AN ENHANCED DESIGN PROCEDURE FOR
MICROSTRIP BAND PASS FILTERS

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(ABSTRACT)

Low cost bandpass filters (less than $100) at microwave frequencies cannot be purchased commercially. However, such filters are essential in the design of RF circuits in communications and radar equipment. Reliable microstrip band pass filters which provide an accurate filter response at microwave frequencies can be easily fabricated with low cost. Equations concerning the design of coupled microstrips and microstrip filters are published in the literature and were implemented in a design procedure for maximally flat microstrip band pass filters. The published equations were theoretical and had not been extensively compared with experimental data. Thus, this work established an enhanced microstrip filter design procedure based on experimental data, for a wide range of frequencies and dielectric substrates.

The result of this work is an enhanced design procedure for microstrip band pass filters. The new procedure includes a correction factor for the length of the filter resonators which controls the center frequency of the filter. This correction factor has been found from the measured responses of over 60 filters, which were designed with two different circuit board materials, three different substrate thicknesses, and frequencies ranging between 0.9 and 6 GHz. The experimentally
determined length correction factor decreases the error in center frequency from 
±5.9% down to ±1.7% of the desired design frequency for a wide range of filter 
designs. The improved procedure has been implemented in a personal computer 
(PC) program which calculates all dimensions necessary to fabricate microstrip 
band pass filters in the low microwave frequency range. The maximally flat 
response obtained is accurate and requires very little tuning. Low cost microstrip 
band pass filters can now be designed and fabricated easily and with greater 
accuracy at microwave frequencies. This thesis describes the development of the 
enhanced design procedure and the results of the filters designed with the new 
procedure.
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1 INTRODUCTION

Several projects at Virginia Tech, including a pulsed RF propagation measurement system, an indoor radio modem, and a microwave satellite receiver, have required inexpensive band pass filters at microwave frequencies. It is well known that microstrip transmission line coupling can provide reliable, inexpensive filters at microwave frequencies [1]. Also, in a university setting, we have access to printed circuit board facilities which can fabricate circuit boards inexpensively, but find it difficult to justify the purchase of radio frequency (RF) design software packages that cost several hundreds or thousands of dollars. A complete design procedure for microstrip band pass filters does not currently exist in the literature [1]. Thus, the basic equations for designing microstrip filters were extracted from the published literature [1-12] and implemented in a design procedure. An easy to use personal computer (PC) program called USTRIP was developed from this design procedure and written in Turbo Pascal 5.0 [13]. The design procedure in USTRIP was continually changed based on new results. A source code listing of the final version of USTRIP is given in Appendix D. USTRIP yields physical dimensions for maximally flat (Butterworth) filters which possess the user specified frequency response characteristics. The inputs include such parameters as center frequency, bandwidth, and stop band attenuation as well as substrate characteristics.

Several filters were designed based on the published equations cited above.
However, measured responses consistently showed errors in center frequency. Therefore, an enhancement of these equations was necessary to relate theoretical to experimental results. A data base of filter responses, based on over 60 designed, fabricated, and tested filters, was obtained and analyzed with statistical techniques to find relationships between center frequency error and design parameters. The data resulted in a more accurate length correction factor for the microstrip filter dimensions. The length correction factor corrects for center frequency errors found in tested filters which were designed with theoretical equations only. With implementation of the correction factor, center frequency accuracy has improved to within ±1.7% on average, with a worst case error of 4.7% away from the desired center frequency. The measured 3 dB bandwidths agreed well with desired bandwidths when only theoretical equations were used in the design procedure. The 10 dB bandwidths of the reflection responses of the filters seem to be dependent on desired bandwidth. Without implementation of the correction factor, center frequency accuracy was within ±5.9% on average, with a worst case error of 15.8%. Thus the well accepted microstrip band pass filter design equations extracted from the literature have been enhanced to facilitate more accurate microstrip filter design. These microstrip bandpass filters present a viable option for low cost, easily fabricated filters that can be used in a variety of applications, including diode multiplier circuits [14,15], receiver front ends, and IF filter stages.
1.1 MOTIVATION FOR THIS RESEARCH

Microstrip band pass filter research was originally started at the Mobile and Portable Radio Group (MPRG) of Virginia Tech due to the need for an inexpensive radio frequency (RF) band pass filter for use in an indoor RF modem (modulator/demodulator). The modem was to carry simplex type data transmission at rates up to 300 kilobits per second (kbps). The basic modem design is now discussed, with reference to microstrip filters.

The transmitter includes an FSK (Frequency Shift Keying) encoder, IF (intermediate frequency) section, RF section, and antenna. A block diagram of the transmitter is given in Figure 1.1.1. FSK is a type of modulation used to encode digital data for radio wave transmission. The digital data (ones and zeros, or marks and spaces) are encoded into two different frequencies, one frequency for digital ones and one for digital zeros. The input to the encoder is digital data (5V = mark, 0V = space). These data voltages toggle a diode switch, which produces one of two oscillator outputs, at 2.000 MHz or 2.4576 MHz. The output of the switch is low pass filtered to smooth phase discontinuities due to switching. The resulting FSK signal is mixed up to 146.00 MHz and 146.4576 MHz in the IF section, and mixed again to 1152.00 MHz and 1152.46 MHz, through the use of local oscillator (LO) signals. The LO signals are produced by a PLL (Phase Locked Loop) frequency synthesizer and multiplier. The synthesizer derives a 144 MHz signal from an 18 MHz oscillator. This 144 MHz LO signal is used in the IF section and also multiplied by 7 to obtain 1008 MHz for use as the LO signal in
Figure 1.1.1: Block diagram of RF modem transmitter.
the RF section.

The receiver (Figure 1.1.2) uses the same synthesizer/multiplier circuit for the RF section, but implements an FSK receiver chip for the IF section and the decoder. The incoming signal received at the antenna is mixed back down in the RF section to the IF frequencies and fed to the commercially available FSK receiver chip. The chip contains an IF mixer section and a quadrature detector which decodes the FSK signal to produce digital output.

The ×7 multiplier used for the RF LO in both the transmitter and receiver is called a diode multiplier (see Figure 1.1.3). This circuit consists mainly of a diode and an RF filter, and can therefore be built with low cost, as long as the RF filter is inexpensive. The input signal must be strong enough to saturate the diode, which produces harmonics of the input frequency. The desired harmonic, 1008 MHz, must be filtered from the other harmonics and any source frequency (144 MHz) feedthrough. This filtering can be done with an inexpensively fabricated filter, such as a microstrip band pass filter. Thus, research was begun concerning design of microstrip filters.

The basics of microstrip theory were researched, and a design procedure was developed from equations found in the literature. This design procedure was expected to yield dimensions for microstrip band pass filters from the desired response parameters. This design procedure is discussed in Chapter 2. Several filters were designed with the original design procedure, fabricated, and tested
Figure 1.1.2: Block diagram of RF modem receiver.
Figure 1.1.3: Diode multiplier circuit. The source signal hits a diode which produces harmonics of the source signal frequency. The microstrip filter attenuates the undesirable harmonics.
with results which agreed with most design parameters. However, the measured center frequencies of the responses differed from those specified in the design and the insertion loss was high. For example, a 4 pole filter built on common PC board had a measured insertion loss of 13 dB, when a maximum of 3 dB loss was expected. Correction of these problems required more research. Microstrip filter research eventually replaced the RF modem research due to the center frequency and insertion loss problems and the large amount of time required to develop the design procedure, fabricate filters, and test the filters accurately. The remainder of this thesis completely documents the design procedures, measurement techniques, experimental results, and analysis that have resulted in an enhanced microstrip band pass filter design procedure.
1.2 THESIS ORGANIZATION

The thesis is divided into 6 chapters. The present chapter discusses the motivation and organization of this thesis. Chapter 2, Microstrip Filter Basics, reviews previous research in microstrip filter design and describes the design procedure for microstrip band pass filters as developed from the literature. Chapter 3, Experimental Procedure, discusses the results of preliminary research performed before a controlled design procedure was developed. A test plan for acquiring an extensive data base is developed and details for fabrication, measurement, and tuning procedures for microstrip filters are also included. The extensive data base and the procedures for obtaining this data are presented in Chapter 4, Experimental Data. Chapter 5, Experimental Results and Analysis, is dedicated to analysis of the data for the development of a length correction factor implemented in an enhanced design procedure. Results of filter designs made with this enhanced procedure are also discussed. Chapter 6, Conclusions, summarizes the thesis and gives recommendations for extensions of this work.
2 MICROSTRIP FILTER BASICS

This chapter discusses the microstrip theory currently available in the literature and describes a design procedure for microstrip band pass filters developed from equations found in the literature. The use of a length correction factor in the design procedure is explained, and theoretical insertion loss and reflection response are discussed.

2.1 LITERATURE REVIEW

The use of microstrip has increased over the past 20 years due to its advantages over other microwave transmission media. A great deal of research has produced many expressions for the properties of microstrip. Microstrip has the advantage of low cost, small size, good performance, and ease of fabrication over waveguide, stripline, coaxial cable, coplanar waveguide, slotline, and coplanar strips [2,5,6]. One disadvantage is higher insertion loss in comparison to other types of filters. This section describes the properties of microstrip lines and coupled lines which are necessary for the development of microstrip filters.

A microstrip transmission line structure consists of a thin strip of conductor material and a conducting ground plane separated by a dielectric medium. The dielectric medium serves as a structural substrate upon which thin metal
conductors are deposited. Conductor material is usually gold, copper, or aluminum [2,5]. The range of dielectric constants, denoted by $\epsilon_r$, typically varies between 2.17 and 12.3, although more modern materials are being developed which have $\epsilon_r$ values in excess of 15 [5]. Substrate materials are classified as hard or soft. Hard substrates include ceramics or semiconductors and facilitate circuits with fine line definition and low loss, but at a higher cost than soft substrates. Soft substrates include teflon-glass and teflon-ceramic filled substrates. The dielectric constants of soft substrates are low (~2) for the teflon-glass filled and high for (~10) for teflon-ceramic types. For larger circuits, these soft substrates can be stiffened by adding an aluminum base plate. These substrates are clad with 0.5, 1.0, or 2.0 oz. of copper per square foot of substrate, which correspond to conductor thicknesses of 0.7, 1.4, and 2.8 mils, respectively [5,16]. One mil equals 0.001 inches.

Propagation of electromagnetic waves in microstrip is TEM (transverse electromagnetic mode) with some dispersion into the air [2,5]. The TEM mode defines transmission with electric and magnetic fields transverse to the direction of energy flow. TE and TM, or transverse electric and transverse magnetic, modes define transmission with only the electric or magnetic fields transverse to the direction of energy flow, respectively. The TE and TM modes are known as higher order modes [6,17,18]. Dispersion occurs because the fields present between the microstrip and ground plane are not completely contained in the substrate (see Figure 2.1.1), and must be accounted for in microstrip analysis. This is the reason for a quasi-TEM propagation analysis instead of a true TEM propagation.
Figure 2.1.1: Cross-sectional view of a typical resonator pair and the propagating electric and magnetic fields.
analysis. The external fields are taken into account by introducing an effective
dielectric constant ($\epsilon_{eff}$) which is lower than $\epsilon_r$. At low microwave frequencies,
$\epsilon_{eff}$ is assumed to be constant. This effective dielectric constant is used in the
determination of the wavelength in microstrip line ($\lambda_s$), given by the following
equation [2,6]:

$$\lambda_s \approx \frac{c}{f\sqrt{\epsilon_{eff}}} \text{ cm}, \quad [2.1.1]$$

where $c$ is the speed of light ($3 \times 10^{10} \text{ m/s}$) and $f$ is the frequency in Hz. At higher
microwave frequencies, the effective dielectric constant and characteristic
impedance of a microstrip line increases with frequency, due to dispersion. The
equation given for the frequency where dispersion begins to take effect is

$$f_d = 0.3 \sqrt{\frac{Z_0}{h \sqrt{\epsilon_r - 1}}}, \quad [2.1.2]$$

where $h$ is the substrate thickness in cm, $f_d$ is in GHz, and $Z_0$ is the characteristic
impedance of the microstrip line.

Experimental attempts to describe the frequency dependence of $\epsilon_{eff}$ are given in
[2,20,21]. However, the empirical expressions lack theoretical foundation [2]. The
theoretical results in [2,3,6,19,20] agree more closely to the experimental results
reported in [20,21], and are given here. The following theoretically derived
equations account for dispersion effects on the effective dielectric constant, now
denoted $\epsilon_{eff}(f)$ [2,6,19,20];

$$\epsilon_{eff}(f) = \epsilon_r - \frac{\epsilon_r - \epsilon_{eff}}{1 + G(f/f_p)} , \tag{2.1.3}$$

where

$$f_p = \frac{Z_o}{8\pi h} , \tag{2.1.4}$$

$$G = 0.6 + 0.009 Z_o , \tag{2.1.5}$$

and

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left[ \left( 1 + 12 \frac{h}{W} \right)^{\frac{1}{2}} + 0.04 \left( 1 - \frac{W}{h} \right)^2 \right] ; \frac{W}{h} \leq 1 \tag{2.1.6}$$

or

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left( 1 + 12 \frac{h}{W} \right)^{-\frac{1}{2}} ; \frac{W}{h} \geq 1 \tag{2.1.7}$$

with $h$ in cm and $f$ in GHz. In equations 2.1.3 through 2.1.7, $f_p$ denotes the cutoff frequency for the next higher order propagation mode in GHz, $Z_o$ is the characteristic impedance of the microstrip line in ohms, and $G$ is a constant. The quantity $\epsilon_{eff}$ is also referred to as $\epsilon_{eff}(0)$ and is the same quantity as that given in equation 2.1.1. The quantity $\frac{W}{h}$ is described subsequently.

Theoretically based equations are also found in [2,3,5,6,19,20,22] which account for dispersion effects on $Z_o$. The frequency dependent characteristic impedance is denoted by $Z_o(f)$. However, frequency dependence of $Z_o$ is negligible for single microstrips (no coupling with adjacent microstrips) according to [2,5].

At much higher microwave frequencies (typically above 12 GHz), the wave
propagation along the microstrip line can no longer be considered TEM due to the existence of higher order modes. The frequency where this begins to occur is denoted $f_t$ given by [2,5,6]

$$f_t = \frac{c}{2 \pi h} \sqrt{\frac{2}{\epsilon_r - 1}} \tan^{-1}(\epsilon_r),$$  \hspace{1cm} [2.1.8]

where $f_t$ is in GHz, $h$ is in cm, and $c$ is $3 \times 10^{10} \text{cm/s}$.

Most microstrip research is based on the quasi-TEM assumption. The complete hybrid-mode models of propagation in a microstrip line result in far more complicated results which involve extensive computations. Closed form expressions for the physical dimensions of microstrip filters are desirable for optimization of a computer-aided design of a microstrip circuit since these increase speed and are easily implemented [2]. These closed form expressions are now discussed.

2.1.1 SINGLE MICROSTRIP LINES

A microstrip filter consists of many microstrip lines with different widths and lengths. In microstrip design, the width ($w$) of a single microstrip line is calculated from the desired characteristic impedance of the microstrip line ($Z_0$), the substrate dielectric constant ($\epsilon_r$), and the substrate height ($h$). Figure 2.1.2 illustrates these dimensions in the microstrip structure. The line width is usually expressed in a design ratio, $\frac{w}{h}$, since it is always dependent on substrate thickness,
Figure 2.1.2: Basic structure of microstrip line showing width \( w \) and spacing \( s \) of resonators, substrate thickness \( h \), conductor thickness \( t \), and the ground plane.
h. Closed form equations for \( \frac{W}{h} \) reported in [2,5,6,23], and compared with data from [24], apply for \( 0.05 \leq \frac{W}{h} \leq 20 \) and \( \epsilon_r \leq 16 \). The following closed form expressions fall within \( \pm 1 \% \) of numerical results given in [24]:

For \( \frac{W}{h} \leq 2 \),

\[
\frac{W}{h} = \frac{8 e^{(A)}}{e^{(2A)} - 2}
\]  \hspace{1cm} [2.1.9]

where

\[
A = \frac{Z_0}{60} \sqrt{\frac{\epsilon_r + 1}{2}} + \left[ \frac{\epsilon_r - 1}{\epsilon_r + 1} \right] (0.23 + 0.11 \frac{\epsilon_r}{\epsilon_r}).
\]  \hspace{1cm} [2.1.10]

For \( \frac{W}{h} \geq 2 \),

\[
\frac{W}{h} = \frac{2}{\pi} \left[ B - 1 - \ln(2B - 1) + \frac{\epsilon_r - 1}{2\epsilon_r} \left( \ln(B - 1) + 0.39 - \frac{0.61}{\epsilon_r} \right) \right]
\]  \hspace{1cm} [2.1.11]

where

\[
B = \frac{377\pi}{2Z_0\sqrt{\epsilon_r}}.
\]  \hspace{1cm} [2.1.12]

Alternate equations, 2.1.13 and 2.1.14, developed by Wheeler [24], and presented in [7,25] gives the \( \frac{W}{h} \) ratio for a single microstrip line given \( \epsilon_r \) and \( Z_0 \). Essentially, the following set of equations, adopted from [24], replace equations 2.1.9 through 2.1.12 given by [2,5]. Equations 2.1.13 and 2.1.14 are simpler than equations 2.1.9 through 2.1.12 and are accurate to within \( \pm 1 \% \) of the numerical results given by [24] over the same wide ranges of \( \frac{W}{h} \) and \( \epsilon_r \) as those for equations 2.1.9 through 2.1.12 [7,26]. Therefore, the equations 2.1.9 through 2.1.12 and equations 2.1.13 and 2.1.14 give equivalent results.
\[ \frac{w}{h} = 8 \sqrt{C \frac{7 + \frac{4}{\varepsilon_r}}{11} + \frac{1 + \frac{1}{\varepsilon_r}}{0.81}} \]  \hspace{1cm} [2.1.13]

where
\[ C = \exp \left[ \frac{Z_o}{42.4} \sqrt{\varepsilon_r + 1} \right] - 1 \]  \hspace{1cm} [2.1.14]

Plots relating characteristic impedance to \( \frac{w}{h} \) are also available [27]. Equations are also reported [2,5,23] which give \( Z_o \) in terms of \( \frac{w}{h} \), and \( \varepsilon_r \). However, these are not of great use in microstrip design since the \( Z_o \) is normally the known quantity, and the design width is required.

The above expressions assume a two-dimensional microstrip conductor. However, the microstrip is three-dimensional in practice, and conductor thickness (t) must be accounted for if any of the following conditions are not true: \( \frac{t}{h} \leq 0.005 \), \( 2 \leq \varepsilon_r \leq 10 \), \( 0.1 \leq \frac{w}{h} \leq 5 \). If all conditions are true, \( \frac{t}{h} \) can be assumed to be zero [2,5,23,28].

The effect of strip thickness can be considered in the calculation of \( \frac{w}{h} \) by replacing \( \frac{w}{h} \) by \( \frac{w_e}{h} \), where \( w_e \) is the effective strip width. Expressions for \( w_e \) are reported in [2,3,5,24] as

\[ w_e = w + \frac{t}{h} \left[ 1 + \ln \frac{2h}{t} \right]; \ \frac{w}{h} \geq \frac{1}{2\pi} \]  \hspace{1cm} [2.1.15]

and
\[ w_e = w + \frac{t}{h} \left[ 1 + \ln \frac{4\pi w}{t} \right]; \ \frac{w}{h} \leq \frac{1}{2\pi} \]  \hspace{1cm} [2.1.16]

These equations are valid if the following conditions are true: \( \frac{t}{h} \leq 0.005 \) and \( t < \)
The units for $w$, $w_e$, $h$, and $t$ are specified by the user, but must be consistent for all four quantities.

2.1.2 COUPLED MICROSTRIP LINES

The above discussion has been confined to single microstrip lines. However, microstrip lines couple energy if they are adjacent to one another. This is called edge coupling because the coupling of energy occurs along the sides of the microstrip. There are two types of edge coupling: parallel coupling and end coupling. Parallel coupling occurs along the length of the resonator whereas end coupling occurs at the end of the resonators. Implementation of parallel coupling produces a smaller device and allows greater flexibility in spacings between the coupling lines [4,6,10]. This thesis discusses microstrip filters which use parallel coupled resonators.

A cross-sectional view of two coupled microstrips and their electric and magnetic coupling fields is shown in Figure 2.1.1. The wave propagation of coupled energy in one line can be of the same, or opposite polarity of the wave propagation in the other coupled line. Odd mode propagation corresponds to energy in each microstrip travelling in opposite polarities, and even mode corresponds to propagation in the same polarity in both coupling microstrips. Accordingly, for a coupled pair of microstrip lines, there exist two characteristic impedances instead of the one characteristic impedance of a single microstrip line. Each of the two impedances corresponds to a mode of propagation. The odd mode impedance is
denoted by $Z_{oo}$ and the even mode impedance by $Z_{ee}$.

Any microstrip line with open ends has a resonant frequency, and is called a resonator. Microstrip filters consist of a number of coupling resonators. A resonator pair, as shown in Figure 2.1.3, consists of two segments of microstrip line, approximately one quarter wavelength long, that couple electromagnetic energy across dielectric material. The coupling length for each half wavelength resonator is half a resonator length, or one quarter wavelength. Each half wavelength resonator couples energy to two adjacent resonators, one on each side. The desired coupling area between the adjacent resonators is shaded in Figure 2.1.3. The width and length of each of the two coupling quarter wavelength resonator segments are equal. Therefore, a resonator pair is completely defined by the coupling length of the resonator pair, the width of the coupling resonator segments, and the spacing between them (see Figures 2.1.2 and 2.1.3). The spacing dimension is expressed as a design ratio, $\frac{s}{h}$, since spacing is dependent on substrate thickness, $h$. Hereafter, the quarter wavelength segment of a resonator pair is called the resonator length, denoted $L_r$. This coupling segment length is the relevant dimension to microstrip filter design, not the total resonator length, which is one half wavelength long.

A microstrip filter consists of a two-sided piece of substrate with one side etched into several coupled resonator pairs and one side left unetched as the ground plane. The resonators are strips of conducting material separated by the substrate dielectric. Therefore, unlike the ground plane, the resonator side of the substrate
Figure 2.1.3: Top view of microstrip band pass filter layout showing resonator pair length ($L_r$), spacing ($s$), width ($w$), and desired coupling area. The characteristic impedance ($Z_0$) input/output (I/O) lines are also labelled.
contains only several places where conducting material exists. These resonators resonate at approximately the center frequency of the filter and couple energy to one another across an area where there is dielectric substrate, but no conductor material. The energy due to signals with frequencies other than the resonant frequency of each microstrip resonator is attenuated. Therefore, several microstrip resonators, when coupled together on a piece of printed circuit board, provide a filter response. The number of poles is equivalent to the number of resonators of the filter. A greater number of poles results in more attenuation of the non-resonant frequencies of the microstrip resonators.

In addition to the resonator pairs, a microstrip filter has two input/output (I/O) characteristic impedance lines. The width of these lines are equal, and can be determined from equations 2.1.9 through 2.1.12, or 2.1.13 and 2.1.14. The end of each of these lines serves as one half a resonator pair. Therefore, the length of these lines is greater than one quarter wavelength. In addition to the resonator length, \( L_r \), the length of I/O line must include length for a connector to be attached to the filter. This allowance is described in Section 3.3, but is normally 0.5 inches. A general layout of a microstrip filter with 4 poles is shown in Figure 2.1.3. A photograph of a fabricated filter is shown in Figure 2.1.4.

Unlike the I/O lines, the resonator pair dimensions depend on the desired filter response. The procedures for finding the resonator lengths, widths, and spacings depend on the desired frequency response parameters. The resonator dimensions vary with each resonator. The determination of a mathematical model for the
Figure 2.1.4: Photograph of a typical microstrip filter (Filter 3B) showing SMA connectors at the input and output.
desired filter response and descriptions of methods for finding the design
dimensions from this mathematical model are now discussed.

A typical band pass frequency response is shown in Figure 2.1.5. The filter passes
frequencies in the pass band which is defined with a 3 dB bandwidth, denoted BW. All frequencies within this bandwidth pass through the filter with less than
or equal to 3 dB of attenuation relative to the maximum of the pass band
response. The center of this bandwidth is the center frequency, \( f_c \), which is half
way between the pass band edges. As frequency deviates farther in either
direction from the 3 dB bandwidth, the amount of attenuation, A, increases. The
amount of attenuation at a certain frequency is determined by the number of
resonators, or poles, of the filter. The non-resonant frequencies of the filter
resonators cannot propagate easily through the resonators of a filter. The function
of attenuation versus frequency of a filter response is the rolloff of the filter
response. The rolloff controls the slope of the response from the 3 dB bandwidth
point to \( f_{stop} \). Three main types of filter responses exist, the Tchebyscheff, the
Bessel, and the Butterworth, or maximally flat response. Maximally flat
responses ideally have zero ripple in the pass band, but a slower rolloff than a
Tchebyscheff response, for a specified number of poles. Bessel filters maintain
linear phase in the pass band. Actual maximally flat responses may exhibit some
ripple due to errors in fabrication. Normally, a fast rolloff, or high number of
poles, is desired. However, the size of a filter increases with the number of poles,
which makes the filter harder to tune. Tuning fabricated filters is described in
Section 3.5.
Figure 2.1.5: Typical band pass filter response showing center frequency ($f_c$), 3 dB bandwidth ($BW$), stop band attenuation ($A$), and stop band frequency ($f_{stop}$).
The filter design method for microstrip filters presented here uses a lumped-element, low-pass prototype filter to derive the necessary band pass transmission characteristics. A mathematical 'mapping' is made between the transmission characteristics of the desired band pass response and those of the low-pass prototype [4]. The low-pass prototype parameters are lumped element values (capacitances and inductances). These capacitances and inductances are paired and act as resonators (see Figure 2.1.6). The prototype parameters are then converted to inverter parameters, which are then converted to mode impedances. Inverter parameters and mode impedances are described subsequently. These mode impedances determine the dimensions of the filter resonator pairs. Lengths are then determined by the wavelength in the microstrip resonator. The following discussion describes each of these steps.

The first step is to convert the desired transmission filter response parameters to those of a low-pass prototype. Maximally flat responses are discussed in this thesis since it was unclear at the time of modem design whether ripple in a Tchebyscheff filter passband would adversely affect the performance of the modem circuits or the linear phase characteristic of a Bessel filter would be very advantageous. A maximally flat filter response is a good compromise between the Tchebyscheff and Bessel filter responses. The maximally flat frequency response of a low-pass prototype is shown in Figure 2.1.7. The frequency, $\omega_c$, is the pass band edge of this response. This response can be expressed mathematically as [4]
Figure 2.1.6: Low-pass prototype circuit showing $g$-parameters corresponding to the lumped elements and the input and output loads.
Figure 2.1.7: Low-pass prototype filter response showing pass band edge, $w'_1$, and attenuation at the passband edge, $L_{ar}$. 
\[ A(\omega') = 10 \log \left[ 1 + e^{\frac{[\frac{\omega'}{\omega_1}]^{2n}}{10^{\frac{L_{ar}}{10}} - 1} \right] \text{ dB} \]  

where \[ e = 10^{\frac{L_{ar}}{10}} - 1 \].

\[ A(\omega') \] is the attenuation at the frequency \( \omega' \), and \( n \) is the number of poles. In this research, as well as in most cases, \( \omega_1 \) is defined as the 3 dB pass band edge point [4]. Therefore, \( L_{ar} \) is 3 dB. The desired band pass response is determined by a low pass to band pass mapping, defined by the following equations:

\[ \frac{\omega'}{\omega_1} = \frac{2(f-f_c)}{(bp) f_c} \]  

with \[ bp = \frac{BW}{f_c} \].

which is sometimes expressed as a percentage, in which case it is referred to as a percentage bandwidth instead of a fractional bandwidth. Equation 2.1.19 is recommended for \( bp \leq 0.15 \). It is important to note that this mapping procedure does not take into account the frequencies on the lower side of the band pass response. However, these can be ignored for \( bp \leq 0.15 \). The \( \frac{\omega'}{\omega_1} \) ratio is the desired result of this mapping and is dimensionless. All frequencies must have the same units.

The low-pass prototype element parameters can be calculated from the following equations when \( L_{ar} = 3 \text{ dB} \), and \( g_0 \) is normalized to 1. The following equations
were derived from an analysis of ladder filters [4]. Ladder filters get their name from the fact that the circuit looks like a ladder turned on its side. Though a 'G' or 'g' is normally used to denote a conductance, the low pass element values described here denote several different element types. The quantity $g_0$ is the source resistance and $g_{n+1}$ is the load resistance if $g_n$ is a capacitance or is the load conductance if $g_n$ is an inductance. The element type of $g_n$ depends on the number of poles of the filter. The quantity $\omega'_n$ is defined as 1 [4] for ease in computation.

$$g_0 = g_{n+1} = 1 \quad [2.1.21]$$

$$g_\xi = 2 \sin \left[ \frac{(2\xi - 1) \pi}{2n} \right]; \quad \xi = 1, 2, 3, \ldots n \quad [2.1.22]$$

The quantities denoted by $g_\xi$ are inductance or capacitance values. These equations result in symmetrical values for the low-pass element values. Thus $g_1 = g_n$, $g_2 = g_{n-1}$, etc., where $n$ is the number of poles and can be obtained from equation 2.1.17.

It is convenient to convert these low-pass prototype element values to forms which use only capacitances or only inductances. Conversion is accomplished with impedance or admittance inverters. An inverter acts like a quarter wavelength line having a characteristic admittance at all frequencies [4]. Thus, a 90° phase shift occurs in the inverter, and in the case of admittance inverters, all inductive elements look like capacitive elements. In the case of filters with parallel-coupled
strip resonators, the admittance inverters, denoted by \( J_{xy} \), are given by the following equations [4], where \( Z_o \) is the characteristic impedance of the I/O lines. The \( x \) and \( y \) subscripts on \( J \) correspond to the numbers of the coupled resonators.

\[
Y_o = \frac{1}{Z_o} \tag{2.1.23}
\]

\[
J_{01} = Y_o \left[ \frac{\pi (bp)}{2 g_0 g_1} \right]^{\frac{1}{2}} \tag{2.1.24}
\]

\[
J_{j,j+1} |_{j=1}^{n-1} = Y_o \frac{\pi (bp)}{2 \sqrt{g_j g_{j+1}}} \tag{2.1.25}
\]

\[
J_{n,n+1} = Y_o \left[ \frac{\pi (bp)}{2 g_n g_{n+1}} \right]^{\frac{1}{2}} \tag{2.1.26}
\]

The admittance inverters just described can be used to derive the parameters needed to develop any parallel-coupled strip filter. In the case of microstrip filters, the admittance parameters defined above each correspond to a resonator pair. The odd and even mode impedances, \( Z_{oo} \) and \( Z_{oe} \), for each resonator pair can be determined from these admittance parameters. The relations for parallel-coupled strip resonators are as follows [7]:

\[
(Z_{oe})_{j,j+1} |_{j=0}^{n} = Z_o \left[ 1 + \frac{J_{j,j+1}}{Y_o} + \left( \frac{J_{j,j+1}}{Y_o} \right)^2 \right] \tag{2.1.27}
\]
\[(Z_{oo})_{j,j+1}^{n} = Z_o \left[ 1 - \frac{J_{j,j+1}}{Y_o} + \left( \frac{J_{j,j+1}}{Y_o} \right)^2 \right] \]  \hspace{1cm} [2.1.28]

If resonator lengths are ignored, these mode impedances are related to the characteristic impedance of a single line by the following expression:

\[Z_o = \sqrt{Z_{oe}Z_{oo}} \]  \hspace{1cm} [2.1.29]

However, \(Z_{oe}\) and \(Z_{oo}\) are dependent on the admittance inverters, denoted by \(J\), which by definition act as a quarter wavelength lines. Therefore, \(Z_{oe}\) and \(Z_{oo}\) are dependent on resonator lengths, and substitution of equations 2.1.27 and 2.1.28 into equation 2.1.29 will not yield an exact value of \(Z_o\).

The resonator pair width and spacing dimensions can now be determined from these mode impedances. As stated earlier, the width and spacing dimensions for each resonator pair are expressed as design ratios, \(\frac{W}{h}\) and \(\frac{S}{h}\). There are several different approaches in the literature for finding the design ratios, \(\frac{W}{h}\) and \(\frac{S}{h}\), from the equivalent even and odd mode impedances. Several references [5,6,8,9,12] discuss a method utilizing graphs of the odd and even mode impedances, \(Z_{oe}\) and \(Z_{oo}\), or amount of coupling [11] plotted against the design ratios. These graphs are obtained from equations relating the various capacitances present in microstrip to the design ratios and mode impedances. The capacitances include those occurring from the top, bottom, and edges of a microstrip resonator to the top and edge of
an adjacent resonator and to the ground plane. The equations are complex and require complicated iterative procedures to solve for the desired design ratios. For this reason, plots are usually made with data obtained from evaluation of these equations for several sets of mode impedances and substrate dielectric constants. A search for design ratios which simultaneously yield both desired mode impedances is required, followed by interpolation between different sets of curves for different substrate dielectric constants [5,6,8,9,12,29]. This is a widely accepted design procedure, though it depended on the human eye to obtain values off of plots and is therefore inaccurate. In addition, the equations implemented in this procedure are based on theoretical results only. The results of this method have not been compared to a large set of experimental data which includes a wide range of substrate thicknesses and frequencies.

