Analysis And Design Of Miniaturized Rf Saw Duplexer Package

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ANALYSIS AND DESIGN OF MINIATURIZED RF SAW DUPLEXER PACKAGE

by

HAO DONG
Eng.D. Zhejiang University, 1997
M.S. Northwestern Polytechnical University, 1994
B.S. Northwestern Polytechnical University, 1991

A dissertation submitted in partial fulfillment of the requirements
for the degree of Doctor of Philosophy
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Major Professor: Thomas X. Wu
ABSTRACT

This dissertation provides a comprehensive methodology for accurate analysis and design of miniaturized radio frequency (RF) surface acoustic wave (SAW) duplexer package. Full-wave analysis based on the three dimensional (3-D) finite element method (FEM) is successfully applied to model the package. The die model is obtained by combining the acoustics and die busbars parasitics models. The acoustics model is obtained using the coupling-of-models (COM) technique. The die busbars, bonding wires and printed circuit board (PCB) are modeled using full-wave analysis. After that, the models of package, die, and bonding wires are assembled together to get the total response. To take into account the mutual couplings, the methodology is extended to model the package, die busbars, and bonding wires together. The advantages and disadvantages of the methodology are also discussed.

Based on the methodology, the Korea personal communication system (KPCS) duplexer is analyzed and designed. The isolation of KPCS duplexer package is significantly improved by redesigning inner ground plane, bonding wire scheme and ground via. A KPCS duplexer package is designed and excellent transmitter to receiver isolation in the transmission band is achieved. Simulation and measurement results are compared, and excellent agreement is found. Although we focus on investigating the methods to improve the isolation, the passband performance is also improved.

The methodology is also successfully used for flip chip duplexer. The simulation results from our assembling method match the measurement results very well. Optimization method is applied to improve the transmit band isolation. With the new package and die design, the transmit band isolation can be improved from -53.6 dB to -65.2 dB. Based on the new package,
the effect of the Rx ground trace on the isolation is investigated and the transmit band isolation can achieve -67.3 dB with the modification of the Rx ground trace.

The technique developed in this dissertation reduces the design cycle time greatly and can be applied to various RF SAW device packages.
ACKNOWLEDGMENTS

I express my sincerest gratitude to my parents for their support and encouragement through my higher education. I am thankful to my wife Xiaomin and my son Matthew for their love and support. Without their contributions, this opportunity would not have been possible.

I would like to thank my advisor, Dr. Thomas X. Wu, for his inspiration and guidance over the past four years. Dr. Wu has had a profound effect on me as a researcher and as a person. I am also thankful to my committee members for their support and assistance. Thanks to my friends and colleagues in High Speed Electronic Systems Group at University of Central Florida for their help.

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<th>Acronym</th>
<th>Description</th>
</tr>
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<tbody>
<tr>
<td>ADS</td>
<td>Advanced Design System</td>
</tr>
<tr>
<td>AMPS</td>
<td>Advanced Mobile Phone Service</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code division multiple access</td>
</tr>
<tr>
<td>COM</td>
<td>Coupling of modes</td>
</tr>
<tr>
<td>FEM</td>
<td>Finite element method</td>
</tr>
<tr>
<td>HFSS</td>
<td>High Frequency Structure Simulator</td>
</tr>
<tr>
<td>IDT</td>
<td>Interdigital transducer</td>
</tr>
<tr>
<td>KPCS</td>
<td>Korea personal communication system</td>
</tr>
<tr>
<td>PCB</td>
<td>Printed circuit board</td>
</tr>
<tr>
<td>RF</td>
<td>Radio frequency</td>
</tr>
<tr>
<td>Rx</td>
<td>Receiver channel</td>
</tr>
<tr>
<td>SAW</td>
<td>Surface acoustic wave</td>
</tr>
<tr>
<td>SMD</td>
<td>Surface mounted device</td>
</tr>
<tr>
<td>Tx</td>
<td>Transmitter channel</td>
</tr>
</tbody>
</table>
CHAPTER ONE: INTRODUCTION

This chapter introduces surface acoustic wave (SAW) technology and the concept of SAW duplexer. The challenges of SAW duplexer package modeling are discussed. At the end of this chapter, the organization of the dissertation is presented.

1.1 Surface Acoustic Wave Technology

Since electronic signal processing by means of the selective manipulation of surface acoustic waves (SAW) on piezoelectric substrates was initiated in 1965 with the invention of the thin-film interdigital transducer (IDT) by White and Voltmer at the University of California, at Berkeley [1], SAW devices have a lot of applications such as cellular phones, pagers, ID tags, cellular base-stations, local area networks, etc. They have realized a rapid growth largely due to the increased use of personal communication devices such as pagers and cellular phones. A good historical review of the development of SAW devices can be found in Morgan’s paper [2].

A SAW device consists of a piezoelectric substrate with metallic structures, such as interdigital transducer, and reflection or coupling gratings deposited on its plain-polished surface. Triggered by the piezoelectric effect, a microwave input signal at the transmitting IDT stimulates a micro-acoustic wave that propagates along the surface of the elastic solid, the amplitude of which decays exponentially with substrate depth [3]. Figure 1 shows the basic SAW delay line fabricated on piezoelectric substrate with metal thin-film input/output interdigital transducers. An ac voltage is launched at the input transducer and manipulates a surface acoustic wave. The receiving transducer detects the incident surface acoustic wave and
converts it back to a suitably filtered electrical one. Finger period at center frequency $f_0$ corresponds to acoustic wavelength $\lambda_0 = v/f_0$, where $v$ is SAW velocity [4].

![SAW delay line diagram](image)

**Figure 1:** Basic SAW delay line fabricated on piezoelectric substrate with metal thin-film input/output interdigital transducers.

SAW technology has a fundamental signal processing advantage because the velocity of the acoustic wave is approximately five orders of magnitude slower than an electromagnetic wave. Because of this, SAW devices can be much smaller than their purely electrical counterparts. This advantage is especially useful in the design of high-performance bandpass filters. The size of typical SAW RF filters shrunk significantly over the last couple of years, and, for example, a footprint of 1 mm$^2$ for an RF filter is feasible in the near future using new packaging techniques. For SAW IF filters (e.g., for IS-95 terminals), the footprint has been reduced by more than 80% over the last four years [5].
1.2 SAW Duplexer

A typical analog or digital mobile phone system consists of a super-heterodyne radio, which communicates between the phone and the base station at two different frequencies. The signal is transmitted from the phone to the base station within the transmitter channel (Tx) with the center frequency $f_T$ and from a base station to the phone within the receiver channel (Rx) with the center frequency $f_R$. A full duplex radio system such as code division multiple access (CDMA) IS95 or IS98 requires a device that separates the transmitting signals from the receiving signals and permits transmission and reception simultaneously. The enabling device is known as a duplexer [6]. The requirements of a general duplexer circuit are listed in Table 1 [7]-[10].

