIONS**

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Definition</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K_{d1}$</td>
<td>Gain of phase detector number 1.</td>
<td>cycles$^{-1}$</td>
</tr>
<tr>
<td>$K_{d2}$</td>
<td>Gain of phase detector number 2.</td>
<td>cycles$^{-1}$</td>
</tr>
<tr>
<td>$K_{d3}$</td>
<td>Gain of phase detector number 3.</td>
<td>cycles$^{-1}$</td>
</tr>
<tr>
<td>$M_f$</td>
<td>Clock frequency of programmable counters.</td>
<td>Hz</td>
</tr>
<tr>
<td>$M_f/K_1$</td>
<td>Gain of programmable divide-by-$K_1$ counter.</td>
<td>cycles</td>
</tr>
<tr>
<td>$M_f/K_2$</td>
<td>Gain of programmable divide-by-$K_2$ counter.</td>
<td>cycles</td>
</tr>
<tr>
<td>$1/N$</td>
<td>Gain of divide-by-$N$ counter.</td>
<td>cycles/cycle</td>
</tr>
<tr>
<td>$1/L$</td>
<td>Gain of divide-by-$L$ counter.</td>
<td>cycles/cycle</td>
</tr>
</tbody>
</table>
REFERENCES


[3] Ibid., page 567.


