ELECTRONICALLY TUNABLE COMBLINE FILTER
USING GALLIUM ARSENIDE
VARACTOR TUNING DIODES

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by

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Abstract

Electronically tunable microwave narrowband bandpass filters have wide applications in many military and commercial microwave systems. The filters are attractive because they exhibit fast tunability, have a reasonably wide tuning band, acceptable passband performance and are low in cost. These same requirements are admirably met by tunable combline bandpass filters where Gallium Arsenide (GaAs) varactors are utilized as the electronic tuning elements.

This thesis presents the systematic design of a fast electronically tunable combline filter using GaAs varactor tuning diodes. These filters suffer from the effect of frequency dependent coupling between the filter resonators. A solution for this problem is presented in this thesis. The design method developed in this thesis also takes into account the effects of varactor parasitic elements. This design method consists of properly characterizing the tuning diodes, calculating the varactor parasitics from the measured results and designing new terminating matching circuits to yield improved and predictable filter performance over the tuning band.
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<td>CVR</td>
<td>Crystal Video Receiver</td>
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<tr>
<td>CW</td>
<td>Continuous Wave</td>
</tr>
<tr>
<td>ESM</td>
<td>Electronic Support Measure</td>
</tr>
<tr>
<td>GaAs</td>
<td>Gallium Arsenide</td>
</tr>
<tr>
<td>MICs</td>
<td>Microwave Integrated Circuits</td>
</tr>
<tr>
<td>Q</td>
<td>Quality Factor</td>
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<td>RF</td>
<td>Radio Frequency</td>
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<td>SSS</td>
<td>Suspended Substrate Stripline</td>
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<tr>
<td>TEM</td>
<td>Transverse Electromagnetic</td>
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<td>YIG</td>
<td>Yittrium-Iron-Garnet</td>
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1. INTRODUCTION

1.1. Background

A number of military and commercial microwave systems require fast electronically tunable microwave filters in which the center frequency can be varied by changing voltage or current. The most well known military application is in passive electronic warfare systems which support military operations by monitoring the electromagnetic spectrum, measuring the intercepted signal parameters and using the data to isolate, identify and locate the intercepted emitter. In military terminology these systems are known as electronic support measures (ESM). Tactical use of ESM to bolster area defence, self-defence (radar warning) and power management for jammer control requires that the data from the passive circuit be available quickly enough to influence the outcome of the hostile engagement [1]. Directional azimuthal coverage may be required if antenna gain is needed to satisfy the required receiver sensitivity. Signal search time is proportional to the product of the number of azimuth and frequency steps. Usually ESM systems must cover a wide frequency band. A small frequency step or narrowband search of each directional segment results in an unacceptably long intercept time that could miss detection of signals which are "on" for a short time only. This will not satisfy tactical military requirements. A wideband receiver covering the full search band almost instantaneously is needed so that the revisit time in the search process is short. The ESM system requirements necessitate that the filter used to select the signal possess a wide tuning bandwidth,
a considerably high quality factor or selectivity, be low in cost, have good environmental
stability, and most importantly, fast tuning speed.

Another military application for a tunable filter is in crystal video receivers (CVRs)
for broadband radar systems. Figure 1.1 shows a block diagram of a typical crystal video
receiver.

![Block Diagram of Crystal Video Receiver](image)

**Figure 1.1:** Typical crystal video receiver.

Most of the operational wideband radar receivers currently used are crystal video
receivers. CVRs are small, simple, and inexpensive to manufacture. The basic crystal
video receiver consists of a radio frequency (RF) bandpass filter, RF preamplifier, square
law video detector and logarithmic video amplifier. The logarithmic video amplifier dis-
plays the logarithmic of the video signal power to obtain a wide display range. The front
end bandpass filter is typically an octave or more in bandwidth and thus provides only
gross frequency resolution. The tunable RF crystal video receiver is simply a CVR with
a tunable bandpass filter in the front end. The bandpass filter, Yttrium-Iron-Garnet
(YIG) or varactor tuned, can be switched in or out as desired and can be tuned across the
broad RF band of the receiver. A receiver using a tunable bandpass filter provides a
means to isolate signals within the broad bandwidth of the CVR and accurately measure
their frequencies, so that receiver sensitivity is higher than the basic CVR. The tunable
notch filter added in the front end (Fig. 1.1) is utilized to eliminate continuous wave
(CW) or any other narrowband interfering signals. Therefore, the CVR is more effective
in noise jamming and high power, high duty cycle pulse environments than the basic CVR.

In commercial applications, fast tunable filters are used in automatic test equipment, which test the system response over different frequency bands. Inexpensive microwave tunable filters are also required for microwave receivers which employ microwave frequency demodulation. However, tuning speed is not important for such receivers.

1.2. Electronically Tunable Filters Used in the Past

The ferromagnetic resonant frequency of YIG spheres can be changed by varying an external DC magnetic field. Thus, filters that employ a YIG sphere as the resonant element may be tuned by varying the DC current [2]. These filters are known as YIG filters and they have a high selectivity. In the past, YIG filters have been the most common choice for systems requiring electronically tunable filters. However, the tuning speed of YIG filters is severely limited by magnetic hysteresis effects. Because of their slow tuning speed, such filters do not meet the fast-tunability requirements of modern systems.

1.3. Fast Electronically Tunable Filters

Fast electronically tunable microwave filters may be realized by using varactor tuned suspended substrate combline structures. Combline printed-circuits with appropriately designed matching networks may be utilized to construct a narrowband bandpass filter with acceptable tuning range and good passband performance. Tuning with varactor diodes provides high tuning speed. Combline realization and varactor characteristics are discussed in detail in the later chapters of the thesis. A brief introduction is given here.
1.3.1. Combline Filter Structure

Combline is the name given to filter structures which are realized using transmission line coupled resonators. The resonator line is grounded at one end while the other end is grounded through a capacitor. Such a structure is shown in Fig. 1.2. The resonators are less than a quarter wavelength long at the filter center frequency. Tuning of these filters can be achieved by varying the capacitors at the end of the resonators.

![Combline Filter Diagram](image)

Figure 1.2: Basic combline filter structure.

1.3.2. Varactor Tuning

A varactor diode is a p-n junction diode which is utilized as a variable capacitor. The variable capacitance is realized as the depletion layer capacitance of the reverse biased p-n junction. As the reverse bias is changed, the width and the associated capacitance of the depletion layer changes as a function of the voltage. There is no hysteresis effect and a very fast variable capacitor can be realized.

Low varactor quality factor (Q) has a detrimental effect on the filter performance as it lowers the resonator quality factor. Silicon p-n junction varactors have a low Q because the resistance associated with the varactor capacitance is high. Because of the higher mobility of carriers in Gallium-Arsenide (GaAs) the resistance of the bulk semiconductor region extrinsic to the junction region is much lower. High Q GaAs tuning varactors are
commercially available, and these varactors have quality factors of up to 5,000 at 50 MHz. A filter tuning speed in excess of 1 GHz/µsec can be achieved using GaAs varactors as the tuning elements. These varactors are used in realizing the proposed fast tunable filter.

1.3.3. Physical Realization of the Filter

Compact and low cost microwave filters are commonly constructed using microstrip and stripline technology (Fig. 1.3).

Figure 1.3: Microstrip, stripline and suspended substrate line structures.
A microstrip line is a printed conductor separated from a ground plane by a dielectric substrate. In a stripline structure, the printed center conductor is sandwiched between two dielectric substrates. The substrates have ground planes on the other sides. Explicit design equations and discontinuity analyses [3, 4, 5] have been developed with acceptable precision for these two transmission line structures. The major problem with these filters is that most of the electromagnetic field is contained in the dielectric substrate. Thus the dielectric constant of the transmission line is influenced by changes in temperature as well as production tolerances of the substrate.

As shown in Fig. 1.3, a suspended substrate structure consists of a center conductor which is printed on a suspended, thin dielectric substrate in an enclosure. The substrate does not have a ground plane. The line impedance is determined by the separation of the substrate from the lower and upper metal ground planes and the width of conductor. Since the field of a suspended substrate line is mainly in the air, the effective dielectric constant is quite close to one. Thus, this realization has high stability with regard to temperature and substrate production variations. Compared to microstrip and stripline, suspended substrate stripline is complicated to design because explicit design equations are not available. Instead, a computer aided design approach must be used [6].

1.4. Thesis Objectives

Conventional combline bandpass filter designs do not address wideband tunability requirements. Because of changes in coupling and matching with frequency, the performance of a filter designed using conventional methods deteriorates over a reasonably wide tuning band and is unacceptable. Other effects which affect performance are transmission loss, imperfect mounting and varactor parasitics. The factors which cause the deterioration have to be investigated and a design procedure which compensates for these
effects has to be developed. The design procedure should yield practical filters with good selectivity and acceptable passband return loss over the full tuning band, and should result in improved performance which is close to that predicted by the design. This thesis is mainly concerned with developing this systematic design procedure. Theoretical modeling and computer aided synthesis are used for developing such a design procedure, computer analysis and experimental work are required for verifying the design procedure. Specific objectives for the thesis are:

1. Investigate the factors which cause performance degradation of a wideband tunable combline filter. If necessary, model and measure these factors.
2. Develop a systematic design procedure for a varactor tuned combline filter. The filter should meet the system requirements of fast tuning speed, good selectivity, reasonable-wideband-tunability, acceptable passband response and return loss as well as high stability.
3. Verify the design procedure by using computer simulations and experimental measurements.

1.5. Thesis Outline

This thesis consists of eight chapters, a list of references and five appendices.

Chapter 1 describes the background of the proposed filter design, states the objectives of the project and outlines the organization of the thesis.

The second chapter is a review of a microwave combline tunable narrowband bandpass filter. It briefly discusses the design of a conventional combline filter and describes the theory of a wideband tunable combline filter.

Chapter 3 describes the characteristics of the microwave varactor diodes used to tune the filter. The equivalent circuit and the detrimental effects to the filter of the diodes are also discussed. An improved method is introduced to measure and calculate the varactor parameters which are to be employed in the filter design.
The tunable combline filter design is presented in Chapter 4. The conventional design procedure must be improved due to the severe performance deterioration when the filter is tuned over a wide range. A new set of explicit design equations is developed in this chapter to meet the system requirements of wideband tuning.

Chapter 5 presents the terminating matching circuit design. It explains the necessity of the terminating matching circuits for reasonably wideband tuning and presents the investigation results of these circuits. A tunable combline filter with a tapped-line terminating matching technique is also described in this chapter.

The sixth chapter discusses the last step of the design procedure, namely the physical realization of the combline filter. Compared to microstrip, suspended substrate stripline can be utilized to meet the proposed system requirements much better.

Chapter 7 presents the filter design example using the new design procedure. The computer simulation and experimental results show good agreement with the performance predicted by the design.

Chapter 8 presents a summary of the thesis and outlines the conclusions and recommendations.

This chapter is followed by a list of references and five appendices.

Appendix A contains two tables in which the measured data of a varactor diode and the computed parameters of diode equivalent circuit are shown.

In Appendix B the relationship between bandwidth and midband frequency for a wideband tunable combline filter is developed.
In Appendix C an equivalent circuit and the related equations for the coupled line structure are derived.

Appendix D consists of a computer program which is written in FORTRAN and is used for the tunable combline filter design.

In Appendix E four computer programs used for the matching circuit design and for the analysis of filter performance are presented.
2. A REVIEW OF COMBLINE TUNABLE BANDPASS FILTERS

2.1. Introduction

Microwave combline bandpass filters have many attractive properties. These filters can be designed to give a narrow bandwidth. But more importantly, they can be tuned over a reasonably wide range of frequencies. These qualities are required in an ESM receiving system [1]. One problem with these filters is the deterioration of filter bandwidth as the filter is tuned over a broad frequency band. This can be solved by the proper design of the terminating matching circuits. This chapter briefly reviews the design of combline filters. Considerations for a wideband tunable combline filter are also presented.