Table 1
General circuit requirement of duplexer

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Transmit Pass-band</th>
<th>Receive Pass-band</th>
</tr>
</thead>
<tbody>
<tr>
<td>Impedance</td>
<td>$Z_{Transmit} = Z_0 &lt;&lt; Z_{Receive}$</td>
<td>$Z_{Receive} = Z_0 &lt;&lt; Z_{Transmit}$</td>
</tr>
<tr>
<td>Reflection Coefficient</td>
<td>Large $\Gamma_{Receive}$</td>
<td>Large $\Gamma_{Transmit}$</td>
</tr>
</tbody>
</table>

The SAW duplexers are implemented using ladder filters. This technique must be used because it handles power better than other types of RF SAW filters (more than 1W) [11]-[13]. Ladder filters also provide a high degree of attenuation at lower and upper side frequency range of passband. Compared with other conventional SAW filter approaches, ladder type filters can reduce the size of front-end modules because they often require little or no matching circuitry to interface with system components.
The duplexer is a three-port device, i.e., the receiver output, the transmitter input and the third port connected to the antenna. SAW duplexers are typically implemented by cascading one-port resonators in a ladder configuration. Both Rx and Tx filters are placed on the same chip with each ladder filter containing either three or four shunt resonators as shown in Figure 2 [14].

For Advanced Mobile Phone Service (AMPS) systems, the transmitter band is lower in frequency compared to the receiver band. The transmitter filter is designed by using a T-type ladder network while the receive filter is implemented by using a Pi-Ladder network [7], [15]. The input and output impedance of transmit and receive filters are designed to meet the generalized requirements of duplexer circuit as shown in Table 1. The matching circuitry which uses the meander delay line is designed for the Rx filter to provide a 90° phase shift.

![Figure 2: Schematic of SAW duplexer.](image)
1.3 Challenges of SAW Duplexer Package Modeling

In the past, the ceramic resonators dominated the duplexer market, but recent advances in the surface acoustic wave technology have enabled a smaller and lighter duplexer for the mobile phone market. Because the main function of the duplexer is to isolate the transmitter and receiver channels, smaller physical size of the package makes it very difficult to implement isolation due to the electrical coupling between the transmitter and the receiver portion of the duplexer.

Most critical electrical requirements for a duplexer are very low loss in the transmitter and receiver filter paths with rejection approaching 60 dB of each filter. Transmitter to receiver isolation in the transmission band is one of the most critical requirements. Because the duplexer is a front-end component in a full duplex radio, the insertion loss for the receiver band sets the noise figure; furthermore, it degrades the sensitivity of the receiver. However, the insertion loss of the transmitter band has direct impact on the battery life of the radio. Poor isolation of the transmitting signal causes the leakage of signal in the receiver chain of the radio, hence posing a jamming threat to the receiver portion of the radio.

The packaging requirements of SAW duplexers are unique compared to the other electronic components. Because a wave travels on the surface of the substrate, any physical contact can severely degrade the electrical performance of a SAW device. Typically, hermetically sealed packages are used to package SAW devices. Ceramic packages with appropriate cavities are mostly used due to small size, frequency stability, and easier handling in assembly [16]. Duplexer package, however, consists of a phase-matching network making it more challenging for electrical characterization.
Effects of package parasitics on the performance of SAW filters have been studied by many researchers. Several papers [6], [17]-[20] describe the equivalent circuit with parasitics. The finite difference and current simulation methods are used for calculating electrical parasitic parameters of SAW packages in [21]. These methods use measurements to fit the equivalent circuit model by minimizing the disagreement of measurements and simulations. The problems of these models are the necessity of measurements and the large effort to determine the relevant parameters. However, the range of validity of such models is limited. Recently, electromagnetic simulation tools are successful used in package modeling [22]-[29]. Yang et al. [22] firstly discussed the equivalent model and full-wave analysis for the SAW package in detail. They found the full-wave analysis can accurately predict the performance of the package at microwave frequencies. In the same year, Finch et al. [23] described several new methods for package simulation using full-wave electromagnetic simulator. The full-wave simulation was first performed on an empty package with inductors used to represent bonding wires. More accuracy can be achieved if the bonding wires are included in the full-wave analysis. Their results demonstrated that full-wave analysis techniques are useful for predicting critical parameters such as the shape and rejection level of a SAW filter. Based on their research, Dong et al. [24] firstly discussed the comprehensive methodology for analysis and design of the RF SAW duplexer. Full-wave simulation was successfully used for the characterization of the RF SAW duplexer package. Using this design method, the isolation of Korea personal communication system (KPCS) duplexer was improved by redesigning package and bonding wire scheme [25]. The detail of the methodology can be found in this dissertation. For the other applications without critical modeling accuracy, 2.5-D EM simulator is also useful. Perois et al. [26] used 2.5-D EM simulator and obtained a good result. Pitschi et al. [27] combined accurate acoustic simulation
tools together with 2.5/3-D EM simulation software packages to predict and optimize the performance of SAW filters and SAW-based front-end modules. They found that depending on the application both 2.5-D and 3-D EM simulations yield good results if used correctly. While 2.5-D and 3-D simulations showed almost identical results with simple configurations, deviations can be observed with complex configurations. For the complex configurations, the simulation using the 3-D EM description of the package agreed better with measurement than the one using 2.5-D description of the package. Lin et al. [28] used full-wave simulator HFSS to investigate the crosstalk effects on SAW substrate and mutual coupling effects between pattern on the SAW substrate and package.

Based on our previous research results, this dissertation gives the detail of comprehensive methodology for accurate analysis and design of miniaturized RF SAW duplexer package. The methodology is extended to take into account the mutual couplings among the package, die busbars, and bonding wires by modeling them together. The improvements of the isolation for KPCS duplexer and flip chip duplexer are also discussed.

1.4 Organization of the Dissertation

The dissertation is organized as follows.

In Chapter 2, the comprehensive methodology for accurate analysis and design of miniaturized RF SAW duplexer package is provided. The full-wave analysis tool, Ansoft HFSS which is based on the finite element method, is used to simulate the duplexer package. The package model simplifications are discussed. The die model is obtained by combining the acoustics and die busbars parasitics models. The acoustics model is obtained using the coupling-
of-models (COM) technique. The die busbars, bonding wires and printed circuit board (PCB) are modeled using full-wave analysis. After that, the models of package, die, and bonding wires are assembled together to get the total frequency response. To take into account the mutual couplings, the methodology is extended to model the package, die, and bonding wires together.

Chapter 3 describes the analysis and design of KPCS duplexer package. The package model is generated from Ansoft HFSS. The die model is obtained by combining the acoustics and die busbars parasitics models. The bonding wires are modeled as ideal inductors with the values extracted from the Ansoft HFSS model. Based on the above methodology, investigation on the improvement of the isolation for the KPCS duplexer package is described. Several novel ideas are proposed to significantly improve the isolation. Simulation and measurement results are compared and excellent agreement is found.

The methodology is also successfully used for flip chip duplexer in Chapter 4. The flip chip technology is briefly introduced at the beginning. Ansoft HFSS is applied to obtain the package model. Momentum is used to get the die busbars parasitics model. The simulation results from our assembling method are compared with the measurement results. To improve the transmit band isolation, the optimization method is applied to select the external inductor values. The topology of the package is investigated to improve the isolation. The effect of the Rx ground trace on the isolation is also discussed.