There is also a method which gives negative results for resonator widths [9], which is unacceptable. Since the equations in [9] are derived only by the author of the paper, and are not referenced by other sources, it is difficult to find any errors that may exist in the equations. The derivation is not given in [9].

The third method [3,8] seems to be the only practical, accurate, and theoretically derived method for finding the design ratios, $\frac{W}{h}$ and $\frac{S}{h}$, available in the literature. This method is referenced by authors of more recent papers. This method is now described.

As an intermediate step, the odd and even mode impedances for each resonator
pair are converted to equivalent single line mode impedances which are the mode impedances of a resonator pair as if it was a single microstrip line. The relations are given by [3,8]

\[ Z_{oee} = \frac{Z_{oee}}{2}, \quad [2.1.30] \]

and

\[ Z_{oee} = \frac{Z_{oo}}{2}. \quad [2.1.31] \]

The equivalent single line impedances are used to calculate an equivalent single line \( \frac{w}{h} \) ratio for each resonator pair, for both odd and even modes of propagation. The resulting \( \frac{w}{h} \) ratios are denoted \( \left[ \frac{w}{h} \right]_{oo} \) and \( \left[ \frac{w}{h} \right]_{oe} \). These ratios are calculated from equations 2.1.9 through 2.1.12, or more easily from 2.1.13 through 2.1.14. The impedance \( Z_o \) is replaced by \( Z_{oee} \) and \( Z_{oo} \), respectively, in both equation sets.

Two simultaneous equations are then solved to determine the design ratios, \( \frac{w}{h} \) and \( \frac{s}{h} \), which are dimensionless. An iterative method can be used to solve the following equations [3,8] for the optimal solution which satisfies the equations within a specified accuracy.

\[ \left[ \frac{w}{h} \right]_{ee} = \frac{2}{\pi} \cosh^{-1} \left[ \frac{2H-G+1}{G+1} \right] \quad [2.1.32] \]

\[ \left[ \frac{w}{h} \right]_{oo} = \frac{2}{\pi} \cosh^{-1} \left[ \frac{2H-G-1}{G-1} \right] + \frac{8}{\pi (\epsilon + 2)} \cosh^{-1} \left[ 1 + 2 \left( \frac{w}{h} \right) \right]; \quad \epsilon, \leq 6 \quad [2.1.33] \]
\[
\begin{align*}
\left[ \frac{w}{h} \right]_{s_\circ} &= \frac{2}{\epsilon_r} \cosh^{-1} \left[ \frac{2H - G - \frac{1}{2}}{G - 1} \right] + \frac{1}{\epsilon_r} \cosh^{-1} \left[ 1 + 2 \cosh \left( \frac{\frac{w}{h}}{2} \right) \right]; \quad \epsilon_r \geq 6 \quad [2.1.34]
\end{align*}
\]

where
\[
G = \cosh \left( \frac{\pi}{2} \frac{s}{h} \right) \quad [2.1.35]
\]

and
\[
H = \cosh \left( \pi \frac{w}{h} + \frac{\pi}{2} \frac{s}{h} \right) \quad [2.1.36]
\]

An approximation to \( \frac{s}{h} \) can be made without use of an iterative method \([12,5]\). This is obtained by ignoring the second term of equations 2.1.33 and 2.1.34, and solving the two truncated equations for \( \frac{s}{h} \). The following equation results:

\[
\frac{s}{h} = \frac{2}{\epsilon_r} \cosh^{-1} \left[ \frac{\cosh[\text{ARG}_{se}] + \cosh[\text{ARG}_{s}] - 2}{\cosh[\text{ARG}_{s}] - \cosh[\text{ARG}_{se}]} \right]; \quad [2.1.37]
\]

where:
\[
\text{ARG}_{se} = \frac{1}{2} \pi \left[ \frac{w}{h} \right]_{s_\circ} \quad [2.1.38]
\]

and
\[
\text{ARG}_{s} = \frac{1}{2} \pi \left[ \frac{w}{h} \right]_{s_\circ} \quad [2.1.39]
\]

The ARG quantities are arguments of the hyperbolic cosine functions. The \( \frac{w}{h} \) ratio can then be determined by 2.1.31, 2.1.32, or 2.1.33.

Both the iterative method and the approximation method result in the design ratios, \( \frac{w}{h} \) and \( \frac{s}{h} \). Multiplying these results by the substrate thickness \( (h) \) yields
the widths and spacing for each resonator pair.

The length of the resonator pair segments, $L_r$, are one quarter wavelength long. The wavelength corresponds to the center frequency of the filter. The length of the resonator is one quarter wavelength calculated with use of a frequency dependent effective dielectric constant. The following equation is similar to equation 2.1.1:

$$L_r = \frac{\lambda_s}{4} = \frac{c}{4f \sqrt{\varepsilon_{eff}(f)}}$$  \hspace{1cm} [2.1.40]

Equations 2.1.24 through 2.1.28 are recommended for fractional bandwidths of up to 0.15 [4]. Parallel coupled microstrip filters can be used at a wide range of frequencies, from hundreds to thousands of megahertz. Resonators designed for the UHF band tend to get too large for practical implementation in small equipment such as portable radios. Other methods exist for realizing microstrip filters at these lower frequencies [1,30,31].

The remainder of this section discusses several equations for effective width and frequency dependent effective epsilon for coupled microstrip lines.

Effective width equations are given in Section 2.1.1 for a single microstrip line (equations 2.1.15 and 2.1.16). There exists another set of equations in the literature for determining $w*$ of coupled microstrip lines. The effective width equations for coupled microstrip lines given in [3,32] include one equation each for
the odd and even propagation modes. The equations from [3,32] can be written in
the following form:

\[ w_{e; \text{even}} = w + w_e \left[ 1 - 0.5 \exp\left[-0.69 \frac{w_e}{\Delta t}\right] \right] \quad [2.1.41] \]

or

\[ w_{e; \text{odd}} = w_{e; \text{even}} + \Delta t, \quad [2.1.42] \]

where

\[ \Delta t = \frac{t \cdot h}{s \cdot \epsilon_r} \quad [2.1.43] \]

and \( w_e \) is calculated from equations 2.1.15 and 2.1.16. Units for \( w_e \) of both modes,
\( s \), and \( \Delta t \) are equivalent to the units of \( w_e \) of a single line, explained in Section
2.1.1. These expressions are valid for \( s \geq 2t \). Spacings less than twice the strip
thickness are impractical.

The equations for \( \epsilon_{eff}(f) \) in [3] are intended for coupled microstrip lines. Two
equations are given, one each for the even and odd modes of propagation. Both
even and odd mode frequency dependent effective dielectric constants are
calculated from these equations. A relation is required which relates both of these
mode dielectric constants to one equivalent dielectric constant which does not
depend on the propagation modes. This relation is required to obtain the
resonator length defined in equation 2.1.40. However, no relation exists which
relates these even and odd mode effective dielectric constants to one equivalent
effective dielectric constant.
2.1.3 LENGTH CORRECTION FACTORS

A length correction factor as discussed in this thesis refers to the addition or subtraction of conductor material on the ends of each resonator of the microstrip filter. The adjustment of the resonator length affects the resonant frequency of the resonator, and therefore the center frequency of the filter.

A correction factor cited in [4] accounts for fringing capacitance at the ends of each resonator. This correction factor is a distance, d, which should be subtracted from each end of all resonators, and is given by

\[ d = 0.165 \, h \]  \hspace{1cm} [2.1.44]

Though [4] does not state this point, equation 2.1.44 is based on experimental results obtained from just one filter of the stripline type [10]. Stripline is similar to microstrip, except there is substrate and a ground plane on both the top and bottom of the filter resonators. The resonators are sandwiched between the dielectric material. Therefore, this correction factor could differ for a microstrip filter. Experimental results published by [11] give a correction factor (equation 2.1.45) for microstrip filters. It seems as though the result of [11] is simply double the result of [10]. However, no information was given for the number of filters that were used to determine this experimentally determined correction factor.

\[ d = 0.33 \, h \]  \hspace{1cm} [2.1.45]
In references [4,10,11] where the two correction factors given above are found, no information is given concerning what equations are used to determine the uncorrected resonator lengths, and therefore it is unknown what type of $\epsilon_{eff}$ was used in the calculations.

Fringing capacitance at the edges of the resonators is accounted for in the iterative equations, 2.1.32 through 2.1.36, developed in [8]. However, these account for the fringing capacitance between resonators, and not the fringing capacitance at the resonator ends.

Another correction factor is based on the actual electrical length of a microstrip line [25]. The additional effective electrical length ($\Delta l$) beyond the mechanical length of any single open circuit microstrip is

$$\frac{\Delta l}{h} = 0.412 \left[ \frac{\epsilon_{eff} + 0.3}{\epsilon_{eff} - 0.258} \right] \left[ \frac{w}{h} + 0.262 \right] \left[ \frac{w}{a} + 0.813 \right].$$ \[2.1.46\]

However, this equation applies for a stand alone microstrip line only, and does not consider coupling between resonators.

It is important to note that many discrepancies exist between different references on microstrip line equations. In cases where there were discrepancies and the original reference was not available, the following decision was made to determine which equation should be used. In most cases, if the equation was used in more
than one reference, that equation was used instead of a slightly different equation given in only one other reference. Many references state the importance of microstrip in microwave circuits, yet it seems not enough attention is given in reporting correct, consistent microstrip design equations.

2.1.4 INSERTION LOSS

The insertion loss is the minimum loss present in the transmission response of a filter (see Figure 2.1.8). The minimum loss value is normally near the center frequency of the filter, and is given in dB. Insertion loss varies according to several factors, including dielectric constant, substrate thickness, frequency of operation, conductor material, and errors in filter fabrication [2,5,6]. This section outlines several procedures for calculating the theoretical insertion loss.

The methods used to calculate the theoretical insertion loss can be divided into two main categories. The first category contains those methods which account for the properties of the substrate and the second category contains those that do not account for these properties. A method that utilizes substrate parameters intuitively makes more sense. However, both types of methods are included in the following discussions.

The quality factor, Q, of a resonator is $2\pi$ times the ratio of the maximum instantaneous energy stored in the circuit divided by the energy dissipated per cycle [27]. The unloaded Q, denoted $Q_u$, of a resonator is the Q when the
Figure 2.1.8: Band pass response showing insertion loss ($L_i$) relative to $S21 \ 0 \ dB$ reference.
resonator is not loaded on either side by, or adjacent to another resonator. The insertion loss of a filter is a function of the $Q_u$ of all filter resonators. The $Q_u$ of the $k^{th}$ reactive element of the low pass prototype is denoted by $(Q_u)_k$. The $k^{th}$ reactive element of the low pass prototype corresponds to the $k^{th}$ resonator of the band pass filter. An equation given by [4] relates $(Q_u)_k$ to the midband unloaded $Q$ of the $k$th resonator of the band pass filter, denoted by $(Q_{u,up})_k$.

\[
(Q_u)_k = bp \times (Q_{u,up})_k \tag{2.1.47}
\]

where $bp$ is the fractional bandwidth. However, $(Q_u)_k$ is used in all theoretical insertion loss calculations. The total insertion loss can be calculated when the $(Q_u)_k$ for all $k$ are known. Two methods, referred to as: METHOD A: $Q_u$ AS A FUNCTION OF SUBSTRATE PARAMETERS and METHOD B: $Q_u$ INDEPENDENT OF SUBSTRATE PARAMETERS are used to calculate $(Q_u)_k$. No advantages are given in the literature for one method over the other. However, the equations implemented in METHOD A are found in several references, whereas the equations in METHOD B are only found in [4].

METHOD A: $Q_u$ AS A FUNCTION OF SUBSTRATE PARAMETERS

When the substrate characteristics are included, the calculation of the theoretical insertion loss includes three factors: conductor loss, dielectric loss, and radiation loss. Conductor loss occurs due to dissipation of electromagnetic energy in the conductor material. Likewise, dielectric loss occurs due to dissipation in the
dielectric. Radiation loss occurs due to dissipation of energy to the air at open stubs, bends, and discontinuities of the microstrip. In the following equations, \( Q_e \) is the quality factor associated with conductive loss, \( Q_d \) is the quality factor associated with dielectric loss, and \( Q_r \) is the quality factor associated with the radiation loss of microstrip line. The following equations from \([2,5,6]\) can be used to calculate \( (Q_u)_k \):

\[
\frac{1}{(Q_u)_k} = \frac{1}{Q_e} + \frac{1}{Q_d} + \frac{1}{Q_r} \tag{2.1.48}
\]

and

\[
\frac{1}{Q_e} + \frac{1}{Q_d} = \frac{\lambda_0(\alpha_e + \alpha_d)}{\pi \sqrt{\varepsilon_{\text{eff}}}}, \tag{2.1.49}
\]

where \( \lambda_0 \) is the free space wavelength in cm corresponding to the center frequency \( (f_c) \), \( \alpha_e \) is the conductive loss, \( \alpha_d \) is the dielectric loss, and \( \varepsilon_{\text{eff}} \) is defined by equation 2.1.6 or 2.1.7. The losses, \( \alpha_e \) and \( \alpha_d \), are \([2,5,6,33]\) given by the following equations.

\[
\alpha_e = \frac{8.68 \frac{R_s PS}{2\pi L_0 h}}{2\pi} \text{ dB/cm} \quad \frac{1}{\frac{w}{h}} < 2 \tag{2.1.50}
\]

where

\[
R_s = \sqrt{\pi f_c \mu_0 \sigma} \quad \Omega, \tag{2.1.51}
\]

\[
P = 1 - \frac{w^2}{4h} \quad ; \quad \frac{w}{h} \leq 0.005, \tag{2.1.52}
\]

and

\[
S = 1 + \frac{h}{w} + \frac{h}{\pi w} \left[ \ln \left( \frac{2h}{w} \right) - \frac{1}{2} \right] \quad ; \quad \frac{w}{h} \leq 0.005 \tag{2.1.53}
\]
\[ \alpha_d = 4.34 \frac{\left( \varepsilon_{\text{eff}} - 1 \right) \sigma}{\sqrt{\varepsilon_{\text{eff}} (\varepsilon_r - 1)}} \sqrt{\frac{\mu_0}{\varepsilon_0}} \text{ dB/cm} \]

\[ = 27.3 \frac{\varepsilon_r (\varepsilon_{\text{eff}} - 1)}{\sqrt{\varepsilon_{\text{eff}} (\varepsilon_r - 1)}} \frac{\tan \delta}{\lambda_0} \text{ dB/cm} \]  \hspace{1cm} \text{[2.1.54]}

where \( R_s \) is called the surface resistivity, \( \sigma \) is the conductor conductivity, \( h \) is in cm, \( \mu_0 \) is the free space permeability which is equal to \( 4\pi \times 10^{-9} \text{ henries/cm} \), \( \tan \delta \) is the loss tangent of the dielectric, and \( P \) and \( S \) are constants. Values of conductivities for several conductor materials are given in Table 2.1.1. The loss tangent is sometimes referred to as the dissipation factor, and is given by \[ 16 \]

\[ \tan \delta = \frac{\sigma_d}{\omega \varepsilon_0} \]  \hspace{1cm} \text{[2.1.55]}

where \( \sigma_d \) is the conductivity of the dielectric in siemens/cm, \( \varepsilon_0 = \varepsilon_r \varepsilon_0 \) in Farads/cm, and \( \omega \) is substituted by the \( 2\pi f_c \) (Hz). The free space permittivity, \( \varepsilon_0 \), is \( 8.854 \times 10^{-14} \text{ farads/cm} \).

The radiation quality factor can be calculated from \[ 2.5, 6, 34 \]

\[ Q_r = \frac{Z_0}{480 \pi (h/\lambda_c)^2 F} \]  \hspace{1cm} \text{[2.1.56]}

where

\[ F = \frac{\varepsilon_{\text{eff}} + 1}{\varepsilon_{\text{eff}}} - \frac{(\varepsilon_{\text{eff}} - 1)^2}{2(\varepsilon_{\text{eff}})^{2/3}} \ln \left[ \frac{\sqrt[3]{\varepsilon_{\text{eff}} + 1}}{\sqrt[3]{\varepsilon_{\text{eff}} - 1}} \right] \]  \hspace{1cm} \text{[2.1.57]}

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Table 2.1.1: Conductivities of some conductor materials

<table>
<thead>
<tr>
<th>Conductor</th>
<th>Conductivity $[\text{siemens/cm}]$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Aluminum</td>
<td>$3.54 \times 10^5$</td>
</tr>
<tr>
<td>Copper</td>
<td>$5.80 \times 10^5$</td>
</tr>
<tr>
<td>Gold</td>
<td>$4.10 \times 10^5$</td>
</tr>
<tr>
<td>Silver</td>
<td>$6.17 \times 10^5$</td>
</tr>
</tbody>
</table>
with $Z_0$ in ohms, and $h$ and $\lambda_0$ have equivalent units. Radiation losses increase with frequency and substrate thickness. Radiation losses are greater than conductor and dielectric losses at higher frequencies ($>10$ GHz) unless the dielectric height is fairly small ($<10$ mils). In general, dielectric losses are less than conductor losses. Conductor loss decreases with dielectric height and increases with the square root of frequency. $(Q_u)_k$ is generally inversely proportional to dielectric thickness [2,5,6].

METHOD B: $Q_u$ INDEPENDENT OF SUBSTRATE PARAMETERS

The following procedure for calculating $(Q_u)_k$ does not account for any substrate parameters. The unloaded $Q$ $(Q_u)$ can be calculated when the external $Q$ $(Q_e)$ of a resonator is known. The external $Q$ of a resonator is defined as the $Q$ of a resonator when $Q_u = \infty$ [4].

Since lumped-elements are difficult to construct at microwave frequencies, the resonators are realized in distributed-element forms rather than the lumped-element form of Figure 2.1.6. As a basis for establishing the resonance properties of a resonator regardless of its form, a resonant frequency and an admittance or susceptance slope parameter are specified. For microstrip resonators, a susceptance slope parameter is specified since microstrip resonators are modelled in terms of capacitances only [3,4]. This model was explained previously in the discussion of admittance parameters, denoted by $J$. The susceptance slope parameter, denoted by $b$, is defined as the quality factor, $Q$, multiplied by the
shunt conductance, $G$. The following equations relate $Q_s$ to $Q_e$ [4].

\[
\frac{J}{G_s} = \sqrt{\frac{b}{G_s Q_e}} \quad [2.1.58]
\]

\[
Q_t = \frac{1}{(Q_s)_1} + \frac{1}{Q_u} + \frac{1}{(Q_s)_2} = \frac{f_e}{(\Delta f)_3_{dB}}, \quad [2.1.59]
\]

where $G_s$ is the shunt conductance of the load (siemens), which is equivalent to $Y_o = \frac{1}{Z_o}$, where $Z_o$ is the characteristic impedance of the microstrip line (ohms). The subscripts 1 and 2 in equation 2.1.59 denote loading on each side of the resonator, $f_o$ is the resonant frequency, $Q_t$ is the $Q$ when the resonator is loaded on both sides, and $(\Delta f)_3_{dB}$ is the 3 dB bandwidth of the resonator with the same units as $f_o$. The susceptance slope parameter, $b$, of all microstrip resonators can be derived as

\[
b = \frac{Y_o \pi}{2} \quad [2.1.60]
\]

and substituting equation 2.1.60 in equation 2.1.58,

\[
\frac{J}{Y_o} = \sqrt{\frac{\pi}{2Q_e}} \quad [2.1.61]
\]

These equations can be extended for all resonators and written as follows:

\[
(Q_s)_{j,j+1} = \frac{\pi}{2 \left[ J_{j,j+1} Z_o \right]^3}; \quad k=0,n \quad [2.1.62]
\]

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\[
\frac{1}{(Q_u)_k} = b p \left[ \frac{1}{(Q_u)_{k-1,k}} + \frac{1}{(Q_u)_{k,k+1}} \right]; \ k = 1 \ to \ n
\]  

[2.1.63]

Therefore, \((Q_u)_{j,j+1}\) for \(j=0\) to \(n\) are calculated from equation 2.1.62 where the \(J_{j,j+1}\) are given by equations 2.1.23 through 2.1.26, and \(Q_u\) are found for all resonators. The quantity \(b p\) is the fractional bandwidth.

The following methods can be used to calculate the total theoretical insertion loss after the \((Q_u)_k\), calculated from METHOD A or B above, for all resonators is known. The total insertion loss is denoted by \(L_i\). There are no advantages given in the literature for one method over another. Differences in resulting theoretical calculations are discussed in Section 2.1.4.

**DISSIPATION FACTOR METHOD:**

The theoretical insertion loss can be derived from the following equations [4]:

\[
L_i = 4.343 \sum_{k=1}^{n} d_k \ g_k \]  

[2.1.64]

\[
d_k = \frac{\omega'_1}{(Q_u)_k} \]  

[2.1.65]

as

\[
L_i = \frac{4.343}{b p} \sum_{k=1}^{n} \frac{g_k}{(Q_u)_k} \]  

[2.1.66]

where \(d_k\) is referred to as the dissipation factor of the \(k^{th}\) resonator with the same
units as $\omega_1$, $g_k$ is the low pass prototype element value, $b_p$ is the fractional bandwidth, $\omega_1$ is the pass band edge of the low-pass prototype response, and $n$ is the number of poles. The quantity $\omega_1$ is defined as 1 rad/sec [4].

TIME DELAY METHOD:

Another method for calculating the theoretical insertion loss is [4]

$$L_i = 20 \log(C_n d + 1) \text{ dB}$$  [2.1.67]

where

$$C_n = \frac{1}{2} \sum_{k=1}^{n} g_k$$  [2.1.68]

and $d$ is the dissipation factor as given by equation 2.1.65 in rad/sec, but is assumed to be the same for all resonators. The quantity $C_n$ is the group time delay in seconds. Group time delay is the time required for a signal to pass through the filter. An approximation to equation 2.1.67 is [4]

$$L_i \simeq 8.686 \ C_n \ d.$$  [2.1.69]

Qe METHOD:

Each resonator has attenuation at resonance. This attenuation is denoted by $(L_i)_k$, and is found as follows [4]:

49
\[(L_i)_k = 10 \log \left[ \frac{(Q_1)(Q_2)}{4 Q_i^2} \right] \text{dB} \quad [2.1.70]\]

where \(Q_i\) is given by equation 2.1.59, and \((Q_1)\) and \((Q_2)\) are described in the TIME DELAY METHOD discussion. This equation can also be extended for all resonators;

\[(L_i)_k = 10 \log \left[ \frac{(Q_e)_{k-1,k} (Q_e)_{k,k+1}}{4 Q_i^2} \right]; \ k=1 \text{ to } n \quad [2.1.71]\]

The total insertion loss is then given by

\[L_i = \sum_{k=1}^{n} (L_i)_k \quad [2.1.72]\]

This concludes the methods for calculation of total insertion loss when the Q of each filter resonator is known. Still other equations exist which predict the insertion loss of filters. For instance, in [3], the loss equations are separated for the even and odd modes of propagation. These equations predict the conductor and dielectric losses for coupled microstrip lines. No equation exists in [3,12] or its references which relates the theoretical losses for the even and odd modes to the total loss for the coupled lines. The authors of [3] maintain that an approximation for the total losses can be calculated as the average loss of the even and odd modes, if the characteristic impedance is 50 \(\Omega\). This is very restrictive and is only an approximation.

Loss equations given by [3,12] are in terms of the capacitances associated with
coupled microstrip lines whereas the equations given above are in terms of the \( \frac{W}{h} \) ratio. A complex design procedure discussed in Section 2.1.2 involves the calculation of these capacitances. The design procedure implemented in this research does not include the calculation of these capacitances. For these reasons, this loss prediction method is not implemented in any design procedure in this thesis.

The effective width, \( w_e \), is substituted for \( w \) in equations 2.1.15 and 2.1.16 in some of the references [2,5]. This substitution does not yield significantly different results since \( w_e \approx w \) as shown in Section 2.2.

2.1.5 REFLECTION RESPONSE

All previous sections of this chapter have discussed parameters which affect the transmission response of the filter. In this section, the reflection response is discussed.

The incident wave to the input of a two port network such as a microstrip filter has two resulting components: a reflected wave and a transmitted wave. The transmitted wave passes through the filter and is attenuated according to the transmission filter response. The reflected wave is the part of the incident wave which is not transmitted through the filter [4]. Over the passband of the filter, the reflected wave is normally small in comparison to the incident or transmitted wave. The passband of the reflection response is ideally equivalent to the
bandwidth of the transmission response. The return loss of a filter is a measure of how much of the incident signal is reflected back to the input. A high return loss corresponds to a low reflected signal magnitude and high transmitted signal magnitude.

The amount of return loss is dependent on the impedance match of the circuit. If the impedance at the input is exactly equal to the impedance at the output, the impedances are said to be matched, and the reflected signal magnitude is zero and the return loss is infinite. And therefore, the transmitted signal is equal to the incident signal. This is an ideal situation which holds for a lossless two port network [4]. If the filter has some insertion loss in the passband, the filter cannot be considered lossless and the transmitted signal is not necessarily equal to the incident signal, even if the source and load impedances are equal. If the insertion loss is due to dissipation in the conductor material and dielectric material only, then the reflected signal magnitude is dependent on the impedance match only. However, the insertion loss experienced in microstrip filters is due to discontinuities as well as conductor and dielectric losses, as described in Section 2.1.4. These discontinuities cause some of the incident signal to be reflected, causing the reflected signal magnitude to increase [4]. Therefore, the reflected signal and return loss are not only a function of the impedance match, but also of the discontinuities present in the filter construction.

The reflected signal is normally measured as a ratio of reflected signal to incident signal, and is called the S11 response (see Section 3.4 for more detail). A typical
S11 response is shown in Figure 2.1.9. The ideal response has no ripples in the pass band. However, if there exist slight errors in fabrication, the S11 response will likely show a downward ripple or spike for each pole.
Figure 2.1.9: Typical band pass $S_{11}$ (reflection) response.
2.2 DESIGN PROCEDURE FOR MICROSTRIP BAND PASS FILTERS

The goal in designing a microstrip bandpass filter is to find the widths, spacings, and lengths of all filter resonator pairs, as well as the width of the I/O lines. These dimensions depend on several substrate characteristics as well as the desired filter response, as described in Section 2.1. The design procedure given here is based on that developed in Section 2.1.2, from equations in the literature. Thus, all dimensions are calculated using equations, or variations (stated below) of equations given in Section 2.1.

The required parameters for designing a microstrip bandpass filter are:

- Center frequency \( f_c \) in MHz
- 3 dB Pass Band Bandwidth \( \text{BW} \) in MHz
- Stop Band Attenuation \( a \) in nepers, or \( A \) in dB
- Stop Band Frequency \( f_{stop} \) in MHz
- Characteristic Impedance \( Z_0 \) in ohms
- Relative Epsilon \( \varepsilon_r \) of the substrate
- Substrate Thickness \( h \) in mils
- Conductor Thickness \( t \) in mils

Two additional parameters, loss tangent \( \tan \delta \) and conductor conductivity \( \sigma \) in siemens/cm, are also included in the design procedure for calculation of theoretical insertion loss, depending on the algorithm used. However, these additional
parameters are not essential to finding the design dimensions.

A typical band pass frequency response is shown in Figure 2.1.5. As described in Section 2.1.2, the desired transmission response is derived from a low pass prototype response. After the desired band pass response is determined, a conversion is made to a low-pass prototype response with equations 2.1.19 and 2.1.20. The frequency variable \( \omega'_1 \) is specified as 1 rad/sec and \( f \) is replaced by \( f_{stop} \) in equation 2.1.19. The variable \( f_{stop} \) is the stop band frequency of the desired band pass response. This is the frequency where the desired attenuation, \( a \) or \( A \) (dB), is needed. This step is referred to as a mapping of filter responses.

The number of poles, or resonators, necessary to achieve the desired rolloff of the filter response is determined by equations 2.1.17 and 2.1.18. \( L_{ar} \) is specified at the pass band edge of the low-pass prototype. Our work uses \( L_{ar} \) equal to 3 dB. The quantity \( a \) is the desired attenuation (not in dB) at \( f_{stop} \), and the ratio \( \frac{\omega'_1}{\omega_1} \) is determined by the mapping function. Equation 2.1.17 can be solved for \( n \), the number of poles required for a desired response, and is rewritten as follows:

\[
n = \text{INT} \left[ \frac{\log \left( \frac{a - 1}{0.9953} \right)}{2 \log \left( \frac{\omega'_1}{\omega_1} \right)} \right] + 1 \tag{2.2.1}
\]

where \( \text{INT} \) denotes integer part and equation 2.1.18 is evaluated with \( L_{ar} = 3 \) dB. See the discussion of equation 2.1.17 in Section 2.1.2 for explanations of the
variables $a$, $\omega'$, and $\omega'$1. The addition of 1 in equation 2.2.1 accounts for the fact that the argument of the INT function will nearly always have a fractional part. Therefore, equation 2.2.1 will evaluate to an $n$ value which will result in an attenuation at least, and never 'almost', as great as the desired attenuation. With the number of poles determined, the low pass prototype element values can be calculated using equations 2.1.21 and 2.1.22. These prototype values, which are capacitances and inductances, are converted to admittance inverters, which define element values in terms of capacitances only. The admittance inverters are obtained with equations 2.1.23 through 2.1.26. There exists one admittance parameter for each resonator pair. Therefore, there are $n+1$ admittance parameters, since there are $n$ resonators in the filter (see Figure 2.1.3).

The odd and even mode impedance values, $Z_{oo}$ and $Z_{eo}$, for a pair of microstrip lines are the equivalent impedance values for the two modes of propagation along a microstrip line. The $n+1$ mode impedance pairs are obtained from the admittance inverters by equations 2.1.27 and 2.1.28. The design ratios, $\frac{W}{h}$ and $\frac{S}{h}$, are then derived from these mode impedances in the following discussion.

The odd and even mode impedances are converted to equivalent single line impedances which are the impedances of a resonator pair as if it were a single microstrip line. Equations for this conversion are 2.1.30 and 2.1.31.

The equivalent single line impedances are used to calculate an equivalent single line $\frac{W}{h}$ ratio for both odd and even modes of propagation, for each resonator pair.
The resulting \( \frac{w}{h} \) ratios are denoted \( \left[ \frac{w}{h} \right]_{se} \) and \( \left[ \frac{w}{h} \right]_{so} \). \( Z_0 \) is replaced by \( Z_{se} \) and \( Z_{so} \) in equations 2.1.13 and 2.1.14 since they are simpler to use than equations 2.1.9–2.1.12. Sample calculations can be made using both sets of equations. A comparison of the results from these two sets yields similar dimension ratios which agree to three decimal places. The maximum thickness substrate used in this research is 100 mils and the minimum accurate width obtainable through the fabrication process is 10 mils. These values correspond to a minimum \( \frac{w}{h} \) ratio of 0.1. Therefore, accuracy to within 0.001 is certainly sufficient. The effective widths are then found with equations 2.1.15 and 2.1.16.

The ratios \( \left[ \frac{w}{h} \right]_{se} \) and \( \left[ \frac{w}{h} \right]_{so} \) are substituted into equations 2.1.32 through 2.1.34 which are then solved simultaneously for the design ratios, \( \frac{w}{h} \) and \( \frac{g}{h} \). Equations 2.1.32 through 2.1.36 are such that they cannot be solved explicitly for either of the two unknowns, or design ratios. Therefore, equations 2.1.32 through 2.1.36 are solved for the design ratios by an iterative method which converges to the optimal solutions of both equations. The iterative method, called the bisection method, is a numerical analysis method which is used to obtain the zeros of a function [35].

The bisection method is chosen for the iteration procedure because it works with all functions as long as a zero exists. If no zero exists within the interval between the two initial limits, the method converges to either of the two initial limits. Other methods, such as the secant or Newton's method [35] may work faster, but do not work for all types of functions, and may not be simple to implement, especially for this set of equations. Regardless of the convergence method used,
the procedure still involves solving essentially two complex equations, for two unknowns.

The bisection method can be explained as follows. Suppose \( f(x) \) is a continuous function on an interval \([a,b]\) and that \( f(a)f(b) < 0 \). Then \( f(x) \) changes sign on the interval \([a,b]\), which implies \( f(x)=0 \) has at least one root on the interval. The procedure for finding the root is to repeatedly halve the interval, and redefine one of the interval limits such that \( f(x) \) still changes sign on the new half interval. This method converges to the root of the function. The method is halted when a desired tolerance, denoted TOL, is reached. The steps for finding the root are as follows:

1. Define \( c = \frac{a+b}{2} \).
2. If \( |b-c| \leq \text{TOL} \) (or \( |a-c| \leq \text{TOL} \)), then accept \( c \) as the root and stop.
3. If \( \text{sign}[f(b)] \text{sign}[f(c)] \leq 0 \), then set the new interval limit as \( a=c \). Otherwise set the new interval limit as \( b=c \).

In this way, the interval \([a,b]\) is halved with each loop through steps 1 to 4. Eventually, the inequality in step 2 is satisfied, and the root is found.

This method is implemented twice in the design procedure, since there are two unknowns to solve for. The two bisection method loops are nested. Two limits, 0.00001 (lower) and 5.0 (upper), are set on both loops for \( \frac{\Delta}{h} \). The tolerance is set
to 0.001. Equation 2.1.35 is evaluated and then equation 2.1.32 is solved for $H$.