2.2. Basic Design of Microwave Combline Filters

As stated above, a tunable microwave filter may be realized using a combline structure. Before developing the design theory for a tunable filter, a brief description of the general design for a combline filter is given.
2.2.1. General

A combline bandpass filter is composed of a commensurate multiwire transmission line with the coupling constrained to be only between adjacent lines. The resonators in this type of filter consist of these transverse electromagnetic mode (TEM-mode) transmission line elements that are RF short-circuited at one end and have a lumped capacitance, $C_k^L$, between the other end of each resonator line element and RF ground, as shown in Fig. 2.1, where $Y_k$ is the even-mode admittance of resonator line and $Y_{k,k+1}$ is the coupling admittance between line $k$ and line $k+1$.

![Lumped capacitor](image)

**Figure 2.1:** The combline filter.

Because of the capacitance $C_k^L$ at the end of the resonator, the resonator lines are less than a quarter wavelength long. This property of combline filters results in the following attractive features.

1. They are compact.
2. They have strong stopbands and the stopband above the primary passband can be made very broad. The shorter the line, the greater the separation between the primary passband and the secondary passband.
3. If desired, they can be designed to have a steep rate of cutoff on the high side of the passband. The closer to a quarter wavelength long the resonators are at the passband centre, the steeper the rate of cutoff above the passband.
4. Finally and most importantly, because of the use of lumped capacitance as an element of the resonator, it is possible to use an element with electronically tunable capacitance such as a reverse-biased varactor in place of the
capacitor. Thus an electronically tunable filter may be realized by varying the varactor bias.

Microwave combline filters can be accurately designed for bandwidths of less than fifteen percent.

2.2.2. Combline Filter Design

The design procedure derived by Matthaei [7] is based on the use of lowpass prototype filter element values to give the bandpass filter its desired characteristics. The element values for lowpass prototype filters have been tabulated for different type of responses such as Butterworth, Chebyshev, Elliptic, etc [8]. The lowpass to bandpass mapping is used to generate a bandpass response from the lowpass prototype. The bandpass filter response has the same type of passband characteristic as the prototype, but the width of the bandpass filter passband can be specified arbitrarily. The lowpass to bandpass mapping is most accurate for filters with a very narrow bandwidth. This technique is well described in the literature [7, 9, 10, 8].

Two types of passband bandwidth are often used in the bandpass filter design. The equal-ripple bandwidth is the width between the attenuation points on the edges with the specified ripple value, while the 3-dB bandwidth is specified as the width within the 3-dB attenuation points on the edges. The fractional bandwidth is defined as the ratio of the filter bandwidth to the filter center frequency. Nevertheless, it must be noticed that the fractional bandwidth of the completed filter may not be exactly as specified when realized because of the approximations made in the design procedure. Unless otherwise specified, the bandwidth mentioned in this thesis is the equal-ripple bandwidth and not the 3-dB bandwidth.

Matthaei [7] developed a transmission line equivalent circuit, shown in Fig. 2.2,
which consists of parallel short-circuited stubs \((Y)_k\), lumped capacitances \(C_k\), \(k = 1, 2, \ldots, N\), and admittance inverters \(J_{k,k+1}\).

![Figure 2.2: Transmission line equivalent circuit of a combline filter which is reorganized to include admittance inverters.](image)

Fig. 2.3 shows the equivalent circuit of an admittance inverter with stubs of electrical length \(\theta\).

![Figure 2.3: An admittance inverter constructed by a \(\Pi\) connection of stubs of electrical length \(\theta\).](image)

\[ J = |Y \cot \theta| \]

The design for combline filters is based on the equivalent circuits given in Figures 2.1, 2.2 and 2.3. The equivalent circuit of combline filter is compared with a lowpass
prototype. Based on this comparison, a lowpass to bandpass transformation is derived. Then, the design equations for calculating the even-mode admittance $Y_k$ and the coupling admittance $Y_{k,k+1}$, as shown in Fig. 2.1, are developed using the lowpass to bandpass transformation and the element values of the lowpass prototype. The design equations for combline filters were shown in [7, 9, 11]. The design of a tunable combline filter is more complicated. This is discussed in the next section.

2.3. Wideband Tuning Realization in Combline Filters

In this section, the problem which arises when a combline filter is tuned over a wideband is introduced and the solution is discussed.

2.3.1. Problem Introduced by Wideband Tuning

Narrowband microwave bandpass filters rely upon the electromagnetic coupling between resonators to provide the required impedance or admittance inverter elements [12]. This can be shown in Fig. 2.4 which gives a general concept of construction of a bandpass filter from the individual resonators. Each resonator line together with the lumped capacitance has its own resonance response. By properly coupling the lines, a bandpass response is achieved with specified bandwidth, passband ripple, etc.

![Figure 2.4: Bandpass response formed from resonance responses of individual comblines.](image)
Coupling between resonators is necessary and is achieved in combline filters by way of fringing fields between adjacent resonator lines. Naturally, such coupling is frequency dependent and thus the impedance or admittance inverters only operate correctly over a narrow bandwidth.

2.3.2. Compensation for the Frequency Dependence of Coupling

Hunter and Rhodes [12] identified the problem discussed above. They introduced terminating matching circuits to compensate for the internal coupling frequency dependence of tunable combline filters. In their method, based on a similar equivalent circuit and lowpass to bandpass transformation, the frequency dependence of a coupling element or admittance inverter was first shifted to the terminating points. The required matching circuits were designed to match these predistorted terminating point impedances over a broad band. Therefore, the wideband tuning problem was reduced to the network design problem for broad band matching. This idea will be partly used in the proposed design procedure given in Chapter 4 and details are not given here. By using this method, Hunter and Rhodes built a suspended substrate combline bandpass filter with over ± 500 MHz tuning band.

2.3.3. The Consideration of Varactor Parasitics

The varactors used to tune the combline filter exhibit some detrimental effects at microwave frequencies. The resultant filter performance is degraded very severely. The effects are included in the design through the varactor parasitics which are experimentally measured. To obtain an acceptable passband performance over a wide tuning band, the method used in [12] is not sufficient. The proposed design procedure not only takes into account the compensation for the frequency dependence of coupling but also considers the varactor parasitics. Improved filter performance from the proposed procedure, which is discussed in the fourth chapter, is expected.
2.4. Summary

A brief description of the tuning capability of a combline filter has been presented. The general description and design of a combline filter have been briefly discussed as well. Also, the problem of frequency dependent coupling and its solution have been introduced in this chapter. Finally, it has been outlined that varactor parasitics have to be considered in the filter design.
3. MICROWAVE TUNING VARACTOR DIODES

3.1. Introduction

Present Gallium Arsenide (GaAs) varactor diodes used for tuning at microwave frequencies create some detrimental effects on the filter. Before the diodes are used in the tunable filter, their characteristics have to be determined experimentally to achieve a more accurate equivalent circuit for the diodes. The equivalent circuit and the measurement results for microwave diodes are discussed in this chapter.

3.2. The Equivalent Circuit and Detrimental Effects of Microwave Varactor Diodes

For small signal levels, the equivalent circuit of a varactor diode is shown in Fig. 3.1.

![Figure 3.1: Equivalent circuit for a varactor diode.](image-url)
The series resistance $R_s$ represents the resistance of the bulk semiconductor region extrinsic to the junction region and also the resistance due to the metal contacts, $C$ and $R_J$ are the junction capacitance and junction resistance within the depletion region, respectively. These three parameters are principally junction properties which vary quite widely depending upon the type of junction employed. The parallel capacitance $C_p$ is the package capacitance due to the mounting structure. The series inductance $L$ is due to the bonding wires attached to the material containing the junction, and as such is a function of the circuit in which the diode is to be placed, but it is affected little by the type of diode involved. $C$, $R_J$ and $R_s$ are in general the functions of bias voltage [13, 14].

This equivalent circuit is accurate for the microwave diode but too complicated to be used for characterization of diode parameters. In fact, the circuit of Fig. 3.2 has been shown to be sufficient to characterize a large class of microwave diodes [14]. For the tuning varactor diode where reverse bias is commonly utilized as tuning voltage, $R_J$ is quite high. As a result, the microwave AC small signal equivalent circuit of Fig. 3.2 may be reduced to that of a series $L$, $C$ and $R_s$ combination.

![Circuit Diagram]

**Figure 3.2:** Diode equivalent circuit used for characterization.

It has to be noticed that although $L$ is categorized as package effect, it exists in chip tuning varactors also. A reference to the experimental results in [12] shows that even though chip varactors with less bondwire inductance were used in that tunable combline filter, it can be found that the center frequency of the tuning band is about ten percent...
lower than the design value. The filter performance would be more severely distorted if packaged varactors were used.

Other degradation factors may be introduced by imperfect mounting, transmission line loss, etc. The experimental results of this study show that these effects may increase the varactor parasitics, especially $L$, to a high level which cannot be neglected. Fortunately, the advent of GaAs technology makes it possible to realize a high $Q$ varactor which can be used to tune a combline filter above 1 GHz. At these frequencies, $R_s$ exhibits much less of an effect than the series inductance $L$. Thus, compensation for the inductance effect is the major concern in the filter design.

As stated above, the inductance $L$ results from a combination of detrimental effects. These effects are quite significant and compensation for these cannot be implemented by simply shortening the resonator line in a combline filter as proposed in [12]. A systematic design procedure which takes into account these effects should be developed. Obviously, such a procedure must include the varactor parameters which result from the diode measurement and should more closely represent the characteristics of the diode. In the next section, the characterization and measurement of microwave tuning diodes is discussed.

### 3.3. Measurement of Microwave Tuning Diodes

In general, a microwave tuning varactor works with a reverse bias voltage to achieve a variable capacitance. The circuit of Fig. 3.2 without $R_f$ may be used as the equivalent circuit for this varactor. When the varactor is embedded in the test circuit of Fig. 3.3, a simple tuned response may be measured using a scalar network analyzer. The characteristic impedance $Z_0$ of the transmission lines in the test circuit is equal to 50 ohm which is the matched termination for the network analyzer. Because of the calibration
problem, however, the characteristics of microwave diodes are usually difficult to determine at microwave frequencies by using standard reflection coefficient measurement techniques. Instead, the measurement technique described in [14] may be used. This technique is only related to power and frequency measurements. This technique is simpler and more accurate at microwave frequencies than the standard reflection coefficient technique which is limited by calibration difficulties.

![Diode measurement circuit.](image)

Figure 3.3: Diode measurement circuit.

Figure 3.4 is a plot of insertion loss versus frequency, where the insertion loss $|S_{21}|$ is given by

$$|S_{21}| = \frac{\text{Voltage Output at Port 2}}{\text{Voltage Input at Port 1}}.$$  

The parameters to be measured are $l_1$, $l_2$, $f_1$ and $f_2$.

![Typical test circuit response.](image)

Figure 3.4: Typical test circuit response.
As shown in this Figure, $l_2$ is the insertion loss (in dB) at the resonant frequency $f_0$ when the varactor is connected. $f_1$ and $f_2$ are the frequencies at 3-dB attenuation points. $l_1$ is the insertion loss (in dB) at $f_0$ when varactor is disconnected. One can then show [14] that:

$$R_s = \frac{0.5Z_0}{\sqrt{P-1}},$$  \hspace{1cm} (3.1)

$$C = \frac{1}{\pi Z_0} \cdot \frac{f_2-f_1}{f_1f_2} \cdot \sqrt{P-2} \cdot \left(1 - \frac{1}{\sqrt{P}}\right),$$ \hspace{1cm} (3.2)

$$L = \frac{1}{4\pi^2 f_1f_2C},$$ \hspace{1cm} (3.3)

where $P$ is given by

$$P = 10^{a_1(l_2-l_1)}.$$ \hspace{1cm} (3.4)

There are some differences between the present measurement method and the technique used in [14]. The differences are:

1. The type of transmission line, e.g. microstrip or suspended substrate stripline, etc., is the same as that used in the filter implementation.
2. The transmission line loss $l_1$ is included in (3.4) in the parameter calculation.
3. Accordingly, the measured data take into account the transmission line loss, mounting effects and varactor parasitics.