Finally, the work is summarized in Chapter 5.
In this chapter, a comprehensive methodology for accurate analysis and design of RF SAW duplexer is discussed. The full-wave analysis tool, Ansoft high frequency structure simulator (HFSS) which is based on the finite element method (FEM), is used to simulate the duplexer package. The package model simplifications are discussed. The die model is obtained by combining the acoustics and die busbars parasitics models. The acoustics model is generated using coupling-of-models (COM) technique. The die busbars, bonding wires and printed circuit board (PCB) are modeled using full-wave analysis. After that, the assembling method is presented to obtain the total frequency response. To take into account the mutual couplings, the methodology is extended to model the package, die, and bonding wires together.

2.1 Full-Wave Analysis of Duplexer Package

For duplexers, a key specification is the rejection that has to reach 60 dB level. To solve this challenging problem, we use the full-wave analysis tool, Ansoft high frequency structure simulator (HFSS) which is based on the finite element method (FEM), to simulate the duplexer package.

The model of the duplexer package is created using the 3-D modeler tool, which can be created in AutoCAD and imported into HFSS using the Standard ACIS Text (SAT) file format. Three main simplifications are made for the package model.

The first simplification is for the embedded trace. The shape of embedded trace cannot be made perfectly in the fabrication process. Figure 3 shows the real shape of the cross section of
the embedded trace. This irregular shape increases the complexity of the modeling. To simply model, the rectangular cross section of the trace is used as shown in Figure 4. In the model, the embedded trace has the width of 101.6 μm and height of 20.32 μm.

To find out how much deviation if we make such simplification, we consider a 5×5 mm² duplexer package with embedded meander delay line. In the simulation setup, we extend the package ground pad on the bottom with the material defined as perfect conductor to reduce the loss of the extended metal pad. Two lumped gap source ports are defined on the external signal pads to get the frequency response of the meander delay line. Figure 5 shows the 3-D structure of this package and port definition. After obtaining the simulation result, it is compared with the measurement results as shown in Figure 6. The solid line shows the measurement results. The dashed line shows the simulation results. From Figure 6, excellent agreement is found between the simulation and measurement results. Therefore this simplification is acceptable.

The second simplification is for the castellations. The real package has curved castellations as shown in Figure 7. The curved castellation causes high mesh density and reduces the simulation speed. Therefore, the rectangular castellations are used in the simulation model to reduce the computation burden as shown in Figure 8. To investigate the effect of castellation, we consider the simplified models as shown in Figure 9. In Figure 9 (a), the ground pad is connected to the package ground through the curved castellation. In Figure 9 (b), the ground pad is connected to the package ground through the rectangular castellation. Based on the same simulation setup, the curved castellation has higher mesh density as shown in Figure 10 and costs more simulation time as we expected. The detailed information is listed in Table 2.
Figure 3: Cross section of the real embedded trace.

metal  dielectric material
Figure 4: Embedded trace with rectangular cross section in our model.
(a) A $5 \times 5$ mm$^2$ duplexer package with embedded meander delay line.

(b) Lumped gap source port with the calibration line and impedance line.

Figure 5: A $5 \times 5$ mm$^2$ duplexer package with embedded meander delay line.
(a) Magnitude of $S_{21}$.

(b) Phase of $S_{21}$.

Figure 6: Comparison of the measurement and simulation results for the meander delay line.
Figure 7: Package model with curved castellations.
Figure 8: Package model with simplified rectangular castellations.
Figure 9: Simplified models for the investigation on the effect of castellation.

(a) Ground pad connected to package ground through curved castellation.

(b) Ground pad connected to package ground through rectangular castellation.
(a) Mesh plot of curved castellation.

(b) Mesh plot of rectangular castellation.

Figure 10: Mesh plots of curved and rectangular castellations.
Table 2
Comparison of curved and rectangular castellations

<table>
<thead>
<tr>
<th></th>
<th># of Tetrahedra</th>
<th>Max. Mag. Delta</th>
<th>Simulation Time*</th>
<th>$S_{11}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Curved castellation</td>
<td>9831</td>
<td>0.00069</td>
<td>6'47&quot;</td>
<td>0.9969/165.98</td>
</tr>
<tr>
<td>Rectangular castellation</td>
<td>6276</td>
<td>0.00083</td>
<td>2'34 &quot;</td>
<td>0.9967/165.98</td>
</tr>
</tbody>
</table>

*Simulation time is calculated on the PC Pentium(R) 4 CPU 3 GHz with 1 GB of RAM

From Table 2, it can be found that the difference of the frequency response between the single curved and rectangular castellations is very small. But the simulation time with rectangular castellation is much shorter than that with curved castellation.

Furthermore, we compare the difference between the duplexer package with curved castellations and the package with rectangular castellations. With the same die model and bonding wire scheme, the frequency responses of the duplexer with curved and rectangular castellations are compared in Figure 11. The solid line shows the frequency response of the duplexer with curved castellations in the package. The dashed line shows the frequency response of the duplexer with simplified rectangular castellations in the package. From Figure 11, the excellent agreement is found. Therefore, the simplification of the castellations is acceptable and can be used in the modeling to reduce the mesh density and improve the simulation speed.
Figure 11: The frequency responses of the duplexer with curved and rectangular castellations.

(a) Frequency response in the Tx channel.

(b) Frequency response in the Rx channel.

(c) Isolation between Tx and Rx channels.
The third simplification is the ground pads on the bottom of the package. The ground pads are separated in the real package. Figure 12 shows the bottom view of the real package. In the simulation model, all the ground pads are united together on the bottom of the package as shown in Figure 13. When the package is set on the PCB, the ground pads are connected to each other through the PCB ground vias and the middle ground pad of the PCB. Therefore, this simplification is valid.

Figure 12: Bottom view of the real package.
After creating the dimensioned model in 3-D modeler, the boundary conditions and excitations should be carefully considered to get good results. Here, a $3.8 \times 3.8 \text{ mm}^2$ cellular duplexer package is selected to demonstrate how to obtain the package model using full-wave analysis. The cellular duplexer package has the transmission center frequency $f_T = 836.5\text{ MHz}$ and receiver center frequency $f_R = 881.5\text{ MHz}$. This is the ceramic surface mounted device (SMD) package. Figure 14 shows the 3-D structure of this package. The ceramic layers and lid are displayed as wireframe in 3-D view to allow insight into the connecting structure. The solder pads are on the bottom side of the first ceramic layer. The fourth ceramic layer comprises bonding pads for the bonding wire and a cavity, wherein the chip is mounted. The electrical
connections between the bonding pads and solder pads are achieved through external castellations. The meander delay line inside the package gives the phase shift at the front of receiver channel.

Figure 14: 3-D view of a $3.8 \times 3.8 \text{ mm}^2$ cellular duplexer package.

Two ladder filters are located inside the package. A cascaded T-network is used for transmitter path and a Pi-network is used to implement the receiver path of the duplexer. The die is mounted using conductive epoxy to ensure ground on the backside of the die. The connections between the die and package are made using gold bonding wires. The schematic of the duplexer assembly is shown in Figure 15.
Figure 15: Schematic of the cellular duplexer assembly.