Equation 2.1.32 can be rewritten for ease in computation as follows:

$$H = \frac{1}{2} \left[ (G+1) \cosh \left[ \frac{G}{2} \left( \frac{W}{h} \right)^{\sigma} \right] - 1 + G \right] \quad [2.2.2]$$

Now, an estimate for $\frac{W}{h}$ corresponding to the $\frac{S}{h}$ outer loop limit can be found from equation 2.1.36 when written in the following form:

$$\frac{W}{h} = \frac{1}{h} \cosh^{-1}[H] - \frac{S}{2h} \quad [2.2.3]$$

This $\frac{W}{h}$ estimate is now substituted into equation 2.1.33 or 2.1.34, depending on the specified dielectric constant. Since $\left[ \frac{W}{h} \right] e_0$ is known, a new estimate for $\frac{S}{h}$ can be found. Equation 2.1.33 or 2.1.34 is now iteratively solved in the inner bisection method loop for the new second estimate of $\frac{S}{h}$ until error is minimized to 0.001. This second estimate is substituted into equations 2.1.35 and 2.1.36 and then equation 2.1.32 is evaluated. This produces a new $\left[ \frac{W}{h} \right] e_0$ which is slightly different from the initial $\left[ \frac{W}{h} \right] e_0$ estimate. The difference between the former and present $\left[ \frac{W}{h} \right] e_0$ ratio was referred to as $\Delta \left[ \frac{W}{h} \right] e_0$. This procedure is repeated for the halved interval value of the outer bisection method loop. A step equivalent to step 3 of the bisection method above is then evaluated with the $\Delta \left[ \frac{W}{h} \right] e_0$ values, and the outer loop continues. This procedure continues until a tolerance of 0.001 is reached for $\frac{S}{h}$ in the outer loop.

In this way, an optimum solution, $\frac{S}{h}$, for the equation set is obtained. During
convergence on $\frac{s}{h}$, $\frac{w}{h}$ is also evaluated for the corresponding $\frac{s}{h}$. Therefore, the equation set is successfully solved for both design ratios, $\frac{w}{h}$ and $\frac{s}{h}$. Multiplying these ratios by the substrate thickness ($h$) yields the widths and spacing for the resonator pair. This iterative procedure is repeated for each resonator pair.

As explained in Section 2.1.2, an approximation to $\frac{s}{h}$ (equations 2.1.37 through 2.1.39) can be made without use of the iterative method described above [3,8]. However, when this approximation is implemented, and the results compared to the more accurate value obtained from the original equations, there is a factor of 3 discrepancy in $\frac{s}{h}$. This discrepancy results when a low dielectric constant value, such as $\epsilon_r = 4.7$, is used which causes the second term of equation 2.1.33 to become significant. Therefore, this approximation is not implemented in the design procedure, and highlights the importance of emphasizing the limits of design equations.

The resonator lengths for each resonator pair, $L_r$, are obtained from equation 2.1.40 for each resonator pair. Finally, the width of the I/O characteristic impedance lines is calculated using equations 2.1.13 through 2.1.14, which are the same equations used to calculate the equivalent single line $\frac{w}{h}$ ratios for a resonator pair. Thus, all dimensions necessary to complete the basic design of a microstrip band pass filter are determined. These dimensions can now be used to determine a filter layout similar to the filter of Figure 2.1.3.

The design procedure given above is intended for filters with fractional
bandwidths less than 0.15. A trial procedure can be made for filters with fractional bandwidths greater than 0.15. In this procedure, equations 2.1.17–2.1.18 and 2.1.21–2.1.22 are replaced by different equations given in [4]. However, when these new equations are implemented in the design procedure, the wider bandwidth designs are not possible because there is no solution found to equations 2.1.32 through 2.1.36. This no solution condition is probably caused by incorrect mode impedances. The trial procedure gives odd and even mode impedances which when used in equation 2.1.29, give $Z_o$ values which are all less than the desired $Z_o$. Normally, the odd mode impedances are smaller and the even mode impedances are larger than $Z_o$. Substitution into equation 2.1.29 should give a $Z_o$ close to the desired $Z_o$. The lower limits on both the inner and outer bisection method loop intervals were changed to the currently used 0.00001 values as an additional check for a no solution condition. These lower limits were implemented in this wider bandwidth design procedure, but still no solution could be obtained. This further points to a problem with incorrect mode impedances, which are determined before the iteration loops.

Thus the original design procedure is not enhanced with this wider bandwidth procedure. It is important to note that the equations for wider bandwidths, and those already being used in the design procedure, are simply for any parallel-coupled lines with half-wavelength resonators. They do not take into account the physical realizability of a certain implementation (microstrip, stripline, etc.) as far as percentage bandwidth is concerned. Physical realizability is determined by the dimensions obtained from the design procedure. For instance, a dimension of less
than 10 mils on a microstrip filter is difficult to obtain accurately when the filter is fabricated. A particular physical implementation of parallel-coupled filters has its own physical realizability limitations. These limitations govern the ability of a particular filter to have narrow or wide bandwidths.

The filter designs of this research use substrates with a maximum substrate thickness, \( h \), of 100 mils and minimum conductor thickness, \( t \), of 0.7 mils. This corresponds to \( \frac{t}{h} = 0.007 \), which is greater than the limit of 0.005 specified in [2,5,23,28] for the zero thickness assumption. Therefore, the conductor thickness is considered in width calculations.

The effective width equations for coupled microstrip lines include one equation each for the odd and even propagation modes (equations 2.1.41–2.1.43). These equations depend on the ratio \( \frac{s}{h} \), which implies that, in the design procedure, these equations could be used after the \( \frac{w}{h} \) and \( \frac{s}{h} \) ratios are found. However, the design procedure requires initial estimates for the \( \left[ \frac{w}{h} \right]_{oe} \) and \( \left[ \frac{w}{h} \right]_{so} \) ratios before finding \( \frac{w}{h} \) and \( \frac{s}{h} \). In other words, these equations depend on two of the inputs, the odd and even mode impedances, and one of the outputs, \( \frac{s}{h} \), of the design procedure described above. The equations for coupled lines can nevertheless be evaluated for a scenario which yields the most width correction for filter parameters used in this research: \( \epsilon_r = 10.5 \), \( t = 1.4 \) mils, \( s = 10 \) mils, \( h = 100 \) mils. The results give a correction of +2.3 mils for the even mode and +3.6 mils for the odd mode. These corrections are close to the correction obtained for a single line from equation 2.1.15, 2.7 mils. Based upon these results and the problem in
implementation of the odd and even mode equations, the single line \( w \) equations, 2.1.15 and 2.1.16, are included in the design procedure.

Two options exist for placement of these effective width equations in the design procedure. As the first option, the equations are placed at the beginning of the procedure to find the effective width of the initial estimate for \( \frac{w}{h} \), and as the second option, are placed at the end of the iteration procedure to find the effective width of the coupled line. Several test filters were designed using the design procedure above, but with implementation of equations 2.1.13 and 2.1.16 after the iteration loops. The results show a decrease in resonator length by less than one mil and an increase in \( \frac{w}{h} \) of about two mils. The resonator length is indirectly a function of \( \frac{w}{h} \), as described in Section 2.1. The ratio \( \frac{s}{h} \) remains unchanged since the \( \frac{w}{h} \) ratio is changed after the iterative loops.

Placing the effective width equations before the iterative loops, and testing the new procedure with the same sample filter set, yields widths and spacings which are all greater by about 3 mils than those given by the original design procedure. The resulting design procedure has the effective width equations placed before the iteration loops since this intuitively seemed to make more sense. Placement before the iteration loops ensures that the \( \frac{s}{h} \) ratio will also be adjusted according to the change in \( \frac{w}{h} \), and makes logical sense since the initial estimate for the equivalent single line design ratio is determined from a single line equation. Placement of the effective width equations before or after the iterative routine does not significantly change in the resulting design dimensions, and therefore the
equations are included for completeness.

Dispersion effects must also be included in the design procedure. Equation 2.1.2 determines the frequency, \( f_d \), where dispersion begins to affect \( \varepsilon_{\text{eff}} \) and \( Z_o \). The value of \( f_d \) is 2.398 GHz for the following typical parameters used in this research: \( \varepsilon_r=10.5 \), \( Z_o=50 \ \Omega \), and \( h=100 \ \text{mils} \). Therefore, dispersion effects are considered in the filter design procedure. The effects of frequency on \( \varepsilon_{\text{eff}} \) are already considered in the equations implemented in the design procedure as explained in Section 2.1.2.

Equations for a frequency dependent characteristic impedance \( Z_o(f) \) are found in the literature [2,5,22], but there is an implementation problem similar to that which arose with the \( w_\ast \) odd and even mode equation set. The equations for \( Z_o(f) \) depend on the width and spacing of the resonators, obtained after the iteration procedure, yet \( Z_o \) is required to be defined before the iterative procedure. In addition, the effect of frequency on \( Z_o \) up to 20 GHz for single microstrip lines is reported as negligible [2,5]. For these two reasons, equations for \( Z_o(f) \) are not included in the design procedure.

Several filters can be cascaded together to obtain a steeper rolloff response and/or greater attenuation at stopband. Due to the definition of 3 dB bandwidth and the non-ideal response of filters, the overall equivalent bandwidth of \( n \) cascaded filters having identical bandwidth, \( \text{BW} \), is less than the bandwidth, \( \text{BW} \), of one of these \( n \) filters. The overall equivalent bandwidth, denoted by \( B_n \), is [27]
\[ B_n = BW \sqrt[2\left(\frac{i}{n}\right) - 1}. \] 

[2.2.4]
3 EXPERIMENTAL PROCEDURE

This chapter first discusses some preliminary research which led to the development of a controlled design plan for finding a length correction factor. This preliminary research was performed to get a better feel for the accuracy of the theoretical design procedure and the effects of desired parameters on measured responses. The correction factor design plan, which was based these preliminary findings, resulted in a database with 60 filter responses and is described in Section 3.2. Fabrication procedures including filter layout, materials, methods used to etch and tape filters, and filter enclosures are discussed in Section 3.3. The methods for measurement of filter responses are detailed in Section 3.4, and methods for tuning the fabricated filters are described in Section 3.5.

3.1 PRELIMINARY RESEARCH

The responses of over 15 different filters which were fabricated before a correction factor design plan (Section 3.2) was developed are discussed in this section. The design procedure used to design the filters of this section was continually updated based on the measured results of the fabricated filter designs. The evolution of this procedure is also described in this section. The results described include the measurement of differences between desired and measured center frequencies, and results of implementation of several theoretical insertion loss algorithms. The
center frequency discrepancies are the basis for the development of a controlled design plan described in Section 3.2.

3.1.1 EVOLUTION OF MICROSTRIP FILTER DESIGN

This section gives a history of the filter designs produced before a controlled experimental plan was devised. The filters described in this section were designed with the published design equations presented in Chapter 2. These different filter designs were fabricated on different substrates, over a wide range of center frequencies and bandwidths. Desired and measured filter characteristics are given in Table 3.1.1. Sample plots of measured transmission and reflection responses are given in Appendix B. The filters are numbered in order of design. The filter discussions which follow refer to specific versions of the design procedure, which is implemented in a PC program called USTRIP. Changes in the design procedure were ongoing, with new results being obtained constantly. The differences in these versions are discussed in tandem with the filter discussions.

Several filters, with center frequencies of 1.333 and 4.000 GHz, are used in a multipath measurement system developed by the Mobile and Portable Radio Research Group (MPRG) at Virginia Tech. This multipath measurement system is a wide band transmitter and receiver system used to measure the amplitudes and excess time delays of scattered multipath components of a transmitted repetitive pulsed signal. These scattered components can cause interference and distortion in a received signal. The results obtained from measurement of these
Table 3.1.1: Characteristics of filters designed in preliminary research.

<table>
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<tr>
<th>Filter Name</th>
<th>Design Response</th>
<th>Measured Response</th>
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<tbody>
<tr>
<td></td>
<td>$f_c$ (MHz)</td>
<td>A (dB)</td>
</tr>
<tr>
<td>1</td>
<td>1350</td>
<td>67</td>
</tr>
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<td>100</td>
</tr>
<tr>
<td>3A</td>
<td>1350</td>
<td>67</td>
</tr>
<tr>
<td>3B</td>
<td>1350</td>
<td>67</td>
</tr>
<tr>
<td>4</td>
<td>4000</td>
<td>400</td>
</tr>
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<td>NA</td>
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<td>350</td>
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<td>2400</td>
<td>240</td>
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<tr>
<td>NCR2A</td>
<td>2400</td>
<td>50</td>
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Table 3.1.1 extended:

<table>
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<th>Filter Name</th>
<th>Program Version</th>
<th>$\epsilon_r$ (mils)</th>
<th>h (mils)</th>
<th>t (mils)</th>
<th>$\varepsilon/T_{\text{Analyzer}}$</th>
<th>$\Delta f_c$ (MHz)</th>
<th>$\Delta BW$ (MHz)</th>
</tr>
</thead>
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<tr>
<td>1</td>
<td>1.0</td>
<td>4.7</td>
<td>62</td>
<td>1.4</td>
<td>E</td>
<td>--</td>
<td>125</td>
</tr>
<tr>
<td>1 Pole</td>
<td>2.0</td>
<td>4.7</td>
<td>62</td>
<td>1.4</td>
<td>T</td>
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<tr>
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<td>10.8</td>
<td>100</td>
<td>0.7</td>
<td>E</td>
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<td>-50</td>
</tr>
<tr>
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<td>10.8</td>
<td>100</td>
<td>0.7</td>
<td>E</td>
<td>8510</td>
<td>-50</td>
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<tr>
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<td>4/7-pole</td>
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<td>100</td>
<td>NA</td>
<td>NA</td>
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<td>1.4</td>
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<tr>
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<td>1.4</td>
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<tr>
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<td>10.5</td>
<td>75</td>
<td>1.4</td>
<td>T</td>
<td>8410 (MAN)</td>
<td>505</td>
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</tbody>
</table>

NOTES ON FILTERS OF TABLE 3.1.1:

1: Used correction factor (d=0.165h) described in Chapter 2; measured with FLUKE Signal Generator and HP Power Meter

1 pole: Rough tape job used strictly to find effects of spacing changes
4 pole: No response; no solution condition
4/7 pole: Altered #4; by-passed first and last resonator pair
5A: Rough tape job
6-1: Reset of #8A which was designed incorrectly (wrong $\epsilon_r$)
6-2: Added 76 mils (66 calculated + 10 for tuning) to each resonator compared to 6-1
7: No solution result

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multipath characteristics will lead to development of improved indoor wireless communications systems and channel models. In addition, several filters with center frequencies of 2400 MHz are designed for use in an impulsive noise measurement system developed by MPRG.

The first prototype filter, named Filter #1, was designed with the first version of the design procedure which was implemented in USTRIP 1.0. This design procedure included equations 2.1.3–2.1.7, 2.1.9–2.1.12, 2.1.17–2.1.28, 2.1.30–2.1.40, 2.1.44, and 2.2.1–2.2.3. The desired filter had four poles with a center frequency of 1350 MHz and a 5% bandwidth, or a 3 dB bandwidth of 67.5 MHz. This filter was etched on 62 mil thick CEM-1 substrate with \(\varepsilon_r=4.7\) (commonly referred to as G-10 PC board) by the Hybrid Microelectronics Laboratory at Virginia Tech. The measured characteristics for FILTER#1 were \(f_c=1475\) MHz and \(BW=58\) MHz. The attenuation in the passband was about 14 dB, which was much higher than the desired value of about 3 dB. Thus, the measured characteristics were acceptable except for \(f_c\) and \(L_c\).

According to theory, the primary dimension which controls the center frequency is the length of the resonators (see Section 2.1). Therefore, since \(f_c\) was higher than desired, the resonator lengths were too short. To confirm this theory, approximately 60 mils of copper tape was added to the end of each resonator to lower the resonator center frequency. The resonator additions increased the resonant wavelength which decreased the resonant frequency to 1366 MHz. Insertion loss did not improve with these copper tape additions.
The resonator widths were not uniform on FILTER#1 due to an inaccurate etching procedure. The large size of the filter (41.25 square inches) made it necessary to apply photo-resist with a brush which resulted in an uneven application (see Section 3.3 for fabrication procedure details). As a result, the amount of etching was uneven across the substrate, and the width of each resonator varied by as much as 5 mils. Bends in the I/O lines were made so the filter would fit on the substrate size available. Subsequent filters in the preliminary research stage were designed without a bend in the I/O characteristic impedance lines, and were smaller due to fabrication on substrates with higher dielectric constants. Fabrication on these substrates decreased insertion loss and alleviated fabrication problems (Section 3.3).

The second prototype, a one pole filter, was designed with the design procedure of USTRIP 2.0. The equations in USTRIP 2.0 were the same as those used in USTRIP 1.0, except equation 2.1.44 (a correction factor) was not implemented. The filter was built with hand cut copper tape on a piece of CEM-1 substrate ($\varepsilon_r=4.7$) without calipers to find the general effects of filter dimensions on bandwidth and insertion loss. This substrate was a piece of regular two-sided PC board with one side etched off completely, resulting in a 'playground' board which could be used repeatedly for various tapered filters. The filter design parameters were: $f_c=1350$ MHz and $BW=100$ MHz. The filter was tested on an HP8410 Network Analyzer. The insertion loss was about 3 dB, a large improvement over Filter #1, though Filter #1 had 4 poles. The measured bandwidth was 100 MHz.
and $f_c$ was 1520 MHz. The resonator strips were adjusted to determine the effect of varying the spacing between the resonators. By moving the strips farther apart, it was observed that an increase in the spacing between the resonators by two times and then three times the original spacing, increased the insertion loss to 14 dB and 25 dB, respectively and decreased the bandwidth to 50 MHz and 42 MHz, respectively. Center frequency remained unchanged throughout the procedure since it is independent of spacing (Section 2.1). The original layout and all adjustments were made by hand, and the results gave only a general idea of the effects of spacing changes. Thus, the effects of resonator spacing on bandwidth and insertion loss had been determined with the one-pole filter.

Since the measured and desired center frequency, which depend on $\epsilon_r$, differed for both Filter #1 and the one pole filter, it was possible that the manufacturer's claim of $\epsilon_r=4.7$ was incorrect. The dielectric constant of the substrate was measured by finding the frequency at which a single microstrip line resonates. The impedance, width, and length of the microstrip line were known. The actual $\epsilon_{eff}$ was calculated from equation 2.1.40, with $\epsilon_{eff}(f)$ replaced by $\epsilon_{eff}$, since the length of the line and resonant frequency were known. The equations for a frequency dependent $\epsilon_{eff}$ were not used since dispersion effects, which cause the frequency dependence, were not accounted for in this part of the research. Dispersion effects were originally thought to be negligible. Since $\epsilon_{eff}$ is a function of $\epsilon_r$ and $\frac{W}{h}$, $\epsilon_r$ was calculated from equations 2.1.6 and 2.1.7. The resulting value, 4.4364, was less than the manufacturer's claim of 4.7. Therefore, the resulting resonator lengths would be longer for a desired response than those designed with
\( \varepsilon_r = 4.7 \) for an identical filter response, and therefore lower center frequencies would result. The values of the response characteristics for FILTER\#1 and the experimental \( \varepsilon_r \) value were used as inputs for the design procedure in USTRIP 3.0. The equations used in the design procedure of USTRIP 3.0 were equivalent to those of USTRIP 2.0. However, USTRIP 3.0 was converted from BASIC to a Turbo Pascal [13] to decrease execution time, which was cut in half. If the new \( \varepsilon_r \) value (4.4364) was the cause for center frequency discrepancy, this design run would yield dimensions which compare well with the original design dimensions. However, the design values did not agree well and therefore the experimental value of \( \varepsilon_r \) did not fix the center frequency discrepancies.

The lower insertion loss of the one-resonator filter in comparison to FILTER\#1 (3 dB versus 14 dB) was attributable to the fewer number of resonators and smaller spacing between the resonators. Closer spacing between resonators resulted in less insertion loss. FILTER\#1 had a greater spacing area due to the greater number of poles. This spacing and the greater number of poles resulted in less coupling between the resonators of Filter \#1 as compared to the one-pole filter, and more insertion loss. An improvement in insertion loss cannot be obtained by designing the resonators closer together without altering the bandwidth of the filter. It is worth noting that the loss for each resonator/pole of a similar type of microstrip filter was about 3 dB on G-10 type substrates [14,15]. Therefore, FILTER\#1 (4 poles \( \times 3 \text{ dB} = 12 \text{ dB} \)) and the one resonator filter (1 pole \( \times 3 \text{ dB} = 3 \text{ dB} \)) exhibited similar insertion loss results.
FILTER#2 was designed using the same filter design parameters as FILTER#1, but with a dielectric constant value of 4.4364. This filter was never fabricated since a smaller filter with the same frequency response parameters, FILTER#3, was designed. The smaller filter was preferred over the larger Filter #2 since smaller size resulted in a decreased fabrication time and more even application of photo-resist.

Substrate with a higher dielectric constant than the CEM-1 substrate was obtained from Rogers Corporation (RT/Duroid teflon, $\epsilon_r=10.8$, $h=100$ mils). The higher dielectric constant reduced the size of the dimensions obtained from the design procedure. A 4-pole filter with $f_c=1350$ MHz and 5% BW of 67.5 MHz (same as FILTER#1) was designed using the design procedure USTRIP 3.0 and fabricated on this substrate. Two filters (FILTER#3A and FILTER#3B) with the same dimensions were produced to determine differences to be expected as a result of etching variations. FILTER#3A and FILTER #3B passband characteristics were essentially equivalent when tested on an HP8410 Network Analyzer (see Table 3.1.1). FILTER#3B was tested again on an HP8510 Network Analyzer with slightly different results than those obtained with the HP8410. No calibration was used for the test on the HP8410 since only a rough idea of the response was desired. Filter #3B had a third harmonic response centered at 3.779 GHz, with 3.7 dB insertion loss and a 100 MHz bandwidth. These microstrip filters, along with most real world fabricated filters had harmonic responses at integer multiples of the center frequency due to multiple resonances inherent in transmission line or cavity resonators [4]. Thus, insertion loss decreased with
Filter #3B, but center frequency discrepancy remained. All other response characteristics were acceptable. Etching variations were found to be negligible and the presence of harmonic responses was verified.

A comparison of the insertion loss responses of FILTER#1 and FILTER#3, yielded 11 dB less insertion loss for FILTER #3B, though both filters were designed with the same desired frequency response. However, Filter #1 was fabricated on CEM-1 ($\varepsilon_r=4.7$, $h=62$ mils) and Filter #3 was fabricated on PTFE (polytetrafluoroethylene), or teflon, substrate ($\varepsilon_r=10.8$, $h=100$ mils). It appeared that the decrease in loss was a function of the electrical permittivity and thickness of the filter substrate. The reason for this dependance is now discussed.

The substrate parameter $\varepsilon_r$ is a measure of how well the electromagnetic waves emanating from the resonators are absorbed into the board. The higher the $\varepsilon_r$ or the thinner the substrate, the less loss of electromagnetic energy to the air. The effect of $\varepsilon_r$ and substrate thickness can be easily explained by the fact that each resonator and the filter ground plane acts like a capacitor [3]. The stored energy of a capacitor is directly proportional to the capacitance and therefore directly proportional to $\varepsilon_r$ and inversely proportional to substrate thickness [17,18]. The stored energy is a measure of how much energy remains near the surface of the filter substrate. If more energy remains near the board surface, there is more energy available to couple to the adjacent resonators on the substrate, and less insertion loss results. This effect can also be seen in the calculation of dielectric loss (equation 2.1.54). The attenuation due to dielectric loss is generally inversely
proportional to the square root of the dielectric constant. Radiation losses (equations 2.1.56 and 2.1.57) also decrease with thinner substrates. To summarize, closer resonator spacing, higher $\epsilon_r$, and thinner substrate result in more energy transfer to adjacent resonators, and therefore less insertion loss. The high insertion loss problem had thus been solved. The discussion now continues with the evolution of filter designs.

A nine pole filter, named FILTER#4, was designed using the design procedure in USTRIP 3.0 with $f_c=4.0$ GHz and $BW=400$ MHz. This filter was originally designed for use in a multipath measurement system developed by MPRG. FILTER#4 was fabricated and tested on a HP8510 Network Analyzer with poor results. The filter response had several ripples around 4.0 GHz with an insertion loss of at least 50 dB which made the response almost undetectable. The widths and spacing dimensions of the first and last resonator pair were calculated to be 11.2 mils wide with a 500 mil spacing between them. These dimensions created a situation where the amount of electromagnetic energy that could couple through these resonator pairs was almost negligible, which therefore caused extremely high insertion loss.

Upon closer inspection, it was found that the results of several design runs with the design procedure in USTRIP 3.0 yielded the same high spacing dimension for the first and last resonator pair. The 500 mil spacing dimension corresponded to a $\frac{g}{h}$ ratio of 5 since $h$ was 100 mils. All desired filter responses which resulted in $\frac{g}{h}=5$ had steeper desired cutoff responses than those that did not result in this
condition. For example, Filter #4 was designed with the following parameters: \( f_c = 4000 \text{ MHz} \), \( BW = 400 \text{ MHz} \), \( f_{\text{stop}} = 4300 \text{ MHz} \), and \( A = 30 \text{ dB} \) at stop band. The ratio of 5 was the same quantity specified as the upper limit on the bisection method intervals in the design procedure. When the upper limit was changed, the 'solution' still converged to the upper limit. Therefore, for certain desired filter responses such as that for Filter #4, no solution existed for the equations implemented in the design procedure. It was then assumed that the desired filter response was not physically realizable using this type of parallel-coupled resonator (microstrip) filter. A new design procedure implemented in USTRIP 3.2 looked for this undesirable 'solution'. Except for this addition, the design procedure of USTRIP 3.2 was identical to that of USTRIP 3.0. If the program gave a 'no solution' condition, the desired input parameters were adjusted to yield a less abrupt cutoff response. For example, the stop band frequency was increased, or the desired attenuation at stop band or 3 dB bandwidth was decreased. Thus all filters designed with the design procedure in USTRIP 3.2 or subsequent versions were physically realizable.

FILTER #4 was altered to bypass the first and last resonator pair, as an experiment to obtain an acceptable filter response. The first and last resonator pair were the only pairs which had dimensions which corresponded to a 'no solution' condition. A copper tape jumper was connected from the input line to the second resonator pair and from the output line to the second to last resonator pair. The resulting quasi-seven pole filter had a filter response with 8 dB ripple in the pass band. This ripple occurred because this altered filter was not a
maximally flat design obtained the design procedure. Center frequency was 3.51 GHz, bandwidth was 510 MHz, and insertion loss was 8 dB measured to the average of the ripple. The first and last resonator pair of the quasi-seven pole filter were then shaved to a thinner width since this was the design pattern seen in earlier filters. The thinnest resonators of earlier filters were those at the first and last resonator pairs. The pass band ripple decreased to 3 dB, bandwidth decreased to 420 MHz, insertion loss measured decreased to 3 dB, and all other parameters remained unchanged. Thus Filter #4 was salvaged by adjusting the widths of several resonators and by-passing the first and last resonator pairs.

Filter #5A was designed using the design procedure in USTRIP 4.0 with a desired center frequency of 4.0 GHz, bandwidth of 350 MHz, and 30 dB stopband attenuation at 4.35 GHz. The design procedure in USTRIP 4.0 was an enhanced version of that in USTRIP 3.2 which included a theoretical insertion loss algorithm. The results of implementation of insertion loss algorithms are explained at the end of this section. Filter #5A was designed as a replacement for Filter #4, for use in the MPRG multipath measurement system. Filter 5A was fabricated with copper tape and tested on an HP8410 network analyzer with the following results: insertion loss=5 dB, f_c=3.77 GHz, and BW=460 MHz. This filter was fabricated without the use of calipers solely to make sure the design would give a true band pass filter response as opposed to the response obtained with Filter #4. Since this rough tape fabrication yielded acceptable results, this same design was used to etch Filters 5-1 to 5-6 in the Hybrid Microelectronics Laboratory. Sample response plots are given in Appendix B. All measured
response characteristics except center frequency agreed with desired characteristics.

The next filter to be designed was Filter #6A, with desired $f_c=1.333$ GHz and $BW=285$ MHz applied to the design procedure in USTRIP 4.0. The filter was taped and tested on the HP8410 Network Analyzer with the following results: $f_c=1.675$ GHz and $BW=350$ MHz. This filter was the first design for a filter with a fractional bandwidth greater than 0.15. The theoretical fractional bandwidth limit in the design procedure was 0.15 [4] (see Section 2.1 and 2.2). Part of the center frequency discrepancy was due to the fact that the filter was built on substrate having $\epsilon_r=10.8$ instead of $\epsilon_r=10.5$. The filter was redesigned for the correct $\epsilon_r$ as FILTER 6-1. The results show $f_c=1475$ MHz and $BW=310$ MHz. FILTERS #6-2 through #6-6 were built with the same design as Filter 6-1. These filters, like Filters #5, were used in the multipath propagation system developed by MPRG. All measured response characteristics except center frequency agreed with desired characteristics.

FILTER#7 was designed for use in a satellite receiver constructed at Virginia Tech. The etched filter was tested with good results. As seen in Table 3.1.1, the desired and experimental values are very close. However the design dimensions were obtained from the design procedure in USTRIP 3.0 which did not have a check for the no solution condition. The same desired filter response parameters which were applied to the design procedure in USTRIP 3.0 were applied to the design procedure of USTRIP 4.0, which had a check for no solution, and a 'no
solution' condition occurred. The measured response parameters Filter #7, including \( f_c \), BW, A, and \( f_{stop} \), were also applied to this new procedure with the same 'no solution' result. The good response was attributed to the fact that the first and last resonator pairs had acceptable physical dimensions which resulted from a true solution to the iterative loop equations. Also, the length of the resonators was approximately one half-wavelength. Therefore, one would expect some type of filter response around the desired center frequency.

All subsequent filters of the preliminary research stage were designed with USTRIP 4.1. The improvements in the design procedure of USTRIP 4.1 excluding insertion loss algorithms are now described, followed by discussions of filters designed with USTRIP 4.1. Insertion loss algorithm results are described at the end of this section.

The tolerance value used for \( \frac{s}{h} \) in the bisection method iteration loops of the design procedure was changed in USTRIP 4.1. The need for a tolerance value was explained in the iterative procedure discussion in Section 2.2. The old tolerance value (5\( \times 10^{-6} \)) was causing the precision of the resulting design ratios to be higher than necessary, since the calculated \( \frac{s}{h} \) ratios were rounded to the nearest tenth. This resulted in wasted calculation time. The dimension accuracy needed was ±0.1 mil which corresponded to an \( \frac{s}{h} \) ratio of \( \frac{0.1}{100} = 0.001 \), where the maximum substrate thickness was assumed to be 100 mils.

The initial lower limits for each of the bisection method iteration loops were
changed to $\frac{s}{h}$ limits of $1 \times 10^{-5}$. This was the smallest value that could be used without causing a divide by zero execution error in USTRIP. The iteration loop limits affected the 'physical realizibility' of the filters. The program also checked for this physical realizibility by checking that all spacings and widths were greater than 10 mils and less than 500 mils. If spacings or widths exceeded this range, the program printed a 'Filter not physically realizable' statement. By changing the lower limit on $\frac{s}{h}$ to $1 \times 10^{-5}$, the number of filters that were physically realizable increased. For instance, it was possible to design wider fractional bandwidth filters when the new lower limit were implemented. Larger percentage bandwidth filters became physically unrealizable because the no solution condition occurred when calculating the $\frac{s}{h}$ ratio for the first (and last) resonator pair. For smaller percentage bandwidth filters, the no solution condition occurred for the middle resonator pair $\frac{s}{h}$ ratios. No theory existed in the literature which explained why widths and spacings vary between resonator pairs.

The expressions implemented in USTRIP 4.1 for $\frac{w}{h}$ were changed from those given by equations 2.1.9 through 2.1.12 to those given by 2.1.13 and 2.1.14. This resulted in a simplification of the design procedure since the second equation set held for all $\frac{w}{h}$ and was not conditional on the resulting $\frac{w}{h}$ ratio.

The equations given in [3,5] for the effective resonator width due to the conductor thickness, equations 2.1.15 through 2.1.16, were added to the design procedure in USTRIP 4.1. These equations were evaluated immediately after the initial $\frac{w}{h}$ ratio calculation, before the iteration loops.
A frequency cutoff where excitation of higher order modes started to affect the usability of microstrip in the realization of a filter was explained in Section 2.1. This cutoff was given by equation 2.1.8 which was evaluated in the design procedure of USTRIP 4.1 from the user specified inputs and compared with f_{stop}. Filters designed with USTRIP 4.1 are now described.

Filter #8 was a wide bandwidth filter designed with USTRIP 4.1 for another test of the design procedure for fractional bandwidths greater than 0.15. The desired bandwidth was 800 MHz at f_{c}=4.0 GHz, which corresponds to a fractional bandwidth of 0.20. The measured response gave a center frequency of 4232 MHz and bandwidth of 853 MHz. Again, there was a center frequency discrepancy in that the measured center frequencies of Filters #6 and #8 were both higher than desired, and both were designed with greater than 0.15 fractional bandwidths. All other filters designed up to this point without length correction factors had measured center frequencies less than desired and fractional bandwidths less than 0.15. Therefore, center frequency discrepancy appeared to depend on fractional bandwidth. This effect had not been reported in the literature.