The diode parameters calculated from the measurement are to be employed in the development of a new design procedure. This design procedure has to include the combined effect of frequency dependent coupling and varactor parasitics. This procedure is developed in the next chapter.
3.4. Example: Measurement of ND3050-3D (NEC) Tuning Diode

This diode is specified as an S to K-band tuning diode. The package and mounting effects are significant for the tunable filter application. The diode was characterized using a microstrip line test circuit on a 0.79mm Teflon substrate with $\varepsilon_r=2.2$. This microstrip line was to be used in the filter implementation afterwards. The data were measured at one volt bias step and are listed in Table A.1. The measurement results are shown in Fig. 3.5.

![Typical measurement results](image)

Figure 3.5: Typical measurement results. Scale – X axis: 100 MHz/Div; Y axis: 3 dB/Div.

The measured results were plotted over a frequency range of 4 GHz to 5 GHz with 100 MHz per division. Both vertical axes were 3 dB per division. The center points for the vertical axes in the left and the right plots were 0.13 dB and -14.87 dB, respectively.

The measured data were used to calculate $R_s$, $C$ and $L$ from (3.1) to (3.4). The results are plotted for different bias values in Fig. 3.6. As shown in this figure, the small
Figure 3.6: The changes of the varactor parameters with the bias voltage.
circles identified the calculated data. The smooth curves showed the estimates of the average values.

From these plots it can be seen that the inductance $L$ does not vary with bias voltages. The average value of $L$ is about 2.1 nH which is a few times higher than the published data. This is a result of the combination of effects discussed above. $R_s$ is very small and has an average value of around 1.7 $\Omega$. In general, the quality factor of a resonant circuit is defined as

$$Q = \frac{\text{Energy Stored in the Resonant Circuit at } \omega_r}{\text{Power Loss in the Resonant Circuit at } \omega_r},$$

where $\omega_r$ is the resonant frequency. For the series resonant circuit of $L$, $C$ and $R_s$,

$$Q = \frac{\omega_r L}{R_s} = \frac{1}{\omega_r CR_s}. $$

Thus the quality factor at 2 GHz and -4.5 volts is given by

$$Q = \frac{1}{2\pi f R_s C} = 67.$$

The diode has a reasonably high $Q$ factor for microwave filter applications. The published data of junction capacitance versus bias voltage is shown as the curve pointed by an arrow in Fig. 3.7 [15], where the reverse voltage and the junction capacitance are represented by $V_R$ and $C_J$, respectively. The capacitance shown in Fig. 3.6 is an effective capacitance which takes into account the package effect of varactor. Due to this package effect, the variable range of the varactor capacitance (Fig. 3.6) is less than the published data.
Figure 3.7: Published data for junction capacitance vs. reverse voltage [15].

3.5. Summary

The equivalent circuit and characteristics of a microwave varactor diode have been described in this chapter. An improved measurement method has been discussed. As shown in the example, the measured parameters which include the transmission line loss, mounting effects and varactor parasitics are to be used in the new design procedure which is to be developed in the next chapter.
4. THE TUNABLE COMBLINE FILTER DESIGN

4.1. Introduction

Reverse biased tuning varactor diodes can be used to tune the combline filter. The combination of the effects of transmission line loss, imperfect mounting and varactor parasitics are severe at microwave frequencies. Based on the measured and calculated varactor parameters which take these effects into account, a new set of design equations is developed in this chapter to compensate for these effects and the frequency dependence of coupling as discussed in Chapter 2.

4.2. Unloaded Q of the Combline Resonator

The unloaded combline without a tuning varactor is a short-circuited transmission line. Rizzi [16] derived a formula to calculate the unloaded Q value for a short-circuited line as follows

\[ Q_{unloaded} = \frac{\beta}{2\alpha} = \frac{\pi}{\alpha\lambda_g}, \tag{4.1} \]

where \( \alpha \) and \( \beta \) are the attenuation constant and phase constant of the transmission line at resonant wavelength \( \lambda_g \), respectively.
Using the typical values given in [16], for an air-insulated short-circuited TEM line with

$$\alpha = 0.005 \text{ Neper/m},$$

at 2 GHz*, an unloaded Q of about 4000 can be expected. This calculation is an optimistic estimation for unloaded Q. For short-circuited microstrip line, the attenuation constant is much higher so that its unloaded Q decreases to about 500.

A combline tunable filter resonator is composed of a short-circuited combline and a tuning varactor. Hence the unloaded Q of the combline resonator is related to the Q factors of these two elements. Typically quality factors of tuning varactor diodes at 2 GHz are less than 150, thus the selectivity of a combline filter is predominantly controlled by the varactor loss.

4.3. The Equivalent Circuits of GaAs Tuning Varactor Diodes

Because of its high quality factor, a reverse biased GaAs varactor diode is used as the tuning capacitance for a wideband tunable bandpass filter. The varactor parasitics have to be considered in the design. As stated in the previous chapter, for small signal levels, the equivalent circuit of a tuning varactor diode is as shown in Fig. 4.1. The series resistance $R_s$ and junction capacitance $C$ are functions of the bias.

![Equivalent Circuit for a Varactor Diode](image)

**Figure 4.1:** Equivalent circuit for a varactor diode.

*1 Neper = 8.686 dB.
The quality factor and admittance of the varactor are given by

\[ Q = (\omega C R_s)^{-1}, \quad \text{and} \]

\[ Y' = G' + jB' = \frac{\omega^2 C^2 R_s}{(1-\omega^2 L C)^2 + \omega^2 C^2 R_s^2} + j \frac{\omega C (1-\omega^2 L C)}{(1-\omega^2 L C)^2 + \omega^2 C^2 R_s^2}, \]

(4.4)

where \( G' \) represents the varactor conductance, \( B' \) symbolizes the susceptance utilized for tuning and \( \omega \) denotes the radian frequency as usual.

4.4. Development of the New Design Equations

For simplicity, the proposed design concentrates on two common cases, a high Q tuning diode with significant parasitics and a chip varactor with low Q.

4.4.1. High Q Tuning Diode with High Inductance

This is the most common case. As stated in the previous chapter, mounting effects and varactor parasitics can increase the inductance \( L \) to a high level which cannot be neglected. For high Q tuning diodes, such as ND3050-3D (NEC) previously measured, \( \omega^2 C^2 R_s^2 < < 1 \) and (4.4) can be simplified as

\[ Y' = G' + jB' = \frac{\omega^2 C^2 R_s}{(1-\omega^2 L C)^2} + j \frac{\omega C}{1-\omega^2 L C}. \]

(4.5)

Using the equivalent circuits of the combline filter shown in Figures 2.1, 2.2 and 2.3, one may write

\[ (Y)_1 = Y_1 + Y_{12} \]

\[ (Y)_k = Y_k + Y_{k,k-1} + Y_{k,k+1} \quad ; \quad k = 2 \text{ to } N-1, \]

\[ (Y)_N = Y_N + Y_{N-1,N} \]

(4.6)
\[ J_{k,k+1} \equiv \frac{Y_{k,k+1}}{\tan \theta} = \frac{Y_{k,k+1}}{\tan (\omega \theta_0/\omega_0)} ; \quad k=1 \text{ to } N-1 , \]  
(4.7)

where \( Y_k \) and \( \theta_0 \) are the even-mode admittance and electrical length of the resonator line at midband frequency \( \omega_0 \), \( Y_{k,k+1} \) is the coupling admittance between line \( k \) and line \( k+1 \) and \( J_{k,k+1} \) is the J-inverter constant as defined in (4.7) [7].

As seen from (4.7), the inverter constant \( J_{k,k+1} \) is frequency dependent. To compensate for this effect which leads to filter performance deterioration with wideband tuning, the ideal transformers shown in Fig. 4.2 are introduced, where \( n \) is the transformer turns ratio, \( Y' \) is the admittance of the varactor diode as given by (4.5). \( G' \) and \( B' \) have the same definitions as before. It is assumed that the same type of varactor is used to tune each filter section.

By calculating the transfer matrix of the circuit within the dashed line box shown in Fig. 4.2, it can be shown that this circuit is equivalent to an admittance inverter with an inverter constant given by

\[ J'_{k,k+1} = n^2 J_{k,k+1} . \]  
(4.8)
By choosing the transformer turns ratio \( n \) such that
\[
n^2 = n_0 \tan \theta ,
\]
where \( n_0 \) is a constant to be determined, the frequency dependence of \( J_{k,k+1} \) is removed so that
\[
J'_{k,k+1} = n_0 J_{k,k+1} .
\]

A typical lowpass filter prototype with admittance inverters [9] is shown in Fig. 4.3.

![Figure 4.3: An admittance inverter type of lowpass prototype.](image)

The inverter constants are given by
\[
J'_{01} = \sqrt{\frac{C_{a1} G_A}{g_0 g_1}} , \quad J'_{k,k+1} = \sqrt{\frac{C_{a(k+1)}}{g_k g_{k+1}}} ,
\]
\[
J'_{N,N+1} = \sqrt{\frac{C_{aN} G_B}{g_N g_{N+1}}} .
\]

Parameters \( g_0, g_1, \ldots, g_{N+1} \) are the lowpass filter elements which are well tabulated in literature [8]. \( G_A \) and \( G_B \) represent source and load conductances, respectively. \( C_{a1}, C_{a2}, \ldots, C_{aN} \) can be chosen arbitrarily. Usually they are chosen to be identical to simplify the design. The first and end inverters transform \( G_A \) and \( G_B \) to the admittances.
respectively. Comparing the filter internal section in Fig. 4.2 to the lowpass prototype and using (4.5) for \( B' \), a lowpass to bandpass transformation can be written as

\[
\omega \rightarrow \frac{n_0}{C_{ak}} \left( \frac{\omega \tan \theta}{1-\omega^2LC} \right) (Y)_k.
\]

At midband frequency,

\[
(Y)_k = \frac{\omega_0 \tan \theta_0}{1-\omega_0^2LC}.
\]

Also, assuming a small fractional bandwidth for the filter, from (4.14) the filter bandwidth \( \Delta \omega \) can be expressed as

\[
\Delta \omega = \frac{2C_{ak}}{n_0(Y)_k} \left( \frac{\omega_0 \tan \theta_0}{\theta_0 + \tan \theta_0 + \theta_0 \tan^2 \theta_0 + 2(Y)_k \omega_0 L} \right).
\]

Details of the derivation of (4.16) can be found in Appendix B. Equation (4.16) shows that bandwidth \( \Delta \omega \) changes with \( \theta_0 \) or resonant frequency \( \omega_0 \). Usually \( C_{ak}, k = 1, 2, \ldots, N \) are chosen to be equal so that \( (Y)_k, k = 1, 2, \ldots, N \) are identical. To achieve the minimum bandwidth change with tuning frequency one may apply

\[
\frac{\partial \Delta \omega}{\partial \theta_0} = 0,
\]

which gives

\[
(Y)_k = \frac{\theta_0^2 \tan^2 \theta_0 \sec^2 \theta_0 - \tan^2 \theta_0 - \theta_0^2 \sec^2 \theta_0}{2\omega_0 L \theta_0 \sec^2 \theta_0}.
\]

From (4.15) and (4.17) the equation for the optimum electrical length \( \theta_0 \) of the resonator line is obtained as shown below.
A step by step design procedure can be developed as shown below.

1. Numerically solve (4.18), to get the optimum value of $\theta_0$.
2. From (4.15) calculate $(Y)_k$.
3. From (4.16) calculate
   \[
   n_0 = \frac{2C_{ak}}{(Y)_k \Delta \omega} \cdot \frac{\omega_0 \tan \theta_0}{\theta_0 + \tan \theta_0 + \theta_0 \tan^2 \theta_0 + 2(Y)_{\omega_0} L}. 
   \]
4. Using (4.10) and (4.11) calculate all mutual admittances $Y_{k,k+1}$.
5. Using (4.6) calculate all line admittances $Y_k$.

### 4.4.2. Chip Varactor with Low Q

In this case, inductance $L$ can usually be neglected. Equation (4.4) can then be simplified as

\[
Y' = G' + jB' = \frac{\omega^2 C^2 R_s}{1 + \omega^2 C^2 R_s^2} + j \cdot \frac{\omega C}{1 + \omega^2 C^2 R_s^2}. 
\]

The lowpass to bandpass transformation is given by

\[
\omega \rightarrow f(\omega) = \frac{n_0}{C_{ak}} \cdot \left( \frac{\omega C \tan \theta}{1 + \omega^2 C^2 R_s^2} - (Y)_k \right). 
\]

Using the procedure in Section 4.4.1, another set of design formulas can be developed.