For the duplexer package simulation, eight ports should be defined for the inner bonding pads that can be connected with die through bonding wires. The other three ports should be defined for the external signal pads. Figure 16 shows the port definition for the 3.8×3.8 mm² cellular duplexer package.
After finishing the duplexer package simulation, a 11 by 11 S-matrix can be obtained. All the useful information for the package is included in the S-parameters. From antenna (port 11) to Rx input (port 8), we can find the effect of meander delay line. Figure 17 shows the magnitude and phase for the meander line from 0.68 to 1 GHz. The results from Figure 17 are helpful to check the trace design. Other scattering parameters can also be found but are not presented.
(a) Magnitude of $S_{11,8}$.

(b) Phase of $S_{11,8}$.

Figure 17: Magnitude and phase of meander delay line from 0.68 to 1 GHz obtained from port 8 to port 11.
2.2 Extraction of the Die Model

The die model is comprised of die busbars parasitics and acoustics models. The acoustics model is obtained using the coupling-of-modes (COM) techniques [30]-[35]. The die busbars parasitics model is generated by Ansoft HFSS or Agilent Momentum.

Figure 18 shows the die busbars model in Ansoft HFSS. The busbars are defined as aluminum with the thickness of 0.48 μm. The substrate is lithium tantalite (LiTaO3) with the thickness of 350 μm. In the simulation, the thickness of the die busbars can be neglected. They are defined as finite conductivity boundary. Combined with die acoustics model, the die model can be obtained. Figure 19 shows the schematic of the die with port assignments.

Based on the die results, we can investigate the electrical response of transmitter and receiver path of the duplexer die and the effect of the phase shift in Agilent advanced design system (ADS). Figure 20 shows a schematic of the connection for the die simulation and the electrical response of each path with an ideal 90º phase shift applied to the input of the receiver path. In Figure 20(b), the solid line shows the frequency response in Tx channel. The dashed line shows the frequency response in Rx channel.
Figure 18: Die busbars model in Ansoft HFSS.
Figure 19: Schematic of the die with port assignments.
(a) Schematic of the connection for the die simulation.

(b) Simulated electrical response of the die.

Figure 20: Schematic of the connection and electrical response of the die with an ideal 90° phase shift applied to the input of the receive path.
2.3 Bonding Wire Modeling

Bonding wires are extensively used in integrated circuit packaging and circuit design in RF applications [36]-[43]. The RF designers usually use approximate analytical formulae for straight wires to estimate bonding wire inductance $L$ and mutual inductance $M$ as shown below [44], [45].

$$L \approx \frac{\mu_0 \cdot l}{2\pi} \left[ \ln\left(\frac{2l}{r}\right) - 0.75 \right] \quad (1)$$

$$M \approx \frac{\mu_0 \cdot l}{2\pi} \left[ \ln\left(\frac{2l}{D}\right) - 1 + \frac{D}{l} \right] \quad (2)$$

where $\mu_0$ is the permeability in free space, $l$ is the wire length, $r$ is the radius of the wire, and $D$ is the distance between two wires.

For the bonding wire with 2-mm long and 1-mil diameter, the formula yields 2 nH for the inductance. So usually we can estimate the inductor value of the bonding wire by 1nH/mm.

As the circuits become more complex, the package and bonding wires become more complex as well. For the high frequency, the parasitics caused by bonding wires, mainly inductance and capacitance, can no longer be ignored and require careful modeling. The inductance of the bonding wire is shape dependent [46]. The general trend is that the larger curvature a wire has, the smaller its inductance. The reason that curved wires have smaller inductance is due to the mutual inductance cancellation of the different segments of a single wire [47].
To get the accurate bonding wire model, we can calculate the S-parameters of the bonding wire in HFSS. Figure 21 shows the dimension of the bonding wire and port assignments. Two lumped gap source ports are defined at both terminators of the bonding wire.

![Figure 21: The dimension of the bonding wire with the unit in mils and the port assignments.](image)

After we get the S-parameters, the inductance can be calculated using the microwave network theory. The bonding wire can be considered as a series RL model. Therefore, for this two-ports network, we have

\[
R + j\omega L = Z_0 \frac{(1+S_{11})(1+S_{22}) - S_{12}S_{21}}{(1-S_{11})(1+S_{22}) + S_{12}S_{21}}
\]  

(3)

From (3), the inductance is obtained as shown in Figure 22. From Figure 22, we can find that the inductance almost keeps constant when the frequency changes from 1.65 to 1.95 GHz. In this case, we can use the constant inductor value for the single bonding wire instead of using the S-parameters.
Due to the high isolation specification for SAW duplexer, the PCB design is very important. PCB should provide excellent isolation among the signal traces. In our design, we use grounded coplanar waveguides (GCPW) which have a number of advantages over microstrip such as low loss, low coupling, low distortion, accommodating three-terminal devices and allowing high density circuitry [48]-[50]. Figure 23 shows the cross section of GCPW.

Figure 22: Inductance for the bonding wire versus the frequency.

2.4 PCB Design and Modeling
The effective dielectric constant $\varepsilon_{\text{eff}}$ and characteristic impedance $Z_0$ can be calculated using the following equations [51].

\[
\varepsilon_{\text{eff}} = \frac{1 + \varepsilon_r K(k') K(k_1)}{1 + \frac{K(k') K(k_1)}{K(k) K(k_1)}}
\]  

(4)

\[
Z_0 = \frac{60\pi}{\sqrt{\varepsilon_{\text{eff}}}} \frac{1}{K(k) + K(k_1)}
\]  

(5)

where $k = \frac{a}{b}$, $k' = \sqrt{1-k^2}$, $k_1 = \frac{\tanh(\pi a/2h)}{\tanh(\pi b/2h)}$, $k_1' = \sqrt{1-k_1^2}$ and
In order to reduce the cross-talk among the signal traces, some ground vias are added along the signal traces. We set two PCB simulations with different number of vias to investigate the effect on isolation. In the first simulation, twenty ground vias are added along the traces as shown in Figure 24.

In the second simulation, we add sixty-one ground vias along the signal traces as shown in Figure 25. The results are compared in Figure 26. The solid line shows the results with twenty ground vias and the dashed line shows the results with sixty-one ground vias.

The other simulations with different ground vias are also accomplished to verify the design idea. From the comparison results, it can be found that the ground vias along the signal traces can help to improve the isolation. The more ground vias, the better the isolation. But when the number of ground vias reaches a certain value, the differences of the isolation are very small as shown in Figure 26. If more than sixty-one ground vias are added, only a little difference can be found. The PCB used for the measurement is shown in Figure 27. It has three metal layers. The dimension is $1 \times 1 \times 0.063$ inch$^3$. 

\[
K(k) = \begin{cases} 
\frac{\pi}{\ln(2+\sqrt{k'})}, & 0 \leq k \leq 0.707 \\
\frac{1}{\pi} \ln(2+\sqrt{k}), & 0.707 \leq k \leq 1 
\end{cases}
\]
Figure 24: PCB with twenty ground vias along the signal traces.
Figure 25: PCB with sixty-one ground vias along the signal traces.
(a) Magnitude of $S_{21}$.

(b) Magnitude of $S_{31}$.