Filter NCR-2A was designed with f_{c}=2400 MHz, BW=50 MHz, and a length correction factor. Correction factors were defined in Section 2.1.3 as subtractions or additions of resonator lengths to account for center frequency discrepancies. The correction factor used in this design is derived in Section 3.1.2 and is given by d=0.25h. The correction factor had been determined from the responses of only
two filters. The resulting center frequency was 2905 MHz which is much higher than the desired 2400 MHz. Therefore, the length correction factor, \( d = 0.25h \), was not implemented in any other filter designs.

Filters NCR-2RT and NCR5 were designed for use in an impulsive noise measurement system being developed by MPRG. NCR-2RT was a re-tape of NCR-2A without a length correction factor. Both had good filter responses, except for deviations in center frequency. These responses are discussed in more detail in Chapter 5.

The above discussion highlights the experimental results of filters designed before a controlled design plan was developed. The work discussed above provided an intuitive feel for the performance of microstrip filters and revealed that further work was needed to develop a more accurate filter design procedure based on the results of filter designs using various substrate materials and frequency responses. The remainder of this section summarizes insertion loss results obtained from implementation of the algorithms of Section 2.1.4.

METHODS A and B were algorithms for calculating the unloaded \( Q \), denoted \( Q_u \), of a resonator with and without consideration for substrate parameters, respectively. The DISSIPATION FACTOR, TIME DELAY, and \( Q_e \) METHODS were algorithms for calculation of the total insertion loss when the \( (Q_u)_k \) of all filter resonators are known. The effective dielectric constant, \( \varepsilon_{eff} \), and width, \( w \), were taken into account in all theoretical insertion loss calculations. These
variables were explained in Section 2.1.1. The measured insertion losses for all filters are given in Table 3.1.1.

The insertion loss algorithm implemented in the design procedure of USTRIP 4.0 was a combination of METHOD A: $Q_u$ AS A FUNCTION OF SUBSTRATE PARAMETERS and the DISSIPATION FACTOR METHOD of Section 2.1.4. A check on the loss equations used in METHOD A indicated that the wrong value for conductivity ($\sigma$) in an equation for dielectric loss (equation 2.1.54) had been used. Though most references did not indicate this, the value for $\sigma$ should have been the conductivity of the dielectric, not the conductor conductivity. Since the dielectric conductivity was not known, an equivalent equation in terms of the loss tangent, $\tan \delta$, was used. Since the loss tangent was given by equation 2.1.55, where $\omega$ was taken to be the center frequency of the filter in rad/sec, and $\epsilon$ was equal to $\epsilon_r\epsilon_0$, the conductivity was calculated with $\omega$, $\epsilon_r$, and the $\tan \delta$ given by the substrate data specified by the manufacturer. Rogers Corporation, the suppliers of most of the substrate material for this research, gave the frequency but not the dielectric constant at which the loss tangent was measured. The loss tangent given by Rogers was specified as a 'typical' value for all of their 6010 Duroid substrates. Since the dielectric constant for which this loss tangent was measured was not specified by Rogers, it was assumed that the dielectric constant was 10.5 for this 'typical' loss tangent value. This 10.5 value was the median value of the dielectric constants available from Rogers in their 6010 Duroid board (10.2, 10.5, and 10.8). Thus, $\sigma$ was calculated to be $1.3435\times10^{-4}$ siemens cm for $\epsilon_r=10.5$ 6010 board. The loss tangent value specified by Norplex/Oak, Inc. was
0.025 measured at 1 MHz on ε_r=4.7 (CEM-1) substrate.

The calculated loss for filter #3B was 10.6 dB, using METHOD A: Q_u AS A FUNCTION OF SUBSTRATE PARAMETERS and DISSIPATION FACTOR METHOD. If the length correction factor determined in Section 3.1.2, d=0.25h, was accounted for on each of the resonators, the loss decreased to 9.8 dB. These values were much higher than the measured value of 1 dB. Therefore this insertion loss procedure did not yield good estimates of the insertion loss.

Theoretical insertion loss algorithms which did not take into effect the substrate parameters were then implemented in experiments to find a better estimate of the theoretical insertion loss. METHOD B: Q_u INDEPENDENT OF SUBSTRATE PARAMETERS of Section 2.1.4 was included in several trial algorithms for the sake of completeness.

Implementation of METHOD B: Q_u INDEPENDENT OF SUBSTRATE PARAMETERS and the DISSIPATION FACTOR METHOD of Section 2.1.4 in the design procedure of USTRIP 4.0 yielded theoretical insertion loss calculations of 12.4 dB and 15.4 dB for filters #3B and #5, respectively. These estimates were much greater than the experimental results. Besides not taking into account substrate parameters, an assumption that Δf_3 dB=BW was incorrect. The term Δf_3 dB was the 3 dB bandwidth for only one resonator, whereas BW was the 3 dB bandwidth of the filter as a whole.
When METHOD B: $Q_u$ INDEPENDENT OF SUBSTRATE PARAMETERS and the TIME DELAY METHOD were implemented, the results were 0.28 dB for both filter #3B and #5. These values were low. However, this procedure did not take into account the dielectric constant or any other substrate parameter into account. In equation 2.1.67, the dissipation constant $d$ was assumed to be constant for all resonators. An average dissipation factor was calculated from $d_k$ over $k=1$ to $n$, and substituted for $d$.

METHOD B: $Q_u$ INDEPENDENT OF SUBSTRATE PARAMETERS and the $Q_v$ METHOD were implemented in the loss calculation routine of USTRIP 4.0. The insertion loss estimates were 56.8 dB and 55.1 dB for filters #3B and #5 respectively. Again, the missing link seemed to be the substrate characteristics.

Substrate parameters must be included insertion loss calculations, based on the insertion loss data obtained for the filters described in this section and based on the discussion in Section 2.1.4. Therefore, METHOD A: $Q_u$ AS A FUNCTION OF SUBSTRATE PARAMETERS and the TIME DELAY METHOD were implemented to include substrate characteristics. This procedure yielded good results when compared to previously implemented procedures. Theoretical insertion loss calculations were 1.6 dB and 1.9 dB for Filters #3B and #5B, respectively. These calculations are similar to the measured values of 2 dB and 1 dB for Filters #3B and #5B respectively. Therefore, the equations corresponding to this method combination, 2.1.48–2.1.54, 2.1.56–2.1.57, and 2.1.67–2.1.68 were included in the design procedure of USTRIP 4.1. The approximation to equation
2.1.67 given by equation 2.1.69 was evaluated for values of $C_n$ and $d$ which correspond to Filters #3B and #5B. The approximated values were 1.8 dB and 2.1 dB, for Filter #3B and #5B respectively. These values are both off by 0.2 dB, and therefore the approximation equation was not implemented in the design procedure.

3.1.2 LENGTH CORRECTION FACTOR EFFECT ON FILTER RESPONSES

This section discusses the effect of length correction factors on center frequency for some filter responses. These adjustments were needed to account for center frequency discrepancies evident in the filters of early designs described in Section 3.1.1. Length correction factors reported previously in the literature were described in Section 2.1.3.

The correction factor given by equation 2.1.44 [4,10] was implemented in the design of Filter #1. Thus the trimming of $0.165 \times 62$ mils = 10.2 mils of conductor material from both ends of all resonators caused some of the discrepancy in the center frequency of this filter. A shorter resonator corresponds to a higher center frequency. An experimental addition of copper tape to the resonator ends (described in Section 3.1.1) offset the effect of this factor. This correction factor was not included in any subsequent filter designs.

Many more filters were designed without a length correction factor and fabricated with good results for most desired response parameters. However, the measured
center frequencies always differed from those desired. It was obvious that a length correction factor was needed. A brief analysis was done to find what effect the correction factor of [4,10], equation 2.1.44, would have on these filters. The effective dielectric constant, \( \epsilon_{eff} \), was calculated for a 50 ohm line with \( \epsilon_r = 10.8 \) and \( \epsilon_r = 10.5 \), since these were the dielectric constant values of all subsequent filters of the preliminary research stage. The corresponding \( \frac{W}{h} \) ratios, 0.886 and 0.912 were obtained from the design procedure of USTRIP 4.0. The resulting \( \epsilon_{eff} \) values were 7.187 and 7.014 respectively, from equations 2.1.6 and 2.1.7. The center frequency of FILTER#5B, from Table 3.1.1, was 3.485 GHz. The quarter wavelength for this measured \( f_c \) was then 320 mils (2.1.39). For the desired \( f_c = 4.0 \) GHz, the quarter wavelength was 279 mils. The difference between these quarter wavelengths was 41 mils which was greater than the 16.5 mil correction factor obtained from equation 2.1.44 with \( h = 100 \) mils. Equation 2.1.44 would theoretically shift the center frequency up to 3.67 GHz, still shy of the desired 4.0 GHz center frequency. The measured center frequency of FILTER#3B, from Table 3.1.1, is 1.3 GHz which corresponded to a quarter wavelength of 847 mils. The desired center frequency of 1.35 GHz gave a quarter wavelength of 816 mils. The difference in quarter wavelength was then 31 mils which was also greater than the correction factor of 16.5 mils obtained from equation 2.1.44. A 16.5 mil correction would theoretically produce a center frequency of 1.326 GHz, for Filter 3B. Therefore, the correction factor of [4,10], and given by equation 2.1.44, would not fully correct the center frequency. The other design and measured parameters for Filters #3B and #5B were shown in Table 3.1.1.
Using the measured response parameters for Filters 3B and 5B (see Table 3.1.1) as inputs to the design procedure of USTRIP 4.0, and comparing the physical filter dimensions to calculated dimensions confirmed the above discussion. For FILTER #3B, the calculated dimensions for the actual response averaged 31 mils greater in length than the original desired dimensions. For FILTER #5B, the measured parameter inputs yielded an average difference of 41 mils, though the calculated number of poles for the actual response was four instead of five. This difference in number of poles does not affect the resonant frequency of the resonator. The measured filter response was not symmetrical about the center frequency. The amount of attenuation for a specific frequency deviation from center frequency was different for the lower and upper sides of the frequency response. Since only one stop band frequency and stop band attenuation are applied to the design procedure, the stop band attenuation was chosen and held fixed. The stop band frequencies on the lower and upper sides of the response which corresponded to the specified stop band attenuation were determined. The deviations from center frequency for each of these stop band frequencies were averaged together. This average was added to the center frequency and used as the stop band frequency. The resulting resonator length differences agreed with those of the previous paragraph.

Filters #4 and #7 were not analyzed in the above discussion since a 'no solution' condition in the design procedure results in a high, incorrect $\frac{w}{h}$ ratio which decreased the resonator lengths. Filter #6 was not used in the analysis above since the desired fractional bandwidth was 0.21 which was greater than the
recommended limit of 0.15 (Section 2.1).

It was obvious that the error in quarter wavelength seems to be frequency dependent since the error in resonator length varied depending on the center frequency. For the filters built thus far, dispersion effects (see Section 2.1) were thought to be negligible, and therefore were not taken into account in the filter design procedure. This turned out to be an incorrect assumption. Dispersion has a more profound effect at higher frequencies, which coincides with the increase in center frequency error for Filters #3B and #5B.

The effects of dispersion at 4.0 GHz for Filter 5B were calculated as follows. Using equations 2.1.3 through 2.1.7, \( \epsilon_{\text{eff}}(4.0) \) was calculated to be 7.763. The corresponding quarter wavelength is 265 mils, which was 14 mils less than the 279 mils calculated originally with \( \epsilon_{\text{eff}}(0) \). The theoretical frequency that would occur if the original quarter wavelength resonator length of 279 mils and \( \epsilon_{\text{eff}}(4.0) = 7.763 \) was 3.798 GHz, using equation 2.1.40. This was 313 MHz higher than the measured center frequency (3.485 GHz), and 202 MHz lower than desired. The conclusion was that when dispersion effects are included, a more accurate resonator length would result, but would not completely correct the center frequency. Actual implementation of the equations for \( \epsilon_{\text{eff}}(f) \), 2.1.3 through 2.1.7, in the design procedure of USTRIP 4.0 resulted in resonator lengths an average of 14 mils shorter than those calculated with the original design procedure at 4.0 GHz. The width and spacing calculations were not a function of \( \epsilon_{\text{eff}} \), and therefore were not affected. The procedure described in this paragraph was also
performed for FILTER#3B. The average difference in resonator lengths was 6 mils shorter than those obtained with the design procedure of USTRIP 4.0 and use of $\varepsilon_{eff}$.

The change in resonator length was a function of frequency, with more correction required at higher frequencies. This relation was similar to the one found for frequency dependence of quarter wavelength. However, the correction of 14 mils was still 27 mils less than needed for FILTER#5B and the correction of 6 mils was still 25 mils less than that needed for FILTER#3B. The corrections still needed were very close (say 25 mils), and corresponded to a length correction factor $d=0.25h$, where the substrate thickness for both filters was 100 mils. This suggested that a new correction factor had been found which, like that of [4,10], was independent of frequency. However, this correction factor was based on the results of only two filters. This length correction factor was implemented in the design of one additional filter, Filter NCR-2A, which had a measured center frequency of 2905 MHz. This was much greater than the desired 2400 MHz center frequency.

It was clear at this point that a controlled experiment was necessary to determine what, if any, effects $f_c$, BW, $\varepsilon_r$, and $h$ had on measured center frequency of fabricated filters. Three filters with the same desired filter response were built on different thickness boards. The dielectric constant and number of poles remained the same for all three filters. This allowed a correction factor to be determined that was a function of substrate thickness. In addition, the fractional bandwidth
did not exceed the recommended limit of 0.15. The previously determined length correction factor, \( d = 0.25h \) was based on 0.05 (Filter #3B) and 0.0875 (Filter #5) fractional bandwidths. The number of poles was greater than one to make sure fairly steep responses would result. Also, no correction factor was used in the design.

With these points in mind, Filters 9A, 9B, and 9C were designed with \( f_c = 2400 \) MHz and \( BW = 240 \) MHz (10%). All were three pole filters built on substrate having \( \varepsilon_r = 10.5 \). Filters 9A-C were built on 50, 75, and 100 mil thickness boards respectively. A fourth filter, named Filter 9D, was also built with the same desired filter response, but on 100 mil thickness board with \( \varepsilon_r = 10.8 \). The response data from these filters was analyzed and tabulated with the data from a more extensive controlled experiment described in Section 3.2. Sample responses are shown in Appendix B and measured parameters are discussed in Chapter 5.
3.2 CORRECTION FACTOR DESIGN PLAN

As described in Section 3.1, many filters were made with a variety of desired responses, fabrication techniques, and testing procedures. However, a reliable data base obtained with controlled design, fabrication, and testing techniques was needed. Analysis of the data base would result in a correction factor for center frequency, determined from the responses of 60 filters. Previously measured filter responses had shown the need for a correction factor for the resonator lengths, which control the center frequency of the filter. This length correction factor would be added or subtracted to each end of all resonators. The data base was also analyzed for differences between theoretical and experimental insertion loss, and properties of return loss. The design plan for the filter data base is outlined in this section.

The design plan was developed with a variation of each of the following design parameters: center frequency, bandwidth, substrate thickness, and dielectric constant. The center frequencies coincide with the industrial, scientific, and medical (ISM) bands reserved by the FCC [36]. Some filters were required at these frequencies for a noise measurement system developed by the Mobile and Portable Radio Research Group at Virginia Tech. The bandwidths for all filters were chosen to obtain a wide range of fractional bandwidths. The minimum fractional bandwidth in the design plan was 50 MHz at 5790 MHz center frequency, or 0.0086. The maximum fractional bandwidth was 400 MHz at a 914 MHz center frequency, or 0.44. Substrate thicknesses chosen were based on the
thicknesses of available samples obtained from Rogers Corporation and Norplex/Oak, Inc. The original plan included the design of the filters listed in Table 3.2.1.

All filters were to be designed for fabrication on PTFE substrate with thicknesses of 50, 75, and 100 mils and an \( \epsilon_r \) of 10.5. The bandwidths accompanied by a 'x' in Table 3.2.1 indicate those filters which were also to be fabricated on CEM-1 substrate, with \( \epsilon_r = 4.7 \) and 62 mil thickness. This plan called for a total of 18 desired filters on 3 different thickness PTFE substrates plus 10 additional filters built on CEM-1 substrate, for a total of 64 filters. The properties of PTFE and CEM-1 substrates were described in Sections 2.1 and 3.3.2.

A naming convention was used to facilitate data analysis. The naming convention for these filters was as follows. The first four digits were the center frequency in MHz (three digits in the case of 914 MHz). The next three digits corresponded to the bandwidth in MHz (two digits in the case of 50 MHz, and 1K2 in the case of 1.2 kHz). The filter name ended with a letter A, B, C, or D which corresponded to the substrate as follows:

A: 100 mil, \( \epsilon_r = 10.5 \)
B: 75 mil, \( \epsilon_r = 10.5 \)
C: 50 mil, \( \epsilon_r = 10.5 \)
D: 62 mil, \( \epsilon_r = 4.7 \)
Table 3.2.1: Original filters to be designed for use in the correction factor research plan. All filters to be designed for substrate thicknesses of 50, 75, and 100 mils. Starred bandwidths indicate additional design on 62 mil CEM-1 substrate.

<table>
<thead>
<tr>
<th>Center Frequency (MHz)</th>
<th>3 dB Bandwidths (MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>914</td>
<td>50*, 100*, 200*, 400</td>
</tr>
<tr>
<td>2440</td>
<td>50*, 100*, 200*, 400*</td>
</tr>
<tr>
<td>4000</td>
<td>50, 100*, 200*, 400*, 800</td>
</tr>
<tr>
<td>5790</td>
<td>50, 100, 200, 400, 1200</td>
</tr>
</tbody>
</table>
For example, Filter 914100C was a 100 MHz bandwidth filter centered at 914 MHz, and fabricated on 50 mil, $\epsilon_r=10.5$ substrate.

The filters were designed with equations 2.1.3–2.1.7, 2.1.13–2.1.16, 2.1.19–2.1.28, 2.1.30–2.1.36, 2.1.40, 2.1.48–2.1.57, 2.1.67–2.1.69, and 2.2.1 which were used in the procedure described in Section 2.2. Sample filter summaries are listed in Appendix A. Filter summaries include the desired parameters: center frequency, 3 dB bandwidth, stop band frequency, characteristic impedance, relative epsilon, substrate thickness, loss tangent, conductor conductivity, and conductor thickness, and the output parameters: number of poles, theoretical insertion loss, and the widths, spacings, and lengths of all resonator pairs. No correction factor was implemented in these designs. For consistency, all filters corresponding to a particular bandwidth and center frequency combination were designed with the same number of poles. For example, the 100 MHz bandwidth, 4000 MHz center frequency filters were all designed with one pole, regardless of the substrate thickness or dielectric constant.

Several filters were not physically realizable. A filter was specified as physically unrealizable if any spacing or width dimension was less than 10 mils or greater than 500 mils. A dimension less than 10 mils was too small to be fabricated. This lower limit also approached the tolerance of the fabrication techniques. Fabrication is explained in more detail in Section 3.3. The maximum dimension of 500 mils corresponded to the upper limit of $5 \frac{\text{an}}{\text{h}}$ used in the iterative method of the design procedure. The unrealizable filters are listed in Table 3.2.2. A total
Table 3.2.2: Physically unrealizable filters not used in the correction factor research plan.

<table>
<thead>
<tr>
<th>Center Frequency (MHz)</th>
<th>Bandwidth/Substrate Code</th>
</tr>
</thead>
<tbody>
<tr>
<td>914</td>
<td>200C, 200D, 400B, 400C</td>
</tr>
<tr>
<td>2440</td>
<td>200D, 400D</td>
</tr>
<tr>
<td>4000</td>
<td>400D, 800C</td>
</tr>
<tr>
<td>5790</td>
<td>1K2C</td>
</tr>
</tbody>
</table>
of 64 - 9 = 55 filters, or 86% of those originally desired, were actually fabricated.

A trade off existed in attempting to create a controlled set of data while producing as large a data base as possible. For each filter design of Table 3.2.1, there existed at least 3 possible filters, depending on whether the particular design was also to be built on CEM-1 substrate, in which case the number of possible filters was 4. As explained earlier, not all filter designs were physically realizable. A desired filter design with fewer poles was more likely to be physically realizable than one with a greater number of poles. The physical realizibility of a filter also depended highly on the specified substrate thickness. For a particular filter design of Table 3.2.1, the filter designed with a 100 mil substrate thickness was more likely to be physically realizable than the filter designed with a 50 mil substrate thickness. Therefore, it was possible that for a particular filter design of Table 3.2.1, a 3 pole filter was physically realizable on 100 mil and 75 mil substrate, but not 50 mil substrate. However, it was also possible that for the same filter design of Table 3.2.1, a 2 pole filter was physically realizable for all 3 substrate thicknesses. Therefore, a tradeoff existed between maximizing the number of physically realizable filters with out regard for number of poles, and maximizing the number of poles with out regard for the number of physically realizable filters, for a particular filter design of Table 3.2.1. In all filter designs of Table 3.2.1, the number of filters that could be designed for a particular center frequency and bandwidth combination was maximized, instead of the number of poles of the filter. If, for instance, one desired filter could be designed on the 100 mil substrate with 3 poles or the same desired response could be designed on 50, 75,
and 100 mil thickness substrates with 2 poles, the 2 pole, 3 filter option was taken. This rule resulted in a more comprehensive database than one with fewer filters. However, if the number of filters that could be designed for a particular desired response was the same whether the design was 2 or 3 poles, the three pole filters were designed, since a higher order filter provides a faster roll off, which was normally desired.

Several filters (Filters #9 A-C) had already been designed with $f_c=2400$ MHz and a 3 dB bandwidth of 240 MHz. These filters were built on PTFE, or teflon, substrate with $\varepsilon_r=10.5$ with thicknesses of 50, 75, and 100 mils. Another filter (Filter #9 D) was built on PTFE substrate with $\varepsilon_r=10.8$, 100 mil thickness board. However, the PTFE substrate with $\varepsilon_r=10.8$ was not used in this research plan because of the tolerance of this substrate's dielectric constant. The PTFE substrate obtained from Rogers Corporation had a dielectric constant tolerance of ±0.25. This resulted in an overlap of the possible actual values of the nominal dielectric constants, 10.5 and 10.8. Two additional filters were also designed and fabricated with center frequencies of 2400 MHz. One filter, NCR2RT, had a desired bandwidth of 50 MHz and was built on 75 mil PTFE substrate with $\varepsilon_r=10.5$. The second, NCR5, had a desired bandwidth of 400 MHz and was built on $\varepsilon_r=10.5$, 100 mil substrate. It was possible these fabricated filters could have been used in the proposed design plan, in place of the 2440 MHz center frequency filters of Table 3.2.1 having the same desired bandwidth, $\varepsilon_r$, and substrate thickness. However, when the 2400 MHz center frequency was replaced by 2440 MHz in the design parameters, the resulting difference in resonator design lengths
was too significant to replace the 2440 MHz center frequency filters with the 2400 MHz center frequency filters already fabricated. Therefore, Filters #9 A-C, NCR2RT, and NCR5 were used for additional data in the correction factor analysis. These filters were designed using the same procedure of Section 2.2 as that used for the design of the filters of Table 3.2.1.

All filters were tested on an HP8510 Network Analyzer. The fabrication procedures, testing methods, responses, data, results, and analysis are given in the remainder of the thesis.
3.3 FABRICATION

This chapter describes two fabrication techniques used to realize a filter design from the dimensions obtained from the design procedure outlined in Section 2.2. First, the filter layout is described, followed by a section describing the materials needed. Methods for etching and taping filters are then explained followed by a section on filter enclosures.

3.3.1 FILTER LAYOUT

The first step in fabricating a microstrip bandpass filter is to make a sketch of the filter layout. This step results in a 'picture' of how the filter will look after fabrication. Once a filter is designed from the procedure in Section 2.2, the physical dimensions are sketched to determine the layout of the filter (see Fig 2.1.3). In addition to the dimensions obtained from the design, provisions must be made for connectors and the filter enclosure. The filter layout sketch includes extra length on each of the input and output lines for connector protrusions which are approximately 0.3" long. It is recommended that this allowance be about 0.5" since some coupling from the connector protrusion to resonators could occur through the air if the protrusion is close enough to the resonators. The strength of the electromagnetic fields is assumed negligible at a distance of 5 times the substrate thickness away from the top of a microstrip [2,5] (see Section 3.3.5). Since the maximum substrate thickness used in this research is 100 mils, this corresponds to a distance of 500 mils, or 0.5". Though this assumption does not
strictly apply to coupling from connector protrusions, this distance is used as a good estimate for the purposes described here.

The filter layout also includes an extra area on each side of the filter to allow sufficient clearance for the width of the connector flanges and a metal enclosure. The microstrip lines closest to the substrate edges are the input/output (I/O) characteristic impedance lines. The distance between the I/O line and the board edge should be 5 times the I/O line width [2,5] and at least 0.25” to account for the connector flange width. The electromagnetic fields emanating from a microstrip line are assumed to be negligible at a distance of 5 times the I/O line width away from the line [2,5]. With these dimensions included, the total width and length of the filter layout are determined. The desired substrate is then cut to these specified dimensions.

3.3.2 MATERIALS

One of the advantages of microstrip filters is the low cost of materials. The method of fabrication controls what materials are needed for the fabrication of a microstrip filter. The materials common to the etching and taping techniques described subsequently are the substrate on which the filter is fabricated and the connectors attached to the filter.

Substrate properties affect the size and loss of a filter. Two types of substrate, CEM-1 and PTFE, were used for the fabrication of the filters in this research.
CEM-1 substrate is a common PC board used in a wide variety of electronic circuits applications. PTFE is a teflon substrate with a high dielectric constant and higher cost. PTFE substrate obtained from Rogers Corporation were donations, but normally cost $500 per 10" X 10" piece. (Rogers Corporation, Microwave Materials Division, 100 South Roosevelt Ave., Chandler AZ 85226, (602)961-1382).

CEM-1 substrate obtained from Norplex/Oak Inc. is a copper clad composite laminate which utilizes a core of cotton linter paper and surface plies of woven glass fabric. (Norplex/Oak Inc., 505 King St. Box 1448, LaCrosse WI 54602-1448, (608)784-6070). The relative dielectric constant, \( \varepsilon_r \), is typically 4.7 and the dissipation factor, or loss tangent, is 0.025 at 1 MHz, or \( 2.5 \times 10^{-5} \) at 1 GHz. CEM-1 substrate is inexpensive, but has high loss characteristics and causes filter designs to be large due to its relatively small dielectric constant. This substrate was used for 8 filter designs of the correction factor design plan.

Typical PTFE substrate specified as Rogers Duroid 6010 has a dielectric constant of around 10. The typical loss tangent is 0.0023 at 10 GHz. This particular brand has a dielectric constant tolerance of \( \pm 0.25 \). This poor tolerance can cause some discrepancies in filter responses. For instance, if the \( \varepsilon_r \) is 10.5 and \( f=4.0 \) GHz, the corresponding electrical wavelength is 2.3146 cm, using equation 2.1.1 with \( \varepsilon_{eff}(f) \) replaced by \( \varepsilon_r \). However, if the \( \varepsilon_r \) is actually 10.25 or 10.75 due to dielectric tolerances, the resulting resonant frequency would be 4.0484 or 3.9531 GHz, respectively. Thus the frequency response of a filter could deviate by almost 50
MHz in either direction due to variations in $\varepsilon_r$. Three different samples of PTFE substrate with a nominal dielectric constant of 10.5 were used to average out these tolerance effects in the design plan of Section 3.2.

N-type connectors were attached directly to the filter boards for FILTER#1. SMA-type connectors were attached to all subsequent filters. The SMA-type connectors have flanges which make attachment to PC board possible (see picture in Figure 2.1.4). These connectors can be purchased with two or four screw holes. The four hole type is recommended since this type provides a more stable connection to PC board or box sides. The impedance of the connectors should equal the characteristic impedance of the I/O lines.

3.3.3 ETCHED FILTER METHOD

Microstrip filters can be physically realized from a filter design layout by etching one side of a two-sided printed circuit board substrate, or by hand taping the design with copper tape on a one-sided 'playground' board which has a ground plane on the bottom side of the substrate.

The etching procedure for the microstrip filters is a common technique for developing printed circuit boards. After the layout is drawn and the substrate is cut, the filter layout is entered into MicroCAD [37], a plotting program used by the Hybrid Microelectronics Laboratory. Alignment markers of about 100 mils square are placed at opposite corners of the design. These alignment markers are
used for positioning in a later step in the etching procedure. The MicroCAD file
is then used to scribe rubylith at ten times actual size. Rubylith material consists
of a clear plastic sheet with a red material coating on one surface. The rubylith
material is removed from its clear plastic backing to leave clear plastic where the
conductor is to remain during the etching process. A photograph is made of the
rubylith pattern by placing the pattern on a backlit screen in a darkroom. The
film is developed, and the resulting mask (the dark areas represent conducting
material) is a reduced (1:10) version of the rubylith pattern (see Figure 3.3.1).
The 10:1 enlargement and reduction allow little error to be present in the
resulting mask.

The PC board is prepared for etching by first cleaning it with acetone, and then
heating it at 90°C for about 15 minutes. Photoresist, a solution which resists
ultra-violet light, is then applied with a dropper to the ground plane side of the
board only. The board is spun to create an even layer of photoresist. The
spinning mechanism is a small motor driven wheel which holds a printed circuit
board in place by suction. The board is heated again for about 15 minutes to soft
bake the photoresist. This process is repeated for the filter side of the board. The
photoresist prevents copper from being removed during the etching process and
remains on the copper after the etching process is completed.

The dimensions of several filter layouts were such that they could not be spun.
The spinning equipment available has a maximum dimension limit of
approximately 4.5". Therefore, larger boards were dipped in a deep container of
Figure 3.3.1: Typical mask for an etched microstrip filter including alignment markers.
photoresist. This dipping method produces a relatively even layer of photoresist as long as the board is pulled from the solution at a constant slow rate. However, this method of producing an even layer of photoresist is much more costly than the spin method due to the large amount of photoresist required.

Once the photoresist is baked onto the board, the mask is then aligned on the filter side of the board, using the alignment markers mentioned earlier. A piece of glass holds the mask flat on the board while it is exposed to ultraviolet light for approximately 5 seconds. The board is then placed in developer for 10-15 seconds. This protects the conductor pattern from being etched during the next step. The next step requires the board to be placed in an etchant solution for about 3 minutes. This solution etches away the undesired copper. The completed filter is cleaned with acetone which removes photoresist from the ground plane and filter resonators.

The etched filter method is highly accurate. The dimensions of the fabricated filter are within 5 mils of the design dimensions. Elapsed times cited in the steps described above are for a 2.1” square filter with a 0.7 mil conductor thickness and may vary with the particular filter being fabricated. For example, a conductor thickness of 0.7 mils requires approximately half the time to be etched than a conductor thickness of 1.4 mils.
3.3.4 TAPED FILTER METHOD

Due to the finite availability of work time at the Hybrid Microelectronics Laboratory, many filter designs were realized by hand taping copper strips onto 'playground boards'. This procedure decreases the fabrication time and cost as compared to the etching fabrication technique. The taping procedure requires cutting the copper tape into resonators according to the specified dimensions of the filter layout and placing them on the board with the specified spacing dimensions. The filter layout makes taping a filter much easier.

Calipers accurate to within ±1 mil, drafter's triangles, and an exacto knife are used to cut the resonators. Once the dimensions are marked on the copper tape, the desired piece is cut with the use of a drafter's triangle to ensure straight cuts. The width of the knife needs to be accounted for in the cutting process. Care is taken to ensure the tip of the knife is kept in contact with the drafter's triangle so that the width of the knife does not affect the resulting dimension accuracy. Parallel lines are drawn on the playground board as a reference for placing the resonator strips in parallel on the substrate. However, the spacing between the resonators is marked at several points along the resonator length with the calipers to ensure the spacing is constant along the full length of the resonator pair. The parallel lines drawn on the substrate are simply a rough reference.

Fabrication by placing hand cut copper tape on one-sided substrate produces a fragile filter which is more easily damaged than an etched filter. The taped lines
can accidently be shifted during handling. However, the filter can be protected by applying a coating of clear spray paint. This paint does not affect the response of the filter [14]. Copper tape can be purchased from Digi-Key, 701 Brooks Ave. South, Thief River Falls MN 56701, (800)344-4539.

3.3.5 FILTER ENCLOSURES

Some fabricated filters were boxed in metal enclosures for ease of handling and use in a multi-path measurement system developed by Dwayne Hawbaker of the Mobile and Portable Radio Research Group at Virginia Tech. The boxes provide protection of the filter from external interference. They also protect other components of a system from any fields produced by the filters.

An object too close to the microstrips causes interference and alters the filter response. The object interferes with the coupling of electromagnetic energy between resonators by blocking and/or reflecting the energy away from the filter. For this reason, it is necessary to adhere to two rules when enclosing the filter in a box. The effects of a metal enclosure can be ignored when the spacing between the resonators and the lid is five times the substrate thickness (h), and the spacing between the conductor edge of the input/output microstrip lines and the side wall is five times the I/O strip width (w) [2,5]. Another set of recommendations for boxing dimensions are given by [6], but these are unclear, not complete, not commonly referenced, and are of questionable origin. A metal enclosure is desired since this will isolate the filter from electromagnetic
interference. The enclosure design rules given by [2,5] were adhered to in all filter enclosures of this research.