The design procedure may then be given as:

1. Numerically solve the equation below to get $\theta_0$:
   \[
   \frac{2\omega_0^2 C^2 R_s^2 \tan^2 \theta_0}{1 + 2\omega_0^2 C^2 R_s^2} = \frac{\tan^3 \theta_0 + \theta_0^2 \tan^2 \theta_0 \sec^2 \theta_0 - \theta_0^3 \tan^2 \theta_0 \sec^2 \theta_0}{\tan \theta_0 + 2\theta_0 \sec^2 \theta_0}. 
   \]
2. Calculate $(Y)_k$ from
\[ (Y)_k = \frac{\omega_0 C \tan \theta_0}{1 + \omega_0^2 C^2 R_s^2} . \]

3. Calculate \( n_0 \) from

\[ n_0 = \frac{2C \alpha_k}{(Y)_k \Delta \omega} \cdot \frac{\omega_0 \tan^2 \theta_0}{\theta_0 \tan \theta_0 + \tan^2 \theta_0 + \theta_0 \tan^3 \theta_0 - 2R_s(Y)_k^2} . \]

4. Using (4.10) and (4.11) calculate all \( Y_{k,k+1} \).
5. Using (4.6) calculate all \( Y_k \).

### 4.5. Input and Output Considerations

The method developed in the previous section does not remove the frequency dependence of the admittance inverter from the filter; it merely transfers this dependence onto the terminating conductances which now have a value of

\[ G_A' = n^2 G_A = n_0 G_A \tan \theta , \tag{4.21} \]

as shown in Fig. 4.4. Octave tuning around \( \theta_0 \) will result in a significant change in these elements and a corresponding drastic deterioration in the filter performance [12]. Thus the terminating matching circuits between the terminating conductance and the first resonator section should be used to compensate for the frequency dependence which now exists in the transformer turns ratio \( n \). The design of the terminating matching circuits to compensate for the frequency dependence of the turns ratio is developed in the next chapter.
4.6. Summary

The varactor model used as the tuning elements of the combline filter has been discussed. The theoretical development resulting in the first half of the design procedure has been carried out. The need to design a matching circuit to compensate for the frequency dependence of the filter has been established.
5. THE TERMINATING MATCHING CIRCUIT DESIGN

5.1. Introduction

The terminating matching circuits can be used to remove the frequency dependence of admittance inverters and to compensate for the effects of the varactor. Several terminating circuits are discussed and their suitability for frequency dependence compensation is evaluated. A new matching circuit with a tapped-line input and output is introduced. This is the configuration selected for the filter implementation.

5.2. Criteria of Designing Terminating Matching Circuits

The input equivalent circuit shown in Fig. 5.1 is the circuit of Fig. 4.4 with a matching circuit connected.

![Input equivalent circuit for the combline filter with matching circuit.](image)

**Figure 5.1:** Input equivalent circuit for the combline filter with matching circuit.
\[ G' \text{ can be calculated from (4.4) as} \]
\[
G' = \frac{\omega^2 C^2 R_s}{(1 - \omega^2 LC)^2 + \omega^2 C^2 R_S^2}.
\] (5.1)

The input matching circuit should be designed to maintain the passband ripples and to achieve a good overall quality factor. Comparing Fig. 5.1 to the input section in Fig. 4.3, and considering a normalized system, i.e.

\[ G_A = 1, \] (5.2)

one can set the conditions
\[
\text{Re} \{ Y_{IN}(\theta) \} \rightarrow Y_{ina} \quad \text{and} \quad \text{Im} \{ Y_{IN}(\theta) \} \rightarrow 0
\] (5.3) (5.4)

over the tuning band as the design criteria. Here \( \theta \) is a function of the tuning frequency and \( Y_{ina} \) is given in (4.12).

As seen from the circuit in Fig. 5.1, the transformer will deteriorate the admittance \( Y_{IN1} \), looking from node 1 to the source, by a factor \( \tan \theta \) which increases with frequency. Thus to satisfy (5.3) and (5.4), a \( Y_{IN2} \) which decreases with frequency has to be achieved by proper design of the matching circuits. Also, it should be noted that \( C \) in (5.1) is a function of the tuning frequency. From (4.15) \( C \) is given by
\[
C = \frac{(Y)_k}{\omega_0 \tan \theta_0 + (Y)_k \omega_0^2 L}.
\] (5.5)

For the output matching circuit the same design criteria as the input can be applied. Generally these two terminating circuits are identical.

At first \( Y_{IN}(\theta) \) will be considered for a number of simple matching circuits commonly used. A systematic synthesis method to obtain the matching circuit design is subsequently developed.
5.3. Parallel-Coupled Lines as Terminating Circuits

Matthaei [17] has given equivalent circuits for the parallel-coupled transmission line sections which may be used for the matching circuit realization. Hunter and Rhodes [12] used one of the parallel-coupled line sections as the terminating matching circuit to achieve compensation for the frequency dependent coupling of the inverters. These matching circuits along with two other circuits are discussed first.

5.3.1. The Parallel-Coupled Line for Combline Filter Matching Circuits

As shown in Fig. 2.1, the combline filter consists of resonator lines which are short-circuited at one end. Thus, if one of the parallel-coupled lines for terminating is grounded at one end, this line can be included in the first resonator line of the combline filter. By using these kinds of parallel-coupled lines one can

1. Decrease the dimension of filter.
2. Increase the realizable impedance range of this grounded line to include some negative value as this value can be absorbed by the first resonator line.

There are three circuits which may be used for coupled line realization. Their schematics and equivalent circuits are shown in Fig. 5.2. These circuits are discussed one by one in the following sections.

5.3.2. Circuit I

As shown in Fig. 5.2(a), this circuit which was used in [12] compensates for the frequency dependence of coupling.
(a) Circuit I and its equivalent circuit.

(b) Circuit II and its equivalent circuit.

(c) Circuit III and its equivalent circuit.

Figure 5.2: Parallel-coupled lines circuits for terminating matching.
If the varactor loss is neglected, the admittance $Y_{in}(\theta)$ in Fig. 5.1 can be calculated as

$$\text{Re}Y_{in}(\theta) = n_0 \cdot \frac{Y_0^2 \tan \theta}{(Y_0 + Y_{o1})^2 + \tan^2 \theta},$$  \hspace{1cm} (5.6)

$$\text{Im}Y_{in}(\theta) = -n_0 \cdot \frac{(Y_0 \tan^2 \theta + (Y_0 + Y_{o1})(Y_0 Y_{o1} + Y_0 Y_{o1} + Y_0 Y_{o1})}{(Y_0 + Y_{o1})^2 + \tan^2 \theta}.$$  \hspace{1cm} (5.7)

For the normalized system with $Y_{ina}=1$ one can set the conditions for $Y_{in}(\theta)$ as follows

$$\text{Re}Y_{in}(\theta_0) = 1,$$  \hspace{1cm} (5.8)

$$\text{Im}Y_{in}(\theta_0) = 0,$$  \hspace{1cm} (5.9)

$$\frac{\partial \text{Re}Y_{in}(\theta)}{\partial \theta} = 0,$$  \hspace{1cm} (5.10)

Using these conditions along with (5.6) and (5.7) one has

$$Y_0 = \tan \theta_0 - \frac{\sqrt{2\tan \theta_0}}{n_0},$$  \hspace{1cm} (5.11)

$$Y_{o1} = \frac{\sqrt{2\tan \theta_0}}{n_0},$$  \hspace{1cm} (5.12)

$$Y_0' = -\frac{1}{\sqrt{2n_0}} \cdot (2\sqrt{\tan \theta_0} - \frac{\sqrt{2}}{n_0}),$$  \hspace{1cm} (5.13)

where $\theta_0$ and $n_0$ are obtained from the first-step of the design described in the previous chapter.

Inclusion of $G'$ makes the derivation of the explicit equations of $Y_0$, $Y_{o1}$ and $Y_0'$ very difficult. In this case a computer synthesis method must be used.
5.3.3. Circuit II

Matthaei [7, 9] used the circuit shown in Fig. 5.2(b) to match the combline filter to 50 ohm. For this circuit one obtains

\[
\text{Re} \, Y_{IN}(\theta) = n_0^2 \frac{\tan \theta + \tan^3 \theta}{(Y_0 + Y_{01})^2 + \tan^2 \theta},
\]

(5.14)

where \(G'\) was ignored. An examination of (5.14) shows that the real part of \(Y_{IN}(\theta)\) increases with frequency. This is against the present design criteria and this circuit is not suitable for this application.

5.3.4. Circuit III

This circuit is shown in Fig. 5.2(c). The real part of \(Y_{IN}(\theta)\) for this case is given by

\[
\text{Re} \, Y_{IN}(\theta) = \frac{n_0 \tan^3 \theta}{N_1^2 (Z_0^2 + \tan^2 \theta)},
\]

(5.15)

where \(Z_0\) and \(N_1\) are frequency independent. Examination of this circuit reveals that it is unsuitable for the proposed design due to the same reason as in circuit II.

The equivalent circuit of parallel-coupled lines may be considered as a special case of \(\Pi\) networks. Thus the search can be extended to other combinations of \(\Pi\) networks although they may not be realized in parallel-coupled lines.

5.3.5. A Special \(\Pi\) Matching Circuit

The \(\Pi\) matching circuit used for compensation is shown in Fig. 5.3.
Figure 5.3: A Π circuit for matching the combline filter.

The real and imaginary parts of $Y_{\Pi}(\theta)$ for this circuit are given by

$$\text{Re}Y_{\Pi}(\theta) = n_0 Y_{01} \cdot \frac{\tan \theta + \tan^3 \theta}{(Y_{01} - Y_0 \tan^2 \theta)^2 + \tan^2 \theta} ,$$

(5.16)

$$\text{Im}Y_{\Pi}(\theta) = n_0 \cdot \frac{-(Y_0 + Y_{01})\tan^2 \theta + (Y_{01} - Y_0 \tan^2 \theta)[(Y_{01} Y_0 + Y_{01} Y_0')\tan^2 \theta - Y_{01} Y_0']}{(Y_{01} - Y_0 \tan^2 \theta)^2 + \tan^2 \theta} .$$

(5.17)

Thus

$$\text{Re}Y_{\Pi}(\theta) = \text{Re}Y_{\Pi1}(\theta) + n_0 G \tan \theta ,$$

(5.18)

$$\text{Im}Y_{\Pi}(\theta) = \text{Im}Y_{\Pi1}(\theta) .$$

(5.19)

By proper selection of $Y_0$, $Y_{01}$ and $Y_0'$, one can design the circuit to approach (5.3) and (5.4) over the required tuning band. This was done using computer synthesis.

A design example using this terminating matching circuit is given in [18] together with the corresponding simulation results. The chip varactor filter example in Chapter 7 uses this matching circuit.

Simulation results show that $Y_{01}$ is a very high impedance line in narrowband (5% to 10% fractional bandwidth) filters. In this case, the effects of the tee junction formed by the input 50-ohm line and $Y_0$ and $Y_{01}$ lines become significant. Also, $Y_0'$ has a large
negative value which makes the first resonator line very narrow and the gap between the first and second resonator lines very small. This increases the effect of the varactor leads. These implementation problems make this circuit difficult to implement and are magnified for filters with fractional bandwidths of less than five percent.

5.4. The Terminating Matching Circuit with Tapped-Line Input

For the physical realization of narrow bandwidth combline filters, the conventional input and output coupling circuits described in [7, 9] are sometimes replaced with direct tapped connections as shown in Fig. 5.4 [19]. The examination of this circuit shows that a tapped-line input and output type circuit is also suitable as a matching circuit with frequency compensation.