(c) Magnitude of $S_{32}$.

Figure 26: Comparison of isolation among the signal traces of PCB.
Figure 27: The PCB used in measurement with the dimension $1 \times 1 \times 0.063$ inch$^3$. 
2.5 Assembling Method

After finishing all the simulations, the S-parameters of the package, die, and bonding wires are assembled in Agilent ADS to get the total frequency response. Figure 28 shows the schematic of the assembled duplexer. In the bonding scheme, if the bonding wires are close to each other and are connected to different bonding pad or a metal pad in the die layout, the mutual inductance is considered. The mutual inductive coupling coefficient is selected according to the distance between the bonding wires. In Figure 28, \( L_{3,5} \) is the mutual inductor between \( L_3 \) and \( L_5 \). \( L_{9,10} \) is the mutual inductor between \( L_9 \) and \( L_{10} \). In the other cases, the mutual couplings are ignored. From the simulation results, this approximation is reasonable and acceptable.

The cellular duplexer is measured in the lab using a 2-port vector network analyzer (HP-8753). Figure 29 compares the simulated response of the cellular duplexer with the measurement of the duplexer. The solid line shows the measurement results. The dashed line shows the simulation results. From Figure 29, excellent agreement is found between the simulation and measurement results. This shows the effectiveness of this methodology.

The above method is effective and flexible. The main advantage is that we can separate the package, die, and bonding wire models. If the die layout changes, we only need to model the die again. It saves a lot of time and is easy to use the optimization method to change the bonding wire scheme. But this method neglects the mutual couplings among the package, die busbars and bonding wires.
Figure 28: Schematic of assembled cellular duplexer.
(a) Frequency response in the Tx channel.

(b) Frequency response in the Rx channel.

(c) Isolation between Tx and Rx channels.

Figure 29: Comparison of the simulation and measurement results for cellular duplexer.
2.6 Consideration of Mutual Coupling

To take into account the mutual couplings, the methodology is extended to model the package, die busbars and bonding wires together. Figure 30 shows the simulation model with the package, die busbars, and bonding wires. The comparison of the simulation and measurement results for $|S_{21}|$ is shown in Figure 31. The solid line shows the measurement results. The dashed line shows the simulation results. An excellent agreement is found between the simulation and measurement results.

Another example is shown in Figure 32. In the simulation model, two lumped gap source ports are defined to check the frequency response in Rx channel. The comparison of simulation and measurement results are shown in Figure 33. The solid line shows the measurement results. The dashed line shows the simulation results.

The main advantage of this modeling method is that it includes all the necessary components. The simulation accuracy is better than that of the previous assembling method. But it is time-consuming and lacks the flexibility. If we only change the die layout or the bonding scheme, we should rerun the whole simulation.

These results show the effectiveness of the methodology. This technique reduces the design cycle time significantly and can be applied to various RF SAW device packages.
Figure 30: Simulation model with the package, die busbars, and bonding wires.
Figure 31: Comparison of the simulation and measurement results for $|S_{21}|$. 
Figure 32: Simulation model for checking the response in the Rx channel.
Figure 33: Comparison of the measurement and simulation results for the Rx channel.

(a) $|S_{21}|$.

(b) $|S_{11}|$.

(c) $|S_{22}|$. 
CHAPTER THREE: ANALYSIS AND DESIGN OF KPCS DUPLEXER PACKAGE

This chapter provides a comprehensive methodology for accurate analysis and design of KPCS duplexer. Full-wave analysis is applied to get the package model. The die model is obtained by combining the acoustics and parasitics models. The modeling of bonding wire is also discussed. The models of package, die, and bounding wires are assembled together to get the total response. Based on this methodology, several novel ideas are proposed to significantly improve the isolation. Simulation and measurement results are compared, and excellent agreement is found.

3.1 Full-Wave Analysis of the Duplexer Package

Here we consider the modeling of 3.8×3.8 mm² KPCS duplexer package that has the transmission center frequency $f_T = 1765$ MHz and receiver center frequency $f_R = 1855$ MHz. This is the ceramic surface mounted device (SMD) package. Figure 34 shows the 3-D structure of this package. The ceramic layers, lid, and inner ground plane are displayed as wireframe in 3-D view to allow insight into the connecting structure. For the same reason, the ceramic layers are displayed as wireframe in the detailed view. The solder pads are on the bottom side of the first ceramic layer. The fourth ceramic layer comprises bonding pads for the bonding wire and a cavity, wherein the chip is mounted. The electrical connections between the bonding pads and solder pads are achieved through external castellations. The meander delay line inside the package gives the phase shift at the front of receiver channel.
Two ladder filters are located inside the package. A cascaded T-network is used for transmitter path and a Pi-network is used to implement the receiver path of the duplexer. The die is mounted using conductive epoxy to ensure ground on the backside of the die. The connections between the die and package are made using gold bonding wires. The schematic of the duplexer assembly is shown in Figure 35.
For the duplexer package simulation, eight ports should be defined for the inner bonding pads that can be connected with die through bonding wires. The other three ports should be defined for the external signal pads. Figure 36 shows the port definition for the 3.8×3.8 mm² KPCS duplexer package. Lumped gap source ports are used for port definition.
After finishing the duplexer package simulation, 11 by 11 S-matrix can be obtained. All the useful information for the package is included in the S-parameters. From antenna (port 11) to Rx input (port 1), we can find the effect of meander delay line. Figure 37 shows the magnitude and phase of the meander line from 1.6 to 2 GHz. Other scattering parameters are not presented.
Figure 37: Magnitude and phase of the meander delay line from 1.6 to 2 GHz obtained from port 11 to port 1.

(a) Magnitude of $S_{1,11}$.

(b) Phase of $S_{1,11}$. 

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### 3.2 Extraction of the Die Model

The die model is comprised of die parasitics model and acoustics model. Agilent Momentum is used to generate the die parasitics model. The acoustics model is obtained using the COM techniques. Figure 38 shows the schematic of the die with port assignments.

![Figure 38: Schematic of the die with port assignments.](image)

Based on the die results, we can investigate the electrical response of transmitter and receiver path of the duplexer die and the effect of the phase shift in Agilent ADS. Figure 39 shows the schematic of the connection for the die simulation and the electrical response of each path with an ideal 90° phase shift applied to the input of the receiver path. In Figure 39(b), the solid line is for Tx. The dashed line is for Rx.
(a) Schematic of the connection for the die simulation.

(b) Simulated electrical response of the die.

Figure 39: Schematic of the connection and electrical response of the die with an ideal 90° phase shift applied to the input of the receiver path.
3.3 Total Response of the Package with Die

After getting the S-parameters of the package, die and bonding wires, we can assemble the results in Agilent ADS. Figure 40 shows the schematics of the assembled duplexer. In the bonding scheme, if the bonding wires are close to each other and are connected to the same bonding pad and metal pad in the die layout, the equivalent inductor is used as $L_{eq1}$ and $L_{eq2}$ in Figure 40. If the bonding wires are close to each other and are connected to different bonding pad or a metal pad in the die layout, the mutual inductance is considered. The mutual inductive coupling coefficient is selected according to the distance. In Figure 40, $L_{23}$ is the mutual inductor between $L_2$ and $L_3$. In the other cases, the mutual couplings are ignored. From the simulation results, this approximation is reasonable and acceptable. Figure 41 shows the simulated total frequency response of the KPCS duplexer.