The first boxing experiment utilized copper waveguide to enclose Filter #5-1. The filter was soldered to the bottom of the box to ensure a good contact between the filter ground plane and the box. The filter was then tested on an HP8410 network analyzer. The 'boxed' response differed quite a bit from the response without the box. The resulting bandwidth had increased by 190 MHz up to 550 MHz and the center frequency was lower (3.825 GHz compared to 4.000 GHz). Since the bandwidth only increased on the lower side of the filter response, and not the upper side of the filter response, it was possible that the extended bandwidth was due to a resonance in the box. To test this theory, one of the copies of Filter #5 (Filter #5-2) was tuned to a higher desired frequency of 4.3 GHz. If the error was not due to a resonance in the box, any filter with approximately the same center frequency would have an extended bandwidth below the center frequency. A desired frequency of 4.3 GHz was the calculated frequency needed to obtain a boxed filter response centered at 4.0 GHz with a 550 MHz bandwidth. If the error was not due to the design equations, and the box simply extended the bandwidth by 190 MHz on the lower side of \( f_c \), the desired bandwidth would have increased to 550 MHz, thereby decreasing the center frequency to 4000 MHz. This desired response would result if the box simply extended the bandwidth by 190 MHz on the lower side of all responses. Filter #5-2 was tuned, boxed, and tested with the following results: \( f_c = 4.305 \text{ GHz} \), \( BW = 470 \text{ MHz} \), and a resonance at 3.6 GHz. This resonance was close enough to
interfere with the lower side of the response and extend the bandwidth of Filter 
#5-1. The resonance most probably occurred due to the distance between the 
filter resonators and the box top. This distance corresponded closely to the 
wavelength at 4.0 GHz.

When the two ends of the box which run parallel with the coupled lines were 
removed, the resonance disappeared. To offset this resonance effect, some 
microwave absorber material was glued to the inside of the box of Filter #5-1 on 
the sides along the length of the filter and on the top above the filter. This boxed 
filter was tested on the HP8410 and had a response essentially the same as that 
before boxing. Absorber material reduced the radiation interference experienced 
with microstrip, as mentioned in [5].

The resonances in the box could have been reduced by building a larger box, but 
this would not have been desirable, and perhaps not practical depending on the 
size needed to offset the resonance effect. The goal was to build as small a box as 
possible without affecting the filter response. The rules discussed above [2,5] 
seemed to be valid. If the resonances were ignored, a comparison of the filter 
responses in and out of a box were equivalent. No attenuation or distortion of the 
filter response was experienced, which met the design goals.

Since waveguide was an expensive boxing material, another method of boxing the 
filters was used. All subsequent boxes were made by soldering 6 pieces of fully 
clad CEM-1 PC board together. Filter #5-1 was re-boxed so it would have the
same boxing characteristics as Filters #5-3 to 5-6 which were boxed with CEM-1 PC board. All boxes contained space allowances of 0.3" for absorber material on either side of the filter and for the height of the box top above the filter. Therefore the shortest distance between the resonators and the PC board walls was 5 times the board thickness plus an additional 0.3", and the distance between the I/O lines of the filter and the PC board walls was 5 times the I/O line width plus 0.3". A good connection was ensured between the filter ground plane and the box. The Filters #5-3 through #5-6 and Filters #6-2 through #6-6 were boxed in the same fashion. Filters #6 also had a resonance present when the box top without absorber was placed over the filter. When the absorber was glued to the box top, and placed over the filter, the resonance disappeared. The height of the box top above the filter resonators must have been a fraction of the wavelength corresponding to the center frequency. All responses were characterized on an HP8510 since better accuracy was desired for the finished filters than the accuracy obtained from the HP8410 (see Appendix B for sample response plots).
3.4 MEASUREMENT OF FILTER RESPONSES

The testing methods used to determine filter responses fabricated filter responses are described in this section. Both an HP8410 and an HP8510 network analyzer were used to measure S-parameters of the filters.

The scattering parameters, or S-parameters, are reflection and transmission coefficients which are ratios of the amount of energy reflected at the input of a device or transmitted through a device versus the amount of energy at the input of the device. The input of the device is referred to as port 1 and the output is referred to as port 2. The reflection S-parameters are denoted S11, the energy measured at port 1 over the energy input at port 1, and S22, the energy measured at port 2 over the energy input at port 2. The transmission S-parameters are denoted S21, the energy measured at port 2 over the energy input at port 1, and S12, the energy measured at port 1 over the energy input at port 2 [3,16]. Most filter response measurements include S11 in addition to S21. Some measurements are also made of S22 to observe any significant differences with S11.

A network analyzer contains a frequency sweep generator which provides for measurement of all four S-parameters at a certain range of frequencies specified by the user. The measured data can then be analyzed for center frequency, insertion loss, bandwidth, etc. The network analyzer contains a phase locked loop which locks onto one frequency at a time and then records the amount of energy present at the ports of the test set of the analyzer. The source energy is also recorded for
both ports, so that the S-parameter ratios can be calculated. Both input and output energy measurements are possible through a directional coupler. Normally, the source is connected directly to each port on a main line. The measurement of received energy at a port is obtained with a secondary line coupled to this main line through the directional coupler. More sensitive measurements can be obtained if the main line is dedicated to the measurement device with the secondary line connected to the source. This also increases the danger of damaging the sensitive port due to static electricity, or energy spikes.

Availability of the HP8510 was limited during the first part of this research. Due to time constraints and limited availability of authorized HP8510 operators, many responses were made with the HP8410. Measurements taken with the HP8410 network analyzer were originally made under manual control with an auto sweep. Due to some problems encountered later in this research (the analyzer would not phase lock correctly), manual control measurements were made with a manual sweep. This allowed the analyzer more time to lock to each desired frequency which resulted in more accurate data. The procedure for calibration of the HP8410 was learned later during the research, and data files could be made from a computer interfaced with the analyzer. The calibration procedure was not required in the beginning of the research since only rough results were desired. Accurate responses, when desired, were made with the HP8510. The HP8410 was used mainly to find rough results.

The computer software which obtains data from the HP8410, called 'MAIN' [38],
contains a calibration procedure and can create corrected data files. Corrected
data files are data files containing the measured data obtained by the network
analyzer corrected according to a set of calibration coefficients. This 'cal set' is an
array of error coefficients which are calculated during the calibration procedure
and which offset any variations from a true zero reflection condition for the
reflection S-parameters and a true complete transmission condition, or 'thru',
between ports 1 and 2 for the transmission S-parameters.

Erroneous data can still be obtained when the analyzer does not phase lock. On
the plots made from the HP8410 data files, some erroneous points were present
and some data points were missing due to the failure of the network analyzer to
phase lock. The erroneous data points were easily observable on data plots, since
they showed a gain greater than 0 dB at a particular frequency. Gain is
impossible on passive devices such as microstrip filters. Several response plots
may have missing points due to this problem. Also, some problems arose with
'MAIN' at times during the research which prevented the ability to save a data
file from the program. The option to write to a disk file had failed due an error in
the program code. In these cases, manual control was used on the HP8410. The
procedures for plotting the saved data are discussed in Chapter 4.

Many filter S-parameter responses were obtained with the help of Ken Baker who
had access to an HP8510 at the Chemistry Department of Virginia Tech. The
HP8510 Network Analyzer there was interfaced with an HP plotter. This made
response plots easy to obtain directly from the analyzer. The HP8510 in the
Electrical Engineering Department could not be interfaced with a plotter due to the fact that a plotter with an HPiB interface was not available. The HPiB (Hewlett-Packard Interface Bus) is the standard communication link for HP equipment.

Some responses, obtained with the help of Dave Fritz of the Electrical Engineering Department, were saved to disk as raw data along with the calibration sets required for each of the S-parameters. This data was to be retrieved from disk with the HP8510 at the Chemistry Department, and plotted on the plotter interfaced with it. This procedure was attempted so that measurement time on the HP8510 could be distributed among the authorized operators. The data was saved as raw data so that manipulation of the data, such as close-ups of a particular filter response section, could be easily obtained. The HP8510 has an option to save corrected data, but this does not allow much flexibility in data manipulation. When the data was retrieved on the Chemistry Department’s HP8510, S11 and S22 could be plotted correctly. However, when the S21 data was retrieved, the corrected response, produced by loading both the cal set and the raw data, was not correct. The somewhat noisy response was 13.5 dB down from the 0 dB reference. This error was traced to a difference in test sets. The test set at the Department of Electrical Engineering was a more sensitive type, similar to the type discussed earlier, with the measurement device for port 2 connected to port 2 on the main line. The raw data and cal sets were retrieved on the HP8510 at the Electrical Engineering Department and saved again as corrected data. These corrected data files were then retrieved by the HP8510 at the Chemistry
Department and plotted. All data saved hereafter was corrected data.

Local training [39] for and access to the HP8510 were obtained through Dr. W. A. Davis and Dr. S. M. Riad of the Electrical Engineering Department. This substantially increased the speed and accuracy with which filters could be measured. The responses obtained on the HP8510 included S21 and S11 over a frequency range of 700 MHz to 7.2 GHz.

The filters used for development of a correction factor (see Section 3.2) were all tested on the HP8510 Network Analyzer. The calibration procedure was a 3.5 mm 'One-Path 2 Port' type calibration. This type of calibration was used since only S11 and S21 were to be calibrated and 3.5 mm connectors were attached to the test ports. Type 3.5 mm connectors are similar to type SMA connectors, with the exceptions that 3.5 mm type connectors have a free space dielectric (no dielectric material), normally are higher quality connectors than the SMA type, and cost more than SMA types. The number of frequencies points used on all tests was 401. The frequency response was a step response, which means the analyzer phase locked and measured the S-parameters at each of 401 frequencies in the band of interest. Over a 1 GHz frequency span, this provides a frequency resolution of 2.5 MHz ±6 Hz. The frequency span was chosen based on a rough filter response obtained by testing without calibration, and the ±6 Hz tolerance was a constant when a stepped frequency response was being used. This ensured a very accurate frequency response.
The HP8510 test set at the Electrical Engineering Department had the following port connections. Port 1 was normally a 3.5 mm female connector, while Port 2 was a 3.5 mm male connector. This setup was for insertable devices, which have one female and one male connector. An insertable device requires no adapters for connection to analyzer ports. Since the filters in this research were non-insertable, or reversible, an extra connector was required to mate the Port 1 connection to the filter (see Figure 3.4.1). Also, SMA connectors were added to both Ports so that gauging of the filter SMA connectors would not be necessary. The gauging procedure was performed to ensure connectors would mate correctly. The male pin of a male connector cannot be allowed to insert too far into the female jack to cause damage to the center conductors of one or both connectors. Gauging was required before each connection between an SMA and 3.5 mm connector to ensure no damage to the 3.5 mm connector occurred.

The following calibration procedure corresponds to Figure 3.4.1. During testing, a male-to-male SMA connector was permanently attached to a female-to-female SMA connector. This combination remained attached to Port 2, hereafter referred to as Port 2', for the entire calibration sequence and throughout all filter tests. The first calibration step was transmission calibration which was made by connecting Port 2' to Port 1. Port 2' was disconnected from Port 1 after transmission calibration was complete. A male-to-male SMA connector was then attached to Port 1, hereafter referred to as Port 1', for the remainder of the calibration sequence (reflection and isolation). This procedure created a calibrated S11 and S21 with male SMA ports, Port 1' and Port 2'. The described
Figure 3.4.1: HP8510 set up for measurement of filters in the correction factor design plan.
calibration procedure caused the actual $S_{21}$ response to include the response of the male-to-male SMA connector at Port 1'. Since magnitude only is being measured, this had a negligible affect on the response.

After calibration of $S_{11}$ and $S_{21}$, a filter was tested, and the corrected data saved to disk in HP format. The corrected data was the corrected raw data according to the calibration. Data file conversion to IBM format and plotting of filter responses are described in Chapter 4. All data for use in analysis was obtained for each filter through the use of markers on the HP8510 and is discussed in Chapter 4.

A calibration technique existed which accounted for the fact that a device under test (DUT) was noninsertable. The technique [40] involved swapping adapters which have connectors of different genders but which are considered equal in attenuation and electrical length, and was called swapping equal adapters. This procedure involved using the 'wrong' adapter for transmission calibration and the 'correct' adapters for reflection calibration. These 'wrong' and 'correct' adapters were assumed to be matched. The 3.5 mm one-path two port calibration was used. Though the technique was investigated, the small increase in measurement accuracy over the method described above was found to be insignificant for our purposes. SMA-type connectors were attached to FILTER#4 and all other filters measured in this research, including all tuned filters, as shown in Figure 2.1.4.
3.5 TUNED RESPONSES

Some tuning was necessary to achieve the desired response on the original tested filters described in Section 3.1.1. In order to achieve the desired center frequency, conductor material was trimmed from or added to the end of each resonator of the filter. This caused a change in the resonant frequency of the resonator. Trimming conductor material resulted in a higher resonant frequency. Since most filters had a lower than desired center frequency, copper was trimmed from each resonator for each of the following filters.

The first tuned filter was FILTER #4. This was an altered nine pole filter which had seven poles. The tuned measured parameters are shown in Table 3.5.1. The filter was tuned by gradually taking off 5 mils at a time from each end of all resonators, and then testing on an HP8410 Network Analyzer. This procedure was repeated until the desired center frequency was obtained.

Filter #3B was then tuned by trimming 31 mils off of each end of all resonators. The 31 mil amount was the previously calculated value of Section 3.1.2 needed to achieve \( f_c = 1350 \text{ MHz} \) from the difference in resonator lengths and \( \varepsilon_{eff} \). The tuned center frequency was 1357.5 MHz. There was a slight curl away from the substrate at the ends of each resonator after tuning due to the trimming. These were flattened, but had a negligible effect on the filter response. The corresponding wavelength on a 50 ohm line at 1357.5 MHz is
\[
816 \times (1350/1357.5) = 811 \text{ mils.}
\]
Therefore, the resonators were cut 5 mils too short.
Table 3.5.1: Tuned filter response characteristics.

<table>
<thead>
<tr>
<th>Filter Name</th>
<th>Measured Response</th>
<th>Network Analyzer</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$f_c$ (MHz)</td>
<td>$BW$ (MHz)</td>
</tr>
<tr>
<td>3B</td>
<td>1359</td>
<td>75</td>
</tr>
<tr>
<td>4</td>
<td>4264</td>
<td>370</td>
</tr>
<tr>
<td>5-1</td>
<td>4000</td>
<td>360</td>
</tr>
<tr>
<td>5-1B</td>
<td>3825</td>
<td>550</td>
</tr>
<tr>
<td>5-2</td>
<td>4305</td>
<td>470</td>
</tr>
<tr>
<td>5-3</td>
<td>3985</td>
<td>370</td>
</tr>
<tr>
<td>5-4</td>
<td>3980</td>
<td>320</td>
</tr>
<tr>
<td>5-5</td>
<td>3970</td>
<td>380</td>
</tr>
<tr>
<td>5-6</td>
<td>3980</td>
<td>350</td>
</tr>
<tr>
<td>5-1B</td>
<td>4055</td>
<td>359</td>
</tr>
<tr>
<td>5-3B</td>
<td>4039</td>
<td>310</td>
</tr>
<tr>
<td>5-4B</td>
<td>4014</td>
<td>356</td>
</tr>
<tr>
<td>5-5B</td>
<td>4032</td>
<td>297</td>
</tr>
<tr>
<td>5-6B</td>
<td>3972</td>
<td>262</td>
</tr>
<tr>
<td>6-2N</td>
<td>1370</td>
<td>300</td>
</tr>
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</tr>
<tr>
<td>6-5T</td>
<td>1350</td>
<td>330</td>
</tr>
<tr>
<td>6-6T</td>
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<td>340</td>
</tr>
<tr>
<td>6-2B</td>
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</tr>
<tr>
<td>6-3B</td>
<td>1338</td>
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</tr>
<tr>
<td>6-4B</td>
<td>1321</td>
<td>310</td>
</tr>
<tr>
<td>6-5B</td>
<td>1322</td>
<td>322</td>
</tr>
<tr>
<td>6-6B</td>
<td>1378</td>
<td>350</td>
</tr>
<tr>
<td>7</td>
<td>1171</td>
<td>16</td>
</tr>
</tbody>
</table>

NOTES ON FILTERS OF TABLE 3.5.1:

3B: C.F.=31 mils
6-2N: Tuned erroneously
6-3 to 6-6: After tuning erroneously
6-2T to 6-6T: Tuned erroneously and then solder added to each resonator to tune back down in frequency.
5-1B: Response with resonance present 33.6 GHz.
5-1: These measurements are out of box; resonance 33.6 GHz present in box#1: C.F.=41 mils
5-2: Over tuned (C.F.=63 mils); Resonance present @ 3.6 GHz
5-3 to 5-6: C.F.=41 mils
5-1B to 6-6B: Boxed
7: C.F.= 29 mils
6-2B to 6-6B: Boxed
4 (7 pole): Tuned by gradually trimming conductor material (no specific C.F.)
This was attributed to the knife cut width. Consequently, all other filters were trimmed by the calculated difference minus 5 mils to account for the knife cut.

Each resonator end on Filter #7 was trimmed by 24 mils (29 mils – 5 mils to account for knife cut width). The resonator lengths corresponding to the desired and measured center frequencies were obtained using equation 2.1.40 and the calculated value for $\varepsilon_{eff}$ for a 50 ohm line. The measured center frequency of 1136 MHz corresponded to a resonator length of 970 mils and the desired center frequency of 1170 MHz corresponded to $L_r=941$ mils. Thus the trimmed length was $970-941=29$ mils minus 5 mils to account for knife cut width. Tuned center frequency was 1171 MHz compared to 1170 MHz desired. All tuned results are shown in Table 3.5.1.

Filter #5-1 was tuned in the same manner as Filter #7, except 41–5=36 mils were removed from each resonator end. The resulting center frequency was 4.00 GHz tested on an HP8410. See Tables 3.1.1 and 3.5.1 for untuned and tuned response parameters.

The calculated difference needed to compensate for the center frequency discrepancy of Filter #6-1 is a 66 mil addition to each resonator. An additional 10 mils to allow for tuning was added to each resonator design length. Therefore, filter 6-2 was fabricated with the same design dimensions as filter 6-1 except each resonator length was increased by 76 mils. The resulting response showed a center frequency of 1.35 GHz and BW=320 MHz. The desired characteristics were
$f_c=1.333 \text{ GHz}$ and $BW=350 \text{ MHz}$. Filters 6-2 through 6-6 were tested and tuned incorrectly. This incorrect tuning procedure involved trimming the ends of the resonators. However, it was later found that the network analyzer was read incorrectly. These filters were then tuned back to responses close to their original responses by adding solder to the ends of the resonators. This crude way of tuning resulted in surprisingly good responses.

Filter #5-1 (5-B) was boxed in copper waveguide. The response after boxing had the following characteristics: $f_c=3.825 \text{ GHz}$ and $BW=550 \text{ MHz}$ (see Table 3.5.1). Therefore, the bandwidth increased dramatically on the lower side of the passband. Since the increase in bandwidth occurred only on the lower side of the center frequency, a resonance was possible. This resonance effect was discussed in Section 3.3.5. Filter 5-2 was tuned to a higher frequency (4.3 GHz) and tested to confirm this resonance possibility. To compensate for extended BW below 4.0 GHz, the filter was tuned to a 4.3 GHz such that the 'extended BW' that would occur when the filter was boxed would result in a filter response with $f_c=4.0 \text{ GHz}$. This center frequency would have resulted only if there were no resonance and the box simply made the bandwidth increase to 550 MHz. Since filters #5 were tuned to 4.0 GHz by trimming 41 mils from each resonator, and the amount of trimming needed to compensate for the measured center frequency of 3.825 GHz is 13 mils, the total tuning necessary was 54 mils. The tuned filter response before boxing had a center frequency of 4.30 GHz and 470 MHz. The tuned filter was boxed in the same fashion as filter 5-1, and tested on an HP8410. The resulting response, shown in Appendix B, clearly showed a resonance near 3.6 GHz.
Tuning a fabricated filter to a higher frequency caused the insertion loss to increase. This was most likely due to less coupling length between resonator pairs (see Figure 3.5.1). A microstrip filter is normally designed so that the middle of one resonator corresponds to the end of two adjacent resonators. This layout provides maximum coupling area between resonators. However, tuning to a higher frequency was accomplished by trimming conductor material from each end of the filter resonators. Therefore, a small region resulted at the middle of each resonator where direct coupling cannot occur. The energy dissipated in this region is essentially lost to the ground plane. Also, the amount of attenuation decreased for the frequencies on the lower side of the center frequency due to the decrease in coupling length. Therefore the rolloff on the lower side of the frequency response was more gradual.
Figure 3.5.1: Lost coupling area due to tuning. Waves emanating from the resonator in the shaded area do not couple to the adjacent resonator and are lost to the ground plane.
4 EXPERIMENTAL DATA

This chapter describes the procedures and programs written for retrieval and conversion of raw data to a usable format for plotting and analysis. This data was the basis for development of an enhanced design procedure which would produce more accurate microstrip band pass filter designs. The raw data was in the form of data files which had been created when filters were tested on the HP8410 and HP8510 Network Analyzers. Measurement of filter responses was described in Section 3.4. The resulting data base of filter response characteristics and the derived data parameters are described and tabulated presently. The plotting procedures and software are also discussed.

4.1 DATA RETRIEVAL

A program developed by Dr. W. A. Davis, 'MAIN', was used to obtain filter responses from the HP8410 Network Analyzer and save the resulting data to disk. The data files produced by MAIN contain the frequency and the corresponding measured S-parameters for each measured frequency of the response. Each S-parameter voltage ratio is given as a magnitude ratio and a phase in degrees. The data files produced by MAIN were converted to files containing S-parameters in decibels (dB), using a short program coded in Pascal, DBCONV, listed in Appendix D. DBCONV 1.0 prompts the user for the name of the source file and
the output file. It then reads the source file created by MAIN and deletes the column headings. All S-parameter voltage ratio magnitudes are then converted to dB by the following equation:

\[
S (\text{dB}) = 20 \log s
\]  

[4.1.1]

where \( s \) is the original S-parameter magnitude ratio. The new file is then saved to disk with a user-specified name. The data files produced by DBCONV 1.0 can then be retrieved by a plotting program for making accurate graphs of the network analyzer data with ordinates in decibels.

A different procedure was required for conversion of response data obtained after testing with the HP8510 Network Analyzer. Some filter responses obtained from the preliminary research described in Section 3.1.1 were plotted directly from the HP8510. In these cases, no data conversion was necessary. Data obtained from responses used in the correction factor design plan described in Section 3.2 were not plotted directly from an HP8510. This was due to access being obtained to the HP8510 in the Electrical Engineering Department. The data saved to disk by the HP8510 is in HPIB (Hewlett Packard Interface Bus) format which is incompatible with the MS-DOS (Microsoft Disk Operating System) format. The HP8510 can write corrected data to disk as an ASCII (American Standard Code for Information Interchange) or non-ASCII (binary) data file. These HP8510 disk files were converted to IBM format by 'HPTOIBM' written by Dr. W. A. Davis. Each file contains a list of complex number pairs for all four S-parameters. The
converted disk file contains lines which pertain to the HP8510 instrument settings made during response measurements. Extraneous characters were also present due to the attempted conversion of HP characters which do not correspond to characters recognized by IBM. The extraneous lines and characters prevent standard plotting programs from retrieving the data correctly.

A program was written in Pascal, 'HPCONV' (program listing in Appendix D), which removed these extraneous lines and characters, converted the data to dB relative to a 0 dB transmission response reference, and produced a chart type file complete with the corresponding frequency for each set of four S-parameters resulting in a 401 X 5 matrix. The number of data points, or measurement frequencies for all responses used in the correction factor design plan was 401. HPCONV 2.0 was designed to convert files which were originally saved by the HP8510 as corrected data, as explained in Section 3.4. HPCONV 2.0 first prompts the user for the source file name, which is the name of the file converted by HPTOIBM. This name should begin with 'DD_' which is the prefix given by the HP8510 when saving corrected data to disk. HPCONV 2.0 then prompts the user for the source file type. This pertains to the format in which the data was originally written to disk by the HP8510, ASCII or non-ASCII (binary). HPCONV 2.0 then asks for the target file name. The program reads the source data, skipping any extraneous characters present in the file. The S-parameter data is then converted to dB with equation 4.1.1 above, and saved to the target file. The frequency for each data point is also calculated and saved with the corresponding S-parameters. The data files produced by HPCONV 2.0 can then
be retrieved by a plotting program which creates accurate graphs of the network analyzer data.

Data pertaining to the development of a length correction factor included characteristics obtained from the filter responses. These characteristics included measured parameters: center frequency ($f_c$), bandwidth (BW), and fractional bandwidth (BPRCNT); and the errors and percentage errors between measured and desired parameters: $\Delta f_c$ (center frequency), $\Delta BW$ (bandwidth), $\Delta L_r$ (resonator length), $\Delta L_i$ (insertion loss), $\%\Delta f_c$, $\%\Delta BW\%$, and $\%\Delta L_r$. The following equations were used to obtain the latter parameters. Measured and desired parameters are subscripted $m$ and $d$ respectively.

\[
\Delta f_c = f_m - f_d \quad [4.1.2]
\]

\[
\Delta BW = BW_m - BW_d \quad [4.1.3]
\]

\[
\Delta L_r = [L_r]_d - [L_r]_m \quad [4.1.4]
\]

\[
\Delta L_i = (L_i)_m - (L_i)_d \quad [4.1.5]
\]

\[
\%\Delta f_c = \frac{f_m - f_d}{f_d} \times 100\% \quad [4.1.6]
\]

\[
\%\Delta BW = \frac{BW_m - BW_d}{BW_d} \times 100\% \quad [4.1.7]
\]
\[
\%\Delta L_r = \frac{[L_r]_d - [L_r]_m}{[L_r]_d} \times 100\% \quad [4.1.8]
\]

All \( f_o \) and BW errors were defined such that an increase in the measured value relative to the desired value corresponds to a positive error value. The \( L_r \) errors were defined such that a positive error value corresponds to the increase in resonator length necessary to achieve the desired center frequency. All parameters were easily calculated by hand except \( \Delta L_r \) and \( \%\Delta L_r \). These two parameters depend on substrate characteristics, and therefore a program called DLAMBDA, was written to calculate them. DLAMBDA was coded in Pascal and is listed in Appendix D. DLAMBDA first calculates the difference between two resonator coupling lengths (\( L_r \)) corresponding to two different user specified frequencies and then calculates the percentage error according to equation 4.1.8. The program was written to find the resonator length error and percentage error between measured and desired center frequencies. The widths of the coupling lines are assumed to be equivalent to the width of a characteristic impedance (\( Z_o \)) line, given by equations 2.1.13 through 2.1.16. The frequency dependent \( \epsilon_{eff} \) is then calculated for both specified frequencies with equations 2.1.3 through 2.1.7. And finally, the resonator length for each frequency is calculated with equation 2.1.40.

All plots made during the course of this research were made with SIGMA- PLOT [41], a commercially available plotting program which can plot data retrieved
from a disk file and perform a brief statistical analysis on the plotted data. The statistical analysis produces a regression equation of user specified order which is plotted superimposed on the data plot. Sigma-Plot provides the regression equation coefficients and correlation coefficient. Sigma-Plot is capable of performing a regression analysis on only one independent variable. A linear regression analysis was used for all statistical analyses. Linear regression is described in Section 5.1. Sigma-Plot and its statistical analysis capability was implemented in the filter response data analysis described in Chapter 5. Sigma-Plot was also used to plot the S11 and S21 data for all filter responses not plotted directly with the HP8510. Data for plotting was obtained from files produced by DBCONV and HPCONV. S11 and S21 responses for all filters in the test plan were plotted on separate graphs. Sample responses are given in Appendix B.
4.2 TABULATED DATA

Files containing data for all filters in the correction factor design plan were created which included all information required to obtain the various plots for use in data analysis. The data in these files is tabulated and described in this section.

Tables 4.2.1 and 4.2.2 contain data derived from the responses of all filters fabricated in the course of this research. Table 4.2.1 contains the desired and measured frequency response parameters, including center frequency ($f_c$), 3 dB bandwidth (BW), attenuation at stop band (A), stop band frequency ($f_{stop}$), and insertion loss ($L_i$), as well as the number of poles (n). All frequencies are in MHz and attenuation and loss are in dB. Stop band attenuation for the measured response equals desired stop band attenuation. Some stop band frequencies are not included in Table 4.2.1. This is due to the fact that some response plots did not cover a sufficiently wide frequency range to make an estimate. In order to create response plots with consistent frequency axes, the same calibration was used for filters having the same desired center frequency. Consistent frequency axes enable an easier comparison between filter responses. Also, this reduced measurement time considerably, since a new calibration was not performed for each filter. The theoretical insertion loss for all filters was obtained with the procedure described in Sections 2.1.4 and 3.1.1.

Table 4.2.2 contains error calculations derived from the filter response parameters of Table 4.2.1 and the S11 10 dB Bandwidth. The error calculations were

### Table 4.2.1 cont'd: Characteristics of filters used in design plan.

<table>
<thead>
<tr>
<th>Filter Name</th>
<th>Design Response</th>
<th>Measured Response</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$f_c$ (MHz)</td>
<td>$\lambda$ (MHz)</td>
</tr>
<tr>
<td>579050C</td>
<td>5790</td>
<td>50</td>
</tr>
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Table 4.2.2: Errors and S11 bandwidths for filters in Table 4.2.1.

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<th>$\Delta B_W$ (MHz)</th>
<th>$\Delta \lambda_r$ (mils)</th>
<th>$% \Delta f_c$</th>
<th>$% \Delta B_W$</th>
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Table 4.2.2 cont'd: Errors and S11 bandwidths for filters in Table 4.2.1.

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<th>$\Delta f_W$ (MHz)</th>
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<th>$% \Delta f_W$</th>
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performed as detailed in Section 4.1. The error and percentage error for center frequency (Δfc and %Δfc), 3 dB bandwidth (ΔBW and %ΔBW), and resonator length (ΔLr and %ΔLr).

S11 data include the measurement of 10 dB bandwidth in MHz. The -10 dB points were measured relative to a 0 dB reference. Some filter responses listed in Table 4.2.2 do not have 10 dB bandwidth values. In these cases, the S11 responses simply did not reach a -10 dB value within the passband. On some S11 plots, the response returns to a value above -10 dB, hereafter referred to as a lobe. In these cases, the bandwidth of the lobe above -10 dB and the gain of the lobe with respect to the -10 dB value are included in Table 4.2.2. The table entry format includes the maximum 10 dB bandwidth followed by a '/' and one or more values of lobe bandwidths. Each lobe bandwidth is followed by a '@' and the peak gain of the lobe with respect to -10 dB. As an example, Filter 2440100C has a 10 dB reflection bandwidth of 86 MHz with a 46 MHz bandwidth lobe. The peak of the lobe is at 4.3 dB above the -10 dB reference, or -5.7 dB from the 0 dB calibrated reference. The S11 response of several filters are in Appendix B.
5 RESULTS AND ANALYSIS

This chapter is devoted to the analysis of the filter response data obtained from the correction factor design plan described in Section 3.2. A statistical analysis was performed on the data which resulted in linear regression approximations to the acquired data. Linear regression is described, followed by a discussion of some theoretical insertion loss and reflection response results, derivation of a correction factor for center frequency, and results of filters designed with an enhanced design procedure.

5.1 LINEAR REGRESSION

A regression problem considers the distribution of one variable when another variable is held at fixed values. The data set used in a regression analysis contains several discrete independent variables and a set of dependent variable values which correspond to each discrete value of the independent variable. The result of a regression analysis is a function which approximates the values of the dependent variable set for all independent variable values. The data obtained from the filter responses form a data set which can be characterized by linear regression. For example, several values of measured center frequency error will correspond to one value of substrate thickness. In the following discussion, the dependent variable, Y, is the variable which deviates while the independent
variable, $X$, holds several discrete values. The deviation of $Y$ at each specific $X$ value is called a distribution. The mean, or average, of the distribution of $Y$ for each $X$ will be denoted by $\mu_{yx}$ and the variance by $\sigma_{yx}^2$. These parameters are constant for each value of $X$, but may vary with $X$. In this analysis, the variance is assumed to be constant for all values of $X$, although this is not necessarily the case.

In an analysis of a set of data by regression techniques, the result is an equation which relates $Y$ to $X$ by a polynomial equation. The order of this equation is specified before the analysis is begun. In the case of linear regression, the order is one, and the equation can be written as follows [42]:

$$
\mu_{yx} = A + B(X - \bar{X})
$$

[5.1.1]

where $A$ and $B$ are called regression coefficients and $\bar{X}$ is the mean value of all $X$. The quantity $(A - B\bar{X})$ is the $y$-intercept and $B$ is the slope of the regression line. The parameters $a$ and $b$ are estimates of $A$ and $B$ and can be obtained from the data by the following equations [42]:

$$
a = \bar{Y}
$$

[5.1.2]

$$
b = \frac{\sum(X_i - \bar{X})(Y_i - \bar{Y})}{\sum(X_i - \bar{X})^2}
$$

[5.1.3]

where $\bar{Y}$ is the mean value of all $Y$, and $X_i$ and $Y_i$ are the $i^{th}$ values of $X$ and $Y$.
respectively.