![](image)

**Figure 5.4:** The combline filter with tapped-line terminating.

5.4.1. The Equivalent Circuit for Tapped-Line Input

Caspi and Adelman [20] suggested an equivalent circuit which allows the derivation of explicit expressions for the tapped-line input and makes the subsequent design procedure very similar to that used in [7, 9]. This equivalent circuit is shown in Fig. 5.5 from node 1 to node 2.
Figure 5.5: Equivalent circuit of a combline with tapped-line input.

The parameters $\beta$ and $\delta$ are given by

$$\beta = \frac{J \sin(\theta - \phi)}{Y_T},$$

$$\delta = \frac{\sin \theta}{\sin \phi},$$

$$Y_T = (Y)_1 - \frac{Y_{12}^2}{(Y)_2},$$

where $(Y)_1$, $(Y)_2$ and $Y_{12}$ have the same definitions as in Chapter 4, $\theta$ and $\phi$ are the electrical lengths as shown in Fig. 5.4. Obviously, if a terminating circuit is applied between the source and node 1 in Fig. 5.5, some of the design equations in [20] would be invalid. For example, the calculation of $\phi_0$ should be based on (5.3) and (5.4).

5.4.2. The Design of A Matching Circuit with Tapped-Line Input

As stated earlier, the matching circuit should satisfy (5.3) and (5.4). Also the practical implementation of this circuit should be feasible. In particular, the line impedances, mutual impedances and the lengths of the lines in the matching circuit should be practically realizable.

Theoretical derivation for tap point $\phi_0$ (where 0 indicates the center frequency) is
very complicated. Computer aided synthesis is utilized to derive the design of the tapped-line input circuit. It has been indicated in [20] and [21] that the tapped-line input response is very similar to the traditional transformer input. This similarity implies that the conclusions derived for different matching circuits in the previous two sections are still valid.

The computer simulation shows that the resultant optimized line impedance values of circuit I (Fig. 5.2(a)) are not physically realizable. However, a circuit similar to the special Π circuit (Fig. 5.3) can be used to achieve a very good match to criteria (5.3) and (5.4). The design using this circuit also yields impedances which are realizable. This circuit is described below.

5.4.3. Proposed Matching Circuit with Tapped-Line Input

The section $Y'_0$ in Fig. 5.3 can be absorbed by the first resonator line of the filter. It can be neglected when (5.3) is applied first, as it only affects the imaginary part of $Y_{IN}$. With the presence of the tapped-line input, the proposed circuit is shown in Fig. 5.6.

![Figure 5.6: Matching circuit for tapped-line input combline filter.](image)
Now the optimization objectives can be set as:

Step 1  \( \text{Re}[n^2 Y_{IN2}] = Y_{\text{ina}} \),

Step 2  \( \text{Im}[Y_{IN}] = 0 \).

Only lines \( Y_0 \) and \( Y_{01} \) are considered in step 1. Short-circuited line \( Y_0 \) and adjustment capacitance \( C_{PC} \) with a very small value are used in step 2. Note that \( \theta_1 \) and \( \theta_2 \) need not be identical.

Another reason in favour of this circuit is its suitability for realization as a coupled line structure when \( \theta_1 \) equals \( \theta_2 \). In this case, the proposed circuit between node 1 and node 2 can be shown to be identical to that shown in Fig. 5.7 by using the basic network model [22]. This extends the realizable range of circuit elements. This is an important consideration in this kind of computer synthesis design when the result of optimization are converted to a physical circuit.

\[\begin{array}{c}
\text{sc} \\
\hline
\bar{1} & \theta_1 = \theta_2 & \bar{2}
\end{array}\]

\(\text{sc} \rightarrow \bar{1} & \theta_1 = \theta_2 & \bar{2} \rightarrow \text{oc}\)

\textbf{Figure 5.7:} Coupled line matching instead of matching circuit part I in Fig. 5.6.

Details of the derivation of the equations related to this equivalence are given in Appendix C. A design example using tapped-line input matching is given in Chapter 7 along with the simulation and measurement results.
5.5. Summary

In this chapter the design of the terminating matching circuits is mainly considered. Both conventional matching and tapped-line input matching have been discussed in detail. The tapped-line input and its corresponding matching circuit for wideband tuning is preferred for the narrowband filter. The design of the matching circuit was the second step in this design procedure for the tunable combline filter.
6. MICROSTRIP AND SUSPENDED SUBSTRATE STRIPLINE

6.1. Introduction

Up to now, the selection of the line structure for physically implementing the filter has not been considered. Two widely used line structures in microwave planar printed-circuits will be discussed. Wideband tunable combline filters can be realized by either of these two structures. However, suspended substrate stripline (SSS) seems more attractive for the filter implementation than microstrip line.

6.2. Planar Configuration in MICs

In order to be suitable as a circuit element in microwave integrated circuits (MICs), transmission structures have to be "planar" in configuration. This implies that the characteristics of the element can be determined by the dimensions in a single plane. The use of planar printed-circuit technology to obtain miniature and possibly low-cost filters has been available for many years [23]. By using low permittivity substrates and combline configurations, most of the side effects of planar printed-circuit such as very low Q and poor stopband rejection, etc., can be alleviated.

There are several transmission line structures which satisfy the requirements of planar microwave circuits. The most common of them are microstrip and suspended substrate stripline.
6.3. Microstrip and Stripline Filter

As shown in Fig. 6.1, microstrip is a two conductor transmission line with a thin dielectric slab inserted between them. Microstrip structure is open on the top. This open configuration makes microstrip very convenient for use in MICs where discrete lumped devices are mounted in the circuit. Also, a slight adjustment or tuning is possible after the circuit has been fabricated [3].

![Printed circuit, Dielectric board, Ground plane]

**Figure 6.1:** Cross-section of microstrip.

Various methods of microstrip analysis may be divided into three groups: the quasi-static approach, the dispersion model and the fullwave analysis. The approaches followed in these three groups increase in order of both rigor and analytical complexity. However, because of the well developed analysis, explicit analysis and synthesis equations for microstrip have been derived with good accuracy. Discontinuity analyses are also available [3].

In the past, many types of microwave filters [23, 24] have been fabricated in microstrip or stripline. However, these have normally been used in situations where the required bandwidth has not been too narrow or where the selectivity requirements have not been severe. Microstrip filters suffer from limitations in stopband loss due to quasi-surface modes, higher in-band dissipation loss, and a low range of impedance values which may be realized. Furthermore, for highly selective or narrowband devices, tem-
temperature stability is insufficient. Stripline (Fig. 6.2) suffers from this last problem as well as higher dielectric loss, the inability to fine tune, and severe size reduction at higher microwave frequencies. The design equations and discontinuity analyses have been well developed for stripline structures [4, 5].

![Figure 6.2: Cross-section of stripline.](image)

### 6.4. Suspended Substrate Stripline

Suspended substrate stripline (Fig. 6.3) provides an alternative transmission medium by which some of the limitations of stripline and microstrip can be overcome.

![Figure 6.3: Cross-section of suspended substrate stripline.](image)

Since most of the enclosure is filled with air, the effective dielectric constant is substantially lower than either microstrip or dielectric loaded stripline. By using recently developed low permittivity thin substrates this constant can be made very close to unity. Therefore, the wavelength is longer at the higher microwave frequencies, which eases the fabrication of the device due to limitations in the etching process used in fabricating...
printed circuit boards. Higher impedance lines can be realized in suspended substrate stripline than in microstrip because the width of the line is larger for a given impedance. The unloaded Q may approach that of air-dielectric stripline.

Suspended substrate stripline is found to be a reliable, high performance, temperature stable microwave transmission medium. It allows almost all common microwave lowpass, highpass, bandpass and bandstop filter configurations to be readily realized.

Nevertheless, the analysis of suspended substrate stripline is more difficult than that of microstrip or stripline. Based upon conformal transformations, Smith [6] derived a set of equations and published a computer program which was written in FORTRAN and can be applied to analyze suspended substrate stripline and broadside-coupled lines. Deriving precise explicit synthesis equations for general SSS seems impossible although this has been attempted in a recent paper [25].

Although there are a number of references in which more accurate but more complicated analysis of SSS has been reported, Smith's program was used to calculate the physical parameters of the proposed filters due to its easy implementation and good approach.

6.5. Summary

The planar configuration and two planar transmission structures have been discussed in this chapter. Compared to microstrip and especially at higher frequencies, suspended substrate stripline is a more attractive alternative to use in the physical realization of microwave narrowband filters.
7. DESIGN AND TEST RESULTS FOR THE PROPOSED FILTERS

7.1. Introduction

A narrowband filter with less than 15% fractional bandwidth is considered in the following design example. The design procedure discussed in the previous three chapters was utilized. The results of computer simulations and hardware measurements are presented and analyzed in this chapter.

7.2. Design Specification and Varactor Properties

7.2.1. Design Specification

A varactor tuned combline filter is designed and constructed to the following specifications:

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter Type</td>
<td>Chebyshev, 0.1dB passband ripple.</td>
</tr>
<tr>
<td>Degree of Filter</td>
<td>2.</td>
</tr>
<tr>
<td>Center Frequency</td>
<td>2 GHz ± 50 MHz.</td>
</tr>
<tr>
<td>Passband Bandwidth</td>
<td>200 MHz ± 15%.</td>
</tr>
<tr>
<td>Tuning Band</td>
<td>&gt; 800 MHz.</td>
</tr>
</tbody>
</table>

It should be stated that the bandwidth is specified between the equal-ripple band edges according to the traditional filter design. Passband bandwidth of 200 MHz is specified at the 0.1dB attenuation points, the 3-dB bandwidth can be calculated in terms of the mapping given in [7, 9] and is approximately 400 MHz.
7.2.2. Varactor Properties

As mentioned in Chapter 3, the tuning elements used in a combline filter are GaAs varactor diodes. ND 3050 S to K-band tuning diodes made by NEC corporation are utilized and have the following published parameters:

1. \( C = 0.45 \text{ pF}, R_s = 1.5 \text{ ohm at } -4.5 \text{ volts bias.} \ (Q = 100 \text{ at } 2 \text{ GHz operating frequency}) \)
2. \( C = 0.25 \text{ pF to } 4.5 \text{ pF over the tuning range.} \)
3. Package capacitance \( C_p = 0.08 \text{ pF} \) and series inductance \( L = 0.7 \text{ nH.} \)

As described in Section 3.4, the diode was measured using a microstrip line test circuit on a 0.79mm Teflon substrate with \( \varepsilon_r = 2.2 \) which was also used for filter implementation. The combined parasitics and mounting effects are significant. The parameters calculated from the measured data are shown in Table A.2. Using Table A.2 the following can be determined:

1. \( C = 0.7 \text{ pF at } -4.5 \text{ volts bias.} \)
2. Average of \( L = 2.1 \text{ nH, average of } R_s = 1.7 \text{ ohm.} \)
3. \( C = 0.38 \text{ pF to } 4.2 \text{ pF over -20 to 0 volts, respectively.} \)

It can be seen that the inductance is three times the published data. This is due to the combination of effects. The range of variable capacitance is less than the published data due to the package capacitance and other parasitic capacitive components. In this case, the cutoff frequency of this diode can be calculated by

\[
\frac{1}{2\pi \sqrt{LC_{\text{min}}}} = 5.6 \text{ GHz.}
\]

This is identical to the value of \( f_0 \) measured and listed in Table A.1. The diode cannot be considered to be a variable capacitance above this frequency.
The quality factor at 2 GHz and - 4.5 volts is given by

\[ Q = \frac{1}{2\pi f R_s C} \approx 67 , \]

the diode has reasonably high Q for microwave filter applications.

**7.3. Software Design**

The values shown below are referred to a 1-ohm system. For simplification, the capacitor \( C_{a1} \) in the equivalent lowpass filter inverter circuit (Fig. 4.3) was chosen to make \( J'_{01} = 1 \), so that \( Y_{ina} = 1 \) for a 1-ohm system. Also, all the capacitors \( C_{ak} \) were chosen to be the same as \( C_{a1} \).