Figure 40: Schematics of the assembled KPCS duplexer.
Figure 41: Simulated total frequency response of the KPCS duplexer.

(a) Simulated frequency response in the Tx channel.

(b) Simulated frequency response in the Rx channel.

(c) Simulated isolation between Tx and Rx channels.
From Figure 41, the insertion loss is less than 3 dB from 1.75 to 1.78 GHz for the Tx band and less than 1.8 dB from 1.84 to 1.87 GHz for the Rx band. But the isolation in the Tx band is about -52 dB from 1.7 to 1.75 GHz and about -55 dB from 1.75 to 1.78 GHz. In order to reach -60 dB, the redesign of the package, die, bonding wire scheme should be considered.

3.4 Novel Methods to Improve the Isolation

As mentioned above, small insertion loss and high isolation in stopband are the most important specifications in the SAW filter and duplexer package design. With higher frequency and smaller size of the duplexer package, we should have good physical understanding of each part of the package. We can easily analyze the influence of each part based on the methodology developed in this dissertation. For the KPCS duplexer package, we have found several novel methods to improve the isolation at the Tx stopband [25].

3.4.1 Cut on Inner Ground Plane

Good isolation is achieved by reducing the coupling between transmitter and receiver channels. For the package structure shown in Figure 34, the chip is mounted on the inner ground plane, which is a big metal piece. If we cut the center part of the inner ground plane, part of the fields above the inner ground plane will be absorbed by dielectric material. The isolation then can be improved. We cut different shapes to investigate the effect and finally we find cutting the “x” shape from the center of inner ground plane can give us better isolation. Figure 42 shows the inner view of the package with the modified inner ground plane. The isolation can be improved
from 1.73 to 1.76 GHz at the Tx band, and 6.3 dB improvement is achieved at 1.75 GHz as shown in Figure 43. The dashed line shows the new simulation results with the modified inner ground plane. The solid line shows the results shown in Figure 41. At the same time, the performance in passband can stay the same.

Figure 42: Inner view of the package with the modified inner ground plane.
(a) Isolation between Tx and Rx channels.

(b) Frequency response in the Tx channel.

Figure 43: Comparison of the previous simulation results shown in Figure 41 and the new results with modified inner ground plane.
3.4.2 New Bonding Wire Scheme

Note that the rejection is also dependent on the bonding wire location due to the self and mutual inductances of the bonding wires. The inductive couplings between the bonding wires generate parasitic effects in the stopband [52]. After combining the results of the package and die in ADS, we can change the positions of bonding wires to investigate the influence of the bonding wires. For the bonding wire scheme shown in Figure 44, the bonding wire positions are changed for the Rx ground, and two more bonding wires are added compared with the bonding wire scheme shown in Figure 35.

![New bonding wire scheme](image)

Figure 44: New bonding wire scheme.
The inductance to ground for the Rx channel is reduced. The isolation can be improved from 1.65 to 1.8 GHz and 4.6 dB improvement is achieved at 1.75 GHz as shown in Figure 45. The dashed line shows the new simulation results with the new bonding wire scheme. The solid line shows the isolation results shown in Figure 41.

![Figure 45: Comparison of the previous isolation results shown in Figure 41 and the new results with the new bonding wire scheme.](image)

3.4.3 Adding Vias Between the Ground Bonding Pad and Inner Ground Plane

The via has big influence on the package performance. Two vias are added between the inner Rx ground bonding pad and the inner ground plane as shown in Figure 46 to reduce the inductance from the Rx ground bonding pad to the ground. The isolation can be improved from 1.74 to 1.79 GHz at the Tx band, and 6 dB improvement is achieved at 1.765 GHz as shown in
Figure 47. The dashed line shows the new results with two more vias shown in Figure 46. The solid line shows the isolation results shown in Figure 41.

Figure 46: Two vias are added between inner bonding pad and inner ground plane to reduce the inductance from the Rx ground bonding pad to the ground.

Figure 47: Comparison of the previous isolation results shown in Figure 41 and the new results with two more ground vias.
3.4.4 Adding Vias Between the Inner Ground Plane and PCB Ground

Next, we add twelve ground vias between the inner ground plane and PCB ground. These vias are distributed around the meander delay line and form a gridded ground wall to stop the electromagnetic coupling between the meander delay line and the other components [53]. This wall also reduces the impedance between the inner ground plane and PCB ground. Figure 48 shows this new structure. Figure 49 shows the comparison of results. The dashed line shows the new results with twelve ground vias between the inner ground plane and PCB ground. The solid line shows the isolation results shown in Figure 41. The isolation can be improved from 1.65 to 1.8 GHz, and 4.7 dB improvement is found at 1.75 GHz. The attenuation for the Tx channel is improved from 1.65 to 1.71 GHz and from 1.84 to 1.95 GHz.

Figure 48: Twelve vias are added between the inner ground plane and PCB ground.
(a) Isolation between Tx and Rx channels.

(b) Frequency response in the Tx channel.

Figure 49: Comparison of the previous simulation results shown in Figure 41 and the new results with twelve ground vias between the inner ground plane and PCB ground.
3.5 Experiment Results

Based on our investigation of improving the isolation, the KPCS duplexer package is designed using the new bonding wire scheme and adding twelve ground vias between the inner ground plane and PCB ground. After fabrication, we measure the new KPCS duplexer package in the lab using a 2-port vector network analyzer (HP-8753). As apparent from the construction of the package, it is difficult to measure all possible port combination. Only the transmission parameters are extracted in the lab, then port extensions are applied to de-embed the line delay of the PCB. Figure 50 shows the package ready in the fixture for measurement. Periodic holes are made in the PCB of fixture to reduce the couplings among the feeding transmission lines. Therefore, the empty fixture has excellent isolation among three ports.

![Figure 50: The package ready in the fixture for measurement.](image)
Figure 51 compares the new simulated response of the duplexer with the de-embedded measurement results of the duplexer. The previous simulated response shown in Figure 41 is also put in Figure 51 to see the improvement clearly. The solid line shows the measurement results on fixture. The dashed line shows the new simulation results. The dotted line shows the previous simulated results shown in Figure 41.

From Figure 51, we find the simulation can match the measurement results very well. Although we focus on investigating the methods to improve the isolation, we find from Figure 51 that the passband performance is also improved. This shows the effectiveness of our methodology and the novel ideas we proposed here. There still are differences between the simulation and measurement results. This may come from the modeling assumptions. Due to the complex structure of the duplexer, we model the package, die, and bonding wires separately, then we combine these models together to obtain the total response. The couplings among these parts are ignored. Although we make these modeling assumptions, the results are good enough to analyze and design the duplexer package.
(a) Frequency response in the Tx channel.

(b) Frequency response in the Rx channel.

(c) Isolation between the Tx and Rx channels.