A residual, or deviation, is a variation away from the mean value. The residual of the $i^{th}$ value of $Y$ at a particular value of $X$ is $(Y_i - \bar{Y}_x)$ where $\bar{Y}_x$ is the estimate of $\mu_{yx}$, given by [42] as

$$\bar{Y}_x = a + b(X - \bar{X}) . \quad [5.1.4]$$

In regression analysis, the smallest value of the sum of squares of residuals is desired. This will result in the best fit of the regression equation to the data. The sum of squares is denoted by [42]:

$$\text{SSE} = \sum (Y_i - \bar{Y}_x)^2 \quad [5.1.5]$$

Equations 5.1.2 and 5.1.3 will result in the smallest sum of squares of residuals. The regression coefficients, $a$ and $b$, are therefore called the least-squares estimates for the data set. The sums of squares of $X$ and $Y$ residuals and the sum of products of the $X$ and $Y$ residuals are [42]

$$\text{SSX} = \sum (X_i - \bar{X})^2 , \quad [5.1.6]$$

$$\text{SSY} = \sum (Y_i - \bar{Y})^2 , \quad [5.1.7]$$

$$\text{SXY} = \sum (X_i - \bar{X})(Y_i - \bar{Y}) . \quad [5.1.8]$$
Therefore, \( b = \frac{S_{XY}}{SSX} \). SSE can now be computed from the following equations [42]:

\[
SSR = b^2SSX \quad [5.1.9]
\]

\[
SSE = SSY - SSR \quad [5.1.10]
\]

where SSR is the reduction in the sum of squares due to regression. An estimate of \( \sigma_{Y^2}^2 \), denoted by \( s_{yx}^2 \), is

\[
s_{yx}^2 = \frac{SSE}{N-2} . \quad [5.1.11]
\]

The \( N-2 \) factor in the denominator reflects the use of the two estimates for A and B. The value \( s_{yx} \) is called the standard error of estimate and reflects the uncertainty in estimating a particular \( Y \) value when the \( X \) value is known.

The correlation coefficient, \( r \), is the reduction of the sum of squares of \( Y \) due to regression relative to the original sum of squares of \( Y \). It attempts to measure how well the linear regression equation approximates a set of data [42].

\[
r = \sqrt{\frac{SSY - SSE}{SSY}} \quad [5.1.12]
\]

The linear regression analysis described above are used when there is only one
independent variable. The equations given above can be extended to two independent variables, $X_1$ and $X_2$, as follows. The linear regression equation is then given by

$$
\mu_{y,z1z2} = A + B_1(X_1 - \bar{X}_1) + B_2(X_2 - \bar{X}_2), \tag{5.1.13}
$$

where $\bar{X}_1$ and $\bar{X}_2$ are the mean values of all $X_1$ and $X_2$, respectively. Equation 5.1.13 describes a linear regression plane instead of the linear regression line of equation 5.1.1. The variance, $\sigma_{y,z1z2}^2$, is assumed constant for $X_1$ and $X_2$. The following equations for sum of squares of residuals and sum of products are similar to those for one independent variable [42].

$$
SSX_1 = \sum(X_{1i} - \bar{X}_1)^2 \tag{5.1.14}
$$

$$
SSX_2 = \sum(X_{2i} - \bar{X}_2)^2 \tag{5.1.15}
$$

$$
SSY = \sum(Y_i - \bar{Y})^2 \tag{5.1.16}
$$

$$
SX_1Y = \sum(X_{1i} - \bar{X}_1)(Y_i - \bar{Y}) \tag{5.1.17}
$$

$$
SX_2Y = \sum(X_{2i} - \bar{X}_2)(Y_i - \bar{Y}) \tag{5.1.18}
$$

$$
SX_1X_2 = \sum(X_{1i} - \bar{X}_1)(X_{2i} - \bar{X}_2) \tag{5.1.19}
$$
The regression coefficients can then be defined by equation 5.1.2 [42] and

\[ b_1 = \frac{(SSX2)(SX1Y) - (SX1X2)(SX2Y)}{D} \quad [5.1.20] \]

and

\[ b_2 = \frac{(SSX1)(SX2Y) - (SX1X2)(SX1Y)}{D} \quad [5.1.21] \]

where

\[ D = (SSX1)(SSX2) - (SX1X2)^2 \quad [5.1.22] \]

The sum of squares of residuals is given by

\[ SSE = \sum(Y_i - \bar{Y}_{x1x2})^2 \quad [5.1.23] \]

where

\[ \bar{Y}_{x1x2} = a + b1(X1 - \bar{X}1) + b2(X2 - \bar{X}2) \quad [5.1.24] \]

or

\[ SSE = SSY - (b1)SX1Y - (b2)SX2Y \quad [5.1.25] \]

The estimate of \( \sigma_{y,x1x2}^2 \) is given by

\[ s_{y,x1x2}^2 = \frac{SSE}{N-3} \quad [5.1.26] \]

Computing formulas can be used to calculate the sum of squares and sum of products given by equations 5.1.14 through 5.1.19 [42].

\[ SX1X2 = \sum X1X2_i - \left( \frac{\sum X1_i}{N} \right) \left( \frac{\sum X2_i}{N} \right) \quad [5.1.27] \]
\[ SX1Y = \sum X_{1i}Y_i - \frac{(\sum X_{1i})(\sum Y_i)}{N} \]  \hspace{1cm} [5.1.28]

Equation 5.1.28 can be used to compute SX2Y when X1 is replaced by X2.

\[ SSX1 = \sum (X_{1i})^2 - \frac{(\sum X_{1i})^2}{N} \]  \hspace{1cm} [5.1.29]

Equation 5.1.29 can be used to compute SSX2 when X1 is replaced by X2.

A multiple-correlation coefficient denoted by R describes the decrease in variability of Y values due to regression relative to the original variability of the Y values. Like \( r \), given by equation 5.1.12, R is a measure of how well the linear regression equation approximates a set of data. R can be calculated with the following equation [42]:

\[ R = \sqrt{\frac{SSY - SSE}{SSY}} \]  \hspace{1cm} [5.1.30]

R varies between 0 and 1. An R value close to 1 would correspond to a situation where the data points are near the linear regression plane, which is another way of stating that the linear regression equation produces a good approximation to the set of data.

The equations pertaining to the linear regression for two independent variables were implemented in a program called LINREG2V written in Pascal and listed in
Appendix D. LINREG2V computes the equation for a linear regression of two independent variables. Input is a data file called "LR2V.DAT" with ordered pairs in the format X1, X2, Y. The user is prompted for the number of data points in LR2V.DAT. The outputs include the correlation coefficient, variance estimate, standard deviation estimate, and the linear regression equation.
5.2 RESULTS AND ANALYSIS USING LINEAR REGRESSION

This section uses single variable linear regression to describe the effect of several independent variables on several dependent variables. It also describes the derivation of a new length correction factor for center frequency, with the use of both single and two variable linear regression. Regression equation coefficients and correlation coefficients are given in Table 5.2.1.

5.2.1 BANDWIDTH

The deviation of the measured bandwidth from the desired bandwidth was not investigated extensively. However, the results obtained show little correlation between bandwidth error and substrate thickness, or center frequency \( r=0.12 \) and \( r=0.46 \), respectively. See Figures 5.2.1A, 5.2.1B, 5.2.2A, and 5.2.2B for plots with superimposed linear regression lines. Plots with an 'A' suffix contain data point symbols which correspond to the center frequencies/dielectricconstants and plots with a 'B' suffix contain data point symbols which correspond to the filter bandwidths. Due to the low value of the correlation coefficients, random error due to the taping method of fabrication was assumed. The error in bandwidth of etched filters was much less than the error in center frequency. Bandwidth error for taped filters was not extensively investigated because all etched filters had bandwidth responses which agreed well with desired parameters (see Table 3.1.1) and bandwidth is a much more complicated function of the widths and spacings of the filter. Also, the design dimensions produced by the design procedure are
Table 5.2.1: Linear Regression Equation ($Y = A_0X + A_1$) Coefficients and Correlation Coefficients

<table>
<thead>
<tr>
<th></th>
<th>Y</th>
<th>X</th>
<th>$A_0$</th>
<th>$A_1$</th>
<th>r/R</th>
</tr>
</thead>
<tbody>
<tr>
<td>%ΔF</td>
<td>T</td>
<td>-0.155</td>
<td>16.6</td>
<td>0.60</td>
<td></td>
</tr>
<tr>
<td>%ΔF</td>
<td>F</td>
<td>-1.75E-3</td>
<td>10.8</td>
<td>0.59</td>
<td></td>
</tr>
<tr>
<td>%ΔF</td>
<td>BP</td>
<td>0.125</td>
<td>3.88</td>
<td>0.15</td>
<td></td>
</tr>
<tr>
<td>%ΔF</td>
<td>P</td>
<td>0.104</td>
<td>4.68</td>
<td>0.02</td>
<td></td>
</tr>
<tr>
<td>914 %ΔL_r</td>
<td>T</td>
<td>-0.076</td>
<td>14.1</td>
<td>0.72</td>
<td></td>
</tr>
<tr>
<td>2400 %ΔL_r</td>
<td>T</td>
<td>-0.162</td>
<td>21.0</td>
<td>0.80</td>
<td></td>
</tr>
<tr>
<td>2440 %ΔL_r</td>
<td>T</td>
<td>-0.118</td>
<td>13.9</td>
<td>0.89</td>
<td></td>
</tr>
<tr>
<td>4000 %ΔL_r</td>
<td>T</td>
<td>-0.161</td>
<td>15.8</td>
<td>0.87</td>
<td></td>
</tr>
<tr>
<td>5790 %ΔL_r</td>
<td>T</td>
<td>-0.268</td>
<td>19.8</td>
<td>0.88</td>
<td></td>
</tr>
<tr>
<td>%ΔL_r</td>
<td>T</td>
<td>-0.164</td>
<td>16.6</td>
<td>0.62</td>
<td></td>
</tr>
<tr>
<td>%ΔL_r</td>
<td>F</td>
<td>-1.86E-3</td>
<td>10.5</td>
<td>0.61</td>
<td></td>
</tr>
<tr>
<td>%ΔL_r</td>
<td>BP</td>
<td>0.139</td>
<td>3.17</td>
<td>0.17</td>
<td></td>
</tr>
<tr>
<td>%ΔL_r</td>
<td>P</td>
<td>-7.14E-3</td>
<td>4.32</td>
<td>0.001</td>
<td></td>
</tr>
<tr>
<td>%ΔL_r*</td>
<td>F</td>
<td>-1.80E-3</td>
<td>10.1</td>
<td>0.60</td>
<td></td>
</tr>
<tr>
<td>%ΔL_r*</td>
<td>T</td>
<td>-0.155</td>
<td>15.8</td>
<td>0.59</td>
<td></td>
</tr>
<tr>
<td>%ΔB</td>
<td>F</td>
<td>3.20E-2</td>
<td>-36.0</td>
<td>0.46</td>
<td></td>
</tr>
<tr>
<td>%ΔB</td>
<td>T</td>
<td>0.748</td>
<td>-15.5</td>
<td>0.12</td>
<td></td>
</tr>
<tr>
<td>ΔL_i (dB)</td>
<td>F</td>
<td>7.49E-4</td>
<td>0.60</td>
<td>0.43</td>
<td></td>
</tr>
<tr>
<td>ΔL_i (dB)</td>
<td>T</td>
<td>-0.056</td>
<td>7.34</td>
<td>0.37</td>
<td></td>
</tr>
</tbody>
</table>

NOTES:
* denotes plot without data points corresponding to CEM-1 substrate
F = Center Frequency
L_r = Resonator Length
B = 3 dB Bandwidth
T = Substrate Thickness
BP = percentage bandwidth
P = Number of Poles
L_i = Insertion Loss
Figure 5.2.1A: Plot of percent bandwidth error versus substrate thickness.
Figure 5.2.1B: Plot of percent bandwidth error versus substrate thickness.
Figure 5.2.2A: Plot of percent bandwidth error versus center frequency.
Figure 5.2.2D: Plot of percent bandwidth error versus center frequency.
derived mainly from the integer number of poles. There exists little dependence
of the design procedure on the specified frequency response after the number of
poles has been determined. If a user specified two frequency responses with two
different bandwidths, which resulted in an equal number of poles, the difference in
the resulting design dimensions was negligible. Therefore, the measured
bandwidth could varied quite a bit since the fabrication procedure had a tolerance
of ±5 mils.

5.2.2 INSERTION LOSS AND REFLECTION RESPONSES

A plot of insertion loss error versus substrate thickness and center frequency
yields a weak correlation of r=0.37 and 0.43, respectively. Figures 5.2.3 and 5.2.4
show the plots of insertion loss error in dB versus substrate thickness in mils and
center frequency in MHz, respectively. However, insertion loss is consistently
higher for filters built on CEM-1 substrate in comparison to filters built on PTFE
substrate.

Due to the imperfect taping fabrication technique, the resonators may not be in
exactly the correct positions as specified by the design dimensions. Therefore,
some small sections of each resonator may not be fully coupled to the adjacent
resonators. Energy radiating from these sections is lost to the ground plane
through the dielectric. If the filter resonators are not in the ideal position as
specified by the design dimensions, the filter bandwidth may vary. If the
bandwidth varies, ideal coupling is not occurring, and more loss will be present in
Figure 5.2.3: Plot of insertion loss error in dB versus substrate thickness.
Figure 5.24: Plot of insertion loss error in dB versus center frequency.
the passband which increases the insertion loss [4].

The insertion loss discrepancies also depend on the results of the algorithm used to calculate the theoretical loss. The algorithm implemented in the design procedure was one of many possible calculation methods described in Section 2.1.4. Due to the wide variation in the results of these methods, a conclusion concerning the effect of parameters on insertion loss cannot be made with great confidence.

Sample plots of S11 responses for filters in the controlled design plan are given in Appendix B. The reflection response characteristics derived from these plots are given in Tables 4.2.2 and 4.3.1. These characteristics indicate a correlation between small design 3 dB transmission response bandwidth and small 10 dB reflection response bandwidth, as expected. A larger desired 3 dB transmission bandwidth corresponds to a larger S11 10 dB bandwidth. Most of the filters which did not have 10 dB reflection bandwidths have small 3 dB transmission response bandwidths (typically 50 MHz). In general, the number of filters which do not have 10 dB reflection bandwidths increases with center frequency. Therefore, since S11 bandwidth decreases with an increase in specified reflection loss, the reflection loss in general increases with an increase in S21 3 dB bandwidth. The presence of lobes, described in Section 4.2, in the S11 response also increases with bandwidth. There seems to be no correlation between S11 bandwidth and substrate thickness.

One would expect that a smaller value of measured insertion loss would
correspond to a larger value of S11 bandwidth. The large amount of reflection present at a particular frequency evident in the S11 response would correspond to a high loss present in the transmission response at that frequency. However, a comparison of these two variables seems to show no such correlation.

In general, the S11 10 dB bandwidth values are lower than 3 dB bandwidths, and many filters had no S11 10 dB bandwidth. These values may be improved through a better connection between microstrip and coaxial line used for measurement. This may involve some tapering of the input and output transmission lines of the filter [4].

5.2.3 CENTER FREQUENCY CORRECTION FACTOR

A correction factor for center frequency is developed in this section. The resulting correction factor equation was derived from the results of linear regression analyses of center frequency and resonator length errors versus substrate thickness, center frequency, percentage bandwidth, and number of poles. The correction factor developed here and is a major result of the thesis research and an important contribution to microstrip filter research.

The development of the correction factor was based on the desire for a factor in the form of an adjustment in resonator length which would be added or subtracted from the resonator lengths found from equations in Section 2.1. A resonator length correction factor would be more easily determined than a center frequency
correction factor. This is due to the fact that the relationship between resonator length and center frequency involves a frequency dependent variable, $\epsilon_{\text{eff}}(f)$. This point is also supported by the fact that the only correction factors developed in the literature, by [10,11], were resonator length adjustment type factors.

A correction factor which was a percentage correction rather than a discrete correction was more desirable. If the correction factor turned out to be dependent on only substrate parameters, then a correction factor which was a discrete correction would be satisfactory. However, if the resulting correction factor was dependent on frequency parameters, then a percentage type of correction factor would lessen the dependence of the correction factor on the frequency parameter. Therefore, the percentage error was used instead of discrete error because the effects of resonator length on the correction factor could be neglected. Likewise, the percentage error was used instead of discrete error for center frequency. Some preliminary plots had been made with discrete error as the dependent variable and were analyzed with a linear regression. The same plots made with percentage error as the dependent variable produced higher values of correlation coefficients when linear regression analysis was implemented. For the reasons discussed above, resonator length percentage error, defined in equation 4.1.8, was declared the dependent variable for correction factor analysis and plotted versus various filter design parameters. Resonator length error and percentage error were obtained with 'DLAMBDA', described in Section 4.1.

A plot with a resonator length error dependent variable versus a certain
parameter will give similar results to a plot with a center frequency error
dependent variable versus the same parameter. Plots were originally made with
percentage center frequency error as the dependent variable, since this error is
easier to visualize in the filter transmission response, DLAMBDA had not been
developed yet, and preliminary results were desired to ensure the data was going
to have fairly strong correlations which could be exploited in the length correction
factor development. Conclusions made from these plots are first discussed,
followed by the development of the correction factor equation from resonator
length percentage error plots.

The only available independent design variables which may affect center
frequency or resonator length error included the following design parameters:
center frequency, 3 dB bandwidth, percentage bandwidth, number of poles,
dielectric constant, and substrate thickness. The reasons for exclusion of the other
specified design variables were as follows: the specified stop band attenuation and
frequency were accounted for in the number of poles, the characteristic impedance
was equal in all filter designs, the conductor thickness had a negligible effect on
design dimensions, and the loss tangent and conductor conductivity were used
only in the calculation of theoretical insertion loss. The linear regression
statistical analysis for one independent variable described in Section 5.1 was used
to find correlations between the dependent variable and these filter design
parameters. The correlation coefficient, \( r \), is an estimate of how well the given
data is approximated by the regression curve and is defined in Section 5.1. For
each desired plot, two graphs were made: one with data point symbols
corresponding to the center frequencies/dielectric constants and one with data point symbols corresponding to the filter bandwidths. The figures corresponding to these two graphs are denoted by the figure number followed by 'A' and 'B', respectively. From these two plots, the center frequency, bandwidth, and dielectric constant can be determined for a particular data point. The linear regression equation coefficients and the correlation coefficient for all plots are given in Table 5.2.1.

Previous research had produced length correction factors which were dependent solely on substrate thickness. However, as will be shown subsequently, the length correction factor should also depend on center frequency. These correction factors were discussed in Sections 2.1.3 and 3.1.2. A correction factor found by [10] was developed on the basis of fringing capacitance at the end of each resonator, and was dependent on substrate thickness only. Since capacitance between two plates is directly proportional to the dielectric constant and inversely proportional to the distance between the plates, a dependence on these two parameters seemed likely. This capacitance also changes with frequency due to dispersion, discussed in Section 2.1. A plot of percentage center frequency error versus substrate thickness, Figure 5.2.5, does indeed show a relatively strong correlation, $r=0.60$, between center frequency error and thickness as compared to plots of center frequency error versus other possible independent variables such as percentage bandwidth and number of poles. The center frequency error decreases with an increase in thickness.
Figure 5.2.5A: Plot of percent center frequency error versus substrate thickness.
Figure 5.2.5B: Plot of percent center frequency error versus substrate thickness.
The center frequency error may also depend on the dielectric constant and/or center frequency. A plot of percentage center frequency error versus center frequency (Figure 5.2.6) with linear regression analysis yields a correlation coefficient of 0.59. Therefore, center frequency seems to also have an important effect on center frequency error. Center frequency error increases with an increase in center frequency. This makes intuitive sense since a given error in resonator length will have a more profound effect on center frequency at higher frequencies. The amount of material trimmed from the end of a resonator has more affect on a higher center frequency filter than a lower center frequency filter. Since the error in center frequency at lower frequencies is smaller than at higher frequencies, it seems there should not be a problem with a correction factor that is dependent solely on substrate characteristics. However, this theory proves false when percentage resonator length error is the dependent variable. This result is discussed subsequently.

The equations given in Section 2.1 for the design procedure depend on the percentage bandwidth (bp) and not on bandwidth (BW) itself. Also, the fractional bandwidth is theoretically limited to 0.15 [4] for the equations implemented in the design procedure. Therefore, percentage bandwidth was included and used instead of bandwidth as an independent variable. However, center frequency error is a weak function of percentage bandwidth, based on the plots in Figure 5.2.7 and the corresponding linear regression analysis (r=0.15). One data point corresponding to Filter 914400A was excluded from the plot and the linear regression analysis. This filter has a percentage bandwidth of 43.8%,
Figure 5.2.6A: Plot of percent center frequency error versus center frequency.
Figure 5.2.6B: Plot of percent center frequency error versus center frequency.
Figure 5.2.7A: Plot of percent center frequency error versus percent bandwidth.
Figure 5.2.7B: Plot of percent center frequency error versus percent bandwidth.
which is much higher than the percentage bandwidths of all other filters. A data point such as this can radically change the value of the correlation coefficient. This is due to the fact that there are not a sufficient number of data points near bp=0.438 to make the assumption of constant variance a valid one [42].

A plot of center frequency error versus another of the possible independent design variables, number of poles, (Figure 5.2.8) gives the following result. The two pole filters seem to show more concentrated errors than the one or three pole filters. Also, the errors for the three pole filters are more concentrated than for the one pole filters. However, the correlation coefficient was a very low 0.02.

As described in section 3.2, two different dielectric constants, \( \varepsilon_r = 4.7 \) and \( \varepsilon_r = 10.5 \), were used in the filter designs of the controlled design plan. The effect of dielectric constant on center frequency error seems to be negligible. Plots of center frequency error versus substrate thickness, percentage bandwidth, and number of poles show a random distribution of data points which correspond to those filters fabricated on CEM-1 substrate (\( \varepsilon_r = 4.7 \)). Also, the number of filters fabricated on the CEM-1 (\( \varepsilon_r = 4.7 \)) substrate was 6, which is small compared to the number of filters fabricated on PTFE (\( \varepsilon_r = 10.5 \)) substrate, 54. Therefore the effect of dielectric constant should be neglected in the development of the correction factor. Filter designs were fabricated on CEM-1 substrate for a simple estimate of the effect of \( \varepsilon_r \) on center frequency error. The primary parameters affecting center frequency error seem to be substrate thickness and center frequency.
Figure 5.2.8A: Plot of percent center frequency error versus number of poles.
Figure 5.2.8B: Plot of percent center frequency error versus number of poles.
Based on the correlation values found in the research just described, the substrate thickness and center frequency were used as independent variables for plots with resonator length percentage error as the dependent variable. However, plots against percentage bandwidth and number of poles are also given for completeness.

Since center frequency error and substrate thickness were shown to have relatively strong correlation ($r=0.60$), a plot of percentage resonator length error vs. thickness was made which included the results of all filters (Figure 5.2.9). As expected, the linear regression analysis resulted in a high correlation which was calculated to be 0.62, which is higher than that calculated for the same plot with center frequency error as the dependent variable. Plots were also made for each center frequency used in the design plan: 914, 2400, 2440, 4000, and 5790 MHz (see Figures 5.2.10 through 5.2.14). This resulted in a linear regression equation for each center frequency as well as for all filters combined. A comparison of the correlation coefficients shows that the filters with a 914 MHz center frequency had a wider margin of error in comparison to the other filters. However, all coefficients are relatively high, as depicted in Table 5.2.1. Single center frequency plots of $\%\Delta L_e$ versus substrate thickness show frequency dependence of percentage resonator length error versus thickness. The slope of the linear regression approximations decreases as center frequency increases, except for the linear approximation corresponding to the 2400 MHz center frequencies. Thus, the dependence of a correction factor which depends on substrate thickness and center frequency is a distinct possibility.
Figure 5.2.9A: Plot of percent resonator length error versus substrate thickness.
Figure 5.2.9B: Plot of percent resonator length error versus substrate thickness.
Figure 5.2.10: Plot of percent resonator length error versus substrate thickness (914XXX).
Figure 5.2.11: Plot of percent resonator length error versus substrate thickness (2400xxxx).
Figure 5.2.12: Plot of percent resonator length error versus substrate thickness (2440XXXX).
Figure 5.2.13: Plot of percent resonator length error versus substrate thickness (4000XXXX).
Figure 5.2.14: Plot of percent resonator length error versus substrate thickness (5790XXX).
Since center frequency error and center frequency were also shown to have relatively strong correlation ($r=0.59$), a plot of percentage resonator length error vs. center frequency was made which shows a decrease in error with an increase in center frequency (see Figure 5.2.15). The correlation coefficient is 0.61, which is an expectedly high value and is slightly higher than that obtained for the same plot with percentage center frequency error as the dependent variable. Therefore, the correction factor should be found to be dependent on the substrate thickness and center frequency. This result shows the need for a linear regression analysis of two variables.

The linear regression analysis of a plot of resonator length percentage error versus percentage bandwidth resulted in a low correlation coefficient of $r=0.17$. Plots were made for completeness, and are shown in Figure 5.2.16. This low correlation was also seen in the same plot with center frequency percentage error as the dependent variable. As explained for the plots of percentage center frequency error versus percentage bandwidth, one data point corresponding to Filter 914400A was excluded from the plot and linear regression analysis.

Also for completeness, a plot of resonator length percentage error versus number of poles (Figure 5.2.17) was made. The linear regression analysis shows little correlation, which is expected since the same plot with center frequency percentage error as the dependent variable also showed little correlation.
Figure 5.2.15A: Plot of percent resonator length error versus center frequency.
Figure 5.2.15B: Plot of percent resonator length error versus center frequency.
Figure 5.2.16A: Plot of percent resonator length error versus percent bandwidth.
Figure 5.2.16B: Plot of percent resonator length error versus percent bandwidth.
Figure 5.2.17A: Plot of percent resonator length error versus number of poles.
Figure 5.2.17B: Plot of percent resonator length error versus number of poles.
As explained previously in the case of center frequency as the dependent variable, the effect of dielectric constant on resonator length error should be neglected. Plots of resonator length error versus substrate thickness, center frequency, and percentage bandwidth show an essentially random distribution of the data points corresponding to the filters fabricated on CEM-1 substrate. In addition, insufficient data exist for these filters to make confident conclusions about dielectric constant effect on resonator length. Plots of percentage resonator length error versus thickness and frequency were made excluding data points corresponding to the filters fabricated on CEM-1 board (see Figures 5.2.18 and 5.2.19). A comparison of correlation coefficients with plots which included these data points shows a negligible difference in correlation coefficients (see Table 5.2.1). Therefore, the effect of \( \varepsilon_r \) was neglected in the development of the length correction factor.

Based on the correlation coefficients found above, the length correction factor was a function of center frequency and substrate thickness. The correction factor was therefore a function of two mutually independent variables. In other words, the center frequency was not a function of substrate thickness and vice versa. This condition is required when performing a linear regression analysis with two independent variables. If for instance, resonator length percentage error was also highly correlated to percentage bandwidth, linear regression could not be used since percentage bandwidth was a function of center frequency.

LINREG2V, described in Section 5.1, was implemented to determine a linear
Figure 5.2.18: Plot of percent resonator length error versus substrate thickness (no ε<sub>4.7</sub> filters).
Figure 5.2.19: Plot of percent resonator length error versus center frequency (no ε<sub>4.7</sub> filters).
regression equation which related resonator length percentage error (\(Y\)) to center frequency (\(X_1\)) and substrate thickness (\(X_2\)). The number of points was 60, since all filters were included in the analysis. The linear regression equation coefficients were calculated by LINREG2V and the resulting equation is

\[
\%\Delta L_r = (-1.8352 \times 10^{-3})f_r + (-1.6146 \times 10^{-1})h + 22.627. \quad [5.2.1]
\]

The correlation coefficient is 0.75 which is significantly higher than the correlation coefficients calculated for the single variable linear regression analysis performed on plots of resonator length error versus center frequency and substrate thickness (0.61 and 0.62, respectively). This relatively high value of correlation confirms the previous assumption that the correction factor is a function of center frequency and substrate thickness. The estimated standard deviation for the data set, called the standard error of estimate and denoted \(s_{y|x}\), is 2.64%. This estimate reflects the uncertainty in estimating a particular \(\%L_r\) value when the center frequency and substrate thickness are known.

The following observations were made concerning this new correction factor. In equation 5.2.1, the coefficient for the dependence on substrate thickness is approximately equal to that given in the correction factor of equation 2.1.44 (0.161 \(\approx\) 0.165). However, equation 5.2.1 is a percentage type correction factor whereas equation 2.1.44 is a discrete correction factor. Also, this comparison does not take into account the constant (third) term in equation 5.2.1. As explained in Section 2.1.3, no information was given in [7,18,23] concerning the procedures used
to determine the original resonator lengths. Therefore, it is unknown whether
dispersion, which causes frequency dependence of the effective dielectric constant,
was taken into account in the wavelength calculations. The frequency dependence
evident in the new correction factor given by equation 5.2.1 may be at least
partially due to the fact that [7,18,23] did not fully consider dispersion. The
frequency dependence of equation 5.2.1 corresponds to a greater decrease in
resonator length as frequency increases. Similarly, the frequency dependence of
the effective dielectric constant when dispersion is taken into account causes a
decrease in resonator length as frequency increases. This is due to the fact that
the effective dielectric constant increases with frequency, according to dispersion
effects.

Equation 5.2.1 is evaluated for each resonator length of the desired filter. The
value of $\%\Delta L_r$ is added to the resonator length, $L_r$, calculated from equation
2.1.40. Implementation of this equation in the microstrip band pass filter design
procedure yielded good results, which are discussed in Section 5.4.

The length correction factor equation will result in additional length for most
cases. This contradicts previously determined correction factors discussed in
Section 2.1.3. As stated in the same section, the references cited there do not
describe their method used to determine resonator length. Since $\varepsilon_{eff}(f)$ increases
with frequency due to dispersion, the resonator lengths decrease with frequency.
It is possible the expressions found for this frequency dependence overshadow any
compensations for fringing capacitance.
5.3 RESULTS USING AN ENHANCED DESIGN PROCEDURE

The design procedure described in Section 2.2 was enhanced by the length correction factor derived in Section 5.2.3 and given by equation 5.2.1. This improved design procedure was implemented in a program called USTRIP 4.2. A user's guide to USTRIP 4.2 is given in Appendix C. The enhanced design procedure was used in the design of filters for the following analysis.

A selection of 10 filter designs from the original 60 filter design plan of Section 3.2 were redesigned with implementation of the length correction factor equation developed in Section 5.2.3. The 10 filters were chosen with care, to assure a quasi-random selection from the set of filters used in the original design plan. Factors which influenced this selection were: center frequency, bandwidth, substrate thickness, dielectric constant. The distribution according to center frequency was: 914 MHz (3), 2440 MHz (2), 4000 MHz (2), and 5790 MHz (3). More emphasis was placed on center frequencies at the ends of the spectrum than those in the middle of the spectrum since the extremes in an independent variable set could adversely affect regression analysis. Consider the situation where the number of dependent variable data points which existed for a specific value of the independent variable, which was at the low or high end of the independent data set, was low in comparison to those for the independent variable values in the middle of the set. Then the variance for the high or low independent variable value could not be considered constant, in comparison to the variance for independent variable values at the middle of the independent variable set. And
therefore, the data points corresponding to low or high independent variable values would have too much emphasis. Thus, the number of data points corresponding to high and low independent variable values must be greater than that for the middle independent variable values.

No 2400 MHz center frequency filters were included since there were only 5 of them in the original design plan and 2400 MHz is close to 2440 MHz, which is a center frequency of filters used in this analysis. For each center frequency, the bandwidths and substrate thicknesses were selected according to the amount of center frequency error. Filters with high center frequency error were preferred to filters with low center frequency error, within each set of filters at a specific center frequency. The reason for this criteria is similar to that discussed for emphasis on high and low center frequency values. The conditions just described were also weighed against the need for a random distribution among substrate thicknesses.

Two filters specified with \( \epsilon_r = 4.7 \) were chosen. All other filters were specified with \( \epsilon_r = 10.5 \). The selected filters were: 91450C, 914100D, 914200A, 244050B, 2440100D, 400050A, 4000400B, 5790100C, 5790200A, and 57901K2B. A suffix, '–42', was added to the names of these filters to indicate design with the enhanced design procedure implemented in USTRIP 4.2.