From the specified data and [8],

\[ g_1 = 0.843, \quad g_2 = 0.622, \quad g_3 = 1.3554 . \]

As mentioned above, for a 1-ohm system

\[ J'_{01} = J'_{23} = 1 \quad \text{and} \quad G_A = G_B = 1 \]

so, from (4.11)

\[ C_{a1} = C_{a2} = 0.843 \quad \text{and} \quad J'_{12} = 1.1642 . \]

Using the design procedure described in Section 4.4.1 and running the related FORTRAN program (given in Appendix D) on a VAX780 computer, the data were calculated step by step as below:

\[ C = 35 \text{ pF}, \quad R_s = 0.034 \text{ ohm for 1-ohm system}, \]

\[ \theta_0 = 58.776^\circ, \quad n_0 = 4.5534, \quad (Y)_1 = 0.945 \text{ siemens}, \]

\[ Y_{12} = 0.256 \text{ siemens}, \quad Y_1 = Y_2 = 0.689 \text{ siemens}. \]

The bandwidth change \( \Delta f \) with tuning frequency \( f_0 \) can be calculated using the same program (from (4.16)) and is shown in Fig. 7.1. For a bandwidth change of less than 15 percent, the tuning band is from 1.6 GHz to 2.4 GHz.
Figure 7.1: Bandwidth change with tuning frequency.

For the tapped-line input, $Y_T$ can be obtained from (5.22) as

$$Y_T = 0.875 \text{ siemens}.$$ 

This completes the first step in the design process.

The parameters of the matching network had to be calculated next. To do this the formulas and the computer optimization method given in Chapter 5 were utilized. The tapped-line input circuit as discussed in Section 5.4 was used in the design. The notations shown below are the same as defined previously.

For a tapped-line input circuit, the computer optimization method was utilized by running Touchstone software [26] on an IBM/AT computer. The related programs are given as the first two programs of Appendix E. The results are given by

- $\phi = 19.5^\circ$;
- 50-Ω line: width = 2.337 mm;
- Line $Y_0$: width = 1.245 mm, length = 10.67 mm;
- Line $Y_1$: width = 1.016 mm, length = 11.91 mm;
\[ Y_0' = -0.2295 \text{ siemens}, \quad C_{PC} = 0.0086 \text{ pF}. \]

The admittance of first resonator is given by

\[ Y_1' = Y_1 + Y_0' = 0.4595 \text{ siemens}. \]

After the design of coupled resonator lines by using the third program in Appendix E, the resultant experimental filter circuit is as shown in Fig. 7.2.

![Experimental tunable combline filter](image)

**Figure 7.2:** Experimental tunable combline filter (all dimensions are in mm).

### 7.4. Computer Simulation and Experimental Results

The resultant filter circuit in the previous section was obtained by calculation and computer optimization. The filter was realized in microstrip on a 0.79mm Teflon substrate with \( \varepsilon_r = 2.2 \). As shown in Fig. 7.2, two 470-ohm chip resistors and two 82-pF chip bypass capacitors constituted the bias circuit for two ND3050-3D (NEC) tuning varactors.
To describe the filter performance, S parameters are commonly used. They are defined as

\[ S_{11} = \text{Reflection Coefficient at Port 1} , \]
\[ S_{12} = \text{Transmission Coefficient from Port 2 to Port 1} , \]
\[ S_{21} = \text{Transmission Coefficient from Port 1 to Port 2} , \]
\[ S_{22} = \text{Reflection Coefficient at Port 2} . \]

In practice, the magnitudes of \( S_{11} \) and \( S_{21} \) are mostly used and are given by

\[ |S_{11}| = \frac{\text{Voltage reflected at port 1}}{\text{Voltage available at port 1}} , \]
\[ |S_{21}| = \frac{\text{Voltage output at port 2}}{\text{Voltage input at port 1}} . \]

Since the filter is symmetrical, one can write

\[ |S_{22}| = |S_{11}| , \text{ and } |S_{12}| = |S_{21}| . \]

The parameters used in the computer simulation and hardware measurement are given by

\[ \text{Return loss (in dB)} = 20 \log |S_{11}| , \]
\[ \text{Insertion loss (in dB)} = 20 \log |S_{12}| = 20 \log |S_{21}| . \]

The filter is expected to present large return loss and small insertion loss in the passband.

Results of computer simulation using the last program in Appendix E are shown in Fig. 7.3.

The return loss is shown by the cross identifiers and the insertion loss is shown by the square identifiers. The results are plotted over a frequency range of 1 GHz to 3 GHz.
Figure 7.3: Computer simulation results of filter insertion loss $|S_{21}|$ and return loss $|S_{11}|$ (in dB) where the reverse bias was: (a) 2V; (b) 4V; (c) 20V. Scale – X axis: 200 MHz/Div; Y axis: 1.5 dB/Div.
with 200 MHz per division. The vertical axes are 1.5 dB per division. PCLF is the name of the computer program used for simulation. Over the tuning range 1.55 GHz to 2.5 GHz (corresponding to -2 and -20 volts, respectively), the return loss was larger than 8 dB and the insertion loss < 2 dB. The center frequency was 2 GHz at -4 volts. These results are close to those predicted by the design.

The measurement was performed on a HP8510B network analyzer (Fig. 7.4).

As mentioned earlier, same type of varactors was used. Same bias voltage was used for the two diodes. The network analyzer is calibrated before the filter performance is measured.

The measured results over the bias range of 1.5 volts to 20 volts are shown in Fig. 7.5. The results in Fig. 7.5 are plotted over a frequency range of 1 GHz to 3 GHz, with 200 MHz per division on the X axis. The center of each vertical axis was -11 dB. The insertion loss (curve 1) was 2 dB per division while the return loss (curve 2) was 3 dB per division.

As shown in Fig. 7.5 (a) and (b), the passband return loss was unacceptable for bias
Figure 7.5: Measured performance of experimental tunable combline filter where the reverse bias was: (a) 1.5v; (b) 2v; (c) 2.5v; (d) 3v; (e) 3.5v; (f) 4v; (g) 4.5v; (h) 5v; (i) 7.5v; (j) 10v; (k) 15v; (l) 20v.
Figure 7.5, continued
Figure 7.5, concluded
voltages less than 2.5 volts. At -2.5 volts (Fig. 7.5(c)), the return loss was > 9.8 dB, with an insertion loss < 3.12 dB. The equal-ripple bandwidth was about 215 MHz. This is within the tolerant range as shown in Section 7.2.1. At voltages more than 3 volts (Fig. 7.5(c-1)), the return loss was > 10.7 dB while the insertion loss was < 2.9 dB. The equal-ripple bandwidths had little variation with the bias voltage although the cutoff rate of the passbands decreased as the voltage increased. This effect resulted from the imperfect matching at the input and output. The passband ripple was less than 0.5 dB. The measured results showed that a tuning range of 1.68 GHz to 2.48 GHz could be obtained. The filter midband frequency was 2.05 GHz at -4.5 volts, which was 50 MHz or 2.5 percent higher than that predicted by the design. The previously published results for a similar tunable combline filter had a 7.5 percent inaccuracy in predicting this frequency [12]. These results establish the validity of the design method developed in this thesis.

As shown in Fig. 7.5, the two peaks of insertion loss curve were found slightly unequal. This was mainly due to the asymmetric soldering and mounting of the two varactor diodes. It can be corrected by using two slightly different bias voltages. In this case, an other voltage supply is required.

The measurement results were slightly worse than the simulation results, this is because the simulation program (given in Appendix E) does not include the tee junction effect of the tap position, the roughness of the conductor and the environmental effect. The effect of tee junction is discussed in [5]. The combined effect of these factors was a 1 dB higher insertion loss for the filter.

It should be noted that although the detrimental effect of varactors used to tune the filter was significant, the measured performance was in good agreement with that predicted by the design. This was achieved because the design took into account the
measured varactor parasitics. It also proved the suitability of the proposed matching circuit with tapped-line input.

7.5. Design Example and Results by Using Chip Varactors

Chip varactors can be utilized to achieve more predictable and better filter performance. Using the design method developed in this thesis a filter was designed using chip varactor diodes. The filter was a two-resonator, 0.1 dB ripple Chebyshev bandpass filter. Its bandwidth was 200 MHz at 4 GHz. The tuning range of this filter was specified as 3.2 GHz to 4.8 GHz. Alpha GaAs chip varactor with $C = 0.5$ pF and $Q = 75$ (corresponding to filter midband frequency 4 GHz) were used to tune the filter. The special $\Pi$ matching circuit was employed. The design equations used were those in Section 4.4.2. The circuit and computer simulation results are shown in Fig. 7.6 and 7.7.

![Filter circuit using Alpha chip varactors and special $\Pi$ matching circuit (all units are in siemens).](image)

The results showed that at the midband frequency, the passband return loss (identified by crosses in Fig. 7.7) was 11 dB and the insertion loss (identified by squares in Fig. 7.7) was about 1.5 dB. Tuning the filter over the full tuning range of 3.2 GHz to 4.8 GHz resulted in a minimum return loss of 8 dB and a maximum insertion loss of 1.7 dB. The change in filter bandwidth was less than 10 percent.
The inductance effect is neglected in this example. This results in much better performance. However, a practical realization will not achieve the performance seen here due to the transmission line loss, mounting effects and varactor parasitics which were considered in the filter design in the first example.

7.6. Summary

Details of the design of a microstrip tunable combline filter with a tapped-line input matching circuit have been given. The results of computer simulation and hardware measurements have been presented and show good agreement with those predicted by the new design procedure. Another example using chip varactors and special Π matching circuit has also been briefly described along with simulation results.

As stated in the previous chapter, suspended substrate stripline (SSS) is a better
transmission medium than microstrip. However, SSS manufacturing has proven to be more complicated, time-consuming and costly. Hence, in order to prove the suitability of the design theory developed in this thesis, a cheaper and faster microstrip realization for the tunable combline filter was made and the filter performance was measured. Better performance using SSS structure can be achieved using the design method described in this thesis.
8. CONCLUSIONS AND RECOMMENDATIONS

This project was concerned with the development of an improved systematic design procedure for a wideband varactor tuned combline narrowband bandpass filter. In this chapter the conclusions from the results of the research work done under this project are drawn and the recommendations are presented.

The factors which degrade the performance of varactor tuned combline bandpass filters were identified to be:

1. The frequency dependent coupling between the adjacent combline resonators.
2. The parasitic components of varactor diodes.
3. The mounting effects.
4. The transmission line loss.

These factors have been modeled and characterized. A new design procedure has been developed which consists of the following design steps:

1. Characterization of the parasitic components of the varactor diode.
2. Development of new design equations for calculating the internal filter sections.
3. Design of a new matching circuit to compensate for the detrimental effects mentioned above.

This design procedure has been verified experimentally using a 2 GHz microstrip tunable filter.

A suspended substrate stripline configuration has higher stability than a microstrip
structure with respect to temperature changes and production tolerances. Because of these reasons it is expected that the suspended substrate combline varactor tuned narrow-band filters, designed using the new design procedure, will perform better in a production environment. Also, due to the large range of realizable impedances in suspended substrate stripline, these filters can be designed to have a very narrow bandwidth and a substantially wider tuning band. Narrow bandwidth and wide tuning range enhance filter performance and are highly desirable for military and commercial systems where such filters are used.
REFERENCES


14. B. Deloach, "A New Microwave Measurement Technique to Characterize Diodes


Appendix A. The Measured Data of ND3050 Diode and The Equivalent Circuit Parameters

Table A.1: Measured data of ND3050-3D (NEC) diode.

<table>
<thead>
<tr>
<th>Bias (volts)</th>
<th>$f_0$ (GHz)</th>
<th>$l_1$ at $f_0$ (dB)</th>
<th>$l_2$ at $f_0$ (dB)</th>
<th>$f_1$ (GHz)</th>
<th>$f_2$ (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>1.674</td>
<td>1.39</td>
<td>24.58</td>
<td>1.603</td>
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<td>5.479</td>
<td>5.606</td>
</tr>
<tr>
<td>-18</td>
<td>5.566</td>
<td>2.86</td>
<td>27.68</td>
<td>5.508</td>
<td>5.622</td>
</tr>
<tr>
<td>-19</td>
<td>5.585</td>
<td>2.85</td>
<td>27.62</td>
<td>5.528</td>
<td>5.643</td>
</tr>
<tr>
<td>-20</td>
<td>5.601</td>
<td>2.87</td>
<td>27.60</td>
<td>5.546</td>
<td>5.661</td>
</tr>
</tbody>
</table>
Table A.2: The diode equivalent circuit parameters computed from measured data.