Figure 51: Comparison of the simulation and measurement results for new KPCS duplexer.
CHAPTER FOUR: ANALYSIS AND DESIGN OF FLIP CHIP DUPLEXER PACKAGE

This chapter discusses the analysis and design of flip chip duplexer package. The flip chip technology is briefly introduced at the beginning. Ansoft HFSS is applied to obtain the package model. Momentum is used to get the parasitics model of the die busbars. The simulation results from the assembling method are compared with the measurement results. To improve the isolation in Tx band, the optimization method is applied to select the external inductor values. The topology of the package is investigated to improve the isolation. The effect of the Rx ground trace on the isolation is also discussed.

4.1 Introduction of Flip Chip Technology

Flip chip technology was introduced by IBM to replace the bonding wires in the early sixties. Flip chip microelectronic assembly is the direct electrical connection of face-down electronic components onto substrates, circuit boards or carriers by means of conductive bumps on the chip bond pads. In contrast, the wire bonding technology uses face-up chip with a wire connection to each pad. The mechanical stability using wire bonding technology requires a minimum package wall thickness and the wire-bonding wasters twice the space for chip and package bonding [16].

The boom in flip chip packaging results from flip chip’s advantages in size, performance, flexibility, reliability and cost over other packaging methods [54]-[56]. In this chapter, we focus on the modeling and design issues of the flip chip duplexer package.
4.2 Full-wave Analysis of Flip Chip Duplexer Package

Compared with the KPCS duplexer package, the flip chip duplexer package is simple as shown in Figure 52. For the flip chip duplexer package, there’s no embedded trace inside the package to provide the phase shift. The phase shift is accomplished by the external inductor or the trace on the PCB. The solder pads are on the bottom side of the first ceramic layer. The second ceramic layer comprises the bonding pads and a cavity. The electrical connections between the bonding pads and solder pads are achieved through the internal vias. The size of this package is $2 \times 2.5 \text{ mm}^2$.

Figure 52: 3-D detailed view of the flip chip duplexer package.
Table 3
Critical specification for flip chip duplexer

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Units</th>
<th>Min</th>
<th>Max</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transmit Band Insertion Loss</td>
<td>dB</td>
<td>-</td>
<td>-2.5</td>
</tr>
<tr>
<td>Receive Band Insertion Loss</td>
<td>dB</td>
<td>-</td>
<td>-3.5</td>
</tr>
<tr>
<td>Transmit Band Isolation</td>
<td>dB</td>
<td>-54</td>
<td>-</td>
</tr>
<tr>
<td>Receive Band Isolation</td>
<td>dB</td>
<td>-43</td>
<td>-</td>
</tr>
</tbody>
</table>

For the package simulation, five ports are defined for the bottom pads, ten ports are defined for the inner bonding pads which can be connected with die through the bonding balls. Figure 53 shows the port definition of the flip chip duplexer package. After finishing the duplexer package simulation, 15 by 15 S-matrix can be obtained. All the useful information for the package is included in the S-parameters.

Figure 53: Port definition of the flip chip duplexer package.
As discussed above, HFSS or Momentum can be used to get the parasitics model of the die busbars. In Momentum, it is easy to define the ports and select the reference. In this model, twelve ports are defined for the connection of the die and package. Twenty-two ports are defined for the connection with the resonators as shown in Figure 54. The die busbars are defined as sheet conductor with the thickness of 1 μm. After getting the simulation results, the die model can be obtained by combining the parasitics and acoustics models.

Figure 54: Port definition in Momentum.
4.4 Assembling Method

After getting the S-parameters of the package and die, we can assemble the results in Agilent ADS. Figure 55 shows the schematics of the assembled flip chip duplexer. In the assembly, the bonding ball is modeled as the ideal inductor L with the value of 0.05 nH. External inductor L\(_1\) is the Tx ground inductor. The phase shift between the antenna and receiver path is achieved by the external inductor L\(_2\). Figure 56 shows the comparison of simulation and measurement results with L\(_1\) = 0.6 nH and L\(_2\) = 9.5 nH. The solid line shows the measurement results. The dashed line shows the simulation results. Excellent agreement can be found. It also shows the effectiveness of the method.

Figure 55: Schematics of the assembled flip chip duplexer.
(a) Frequency response in the Tx channel.

(b) Frequency response in the Rx channel.

(c) Isolation between Tx and Rx channels.

Figure 56: Comparison of the simulation and measurement results for flip chip duplexer.
4.5 Improvement of the Isolation

Compared with the specification in Table 3, Figure 57 shows the frequency response with the marks and values at the critical frequency points.

(a) Frequency response in the Tx channel.

(b) Frequency response in the Rx channel.

(c) Isolation between Tx and Rx channels.

Figure 57: Frequency response with the marks and values at the critical frequency points.
The insertion losses in both transmit and receiver bands can meet the specifications. The receive band isolation is less than -43 dB. But the transmit band isolation at 824 MHz is only -53.612 dB, a little bit higher than the specification -54 dB.

To improve the isolation, the package and die should be redesigned. The external inductor values can also be optimized to improve the isolation in Tx band. Because the isolation is sensitive to the phase shift, the external inductor $L_2$ is optimized first to improve the isolation. With the optimized $L_2 = 11.4 \, \text{nH}$, the isolation in Tx band is improved and can meet the specification. The receive band isolation at 892.6 MHz is also improved as shown in Figure 58.

![Figure 58: Improved isolation with optimized external inductor $L_2 = 11.4 \, \text{nH}$](image)

<table>
<thead>
<tr>
<th>Band</th>
<th>Frequency (MHz)</th>
<th>dB(S(3,2))</th>
</tr>
</thead>
<tbody>
<tr>
<td>$m_1$</td>
<td>824.0 MHz</td>
<td>-54.897 dB</td>
</tr>
<tr>
<td>$m_2$</td>
<td>841.5 MHz</td>
<td>-55.306 dB</td>
</tr>
<tr>
<td>$m_3$</td>
<td>869.0 MHz</td>
<td>-51.967 dB</td>
</tr>
<tr>
<td>$m_4$</td>
<td>892.6 MHz</td>
<td>-44.034 dB</td>
</tr>
</tbody>
</table>
The Tx ground inductor $L_1$ has the effect on the isolation too. Therefore, the external inductors $L_1$ and $L_2$ can be optimized to obtain the best isolation. For this two variables problem, ADS optimization function can help to obtain the optimized $L_1$ and $L_2$ for the best isolation. The target is to improve the isolation in Tx band. Therefore, the multiple goals can be defined as follows:

1) $S_{23} \leq -56$ dB from 824 MHz to 849 MHz
2) $S_{23} \leq -45$ dB from 869 MHz to 894 MHz
3) $S_{12} \geq -2.5$ dB from 824 MHz to 849 MHz
4) $S_{13} \geq -3.5$ dB from 869 MHz to 894 MHz

where port 1 is the antenna port, port 2 is the Tx input and port 3 is the Rx output. From the optimization result, the optimized $L_1$ is 1.383 nH and $L_2$ is 12.98 nH. The improvement can be found in Figure 59. Compared with the results in Figure 58, the transmit band isolation is improved by 0.952 dB at 824 MHz and 0.586 dB at 841.5 MHz. The receive band isolation is also improved by 1.094 dB at 892.6 MHz. The receive band insertion loss gets worse within the specification as the trade-off result. In Agilent ADS, it’s very flexible to change the goals to meet different targets.
(a) Frequency response in the Tx channel.