These 10 filters were fabricated on the same playground boards used in fabrication of the filters of the original design plan. They were all tested on the HP8510 Network Analyzer. The response characteristics are listed in Table 5.3.1. Sample filter design summaries and measured responses are given in Appendices A and B.
<table>
<thead>
<tr>
<th>Filter Name</th>
<th>Measured Response</th>
<th>Bandwidth</th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$f_c$ (MHz)</td>
<td>$BW$ (MHz)</td>
<td>$A$ (dB)</td>
<td>$f_{sup}$ (MHz)</td>
<td>$L_{11}$ (dB)</td>
</tr>
<tr>
<td>91450C-42</td>
<td>915</td>
<td>45</td>
<td>20</td>
<td>842/1004</td>
<td>2.6</td>
</tr>
<tr>
<td>9141000-42</td>
<td>920</td>
<td>112</td>
<td>20</td>
<td>752/-</td>
<td>4.3</td>
</tr>
<tr>
<td>914200A-42</td>
<td>930</td>
<td>184</td>
<td>15</td>
<td>748/-</td>
<td>0.7</td>
</tr>
<tr>
<td>244050B-42</td>
<td>2416</td>
<td>82</td>
<td>15</td>
<td>2218/2706</td>
<td>2.6</td>
</tr>
<tr>
<td>2440100D-42</td>
<td>2460</td>
<td>98</td>
<td>25</td>
<td>2270/2646</td>
<td>12.9</td>
</tr>
<tr>
<td>400050A-42</td>
<td>3950</td>
<td>160</td>
<td>10</td>
<td>3715/4269</td>
<td>6.2</td>
</tr>
<tr>
<td>400040B-42</td>
<td>3988</td>
<td>523</td>
<td>15</td>
<td>3579/4433</td>
<td>2.4</td>
</tr>
<tr>
<td>5790100C-42</td>
<td>5921</td>
<td>251</td>
<td>10</td>
<td>5669/-</td>
<td>4.6</td>
</tr>
<tr>
<td>5790200A-42</td>
<td>5653</td>
<td>223</td>
<td>10</td>
<td>5293/5902</td>
<td>8.5</td>
</tr>
<tr>
<td>57901K2B-42</td>
<td>6036</td>
<td>332</td>
<td>10</td>
<td>4997/6921</td>
<td>2.8</td>
</tr>
<tr>
<td>91450C-42P</td>
<td>903</td>
<td>54</td>
<td>20</td>
<td>820/1006</td>
<td>4.3</td>
</tr>
<tr>
<td>9141000-42P</td>
<td>918</td>
<td>114</td>
<td>20</td>
<td>747/-</td>
<td>4.3</td>
</tr>
<tr>
<td>914200A-42P</td>
<td>930</td>
<td>174</td>
<td>15</td>
<td>752/-</td>
<td>9.7</td>
</tr>
<tr>
<td>244050B-42P</td>
<td>2450</td>
<td>102</td>
<td>15</td>
<td>2231/2714</td>
<td>2.4</td>
</tr>
<tr>
<td>2440100D-42P</td>
<td>2456</td>
<td>101</td>
<td>25</td>
<td>2267/2626</td>
<td>13.7</td>
</tr>
<tr>
<td>400050A-42P</td>
<td>3921</td>
<td>145</td>
<td>10</td>
<td>3695/4354</td>
<td>6.7</td>
</tr>
<tr>
<td>400040B-42P</td>
<td>3993</td>
<td>457</td>
<td>15</td>
<td>3614/4450</td>
<td>1.9</td>
</tr>
<tr>
<td>5790100C-42P</td>
<td>5962</td>
<td>346</td>
<td>10</td>
<td>5641/-</td>
<td>5.2</td>
</tr>
<tr>
<td>5790200A-42P</td>
<td>5603</td>
<td>275</td>
<td>10</td>
<td>--/-5867</td>
<td>8.4</td>
</tr>
<tr>
<td>57901K2B-42P</td>
<td>6060</td>
<td>1323</td>
<td>10</td>
<td>4988/6953</td>
<td>2.0</td>
</tr>
</tbody>
</table>

193
respectively.

A resonator length difference, $\Delta L_{r, 42-41}$, was defined as follows:

$$\Delta L_{r, 42-41} = L_{r, 42} - L_{r, 41}$$ \hspace{1cm} [5.3.1]

where $L_{r, 42}$ is the resonator length as designed by the enhanced design procedure in USTRIP 4.2 and $L_{r, 41}$ is the resonator length as designed by the design procedure of USTRIP 4.1. This difference was defined so a check could be made on the correction factor. If the correction factor equation is a good approximation for substrate thickness and center frequency dependence, then the difference defined above will not be correlated to these two parameters. A linear regression analysis of the plots of $\Delta L_{r, 42-41}$ versus substrate thickness and center frequency yields correlation coefficients of 0.13 and 0.19 (see Figures 5.3.1 and 5.3.2). These results show that correlation of resonator length error with substrate thickness and center frequency is now negligible with the enhanced design procedure.

A comparison of the measured center frequencies of this 10 filter set designed with the original design procedure (USTRIP 4.1) and the measured center frequencies of filters designed with the enhanced procedure (USTRIP 4.2) show a decrease in error for all filters except 57901K2B. This exception can be explained by the fact that at this high frequency, the center frequency depends highly on the accuracy with which the resonators are fabricated. Therefore, a slight deviation from the desired decrease in error is acceptable.
Figure 5.3.1: Plot of resonator length difference in mils versus substrate thickness.
Figure 5.3.2: Plot of resonator length difference in mils versus center frequency.
These 10 filters fabricated with the enhanced design procedure were re-fabricated. This created an almost identical set of 10 filters. This re-fabrication provided an estimate of the repeatability of accurate filter designs. Several resulting filter responses are given in Appendix B. The filter names remained the same as those used in the original design plan, except for an additional '-42P' suffix, denoting the second fabrication of designs with the design procedure in USTRIP 4.2. The results again show a decrease in center frequency error for all filters except 57901K2B.

The percentage center frequency error for both sets of 10 filters fabricated from the enhanced design procedure is tabulated in Table 5.3.2. Percentage center frequency error was defined by equation 4.1.6. The average percentage error for both sets is 0.46% of desired center frequency. The average percentage center frequency error obtained from the original 60 filter data base was 4.57%. Therefore, the average percentage frequency error has improved by approximately 4.11%, or by a factor of 10. The difference in the percentage errors of the filters in the two 10 filter sets is also given in Table 5.3.2. The average difference in these percentage errors is −0.06%. Therefore, the consistency in the fabrication procedure is very good for these microstrip filters. The average percentage center frequency error for the negative values is −1.4% and for the positive values is 1.7%. Therefore, center frequency accuracy has improved to an average of within ±1.7%, with a worst case error of 4.7% of desired center frequency. Without implementation of an improved correction factor, the average percentage error for
Table 5.3.2: Center frequency percentage errors for both 10 filter sets designed with the enhanced design procedure.

<table>
<thead>
<tr>
<th>Filter Name</th>
<th>'-42'</th>
<th>'-42P'</th>
<th>ΔΔf_c</th>
</tr>
</thead>
<tbody>
<tr>
<td>91450C</td>
<td>0.1</td>
<td>-1.2</td>
<td>-1.3</td>
</tr>
<tr>
<td>914100D</td>
<td>0.7</td>
<td>0.4</td>
<td>-0.3</td>
</tr>
<tr>
<td>914200A</td>
<td>1.7</td>
<td>1.7</td>
<td>0.0</td>
</tr>
<tr>
<td>244050B</td>
<td>-1.0</td>
<td>0.4</td>
<td>1.4</td>
</tr>
<tr>
<td>2440100D</td>
<td>0.8</td>
<td>0.7</td>
<td>-0.1</td>
</tr>
<tr>
<td>400050A</td>
<td>-1.2</td>
<td>-2.0</td>
<td>-0.8</td>
</tr>
<tr>
<td>4000400B</td>
<td>-0.3</td>
<td>-0.2</td>
<td>0.1</td>
</tr>
<tr>
<td>5790100C</td>
<td>2.3</td>
<td>3.0</td>
<td>0.7</td>
</tr>
<tr>
<td>5790200A</td>
<td>-2.4</td>
<td>-3.2</td>
<td>-0.8</td>
</tr>
<tr>
<td>57901K2B</td>
<td>4.2</td>
<td>4.7</td>
<td>0.5</td>
</tr>
</tbody>
</table>
the 60 filter data base was $-4.0\%$ and $5.9\%$. Therefore, center frequency accuracy was an average of within $\pm 5.9\%$, with a worst case error of $15.8\%$.

As a further test of the enhanced design procedure, 12 filters were fabricated for use in the noise measurement system developed by MPRG. These 12 filters included 4 different designs with center frequencies, bandwidths, and insertion losses given in Table 5.3.3.

All filters were built with the taping fabrication procedure and were tested on the HP8510 Network Analyzer with the same procedure used for testing the filters used in the correction factor design plan. Due to time constraints, the original filter responses were not saved. These filters needed only slight tuning, which was done during testing. Sample filter design summaries and tuned filter responses are given in Appendix A and B, respectively. The filter responses are named with the same naming convention as that used in the original design plan, except a 'K' suffix has been added.
Table 5.3.3: Tuned filters designed with enhanced design procedure for use in noise measurement system. (frequencies in MHz/attenuation in dB)

<table>
<thead>
<tr>
<th>Filter Name</th>
<th>Design Response</th>
<th>Measured Response</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$f_c$ (MHz)</td>
<td>BW (MHz)</td>
</tr>
<tr>
<td>91450-K</td>
<td>914</td>
<td>50</td>
</tr>
<tr>
<td>91450-K</td>
<td>914</td>
<td>50</td>
</tr>
<tr>
<td>228400-K</td>
<td>2280</td>
<td>400</td>
</tr>
<tr>
<td>228400-K</td>
<td>2280</td>
<td>400</td>
</tr>
<tr>
<td>244050-K</td>
<td>2440</td>
<td>50</td>
</tr>
<tr>
<td>244050-K</td>
<td>2440</td>
<td>50</td>
</tr>
<tr>
<td>244050-K</td>
<td>2440</td>
<td>50</td>
</tr>
<tr>
<td>244050-K</td>
<td>2440</td>
<td>50</td>
</tr>
<tr>
<td>400050-K</td>
<td>4000</td>
<td>50</td>
</tr>
<tr>
<td>400050-K</td>
<td>4000</td>
<td>50</td>
</tr>
<tr>
<td>400050-K</td>
<td>4000</td>
<td>50</td>
</tr>
<tr>
<td>400050-K</td>
<td>4000</td>
<td>50</td>
</tr>
</tbody>
</table>
6 CONCLUSIONS

This chapter concludes the thesis with a summary and recommendations for possible future research related to the material discussed.

6.1 SUMMARY

A microstrip band pass design procedure has been developed based on equations available in the published literature. Included in the design procedure are allowances for dispersion effects at microwave frequencies. Also included are the effects of a non-ideal conductor thickness. Theoretical insertion loss algorithms were also developed from equations found in the literature. Preliminary research was performed which gave important, though limited, results. These results included effects of dielectric constant on insertion loss and filter size, discrepancies in center frequency, and effects of length correction factors on filter responses. The preliminary research included the design, fabrication, and measurement of over 15 filters. These designs were made for a variety of purposes, including implementation in RF sections of an indoor modem, multipath measurement system, satellite receiver, and noise measurement system.

The preliminary research provided a good basis from which a controlled design plan for a data base of filter responses could be developed. This design plan was
produced with emphasis on finding a length correction factor for errors in center frequency. The preliminary research had consistently shown discrepancies between measured and desired center frequency.

Fabrication methods were then discussed, including development of a filter layout, discussion of materials required, fabrication by etching and taping, and the necessary considerations for fabricated filter enclosures. Testing methods were then described, including methods for measurement of filter responses on both the HP8410 and HP8510 Network Analyzers. Filters obtained in the preliminary research required tuning to correct for center frequency discrepancy. These tuning methods were described in the last section on fabrication.

Several methods for data retrieval, including the development of several programs useful in obtaining filter response data were then explained. The data pertaining to the correction factor design plan was then analyzed. The analysis discussion included a detailed explanation of linear regression, the results of which were used to determine a new length correction factor for center frequency. This new correction factor was based on the data of 60 fabricated filter responses and was found to be highly correlated to center frequency and substrate thickness. Insertion loss results were discussed.

The length correction factor has been implemented in the microstrip band pass filter design procedure developed in Chapter 2. The enhanced design procedure has been implemented in an easy to use program, USTRIP 4.2, which calculates
all dimensions needed for the realization of a microstrip band pass filter. Several filters were designed using the enhanced design procedure in USTRIP 4.2. The responses of these filters show center frequency accuracy has improved to an average of within ±1.7%, compared to an average accuracy within ±5.9% obtained without the new correction factor.

More accurate filter responses than those obtainable without the length correction factor can now be obtained at microwave frequencies with microstrip bandpass filters. USTRIP 4.2 calculates all dimensions needed to fabricate a microstrip bandpass filter and also attempts to find the theoretical insertion loss based on the desired frequency response characteristic and the substrate parameters. The filters designed by the program will yield reliable, inexpensive filters for use in a wide range of microwave applications.
6.2 RECOMMENDATIONS FOR EXTENSIONS OF THIS THESIS

Included in the suggestions for future research in microstrip band pass filters are an improvement in the speed of the design procedure, an option for Tchebyscheff filter response designs, a more accurate theoretical insertion loss algorithm, and an improvement in filter reflection characteristics.

A numerical analysis method may be found which yields a faster simultaneous solution to the equations given by [5,12] (equations 2.1.32 through 2.1.36). For instance, a variation on Newton’s method [1] is usually faster than the bisection method implemented here, but does not work for all conditions. Several methods could be attempted so that an optimal algorithm which yields faster, but still accurate solutions.

The design procedure could be enhanced by including an option for Tchebyscheff type filter responses. These responses have steeper rolloffs, but also have ripple in the pass band. The amount of ripple can be specified by the designer. Equations 2.1.17, 2.1.18, 2.1.21, and 2.1.22 of the original literature review would have to be altered to account for differences in the characteristics of the low-pass prototype. It is unknown whether the correction factor equation derived previously could also apply for Tchebyscheff type filters. However, the equations that would be altered do not directly affect the equations used to find resonator lengths. Therefore, it seems the length correction factor could be used in a design procedure for Tchebyscheff type filters.
A better estimate for the theoretical insertion loss may be found. The theoretical insertion loss calculations implemented in USTRIP 4.2 compose one set of many different combinations of the methods developed in Section 2.1.4 from the literature. It is assumed that some of the discrepancy between theoretical and experimental results is due to the method of fabrication. Therefore, a data base of etched filters may yield more accurate data for use in insertion loss analyses.

And finally, a method of predicting the reflection characteristics (S11 and S22) needs to be established. More insertion loss was desired in the pass band of the reflection responses. The majority of research discussed here was based on the transmission characteristics (S21) of the filters. However, reflection response characteristics have been tabulated in Table 4.2.2. Therefore, some preliminary data found in this thesis could be used in an analysis of reflection characteristics of microstrip band pass filters.
REFERENCES


[38] W. A. Davis, “MAIN”, Virginia Polytechnic Institute and State University.


APPENDIX A: Filter Design Summaries

The dimensions given in the following filter summaries are used to sketch a filter layout similar to that in Figure 2.1.3.
FILTERS DESIGNED WITH ORIGINAL DESIGN PROCEDURE

(USTRIP 4.1)
FILTER SUMMARY (frequencies in MHz, dimensions in mils):

NCR2RT

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Center Frequency:</td>
<td>2400.0</td>
</tr>
<tr>
<td>3 dB Pass Band Bandwidth:</td>
<td>50.0</td>
</tr>
<tr>
<td>Stop Band Attenuation (dB):</td>
<td>15.0</td>
</tr>
<tr>
<td>Stop Band Frequency (FSTOP&gt;FCNTR):</td>
<td>2600.0</td>
</tr>
<tr>
<td>Characteristic Impedance:</td>
<td>50.0</td>
</tr>
<tr>
<td>Relative Epsilon:</td>
<td>10.5</td>
</tr>
<tr>
<td>Substrate Thickness:</td>
<td>75.0</td>
</tr>
<tr>
<td>Loss Tangent:</td>
<td>0.009580</td>
</tr>
<tr>
<td>Conductor Conductivity (mhos/cm):</td>
<td>580000.0</td>
</tr>
<tr>
<td>Conductor Thickness:</td>
<td>1.4</td>
</tr>
<tr>
<td>Number of Poles:</td>
<td>1</td>
</tr>
<tr>
<td>Theoretical Insertion Loss:</td>
<td>0.5 dB</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>69.2</td>
<td>74.5</td>
<td>456.4</td>
</tr>
<tr>
<td>1, 2</td>
<td>69.2</td>
<td>74.5</td>
<td>458.4</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>70.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

FILTER SUMMARY (frequencies in MHz, dimensions in mils):

NCR5

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Center Frequency:</td>
<td>2400.0</td>
</tr>
<tr>
<td>3 dB Pass Band Bandwidth:</td>
<td>400.0</td>
</tr>
<tr>
<td>Stop Band Attenuation (dB):</td>
<td>10.0</td>
</tr>
<tr>
<td>Stop Band Frequency (FSTOP&gt;FCNTR):</td>
<td>2700.0</td>
</tr>
<tr>
<td>Characteristic Impedance:</td>
<td>50.0</td>
</tr>
<tr>
<td>Relative Epsilon:</td>
<td>10.5</td>
</tr>
<tr>
<td>Substrate Thickness:</td>
<td>100.0</td>
</tr>
<tr>
<td>Loss Tangent:</td>
<td>0.009580</td>
</tr>
<tr>
<td>Conductor Conductivity (mhos/cm):</td>
<td>580000.0</td>
</tr>
<tr>
<td>Conductor Thickness:</td>
<td>1.4</td>
</tr>
<tr>
<td>Number of Poles:</td>
<td>3</td>
</tr>
<tr>
<td>Theoretical Insertion Loss:</td>
<td>1.0 dB</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>51.9</td>
<td>16.5</td>
<td>461.8</td>
</tr>
<tr>
<td>1, 2</td>
<td>87.4</td>
<td>66.9</td>
<td>455.2</td>
</tr>
<tr>
<td>2, 3</td>
<td>87.4</td>
<td>66.9</td>
<td>455.2</td>
</tr>
<tr>
<td>3, 4</td>
<td>51.9</td>
<td>16.5</td>
<td>461.8</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>93.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
FILTER SUMMARY (frequencies in MHz, dimensions in mils):

91450D
Center Frequency: 914.0
3 dB Pass Band Bandwidth: 50.0
Stop Band Attenuation (dB): 20.0
Stop Band Frequency (FSTOP>FCNTR): 1000.0
Characteristic Impedance: 50.0
Relative Epsilon: 4.7
Substrate Thickness: 62.0
Loss Tangent: 0.000027
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 2
Theoretical Insertion Loss: 0.1 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>100.4</td>
<td>19.1</td>
<td>1728.2</td>
</tr>
<tr>
<td>1, 2</td>
<td>118.1</td>
<td>119.5</td>
<td>1716.8</td>
</tr>
<tr>
<td>2, 3</td>
<td>100.4</td>
<td>19.1</td>
<td>1728.2</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td></td>
<td>115.0</td>
<td></td>
</tr>
</tbody>
</table>

FILTER SUMMARY (frequencies in MHz, dimensions in mils):

914100A
Center Frequency: 914.0
3 dB Pass Band Bandwidth: 100.0
Stop Band Attenuation (dB): 20.0
Stop Band Frequency (FSTOP>FCNTR): 1100.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.5
Substrate Thickness: 100.0
Loss Tangent: 0.025200
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 2
Theoretical Insertion Loss: 1.3 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>70.5</td>
<td>28.7</td>
<td>1226.6</td>
</tr>
<tr>
<td>1, 2</td>
<td>92.4</td>
<td>104.4</td>
<td>1215.0</td>
</tr>
<tr>
<td>2, 3</td>
<td>70.5</td>
<td>28.7</td>
<td>1226.6</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>93.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
FILTER SUMMARY  (frequencies in MHz, dimensions in mils):

2440200A
Center Frequency: 2440.0
3 dB Pass Band Bandwidth: 200.0
Stop Band Attenuation (dB): 20.0
Stop Band Frequency (FSTOP>FCNTR): 2700.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.5
Substrate Thickness: 100.0
Loss Tangent: 0.009430
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 3
Theoretical Insertion Loss: 1.0 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>69.3</td>
<td>27.6</td>
<td>450.6</td>
</tr>
<tr>
<td>1, 2</td>
<td>94.4</td>
<td>137.2</td>
<td>446.3</td>
</tr>
<tr>
<td>2, 3</td>
<td>94.4</td>
<td>137.2</td>
<td>446.3</td>
</tr>
<tr>
<td>3, 4</td>
<td>69.3</td>
<td>27.6</td>
<td>450.6</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>93.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

FILTER SUMMARY  (frequencies in MHz, dimensions in mils):

2440200B
Center Frequency: 2440.0
3 dB Pass Band Bandwidth: 200.0
Stop Band Attenuation (dB): 20.0
Stop Band Frequency (FSTOP>FCNTR): 2700.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.5
Substrate Thickness: 75.0
Loss Tangent: 0.009430
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 3
Theoretical Insertion Loss: 0.9 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>52.2</td>
<td>20.7</td>
<td>454.9</td>
</tr>
<tr>
<td>1, 2</td>
<td>71.0</td>
<td>102.9</td>
<td>450.3</td>
</tr>
<tr>
<td>2, 3</td>
<td>71.0</td>
<td>102.9</td>
<td>450.3</td>
</tr>
<tr>
<td>3, 4</td>
<td>52.2</td>
<td>20.7</td>
<td>454.9</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>70.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
FILTER SUMMARY (frequencies in MHz, dimensions in mils):

400050A
Center Frequency: 4000.0
3 dB Pass Band Bandwidth: 50.0
Stop Band Attenuation (dB): 10.0
Stop Band Frequency (FSTOP>FCNTR): 4100.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.5
Substrate Thickness: 100.0
Loss Tangent: 0.005750
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 1
Theoretical Insertion Loss: 0.6 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>93.9</td>
<td>126.8</td>
<td>264.7</td>
</tr>
<tr>
<td>1, 2</td>
<td>93.9</td>
<td>126.8</td>
<td>264.7</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>93.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

FILTER SUMMARY (frequencies in MHz, dimensions in mils):

400050B
Center Frequency: 4000.0
3 dB Pass Band Bandwidth: 50.0
Stop Band Attenuation (dB): 10.0
Stop Band Frequency (FSTOP>FCNTR): 4100.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.5
Substrate Thickness: 75.0
Loss Tangent: 0.005750
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 1
Theoretical Insertion Loss: 0.5 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>70.6</td>
<td>95.2</td>
<td>269.6</td>
</tr>
<tr>
<td>1, 2</td>
<td>70.6</td>
<td>95.2</td>
<td>269.6</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>70.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
FILTER SUMMARY (frequencies in MHz, dimensions in mils):

5790200A
Center Frequency: 5790.0
3 dB Pass Band Bandwidth: 200.0
Stop Band Attenuation (dB): 10.0
Stop Band Frequency (FSTOP>FCNTR): 5950.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.5
Substrate Thickness: 100.0
Loss Tangent: 0.003970
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 3
Theoretical Insertion Loss: 1.7 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>83.0</td>
<td>50.5</td>
<td>177.5</td>
</tr>
<tr>
<td>1, 2</td>
<td>98.7</td>
<td>363.1</td>
<td>176.9</td>
</tr>
<tr>
<td>2, 3</td>
<td>98.7</td>
<td>363.1</td>
<td>176.9</td>
</tr>
<tr>
<td>3, 4</td>
<td>83.0</td>
<td>50.5</td>
<td>177.5</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>93.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

FILTER SUMMARY (frequencies in MHz, dimensions in mils):

5790200B
Center Frequency: 5790.0
3 dB Pass Band Bandwidth: 200.0
Stop Band Attenuation (dB): 10.0
Stop Band Frequency (FSTOP>FCNTR): 5950.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.5
Substrate Thickness: 75.0
Loss Tangent: 0.003970
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 3
Theoretical Insertion Loss: 1.2 dB

<table>
<thead>
<tr>
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</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>62.5</td>
<td>37.9</td>
<td>182.3</td>
</tr>
<tr>
<td>1, 2</td>
<td>74.3</td>
<td>273.1</td>
<td>181.4</td>
</tr>
<tr>
<td>2, 3</td>
<td>74.3</td>
<td>273.1</td>
<td>181.4</td>
</tr>
<tr>
<td>3, 4</td>
<td>62.5</td>
<td>37.9</td>
<td>182.3</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>70.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
FILTERS DESIGNED WITH THE ENHANCED DESIGN PROCEDURE

(USTRIP 4.2)
FILTER SUMMARY (frequencies in MHZ, dimensions in mils):

914200A
Center Frequency: 914.0
3 dB Pass Band Bandwidth: 200.0
Stop Band Attenuation (dB): 15.0
Stop Band Frequency (FSTOP>FCNTR): 1200.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.5
Substrate Thickness: 100.0
Loss Tangent: 0.025200
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 2
Theoretical Insertion Loss: 1.3 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
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</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>54.0</td>
<td>17.4</td>
<td>1295.7</td>
</tr>
<tr>
<td>1, 2</td>
<td>82.0</td>
<td>47.8</td>
<td>1279.0</td>
</tr>
<tr>
<td>2, 3</td>
<td>54.0</td>
<td>17.4</td>
<td>1295.7</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td></td>
<td>93.9</td>
<td></td>
</tr>
</tbody>
</table>

FILTER SUMMARY (frequencies in MHZ, dimensions in mils):

244050B
Center Frequency: 2440.0
3 dB Pass Band Bandwidth: 50.0
Stop Band Attenuation (dB): 15.0
Stop Band Frequency (FSTOP>FCNTR): 2600.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.5
Substrate Thickness: 75.0
Loss Tangent: 0.003430
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 1
Theoretical Insertion Loss: 0.5 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>69.2</td>
<td>75.1</td>
<td>477.9</td>
</tr>
<tr>
<td>1, 2</td>
<td>69.2</td>
<td>75.1</td>
<td>477.9</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>70.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
FILTER SUMMARY (frequencies in MHz, dimensions in mils):

4000400B
Center Frequency: 4000.0
3 dB Pass Band Bandwidth: 400.0
Stop Band Attenuation (dB): 15.0
Stop Band Frequency (FSTOP>FCNTR): 4400.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.5
Substrate Thickness: 75.0
Loss Tangent: 0.005750
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 3
Theoretical Insertion Loss: 0.9 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>49.0</td>
<td>17.9</td>
<td>281.1</td>
</tr>
<tr>
<td>1, 2</td>
<td>70.1</td>
<td>85.5</td>
<td>278.3</td>
</tr>
<tr>
<td>2, 3</td>
<td>70.1</td>
<td>85.5</td>
<td>278.3</td>
</tr>
<tr>
<td>3, 4</td>
<td>49.0</td>
<td>17.9</td>
<td>281.1</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>70.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

FILTER SUMMARY (frequencies in MHz, dimensions in mils):

5790100C
Center Frequency: 5790.0
3 dB Pass Band Bandwidth: 100.0
Stop Band Attenuation (dB): 10.0
Stop Band Frequency (FSTOP>FCNTR): 5950.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.5
Substrate Thickness: 50.0
Loss Tangent: 0.003970
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 1
Theoretical Insertion Loss: 0.4 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>46.8</td>
<td>54.5</td>
<td>194.0</td>
</tr>
<tr>
<td>1, 2</td>
<td>46.8</td>
<td>54.5</td>
<td>194.0</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>48.0</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
APPENDIX B: Filter Responses
PLOTS OF DATA OBTAINED FROM HP8410 NETWORK ANALYZER
PLOTS OBTAINED DIRECTLY FROM HP8510 NETWORK ANALYZER
PLOTS OF DATA OBTAINED FROM CONVERTED HP8510 DISK FILES
FILTER #400050A-42 S11
APPENDIX C: USTRIP 4.2 User's Guide

USTRIP 4.2 designs a microstrip maximally flat (Butterworth) bandpass filter for a user-specified filter response and substrate characteristics. USTRIP calculates the theoretical insertion loss and all dimensions needed for fabrication of the microstrip filter. USTRIP 4.2 uses a length correction factor for center frequency in addition to equations found in the literature. The usage of USTRIP 4.2 is explained, followed by an example filter design run, and a list of constraints on the specified filter designs.

USING USTRIP 4.2

USTRIP 4.2 is contained in one executable file compiled from Pascal. To begin running USTRIP 4.2, type 'USTRIP42' at the DOS prompt. The screen will display a short message (shown in Figure C.1) concerning the use of the program. USTRIP 4.2 then prompts the user to press any key to begin entering filter response parameters and substrate characteristics. A typical band pass filter response is shown in Figure 2.1.5.

The data entry screen prompts the user for the following values. A brief statement follows each parameter with information concerning the input of that parameter. All frequencies are in MHz and all dimensions are in mils (1
USTRIP 4.2 MICROSTRIP BANDPASS FILTER DESIGN PROGRAM

written by Alan S. Fox
Virginia Polytechnic Institute and State University

USTRIP calculates the physical dimensions for maximally flat (Butterworth) band pass filters given the desired filter response characteristics and substrate parameters. The program produces an approximation for the resonator widths (W) and lengths (L), and the spacings (S) between resonators. It also calculates the theoretical insertion loss.

Press a key to enter design parameters

Figure C.1: USTRIP 4.2 title screen.
mil = 0.001 inch).

Filter Title: 20 character maximum.

Center Frequency (MHz)

3 dB Pass Band Bandwidth (MHz)

Stop Band Attenuation (dB): Attenuation value at the stop band frequency.

Stop Band Frequency (MHz): Must be greater than the center frequency, otherwise a run time error will occur.

Characteristic Impedance (Ω): Refers to the input/output transmission lines.

Relative Epsilon (ε_r): Does not include permittivity of free space (ε_0).

Substrate Thickness (mils): Does not include conductor thickness.

Loss Tangent (unitless): Must be adjusted for the specified center frequency. The loss tangent is inversely proportional to frequency and dielectric constant.

Conductor Conductivity (siemens cm⁻¹ or mhos cm⁻¹): Refers to the resonator conducting material. It is possible to have a different conducting material for the ground plane. Some conductivity values for various conductors were given in Table 2.3.1.

Conductor Thickness (mils): Refers to the resonator conducting material. It is possible to have a different conducting material for the ground plane. Conductor thickness is directly proportional to the weight of copper cladding per square foot. For copper, 1.0 oz. per square foot is equivalent to 1.4 mils, or 35 microns.

A typical data entry screen is shown in Figure C.2. When all entries are made, USTRIP determines the number of poles required to realize the desired response. If the number is greater than 20, the message ‘Pole Limit is 20’ is displayed and
Enter the following parameters (frequencies in MHz, dimensions in mils):

Filter Title: EXAMPLE
Center Frequency: 1350
3 dB Pass Band Bandwidth: 67.5
Stop Band Attenuation (dB): 40
Stop Band Frequency (FSTOP>FCNTR): 1500
Characteristic Impedance: 50
Relative Epsilon: 10.8
Substrate Thickness: 100
Loss Tangent: 0.017
Conductor Conductivity (mhos/cm): 5.8E5
Conductor Thickness: 1.4

Number of Poles Required: 4

Press R to re-enter parameters or another key to start calculation

Figure C.2: USTRIP 4.2 data entry screen with entries for filter design example.
the user is prompted for a filter summary save or printout, discussed subsequently. If the number of poles is not acceptable or an error was made during entry of the parameters, press 'R' to Re-enter the values. When all entries are acceptable, press any key to commence calculation. The message 'Working...' will appear several times during calculation which takes approximately 5 seconds per pole when executing on a 80286 based machine with no math co-processor.

When calculations are completed, USTRIP displays the width, spacing, and length for each resonator pair. It also displays the width of the characteristic impedance line used for the input and output of the filter. All dimensions are in mils. Figure C.3 shows a sample output of USTRIP 4.2.

USTRIP 4.2 then prompts the user for a filter summary save or printout. The summary includes all of the previously entered design parameters, number of poles, theoretical insertion loss, and dimensions for all resonator pairs and the characteristic impedance line. A sample filter summary is shown in Figure C.4. If a summary save or printout is desired, enter 'Y' at the prompt. USTRIP will then prompt the user for a filename or device name. If a summary save is desired, enter the destination filename, and the filter summary will be saved. If a printout is desired, enter the appropriate device name for the desired output device (usually 'LPT1' or 'PRN'), and the filter summary will be printed. Lastly, USTRIP will then prompt the user for another filter design. A 'Y' will restart USTRIP and any other key will exit the program.
**EXAMPLE**

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>71.6</td>
<td>33.1</td>
<td>848.3</td>
</tr>
<tr>
<td>1, 2</td>
<td>93.4</td>
<td>186.3</td>
<td>840.4</td>
</tr>
<tr>
<td>2, 3</td>
<td>95.5</td>
<td>309.8</td>
<td>839.7</td>
</tr>
<tr>
<td>3, 4</td>
<td>93.4</td>
<td>186.3</td>
<td>840.4</td>
</tr>
<tr>
<td>4, 5</td>
<td>71.6</td>
<td>33.1</td>
<td>848.3</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>91.4</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Dimensions in mils

Theoretical insertion loss: 1.7 dB

Save or Print Filter Summary?
Enter File Name or Device Name for Printout: PRN

Another Filter?

Figure C.3: USTRIP 4.2 output screen for filter design example.
USTRIP 4.2 written by Alan S. Fox
Virginia Polytechnic Institute and State University

FILTER SUMMARY (frequencies in MHz, dimensions in mils):

EXAMPLE
Center Frequency: 1350.0
3 dB Pass Band Bandwidth: 67.5
Stop Band Attenuation (dB): 40.0
Stop Band Frequency (FSTOP-FCNTR): 1500.0
Characteristic Impedance: 50.0
Relative Epsilon: 10.8
Substrate Thickness: 100.0
Loss Tangent: 0.017000
Conductor Conductivity (mhos/cm): 580000.0
Conductor Thickness: 1.4
Number of Poles: 4
Theoretical Insertion Loss: 1.7 dB

<table>
<thead>
<tr>
<th>Resonator Pair</th>
<th>Width</th>
<th>Spacing</th>
<th>Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>0, 1</td>
<td>71.6</td>
<td>33.1</td>
<td>848.3</td>
</tr>
<tr>
<td>1, 2</td>
<td>93.4</td>
<td>186.3</td>
<td>840.4</td>
</tr>
<tr>
<td>2, 3</td>
<td>95.5</td>
<td>309.8</td>
<td>839.7</td>
</tr>
<tr>
<td>3, 4</td>
<td>93.4</td>
<td>186.3</td>
<td>840.4</td>
</tr>
<tr>
<td>4, 5</td>
<td>71.6</td>
<td>33.1</td>
<td>848.3</td>
</tr>
<tr>
<td>50.0 ohm</td>
<td>91.4</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Figure C.4: USTRIP 4.2 filter summary for filter design example.
The dimensions obtained from USTRIP can then be used to make a filter layout as shown in Figure 2.1.3. This layout is then used to fabricate the filter. See chapter 3 for more information regarding fabrication and testing.