<table>
<thead>
<tr>
<th>Bias (volts)</th>
<th>$R_s$ (ohm)</th>
<th>$C$ (pF)</th>
<th>$L$ (nH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>1.8604</td>
<td>4.2379</td>
<td>2.1405</td>
</tr>
<tr>
<td>-1</td>
<td>1.9000</td>
<td>2.6898</td>
<td>2.2183</td>
</tr>
<tr>
<td>-2</td>
<td>2.0368</td>
<td>1.9024</td>
<td>2.0505</td>
</tr>
<tr>
<td>-3</td>
<td>1.8953</td>
<td>1.0878</td>
<td>2.1654</td>
</tr>
<tr>
<td>-4</td>
<td>1.8490</td>
<td>0.8000</td>
<td>2.0252</td>
</tr>
<tr>
<td>-5</td>
<td>1.7383</td>
<td>0.5883</td>
<td>2.2589</td>
</tr>
<tr>
<td>-6</td>
<td>1.7729</td>
<td>0.5495</td>
<td>2.1471</td>
</tr>
<tr>
<td>-7</td>
<td>1.7003</td>
<td>0.5097</td>
<td>2.1227</td>
</tr>
<tr>
<td>-8</td>
<td>1.6307</td>
<td>0.4648</td>
<td>2.1892</td>
</tr>
<tr>
<td>-9</td>
<td>1.6388</td>
<td>0.4738</td>
<td>2.0454</td>
</tr>
<tr>
<td>-10</td>
<td>1.6228</td>
<td>0.4223</td>
<td>2.2157</td>
</tr>
<tr>
<td>-11</td>
<td>1.6795</td>
<td>0.4522</td>
<td>2.0027</td>
</tr>
<tr>
<td>-12</td>
<td>1.5991</td>
<td>0.4389</td>
<td>2.0112</td>
</tr>
<tr>
<td>-13</td>
<td>1.5566</td>
<td>0.3983</td>
<td>2.1806</td>
</tr>
<tr>
<td>-14</td>
<td>1.4698</td>
<td>0.3680</td>
<td>2.3232</td>
</tr>
<tr>
<td>-15</td>
<td>1.6069</td>
<td>0.4142</td>
<td>2.0372</td>
</tr>
<tr>
<td>-16</td>
<td>1.6069</td>
<td>0.4286</td>
<td>1.9446</td>
</tr>
<tr>
<td>-17</td>
<td>1.6148</td>
<td>0.4060</td>
<td>2.0312</td>
</tr>
<tr>
<td>-18</td>
<td>1.5227</td>
<td>0.3835</td>
<td>2.1329</td>
</tr>
<tr>
<td>-19</td>
<td>1.5320</td>
<td>0.3817</td>
<td>2.1274</td>
</tr>
<tr>
<td>-20</td>
<td>1.5395</td>
<td>0.3774</td>
<td>2.1379</td>
</tr>
</tbody>
</table>
Appendix B. The Derivation of (4.16)

The lowpass to bandpass transformation of (4.14) is rewritten below

\[ \omega \rightarrow n_0 \cdot \frac{\omega \tan \theta}{1-\omega^2LC} (Y)_k \]  \hspace{1cm} (B.1)

Assuming \( \theta = \alpha \omega \) where \( \alpha \) is constant, and equating \( \omega = \pm 1 \) in the lowpass prototype (these correspond to the equal-ripple edges of the bandpass filter), we have

\[ \begin{align*}
-1 &= \frac{n_0}{C_{ak}} \left[ \frac{\omega_1 \tan (\alpha \omega_1)}{1-\omega_1^2LC} (Y)_k \right], \\
+1 &= \frac{n_0}{C_{ak}} \left[ \frac{\omega_2 \tan (\alpha \omega_2)}{1-\omega_2^2LC} (Y)_k \right],
\end{align*} \]  \hspace{1cm} (B.2, B.3)

where

\[ \begin{align*}
\omega_1 &= \omega_0 - \frac{\Delta \omega}{2}, \\
\omega_2 &= \omega_0 + \frac{\Delta \omega}{2},
\end{align*} \]  \hspace{1cm} (B.4, B.5)

\( \Delta \omega \) denotes the passband bandwidth which is much less than \( \omega_0 \) in narrowband case. Thus (B.2) and (B.3) can be simplified as
\[
[-1 + \frac{n_0(Y)_k}{C_{ak}}][1 - (\omega_0^2 - \omega_0\Delta\omega)L C][1 + \frac{\alpha\Delta\omega}{2}\tan \theta_0]
\]

\[
= \frac{n_0 C}{C_{ak}} \cdot (\omega_0 - \frac{\Delta\omega}{2})(\tan \theta_0 - \frac{\alpha\Delta\omega}{2}) , \quad (B.6)
\]

\[
[1 + \frac{n_0(Y)_k}{C_{ak}}][1 - (\omega_0^2 + \omega_0\Delta\omega)L C][1 - \frac{\alpha\Delta\omega}{2}\tan \theta_0]
\]

\[
= \frac{n_0 C}{C_{ak}} \cdot (\omega_0 + \frac{\Delta\omega}{2})(\tan \theta_0 + \frac{\alpha\Delta\omega}{2}) . \quad (B.7)
\]

Subtracting (B.6) from (B.7) and rearranging, we have

\[
\Delta \omega = \frac{2C_{ak}(1 - \omega_0^2 L C)}{n_0 C(\theta_0 + \tan \theta_0) - n_0(Y)_k[-\alpha\tan \theta_0(1 - \omega_0^2 L C) - 2\omega_0 L C]} \quad (B.8)
\]

From (4.15) we have

\[
1 - \omega_0^2 L C = \frac{\omega_0 C\tan \theta_0}{(Y)_k} \quad (B.9)
\]

Putting (B.9) into (B.8), we obtain

\[
\Delta \omega = \frac{2C_{ak}\omega_0\tan \theta_0}{n_0(Y)_k[\theta_0 + \tan \theta_0 + \theta_0\tan^2 \theta_0 + 2(Y)_k\omega_0 L]} . \quad (B.10)
\]

This is the same as (4.16).
Appendix C. The Equivalence between the Fig. 5.7 and The Matching Circuit Part I of Fig. 5.6

As shown in Fig. C.1, the top left circuit is identical to Fig. 5.7. After using the network model and doing simplifications [22], we obtain the equivalent circuit shown in the bottom left figure which is the same as the matching circuit part I of Fig. 5.6. The square boxes $Z_0$ and $Z_{01}$ shown in Fig. C.1 represent two ideal transmission lines with characteristic impedances $Z_0$ and $Z_{01}$, respectively. Both the electric lengths of these lines are $\theta_1$.

Figure C.1: Application of the network model for obtaining the equivalent circuit of Fig. 5.7.
From [22] we can write the coupled line capacitances as

\[ C_a = \eta \left( \frac{1}{Z_{01}} \left( \frac{n-1}{Z_0} \right) \right), \]  
\[ C_{ab} = \eta \left( \frac{n}{Z_0} \right), \]  
\[ C_b = \eta \left( \frac{n(n-1)}{Z_0} \right), \]

where

\[ \eta = 377 \text{ ohm}, \]  
\[ \varepsilon_r = \text{Dielectric constant}, \]  
\[ \varepsilon = \varepsilon_r \varepsilon_0, \varepsilon_0 = \text{Permittivity of free space}, \]  
\[ C_a = \text{Capacitance of line a to ground}, \]  
\[ C_b = \text{Capacitance of line b to ground}, \]  
\[ C_{ab} = \text{Coupling capacitance between line a and b}, \]  
\[ n = 1 + \frac{C_b}{C_{ab}}, \text{reasonably arbitrary constant}. \]

The next step would be to relate these capacitances to the line dimensions. From the even- and odd-mode point of view, we can write

\[ C_{(a,b)} = 2C_p(a,b) + 2C_f + 2C_{fe}, \]  
\[ C_{ab} = C_{fo} - C_{fe}, \]

where \( 2C_p \) is the equivalent parallel-plate capacitance, \( 2C_f \) and \( 2C_{fe} \) are the fringing-field capacitances of the uncoupled and coupled edges, respectively.

The solution of (C.1) to (C.4) would become unique if we use equal width coupled lines. In this case we have
$C_a = C_b$, or

$$n^2 = 1 + \frac{Z_0}{Z_{01}}.$$  \hfill (C.7) \hfill (C.8)

This is often the case in practice.
Appendix D. The FORTRAN Program Used for Filter Design

This program is used to calculate the parameters, $\theta_0$, $(Y)_k$, $Y_{k+1}$, and $Y_k$ in the filter design. The program exactly follows the steps of design procedure described in Section 4.4.1 and can be run in any computer supporting FORTRAN language.

```fortran
REAL*4 AO, A1, A2, A3, A4, A5, VEO, VOK
REAL*4 RV, CV, THETA0, FMID, YYK, XITAO, QV, LV
REAL*4 NO, BANDW, CAK, JM(10), YM(10), Y(10), Y0, Y01, Y1P
COMMON RV, CV, FMID, BANDW, YYK, YY1, LV
DATA PI/3.14159/, VOK/1000.0/

C DESIGN THETA0 AND (Y)_k INCLUDING VARACTOR EFFECT
C RV, CV at -4.5 volts and LV calculated from measured
C data. RV(ohm), CV(pF), LV(nH), f(GHz)

C REAL*4 A0, A1, A2, A3, A4, A5, VEO, VOK
C REAL*4 RV, CV, THETA0, FMID, YYK, XITAO, QV, LV
C REAL*4 NO, BANDW, CAK, JM(10), YM(10), Y(10), Y0, Y01, Y1P
C COMMON RV, CV, FMID, BANDW, YYK, YY1, LV
C DATA PI/3.14159/, VOK/1000.0/

n=2 Chebyshev with 0.1dB ripple and 10% bandwidth
   g1 = 0.843, g2 = 0.622, g3 = 1.3554

DATA RV/0.034/, CV/35.0/, LV/0.042/, FMID/2.0/, N/1/
DATA CAK/0.843/, JM/1.164176, 9*0.0/, BANDW/200.0/
IF (RV.GT.1.E-4) QV=1.0/(2.*PI*FMID*RV*CV*1.E-3)
OPEN (2, STATUS='NEW', FILE='DES2A.DAT')
OPEN (1, STATUS='NEW', FILE='F7-1.DAT')
IF (RV.LE.1.E-4) WRITE (2, 909)
IF (RV.GT.1.E-4) WRITE (2, 901) QV, FMID
AN=N
WRITE (2, 910) RV, CV, FMID, BANDW, AN, CAK, (JM(I), I=1, N)
A0=2.0*PI*FMID
DO 100 I=1, N
   THETA0=20.0+(I-1)*0.001
   A1=THETA0*PI/180.
   A2=TAN(A1)
   A3=1+A2*A2
   A4=CV*AO*1.0E-3*A2/(1.0-LV*CV*AO*A0*1.0E-3)
   VEO=ABS(2.0*LV*AO*A4-A5)
```
IF(VE0.GE.VOK) GOTO 100
VOK=VE0
XITA0=THETA0
YYK=A4
100 CONTINUE
WRITE(2,902) XITA0,YYK,VOK

C C C
C Design the other parameters NO, Y(k,k+1) and Y(k)
C
A1=XITA0*PI/180.
A2=TAN(A1)
A4=2.0*CAK*FMID*1000.0*A2
A5=BANDW*YYK*(A1+A2+A1*A2*A2+2.0*YYK*LV*A0)
NO=A4/A5
DO 120 I=1,N
120 YM(I)=JM(I)/NO
DO 140 I=1,2*N-1
IF(I.EQ.1) Y(I)=YYK-YM(I)
IF(I.NE.1 .AND. I.NE.(2*N-1)) Y(I)=YYK-YM(I-1)-YM(I)
140 IF(I.EQ.(2*N-1) .AND. I.NE.1)
YY1=Y(I)
WRITE(2,903) NO, (YM(I),I=1,N), (Y(I),I=1,2*N-1)
DO 200 I=1,2*N-1
IF(I.LE.N) YM(I)=YM(I)/50.0
200 Y(I)=Y(I)/50.0
WRITE(2,907)
WRITE(2,904) (YM(I),I=1,N)
WRITE(2,904) (Y(I),I=1,2*N-1)
WRITE (2, 905)