(b) Frequency response in the Rx channel.

(c) Isolation between Tx and Rx channels.

Figure 59: Improved performance with optimized external inductor $L_1$ and $L_2$. 
It’s helpful for the isolation improvement if the Rx part is far away from Tx part. For the package shown in Figure 53, it is realized that the distance between Rx out and Tx input doesn’t reach the maximum value within the size limit. Therefore, the new topology of the flip chip duplexer package is proposed as shown in Figure 60. This configuration can also provide more room for the die optimization. Figure 61 shows the simulation results with the new package, new die, optimized $L_1 = 0.9 \, \text{nH}$ and $L_2 = 15 \, \text{nH}$.

Figure 60: Port definition of the new topology of the flip chip duplexer package.
(a) Frequency response in the Tx channel.

(b) Frequency response in the Rx channel.

(c) Isolation between Tx and Rx channels.

Figure 61: Simulation results with the new package, new die, optimized $L_1 = 0.9$ nH and $L_2 = 15$ nH.
With the new design, the transmit band isolation can be improved by 11.153 dB compared with the specification. Figure 62 shows the comparison of new simulated isolation results and the measurement results based on the previous design shown in Figure 56. The dashed line shows the simulation results based on the new package, new die, and optimized external inductors $L_1$ and $L_2$. The solid line shows the measurement results based on the previous design shown in Figure 56. With the new package and die, the isolation can be improved in the whole frequency range. About 13 dB improvement of the isolation is achieved in the Tx band.

Figure 62: Comparison of new simulated isolation results and measured isolation results based on the previous design shown in Figure 56.
Based on the package shown in Figure 60, the effect of Rx ground trace on the isolation is investigated. In Figure 63, a “τ” shape Rx ground trace is used instead of the “T” shape Rx ground trace as shown in Figure 60. With optimized $L_1 = 0.742 \, \text{nH}$ and $L_2 = 15 \, \text{nH}$, the transmit band isolation is improved by 13.217 dB at 824 MHz and it can reach -66.42 dB in the frequency range from 824 MHz to 849 MHz as shown in Figure 64.

Figure 63: Modified Rx ground trace based on the package shown in Figure 60.
Figure 64: Improved transmit band isolation with the modification of the Rx ground trace.

Compared with the result in Figure 61 (c), 1.267 dB more isolation in transmit band can be found in the frequency range from 824 MHz to 849 MHz. In Figure 65, the solid line shows the isolation with “T” shape Rx ground trace. The dashed line shows the isolation with “τ” shape Rx ground trace.
Furthermore, one ground via is added between ground pad which is labeled “8” and the ground as shown in Figure 66. With the optimized $L_1 = 0.786 \, \text{nH}$ and $L_2 = 15.45 \, \text{nH}$, the transmit band isolation can be improved by 3.219 dB at 824 MHz and it can reach -67.321 dB in the frequency range from 824 MHz to 849 MHz as shown in Figure 67.
Figure 66: Adding one ground via between the ground pad labeled “8” and the ground.
Figure 67: Improved transmit band isolation with one ground via between the Rx ground pad labeled “8” and the ground.

Compared with the result in Figure 61 (c), 2.168 dB more isolation in transmit band can be found in the frequency range from 824 MHz to 849 MHz. In Figure 68, the solid line shows the isolation with “T” shape Rx ground trace as shown in Figure 61. The dashed line shows the isolation with “τ” shape Rx ground trace with one more ground via between the Rx ground pad labeled “8” and the ground.
Figure 68: Comparison of the isolation with the “T” shape Rx ground trace and the “τ” shape Rx ground trace with one more ground via between the ground pad labeled “8” and the ground.
CHAPTER FIVE: CONCLUSION

In this dissertation, a comprehensive methodology for analysis and design of the RF SAW duplexer package is established. The full-wave analysis tool, Ansoft HFSS, is successfully used for the characterization of the RF SAW duplexer package. The boundary conditions and excitations should be carefully considered to get the accurate results. The die model is obtained by combining the acoustics and die busbars parasitics models. The die acoustics model is generated using coupling-of-models technique. The die busbars can be modeled by Ansoft HFSS or Agilent Momentum. The accurate bonding wire model can be generated from Ansoft HFSS. The inductor value of the bonding wire can also be extracted from the S-parameters. GCPW is used in PCB design. The ground vias are added around the signal traces to reduce the cross-talk. After that, all the results are assembled in Agilent ADS to get the total response. Excellent agreement is found between the simulation and measurement results.

The main advantage of the methodology is that we can separate the package model, die model and bonding wire model. If the die layout changed, we only need to model the die again. It saves a lot of time and is easy to use the optimization method to change the bonding scheme. But this method neglects the mutual couplings among the package, die busbars, bonding wires and PCB. To take into account the mutual couplings, the methodology is extended to model the package, die busbars, and bonding wires together. Because it includes all the necessary components, the accuracy of the simulation results is improved. However it is time-consuming and lack of flexibility. If we only change the die layout or the bonding scheme, we should rerun the whole simulation. It also needs a lot of computer resources due to the complex structure. Sometimes simplifications are needed to reduce the complex of model. For example, we can
simplify the PCB by only considering the signal traces and nearby ground vias. Changing the vias to rectangular shape is also helpful to reduce the mesh density.

Based on the above methodology, we have analyzed and designed the KPCS duplexer package. The package model is obtained from Ansoft HFSS. The bonding wires are models as ideal inductors with the values extracted from S-parameters. In the bonding scheme, if the bonding wires are close to each other and are connected to the same bonding pad and metal pad in the die layout, the equivalent inductor is used. If the bonding wires are close to each other and are connected to different bonding pad or metal pad in the die layout, the mutual inductance is considered. The mutual inductive coupling coefficient is selected according to the distance. In the other cases, the mutual couplings are ignored. From our results, this approximation is reasonable and acceptable. After we obtain the total frequency response, the improvement of isolation is investigated. The isolation of KPCS duplexer package can be significantly improved by redesigning inner ground plane, bonding wire scheme and ground vias. Based on our investigation, a KPCS duplexer package is designed and excellent transmitter to receiver isolation in the transmission band is achieved. Simulation and measurement results are compared and excellent agreement is found. Although we focus on investigating the methods to improve the isolation, the passband performance is also improved.

The methodology is also successfully used for flip chip duplexer. The package model is obtained from Ansoft HFSS. The die busbars are modeled by Agilent Momentum. The bonding ball is modeled as the ideal inductor L with the value of 0.05 nH. The simulation results from the assembling method can match the measurement results very well. Optimization method is applied to improve the transmit band isolation. With the new package and die design, the transmit band isolation can be improved from -53.6 dB to -65.2 dB. Based on the new package,
the effect of the Rx ground trace is investigated and the transmit band isolation can achieve -67.3 dB with the modification of the Rx ground trace. 13.7 dB improvement of transmit band isolation is achieved in the simulation.

The technique developed in this dissertation reduces the design cycle time greatly and can be applied to various RF SAW device packages.
LIST OF REFERENCES