EXAMPLE

A bandpass filter response is desired with a center frequency of 1350 MHz and 3 dB bandwidth of 67.5 MHz (a 0.05 fractional bandwidth). The desired attenuation at 1500 MHz is 40 dB, and the I/O impedance is 50 Ω. The substrate is a PTFE material with \( \epsilon_r = 10.8 \), 100 mil thickness, 0.0022 loss tangent at 10 GHz, and 1 oz. copper cladding on both sides of the board. Therefore, the following desired response characteristics and substrate parameters were entered at their respective prompts (see Figure C.2):

Filter Title: Design Example

Center Frequency = 1350 MHz
3 dB Pass Band Bandwidth = 67.5 MHz
Stop Band Attenuation = 40 dB
Stop Band Frequency = 1500 MHz
Characteristic Impedance = 50 Ω
Relative Epsilon = 10.8
Substrate Thickness = 190 mils
Loss Tangent = 0.017

270
Conductor Conductivity = \(5.8 \times 10^5 \text{ siemens/cm}\)

Conductor Thickness = 1.4 mils

After calculation, USTRIP 4.2 displays the dimension table shown in Figure C.3. Typing a 'Y' at the filter summary save/printout prompt, and entering 'PRN' at the filename/device name prompt, yields the filter summary shown in Figure C.4. At the prompt for another filter calculation, an 'N' was entered which ends the program.

CONSTRAINTS

The pole limit is 20. It is assumed that no solution will be found for any filter with this many poles.

USTRIP 4.2 gives a no solution condition if any spacing or width dimension is less than 10 mils or greater than 500 mils. Under this condition, USTRIP 4.2 will display the message, 'Filter not physically realizable'.

USTRIP also calculates the theoretical maximum frequency at which microstrip can be used without exciting surface wave modes. A warning message is displayed if the specified stop band frequency is greater than this calculated frequency limit.
PROGRAM DSCONV10;
USES
CRT;
LABEL
0;
VAR
  S11,S21,K1,K2,K3: REAL;
  I,POINTS: INTEGER;
  ANS: CHAR;
  C: STRING[80];
  FNAME,FNAME2: STRING[20];
  FN,FN2: TEXT;

BEGIN
  0;
  CLS;
  (Open files)
  WRITELN('DSCONV 1.0');
  WRITELN('');
  WRITELN('This program converts a data file made from `MAIN` to a file');
  WRITELN('containing all dB magnitude values. It assumes the original file');
  WRITELN('contains only two S-parameters.');
  WRITELN('');
  WRITELN('Enter name of file to be converted: ');;
  READLN(FNAME);
  WRITELN('');
  WRITELN('Enter name of new file: ');;
  READLN(FNAME2);
  WRITELN('');
  WRITELN('Enter number of data points: ');;
  READLN(POINTS);
  ASSIGN(FN,FNAME);
  ASSIGN(FN2,FNAME2);
  RESET(FN);
  REWRITE(FN2);
  (Skip column headings)
  FOR I:=1 TO POINTS DO BEGIN
    READLN(FN,C);
  END;
  (Main loop for converting magnitude ratios into dB)
  FOR I:=1 TO POINTS DO BEGIN
    READ(FN,K1,S11,K2,S21,K3);
    S11:=20*LN(S11)/LN(10);
    S21:=20*LN(S21)/LN(10);
    WRITE(FN2,K1:9:3,S11:9:3,K2:9:3,S21:9:3,K3:9:3);
  END;
  (Close files and ask for another file)
  CLOSE(FN);
  CLOSE(FN2);
  WRITELN('');
  WRITELN('Conversion complete. Convert another file? ');
  ANS:=UPCASE(READKEY);
  IF (ANS='Y') THEN GOTO 0;
END.
PROGRAM HPCONV2;
USES
CRT;
TYPE
FAarray=ARRAY[1..80] OF REAL;
SAarray=ARRAY[1..4,1..80] OF REAL;
LABEL
0,2,3;
VAR
POINTS,FBEGIN,FEND,INCI,MAC,RE,IM: REAL;
I,G,J,L,M,SI,FT: INTEGER;
F: FAarray;
S: SAarray;
ANS,NUMCHAR,CHAR1: CHAR;
C: STRING[80];
FNAME: STRING[20];
K4: STRING[4];
FN,FN2: TEXT;

PROCEDURE LOOKFORM(VAR FN: TEXT);
(Looks for two n's)
VAR
1,COUNTER: INTEGER;
CHAR1: CHAR;
LABEL
00;
BEGIN
COUNTER:=0;
FOR I:=1 TO 50 DO BEGIN
READ(FN,CHAR1);
IF (CHAR1='N') THEN BEGIN
COUNTER:=COUNTER+1;
IF (COUNTER=2) THEN GOTO 00;
END;
END;
00:;
EMP;

FUNCTION NUMWORD(VAR FN: TEXT): STRING;
LABEL
11;
VAR
WORD: STRING[20];
NUMCHAR: CHAR;
J: INTEGER;
BEGIN
WORD:='';
FOR J:=1 TO 15 DO BEGIN
READ(FN,NUMCHAR);
IF (NUMCHAR='.') THEN GOTO 11;
WORD:=WORD+NUMCHAR;
END;
11:
NUMWORD:=WORD;
END;

BEGIN
0;
CLRSCR;
(Open Files)
WRITELN('HPCONV 2.0');
WRITELN('');
WRITELN('This program converts a DATA file produced by "OPTISM" written');
WRITELN('by Dr. Davis to a file containing only data points. The frequency');
WRITELN('for each data point is also included. The program assumes the');
WRITELN('original data file made by the HP8510 was a type DATA file.');
WRITELN('');
WRITELN('Enter name of source file: ');
READLN(FNAME);
ASSIGN(FN,FNAME);
WRITELN('Enter type of file; 1 for non-ASCII or 2 for ASCII: ');
READLN(FT);
WRITELN('Enter name of target file: ');
READLN(FNAME);
ASSIGN(FN2,FNAME);
RESET(FN);
REWRITE(FN2);
(Skip Preliminary Lines and Get # of Data Points)
G:=0;
FOR I:=1 TO 300 DO BEGIN
  READ(FN,NUMCHAR);
  IF (NUMCHAR='G') THEN G:=G+1;
  IF (G=5) THEN GOTO 3;
END;
3:;
READ(FN,FBEGIN,FEND,POINTS);
INC:=(FEND-FBEGIN)/(POINTS-1);
FOR I:=1 TO TRUNC(POINTS) DO BEGIN
  F[I]:=I*INC+FBEGIN/FEND;
END;
WRITELN(F[1],F[TRUNC(POINTS)],INC);
FOR SI:=1 TO 4 DO BEGIN
  LOOKFORM(FN);
  (FT=1, non-ASCII type file)
  IF (FT=1) THEN BEGIN
    FOR I:=1 TO TRUNC(POINTS) DO BEGIN
      FOR M:=1 TO 5 DO BEGIN
        READ(FN,CHAR1);
        IF (CHAR1<CHR(0)) THEN GOTO 2;
      END;
    END;
    2:;
    VAL(NUMWORD(FN),RE,L);
    READLN(FN,IM);
    WRITELN(RE,' ',IM);
    MAG:=SORT(SQR(RE)+SQR(IM));
    S[S1]:=20*LOG(MAG)/LOG(10);
  END;
  WRITELN('END,BEGIN');
  END
ELSE
  (FT=2, ASCII type file)
BEGIN
  READLN(FN,CHAR1);
  FOR I:=1 TO TRUNC(POINTS) DO BEGIN
    VAL(NUMWORD(FN),RE,L);
    READLN(FN,IM);
    WRITELN(RE,' ',IM);
    MAG:=SORT(SQR(RE)+SQR(IM));
    S[S1]:=20*LOG(MAG)/LOG(10);
  END;
  WRITELN('END,BEGIN');
END;
END;
FOR I:=1 TO TRUNC(POINTS) DO BEGIN
  WRITE(FN2,F[I]:9:5);
  FOR J:=1 TO 4 DO WRITE(FN2,' ',S[J,I]:9:3);
  WRITE(FN2,' ');
END;
(Close files and ask for another conversion)
CLOSE(FN);
CLOSE(FN2);
WRITE('"");
WRITE('Conversion complete. Convert another file? ');
ANS:=UPCASE(READKEY);
IF (ANS='Y') THEN GOTO 0;
END.
PROGRAM DLAMBD,A;
USES
CRT;
CONST
PI=3.141592654;
EO=4E-9*PI;
EO=8.854E-14;
LABEL
O;
VAR
EEFO,GEFF,FP,
F,C,E,PU,AP,C,D,E,R,
THKNS,EEFF1,EEFF2,
RS,WHZO,GEFF,THNDA1,LAMBD,A2,RESLN1,RESLN2,F1,F2: REAL;
COUNTER,1: INTEGER;
ANS: CHAR;
FUNCTION WHINIT(Z,ER: REAL): REAL;
{Finds initial value of w/h [Hinton]}
VAR
PREL: REAL;
BEGIN
PREL:=EXP(2/42.4*SQR(T(ER+1))-1;
WHINIT:=8*SQR(T(PREL*(74/ER))/(1+((1/ER)/0.81))/PREL;
END;
FUNCTION WHEFF(WH,CT,THNDA,PI: REAL): REAL;
{Finds effective w/h due to conductor thickness}
BEGIN
IF (WH<(1/2/PI)) THEN WHEFF:=WH+CT/THNDA/PI*(1+LN(4*PI*WH/THNDA/CT));
IF (WH>(1/2/PI)) THEN WHEFF:=WH+CT/THNDA/PI*(1+LN(2/THNDA/CT));
END;
BEGIN
{Initialize variables and get inputs}
O;
CLSCR;
Z0:=50;
CT:=1.4;
ER:=4.7;
WRITELN('DLAMBD,A');
WRITELN('');
WRITELN('This program finds the difference between two resonator lengths');
WRITELN('for two given frequencies. Z0=50 ohms, CT=1.4');
WRITELN('');
WRITELN('Enter the following parameters (frequencies in MHz, dimensions in mils):');
WRITELN('');
WRITELN('ER= ',ER:5:1);
WRITELN('');
WRITELN('Desired frequency: '); READLC(F1);
F1:=F1*1E6;
WRITE('Actual frequency: '); READLC(F2);
F2:=F2*1E6;
WRITE('Substrate Thickness: '); READLC(THNDA);
WRITE('Characteristic impedance'); WHEFO:=(ER+1)/2*(ER-1)/2/SQR(1+12/WHZO);
WHZO:=WHEFF(WH,ER,CT,THNDA,PI);
{Find length of ZO resonator using frequency dependent effective epsilon}
EEFFO:=(ER+1)/2*(ER-1)/2/SQR(1+WHZO);
THNDA:=0.6+0.005*ZO;
FP:=Z0/8/PI/THNDA/0.00254;
EEFF:=ER-(ER-EEFFO)/(1+GEFF*SQR(F1/1E9/FP));

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EEFF2:=ER-(ER-EEFF)/(1+GEFF*SQRT(F2/1E9/FP));
LAMBDA1:=3E6/F1/SQRT(EEFF);
LAMBDA2:=3E6/F2/SQRT(EEFF2);
RESLNGTH1:=(LAMBDA1*1E5/2.54/4);
RESLNGTH2:=(LAMBDA2*1E5/2.54/4);
DELTA:=RESLNGTH1-RESLNGTH2;
WRITE('DELTA LAMBDA (QUARTER WAVELENGTH): ',DELTA:10:1);
WRITE('');
WRITE('DELTA LAMBDA PERCENTAGE: ',DELTA*100/RESLNGTH1:10:1);
WRITE('');
WRITE('Another run? ');
ANS:=UPCASE(READKEY);
IF (ANS='Y') THEN GOTO 0;
CLRSRC;
END.
PROGRAM LINREG2V;
USES
CRT;
TYPE
DARRAY=ARRAY[1..100] OF REAL;
VAR
D,SSE,SSX1,SSY,SSX1Y,SYM1,SYM1SQ,SYM2,SYM2SQ,SYM1SQ,ALEY,AVEX1: REAL;
SUM2,SSX2SQ,SSX2Y,SSX1X2,B1,B2,SYM1X2,RVAL,YINT: REAL;
SSQ,ROOTSSQ: REAL;
X1,X2,Y: DARRAY;
1,N: INTEGER;
ANS: CHAR;
FNAME: STRING[20];
FN1: TEXT;

BEGIN
{Title screen}
CLSCR;
WRITELN('LINREG2V: Linear regression in two variables');
WRITELN('');
WRITELN('Computes the equation for a linear regression of two independent');
WRITELN('variables. Equations are from "Introduction to Statistical Analysis"');
WRITELN('by Dixon and Massey, Jr. Format of data file, "LR2V.DAT": X1 X2 Y.');
WRITELN('');
WRITE('Enter # of data points: ');
READLN(N);
{Open file and get data}
FNAME:='B:LR2V.DAT';
ASSIGN(FN1,FNAME);
RESET(FN1);
FOR I:=1 TO N DO READLN(FN1,X1[I],X2[I],Y[I]);
{Compute sums and sums of squares}
SUM1:=0; SUM1SQ:=0; SUM2:=0; SUM2SQ:=0; SUMY:=0; SUMYSQ:=0;
SUM1Y:=0; SUMX2Y:=0; SUMX1X2:=0;
FOR I:=1 TO N DO BEGIN
  SUM1:=SUM1+X1[I];
  SUM1SQ:=SUM1SQ+SOR(X1[I]);
  SUM2:=SUM2+X2[I];
  SUM2SQ:=SUM2SQ+SOR(X2[I]);
  SUMY:=SUMY+Y[I];
  SUMYSQ:=SUMYSQ+SOR(Y[I]);
  SUMX1Y:=SUMX1Y+X1[I]*Y[I];
  SUMX2Y:=SUMX2Y+X2[I]*Y[I];
  SUMX1X2:=SUMX1X2+X1[I]*X2[I];
END;
{Calculate averages}
AVEX1:=SUM1/N;
AVEX2:=SUM2/N;
AVEY:=SUMY/N;
{Calculate sum of squares of residuals}
SSX1:=SUM1SQ-SQR(SUM1)/N;
SSX2:=SUM2SQ-SQR(SUM2)/N;
SSY:=SUMYSQ-SQR(SUMY)/N;
SX1Y:=SUMX1Y-SUM1*SUMY/N;
SK2Y:=SUMX2Y-SUM2*SUMY/N;
SX1X2:=SUMX1X2-SUM1*SUM2/N;
{Calculate regression and correlation coefficients}
D:=SSX1*SSX2-SQR(SSX1X2);
B1:=(SSX2*SX1Y-SX1X2*SX2Y)/D;
B2:=(SSX1*SX2Y-SX1X2*SX1Y)/D;
SSE:=(SSY-B1*SX1Y-B2*SX2Y)/N;
YINT:=AVEY-B1*AVEX1-B2*AVEX2;
RVAL:=(SSE/SSE)/(N-3);
ROOTSSQ:=SQR(TSSQ);
WRITELN('Write output');
WRITELN('Correlation coefficient: RV:RVAL:5:3);
WRITELN('Variance estimate: SSQ:5:3);
WRITELN('Standard deviation estimate: ROOTSSQ:5:3');
WRITELN('Linear regression equation is:');
WRITELN('Y = B1:11:':X1' + B2:11:':X2' + INT:11');
CLOSE(FN1);
ANS:=READKEY;
END.
PROGRAM USTRIP42:
USES
CRT, PRINTER;
TYPE
  PARRAY=ARRAY[0..20] OF REAL;
CONST
  PI=3.141592654;
  UO=4E-9*PI;
  EO=8.854E-14;
LABEL
  0, 1, 2, 3;
VAR
  ALPHAC, ALPHAD, CT, EEF0, LAMBDA, LAMDAFS, EEEF0, EEEFLO, FMAX, GEEF, GEEF, GEEFL, FP, FPL,
  WSHE, WSHE2, FCNTR, BW, CN, TSTOP, ZD, ER, F, G, H, LA, LI, LISUM, P, QCDIN, OR, QUNY,
  CFPRCNT, THKNS, BPRCNT, CHECK, SHCK, WHCK, WPWP1P, ZO, ZO, ZO, WHSO, WSHE, WHECALC, WHECALC,
  RS, S1, S2, SHLOW, SHHI, SHC, SIGMA, WIL, WHZ0, EEEF, QAMDA, LOSSTAN, ZOO, ZOE: REAL;
  GPRM, JPRM, WSH, WSH, RESLENGTH: PARRAY;
  COUNTER, POLES, I: INTEGER;
  PX, PY: BYTE;
  ANS: CHAR;
  A, TITLE, PTITLE, FN: STRING[20];
  FN: TEXT;

FUNCTION COSHINV(X: REAL): REAL;
{Inverse hyperbolic cosine}
BEGIN
  COSHINV:=LN(X+SQRT(SQR(X)-1))
END;

FUNCTION COSH(X: REAL): REAL;
{Hyperbolic cosine}
BEGIN
  COSH:=(EXP(X)+EXP(-X))/2
END;

FUNCTION WHSOCALC(SH, WHSO, PI, ER, H, G, WHECALC: REAL): REAL;
{Single odd mode w/h calculated using [3.121; CAD]}
VAR
  PREL: REAL;
BEGIN
  PREL:=2/PI*COSHINV((2*H-G-1)/(G-1));
  IF (ER<=6) THEN
    WHECALC:=PREL+8/PI*(ER+2)*COSHINV((1+2*WHECALC)/SH)-WHSO
  ELSE
    WHECALC:=PREL+1/PI*COSHINV((1+2*WHECALC)/SH)-WHSO;
END;

FUNCTION WHINIT(Z, ER: REAL): REAL;
{Finds initial value of w/h [Hinton]}
VAR
  PREL: REAL;
BEGIN
  PREL:=EXP(Z/42.4*SQRT(ER+1))-1;
  WHINIT:=8*SQRT(PREL*(74/ER))/11*(1+1/ER)/0.81)/PREL;
END;

FUNCTION WHEFF(WH, CT, THKNS, PI: REAL): REAL;
{Finds effective w/h due to conductor thickness}
BEGIN
  IF (WH<(1/2/PI)) THEN WHEFF:=WH+CT/THKNS/PI*(1+LN(4*PI*WH/THKNS/CT));
  IF (WH<(1/2/PI)) THEN WHEFF:=WH+CT/THKNS/PI*(1+LN(2*THKNS/CT));
END;

PROCEDURE TESTWHESECALC(SHC, PI, WHSE, ER, CHECK: REAL; VAR WHECALC, WHECALC: REAL);
{Solves for new whse}
VAR C,D,I,L,G,H: REAL;
   COUNTER: INTEGER;
BEGIN
   (Find G from [3.122; CAD])
   G:=COSH(P1/2*SHC);
   (Solve for H using [3.120; CAD] and given whse)
   H:=(G+1)*COSH(P1/2*WHSE)-1)/G/2;
   (Solve for w/h using [3.123; CAD])
   WHCALC:=COSH(NV(H)/P1-SHC/2);
   (Solve [3.121; CAD] for s/h using given whso and bisection method)
   C:=0.00001;
   D:=CHECK;
   COUNTER:=0;
   WHILE (0.001<ABS(C-D)) AND (COUNTER<100) DO BEGIN
      L:=WHSCALC(C,WHSO,P1,ER,H,G,WHCALC);
      E:=(C+D)/2;
      IF (WHSCALC(E,WHSO,P1,ER,H,G,WHCALC)*L<0) THEN D:=E ELSE C:=E;
      COUNTER:=COUNTER+1
   END;
   (Find new whse from [3.122,3.123,3.120; CAD])
   G:=COSH(P1/2*E);
   H:=COSH(P1*WHCALC+P1/2*E);
   WHSCALC:=2/P1*COSH(NV((2*H-G+1)/(G+1))
END;

PROCEDURE ANSWER(CORR: REAL; F1,F2: INTEGER; VAR X: REAL);
(Reads value from prompt and keeps old value if answer is a carriage return)
VAR PX,PY: BYTE;
   L: INTEGER;
   A: STRING[20];
BEGIN
   PX:=WHEREX;
   PY:=WHEREY;
   READLN(A);
   IF (A=' ') THEN
      (Make new value)
      BEGIN
         VAL(A,X,L);
         X:=X*CORR;
      END
   ELSE
      (Keep old value)
      BEGIN
         GOTOXY(PX,PY);
         WRITELN(X/CORR:F1:F2);
      END
END;

BEGIN
   TITLE:='';
   (Title screen)
   CLRSCR;
   WRITELN('USTRIP 4.2 MICROSTRIP BANDPASS FILTER DESIGN PROGRAM');
   WRITELN('');
   WRITELN('written by Alan S. Fox');
   WRITELN('Virginia Polytechnic Institute and State University');
   WRITELN('');
   WRITELN('USTRIP calculates the physical dimensions for maximally flat');
   WRITELN('(Butterworth) band pass filters given the desired filter response');
   WRITELN('approximation for the resonator widths (W) and lengths (L), and');
   WRITELN('the spacings (S) between resonators. It also calculates the');
   WRITELN('theoretical insertion loss.');

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WRITELN('');
WRITELN('Press a key to enter design parameters');
ANS:=READKEY;
0;;
(Enter filter design parameters)
CLRSCR;
WRITELN('Enter the following parameters (frequencies in MHz, dimensions in mils):');
WRITELN('');
WRITE('Filter Title: ');
{Option to keep previous title}
PTITLE:=TITLE;
PX:=WHEREX;
PY:=WHEREY;
READLN(TITLE);
IF (TITLE<>') THEN GOTO 3;
IF (TITLE='') THEN BEGIN
GOTOXY(PX, PY);
TITLE:=PTITLE;
WRITELN(TITLE);
END;
3;;
WRITE('Center Frequency: ');
ANSWER(1E6,0,1,FCNTR);
WRITE('3 dB Pass Band Bandwidth: ');
ANSWER(1E6,0,1,BW);
WRITE('Stop Band Attenuation (dB): ');
ANSWER(1,6,1,LA);
WRITE('Stop Band Frequency (FSTOP>FCNTR): ');
ANSWER(1E6,0,1,FSTOP);
WRITE('Characteristic Impedance: ');
ANSWER(1,6,1,20);
WRITE('Relative Epsilon: ');
ANSWER(1,6,1,ER);
WRITE('Substrate Thickness: ');
ANSWER(1,6,1,THKNS);
CHECK:=510/THKNS;
WRITE('Loss Tangent: ');
ANSWER(1,7,6,LOSSLAN);
WRITE('Conductor Conductivity (ohms/cm): ');
ANSWER(1,6,1,SIGMA);
WRITE('Conductor Thickness: ');
ANSWER(1,6,1,CT);
{Check for excitation of surface wave modes}
FMAX:=3E10/(2*PI/THKNS/0.00254*sqrt(2/(ER-1)))*arctan(ER);
IF (FSTOP>FMAX) THEN BEGIN
WRITELN('');
WRITELN('WARNING: Excitation of surface wave modes will occur at ',FMAX/1E9:4:1,');
END;
{Find # of poles}
BPRCNT:=BW/FCNTR;
WPWIP:=2*(FSTOP-FCNTR)/(BPRCNT/FCNTR);
P:=ln(exp(LA/10-LN(10))-1)/0.99526232/ln(WPWIP)/2;
POLES:=TRUNC(P);
IF (FRAC(P)>0.00) THEN POLES:=POLES+1;
WRITELN('');
WRITELN('Number of Poles Required: ',POLES);
IF (POLES>20) THEN BEGIN
WRITELN('Pole limit is 20');
GOTO 1;
END;
WRITELN('');
WRITELN('Press R to re-enter parameters or another key to start calculation');
ANS:=UPCASE(READKEY);
IF (ANS='R') THEN GOTO 0;
{Find g-parameters and Cn for loss calculation)
GPRM[0]=1;
GPRM[POLES+1]=1;
CN=0;
FOR I:=1 TO POLES DO BEGIN
  GPRM[I]=2*Sin((2*I-1)*PI/2/POLES);
  CN:=CN+GPRM[I];
END;
CN:=CN/2;
{Find J-parameters}
JPRM[0]:=SORT(P1*BPRCNT/2/GPRM[0]/GPRM[1]/ZO);
FOR I:=1 TO (POLES-1) DO JPRM[I]:=P1*BPRCNT/SORT(GPRM[I]*GPRM[1]+1)/2/ZO;
JPRM[POLES]:=SORT(P1*BPRCNT/2/GPRM[POLES]/GPRM[POLES+1])/ZO;
{Find w/h for characteristic impedance}
WHZO:=WEFF(WHINIT(ZO,ER),CT,THKNS,PI);
{Main loop for all poles}
FOR I:=0 TO TRUNC(POLES/2) DO BEGIN
  ZOE:=ZO*(1+JPRM[I]^2/2+SOR((JPRM[I]^2)/ZO));
  ZZO:=ZO*(1+JPRM[I]^2/2+SOR((JPRM[I]^2)/ZO));
  ZOED:=ZOE/2;
  ZOZD:=ZOE/2;
  WHSE:=WEFF(WHINIT(ZOEO,ER),CT,THKNS,PI);
  WHSD:=WEFF(WHINIT(ZOZO,ER),CT,THKNS,PI);
  WRITE('Working...');
{Two nested bisection method iteration loops to find s/h and w/h}
SHLOW:=0.00001;
SHHI:=CHECK;
COUNTER:=0;
WHILE ((0.001<ABS(SHHI-SHLOW)) AND (COUNTER<100)) DO BEGIN
  SHC:=SHLOW;
  TESTWHSECALC(SHC,P1,WHSE,ER,CHECK,WHCALC,WHSECALC);
  WHSE1:=WHSECALC-WHSE;
  SHC:=(SHLOW+SHHI)/2;
  TESTWHSECALC(SHC,P1,WHSE,ER,CHECK,WHCALC,WHSECALC);
  WHSE2:=WHSECALC-WHSE;
  IF ((WHSE1*WHSE2)<0) THEN SHHI:=SHC ELSE SHLOW:=SHC;
  COUNTER:=COUNTER+1
END;
SH[I]:=SHC;
SH[POLES-I]:=SHC;
WH[I]:=WHCALC;
WH[POLES-I]:=WHCALC;
{Check for realizability}
SHCK:=ROUND(SH[I]*THKNS);
WHCK:=ROUND(WH[I]*THKNS);
IF ((SHCK<10) OR (WHCK<10) OR (SHCK>500) OR (WHCK>500)) THEN BEGIN
  WRITE('Filter not physically realizable.');
GOTO 1;
END;
END;
{Write W, S, and L for each resonator pair}
CLRSCR;
WRITE(TITLE);
WRITE('Resonator Pair', ' Width', ' Spacing', ' Length');
{Define correction factor}
CFPRCNT:=(-1.6352E-3*FCNTR/1E6 - 1.6146E-1*THKNS + 2.2627E1)/100;
{Find resonator lengths and write output}
FOR I:=0 TO POLES DO BEGIN
  {Find length of resonator using frequency dependent effective epsilon}
  EEFF0:=ER(ER+1)/2*(ER-1)/2/SORT(I+12/WH[I]);
  IF (WH[I]<1) THEN EEFF0:=EEFF0*(ER[I]/2*0.04*SQR(1-WH[I]));
  GEEFF:=0.6*0.009*ZO;
  FP:=ZO/ZP/I/THKNS/0.00254;
  EEFF:=ER(ER-EEFF)/I+GEEFF*SQR(FCNTR/1E6/FP));
END;
LAMBDA:=3E8/FCNTR/SORT(EFF);
RESLENGTH[1]:=(LAMBDA*1E5/2.54/4);
(Add correction factor)
RESLENGTH[1]:=RESLENGTH[1]*(CFPRCNT+1);
WRITE(12,1,'+',I=12,'
','WH1=THKNS:10:1,');
WRITELN(SH1=THKNS:10:1,'+
','RESLENGTH[1]=10:1);
END;
(Write width for characteristic impedance line)
WRITELN(Z0=5:1, 'hm', 'WH2=THKNS:10:1);
WRITELN('');
WRITELN('Dimensions in mils');
WRITELN('');
WRITELN('');
(Find theoretical insertion loss)
LISUM=0;
FOR 1:=1 TO POLES DO BEGIN
WHL:=WH[1-1]+WH[1]/2;
(Find effective epsilons)
EffeFLO:=EQ(ER[1]/2+(ER[1]-1/2)/SORT(1+12/WHL);
IF (WHL<1) THEN EffeFLO:=EffeFLO+(ER[1]-1/2)*0.04*SQR(1-WHL);
GEEFLO:=0.6+0.009*Z0;
FPL:=Z0/PI/THKNS/0.00254;
Eefflo:=ER-(ER-EffeFLO)/(1+GEEFLO*SQR(FCNTR/1E9/FPL));
LAMBAFS:=3E10/FCNTR;
(Find conductor loss)
RS:=SQR(PI*FCNTR*UO/SIGMA);
S1:=1-SQR(WHL/4);
S2:=1+1/WHL+LN2(THKNS/CT)-CT/THKNS/PI/WHL;
ALPHAC:=6.8*RS*S1/S2/2/PI/10/THKNS/0.00254;
(Find dielectric loss)
ALPHAD:=27.3*ER*(EffeFLO-1)/SQR(EffeFLO)/(ER-1)*LOSSTAN/LAMBAFS;
(Find radiation Q)
F:=(EffeFLO+1)/EffeFLO/QR(EffeFLO-1)/2/EXP3/2*LN(EffeFLO)*LN(SQR(1-EffeFLO)+1)/(SQR(EffeFLO-
QR=20/80/PI/10/THKNS/2*0.00254/LAMBAFS)/F;
(Find total loss for resonator pair)
QCQDINV:=LAMBAFS/2*(ALPHAC+ALPHAD)/PI/SQR(EffeFLO);
QUINV:=(QCQDINV+1/QR);
LISUM:=LISUM+QUINV;
END;
L1:=20*LN(CN*LISUM/POLES+1)/LN(10);
WRITELN('Theoretical insertion loss: ',LI:3:1,' dB');
1;((Filter summary save/print)
WRITELN('');
WRITE('Save or Print Filter Summary? ');
ANS:=UPCASE(READKEY);
WRITELN('');
IF (ANS='Y') THEN BEGIN
WRITE('Enter File Name or Device Name for Printout: ');
READLN(FNAME);
ASSIGN(FN,FNAME);
REWRITE(FN);
WRITEFN(FN,'USTRI 4.2 written by Alan S. Fox');
WRITEFN(FN,'Virginia Polytechnic Institute and State University');
WRITEFN(FN,'');
WRITEFN(FM,'FILTER SUMMARY (frequencies in MHz, dimensions in mils):');
WRITEFN(FM,'');
WRITEFN(FM,'Center Frequency: ',FCNTR/1E6:6:1);
WRITEFN(FM,'3 dB Pass Band Bandwidth: ',BW/1E6:6:1);
WRITEFN(FM,'Stop Band Attenuation (dB): ',LA:6:1);
WRITEFN(FM,'Stop Band Frequency (FSTOP:FCNTR): ',FSTOP/1E6:6:1);
WRITEFN(FM,'Characteristic Impedance: ',Z0:6:1);
WRITEFN(FM,'Relative Epsilon: ',ER:6:1);
WRITEFN(FM,'Substrate Thickness: ',THKNS:6:1);
WRITEFN(FM,'Loss Tangent: ',LOSSTAN:7:6);
WRITE(FN,'Conductor Conductivity (mhos/cm) : ',SIGMA:6:1);
WRITE(FN,'Conductor Thickness : ',CT:6:1);
WRITE(FN,'Number of Poles : POLES:4);
WRITE(FN,'Theoretical Insertion Loss : ',LI:4:1,' dB');
WRITE(FN,'');
IF ((SHCK<10) OR (WHCK<10) OR (SHCK>500) OR (WHCK>500)) THEN BEGIN
  WRITE(FN,'Filter not physically realizable.');
  CLOSE(FN);
GOTO 2;
END;
WRITE(FN,'');
WRITE(FN,'Resonator Pair'' Width'' Spacing'' Length');
FOR I:=0 TO POLES DO BEGIN
  WRITE(FN,I:2,'/''I+1:2,'''WH[I]*THKNS:10:1,'''
  WRITE(FN,SH[I]*THKNS:10:1,'''RESLENGTH[I]:10:1);
END;
WRITE(FN,ZO:5:1,' ohm''WHZO*THKNS:10:1);
CLOSE(FN);
END;
2:
(Ask for another run)
WRITE('');
WRITE('Another Filter? ');
ANS:=UPCASE(READKEY);
IF (ANS='Y') THEN GOTO 0;
CLRSCR;
END.
VITA

Alan Sherwood Fox was born on April 25, 1967, in Alexandria, VA. He has attended college continually since Fall 1985. In December 1988, he received his Bachelor of Science degree in Electrical Engineering with a math minor from Virginia Polytechnic Institute and State University, Blacksburg, VA. Since then, he has worked in the area of mobile and portable communications. His graduate work included design of an indoor RF modem and microstrip band pass filters.

[Signature]

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