C C C C
C Generate the data for the figure of bandwidth
C vs tuning frequency from (4.16)
C
FSTART=FMID-1.
DO 300 J=1,21
FREQ=FSTART+(J-1)*2.0/20.
A1=(XITA0*PI/180.)*FREQ/FMID
A2=TAN(A1)
A4=2.0*CAK*FREQ*1000.0*A2
A5=NO*YYK*(A1+A2+A1*A2*A2+4.*YYK*LV*3.14159*FREQ)
DELW=A4/A5
WRITE(1,906) FREQ, DELW
CONTINUE
300 FORMAT(60('"')/8X,'VARACTOR Q =',F5.1,' AT ',F3.1,
# ' GHZ'/60('"')/' RV(ohm),CV(pf)'
# ',FMID(GHz),BANDW(MHz),N,CAK,JM (1-ohm system)'
901 FORMAT(60('"')/" THETA0,(Y)K,ERR = ",
# 3(1X,F10.6)/60('"'))
902 FORMAT(' NO,Y(k,k+1),Y(k) = ",
903 FORMAT(60('"'))
The results of 'DES2A.DAT' are shown as

<table>
<thead>
<tr>
<th>RV(ohm), CV(pf), FMID(GHz), BANDW(MHz), N, CAK, JM (1-ohm system)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.034000 35.000000 2.000000 200.000000</td>
</tr>
<tr>
<td>1.000000 0.843000 1.164176</td>
</tr>
</tbody>
</table>

| THETA0, (Y)K, ERR = 58.776001 0.944885 0.000068 |

| N0, Y(k, k+1), Y(k) = 4.553384 0.255673 0.689212 |

<table>
<thead>
<tr>
<th>MUTUAL ADMITTANCES AND LINE ADMITTANCES (50-ohm system)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.005113</td>
</tr>
<tr>
<td>0.013784</td>
</tr>
</tbody>
</table>

The results of 'F7-1.DAT' are shown as

1.0000, 127.0217
1.1000, 138.3612
1.2000, 149.1930
1.3000, 159.4109
1.4000, 168.8896
These data have been used for Fig. 7.1.
Appendix E. The Programs Used for Matching Circuit Design and Analysis

Touchstone user manual [26] gives complete information regarding the software. The dimension (DIM) block is used to assign the dimensions for the various electrical and physical parameters. The variable (VAR) block lists the variables in the circuit file which are varied to meet the objectives of the design. Equation (EQN) block defines the formulas. Circuit (CKT) block describes the circuit. Frequency (FREQ) block is used to choose the range of frequencies for which the analysis or design has to be done. Optimization (OPT) block specifies the design goals. A gradient algorithm is used for minimizing the error function.

E.1. Synthesis Programs for Design of Matching Circuit with Tapped-Line Input

The first program is used to design the matching circuit with tapped-line input to satisfy (5.3). Then the second one can be used to achieve the goal defined by (5.4).

! SYNTHESIS PROGRAM I.
! MATCHING CIRCUIT I (FIG. 5.6) TO BE DESIGNED.
! ND3050-3D (NEC) VARACTOR DIODES USED.
DIM
FREQ GHZ
ANG DEG
RES OH
IND NH
CAP NF
LNG MIL
! 1 mil = 0.001 inches.

VAR
N00 = 4.5534
THETA0 = 58.776
F0 = 2
YYK = 0.0189
YT = 0.0175
YA = 0.02
W50 = 92.464 ! width of 50-ohm line.
RVT = 1.7 ! varactor resistance.
LV = 2.1 ! varactor inductance.
FIO # 10 25 40
WOC # 20 50 80
LEN1 # 100 400 600
WLINE # 20 50 80
LEN2 # 100 400 600

EQN
OMIGA0=2.*3.14159*F0
THETA=FREQ*THETA0/F0
NN=(N00*(TAN(THETA)))**0.5 ! (4.9)
FI=FREQ*FIO/F0
ZA=1./YA
ZT=1./YT
DELTA=SIN(THETA)/SIN(FI) ! (5.21)
BETA=ZT*SIN(THETA-FI) ! (5.20)
ND=1./DELTA
L1=BETA*DELTA/(2.*3.14159*FREQ)
X1=TAN(THETA)
OMIGA=2.*3.14159*FREQ
CVT=YYK/(OMIGA*X1+LV*YYK*(OMIGA**2)) ! (5.5)

CKT
MSUB ER=2.205 H=31 T=2 RHO=0 RGH=0
RES 9 0 R^ZA
MLIN 9 8 W^W50 L=200
MTEE 8 7 6 W1^W50 W2^WLINE W3^WOC
MLEF 6 W^WOC L^LEN1
MLIN 7 1 W^WLINE L^LEN2
XFER 1 2 0 0 N^ND
IND 2 3 L^L1
DEF1P 3 YIN1
YIN1 3
RES 3 8 R^RVT
CAP 8 9 C^CVT
IND 9 0 L^LV
XFER 3 4 0 0 N^NN
DEF1P 4 YIN2

OUT
YIN2 RE[Y1] GR1
YIN2 IM[Y1] GR2
FREQ
   SWEEP 1.0 3.0 0.05

GRID
   RANGE 1.0 3.0 0.05
   GR1 0.01 0.03 0.002
   GR2 0.5 -0.5 0.1

OPT
   RANGE 1.5 2.5
   YIN2 RE[Y1]=0.02 3000
   RANGE 1.0 1.5
   YIN2 RE[Y1]=0.02 1000
   RANGE 2.5 3.0
   YIN2 RE[Y1]=0.02 1000

Optimization results:

   F10   = 19.5
   WOC   = 49
   LEN1  = 420
   WLINE = 40
   LEN2  = 469

! SYNTHESIS PROGRAM II.
! MATCHING CIRCUIT II (FIG. 5.6) TO BE DESIGNED.
! ND3050-3D (NEC) VARACTOR DIODES USED.

DIM
   FREQ GHZ
   ANG DEG
   RES OH
   IND NH
   CAP PF
   LNG MIL

VAR
   NO0  = 4.5534
   THETA0 = 58.776
   F0    = 2
   YYK   = 0.0189
   YT    = 0.0175
   YA    = 0.02
   W50   = 92.464
   RVT   = 1.7
   LV    = 2.1
FIO = 19.5
WOC = 49
LEN1 = 420
WLINE = 40
LEN2 = 469
YOP  # -0.5 0.1 0.5
CPC1 # 0.0 0.004 0.025

EQN
OMIGA0=2.*3.14159*F0
THETA=FREQ*THETA0/F0
NN=(N00*(TAN(THETA)))**0.5
FI=FREQ*FIO/F0
ZA=1./YA
ZT=1./YT
DELTA=SIN(THETA)/SIN(FI)
BETA=ZT*SIN(THETA-FI)
ND=1./DELTA
L1=BETA*DELTA/(2.*3.14159*FREQ)
X1=TAN(THETA)
OMIGA=2.*3.14159*FREQ
CVT=YYK/(OMIGA*X1+LV*YYK*(OMIGA**2))
ZOP=1./YOP
CPC=CPC1/50.
XC2=OMIGA**2*CVT
REQV=(1.-XC2*LV)**2/(XC2*CVT*RV)

CKT
MSUB ER=2.205 H=31 T=2 RHO=0 RGB=0
RES 9 0 R^ZA
MLIN 9 8 W^W50 L=200
MTEE 8 7 6 W1^W50 W2^WLINE W3^WOC
MLEF 6 W^WOC L^LEN1
MLIN 7 1 W^WLINE L^LEN2
XFER 1 2 0 0 N^ND
IND 2 3 L^L1
DEF1P 3 YIN1
YIN1 3
TLSC 3 0 Z^ZOP E^THETA0 F^F0
CAP 3 0 C^CPC
RES 3 8 R^REQV
XFER 3 4 0 0 N^NN
DEF1P 4 YIN2

OUT
YIN2 RE[Y1] GR1
YIN2 IM[Y1] GR2

FREQ
SWEEP 1.0 3.0 0.05

GRID
  RANGE 1.0 3.0 0.05
  GR1 0.01 0.03 0.002
  GR2 0.5 -0.5 0.1

OPT
  RANGE 1.5 2.5
  YIN2 IM[Y1]=0.0 3000
  RANGE 1.0 1.5
  YIN2 IM[Y1]=0.0 1000
  RANGE 2.5 3.0
  YIN2 IM[Y1]=0.0 1000

Optimization results:
  YOP = -0.2295
  CPC1 = 0.0086

E.2. Synthesis Program for Design of Coupled Resonators

This program is used to design the coupled-line based on the results from the last program.

! SYNTHESIS PROGRAM III.
! COUPLED RESONATORS TO BE DESIGNED.
DIM
  FREQ GHZ
  ANG DEG
  RES OH
  IND NH
  CAP PF
  LNG MIL

VAR
  NO0 = 4.5534
  THETA0 = 58.776
  F0 = 2
  Y1 = 0.00919
  Y12 = 0.00512
  RVT = 1.7
  LV = 2.1
  CVT = 0.7
  W50 = 92.464  ! varactor capacitance.
WCOU # 15 30 80
GAP  # 20 40 80
LEN3  # 250 600 800

EQN
   Z1=1./Y1
   Z12=1./Y12

CKT
   CAP 1 2 C^CVT
   RES 2 3 R^RVT
   IND 3 0 L^LV
   DEF1P 1 VARA
   TLSC 3 0 Z^Z1 E^THETA0 F^F0
   VARA 3
   DEF1P 3 HALF
   HALF 1
   TLSC 1 2 Z^Z12 E^THETA0 F^F0
   HALF 2
   DEF2P 1 2 TX_F

MSUB ER=2.205 H=31 T=2 RHO=0 RGH=0 VARA 1
   MCLIN 1 2 0 0 W^WCOU S^GAP L^LEN3 VARA 2
   DEF2P 1 2 MC_F

OUT
   TX_F MAG[S21] GR1
   MC_F MAG[S21] GR1

FREQ
   SWEEP 1.0 3.0 0.05

OPT
   MC_F MODEL TX_F

Optimization results:
   WCOU = 24
   GAP  = 41
   LEN3 = 585
E.3. Simulation Program for Filter Performance

This program is used to calculate the insertion loss $S_{21}$ and the return loss $S_{11}$ of the filter designed by the three programs above. It gives the computer simulation results of the proposed microstrip combline tunable filter over the tuning band.

! SIMULATION PROGRAM.
! ND3050-3D (NEC) VARACTOR DIODES USED.
DIM
  FREQ GHZ
  RES OH
  IND NH
  CAP PF
  LNG MIL
VAR
  W50 = 92.464
  WOC = 49
  LEN1 = 420
  WLINE = 40
  LEN2 = 469
  WCOU = 24
  GAP = 41
  LEN3 = 585
  LTAP = 204
  CJ = 0.7
  RS = 1.7
  LV = 2.1
EQN
  L1B=LEN3-LTAP
CKT
  MSUB ER=2.205 H=31 T=2 RHO=0 RGH=0
  MLIN 1 2 W^W50 L=200
  MTEE 2 3 4 W1^W50 W2^WLINE W3^WOC
  MLEF 4 W^WOC L^LEN1
  MLIN 3 6 W^WLINE L^LEN2
  DEF2P 1 6 MAT1
  MAT1 1 2
  MCLIN 2 3 0 0 W^WCOU S^GAP L^LTAP
  MCLIN 2 3 4 5 W^WCOU S^GAP L^L1B
  MAT1 6 3
  DEF4P 1 5 4 6 LOUT_T
  IND 1 2 L^LV
The results have been shown in Fig. 7.5.