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High Rejection Low-Pass-Filter Design Using Integrated Passive Device Technology for Chip-Scale Module Package

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Abstract
Currently, there is widespread adoption of silicon-based technologies for the implementation of radio frequency (RF) integrated passive devices (IPDs) because of their low-cost, small footprint and high performance. These devices are receiving increased attention for developing front-end-module (FEM) applications in mobile communication systems. This paper discusses the design of low pass filters (LPF) for use in the GHz range with high harmonic rejection. In large modules, these filters make use of co-planar ground planes and lumped IPD technology on a silicon substrate CSMP (Chip Scale Module Package). The use of coplanar ground in such modules introduces unique performance constraints, especially in filters requiring high harmonic rejection. These problems are discussed, along with design approaches to help mitigate them.

Introduction
With the rapid growth of wireless communication markets, silicon is recognized as an especially promising material to meet the demands of low cost, high integration for RF passive devices [1]. A key requirement is the fabrication of high quality factor (Q) inductors. Typically, silicon substrates have higher substrate losses than GaAs or ceramic materials. In order to overcome this difficulty, many approaches have been tried. In the early development of high quality factor (Q) inductors on silicon substrates, good results have been reported using high resistivity silicon wafer [2].

RF passive networks are predominantly used in the front end of wireless communication systems (i.e. between the active RFIC and the antenna.) Conventional discrete passive components consume 60%-70% of the footprint area of the system. These components are typically made using ceramic technology [3], because of ceramic’s good electrical and thermal characteristics. However, integrated passives devices (IPDs) based on semiconductor processes offer the advantage of excellent parameter control, and allow simplified and compact module designs. IPD processes can be used to make high density capacitors, high Q inductors and large value resistors [4].

The general trend in packaging for RF wireless products, especially for consumer applications, is to reduce form-factor and, consequently, cost. This is leading to the adoption of 3-D packaging technology in front-end-modules. Limitations in silicon-based technologies impose some constraints on the performance of RF passive networks, especially in filters requiring high levels of harmonic rejection. This paper describes some of these constraints, and design approaches aimed at overcoming them.

Simple low-pass filter IPDs
Figure 1 shows two example low-pass filter designs for DCS (1710-1910MHz) cellular applications. These devices are designed as stand-alone discrete circuits. They have the same circuit architecture, similar component values and characteristics. The only substantial difference between the two is the method of attachment. For flip-chip attachment the component is mounted face-down on a system board or SiP assembly, whereas for wire-bond the IPD is mounted face-up with connections to the underlying system board being made by bond wires. These two attachment methods have different parasitic impedance characteristics, leading to some minor differences in the device performance.

Figure 1: micrographs of example discrete IPD low-pass filters.

An important constraint in silicon IPD technology is the lack of vertical connection to the backside of the silicon substrate. Emerging thru-silicon-via (TSV) technology may change this in the future, but at present IPDs (especially wire-bond IPDs) are constrained to make all of their connections to the underlying board from their periphery. In small IPDs, like the examples shown in Figure 1, this is not a severe constraint since the distance from any point in the IPD to the edge is not very great. The wire-bond arrangement for the low-pass filter in this example is shown in Figure 2. Typical wire lengths are 0.5 to 0.7mm. Note in this case the use of dual wire-bonds for the ground connections to minimize the effects of inductance. In the frequency range most widely used for IPDs like these, from 1 to 6GHz, interconnections of length less than 1mm do not present insurmountable problems of parasitic impedance.

However, even for such small devices, the differences between wire-bond and flip-chip parasitic inductance have a noticeable effect on the characteristics. This is especially true for the stop-band characteristics of filters at high frequencies.
Figure 3 compares the characteristics of the two DCS filters shown above. The pass-band response of the filters, around 1.8GHz, is very similar, but appreciable differences can be seen in the high frequency harmonic attenuation. These differences are attributable to the added parasitic inductance of the wire bonds in the ground connections. In the frequency range around 6-7GHz, this inductance adds attenuation poles to the response. These added poles can be used to advantage in the design, but also generally lead to spurious resonant peaks in the response like the one seen just below 8GHz.

In larger modules, especially 3D stacked structures, the problems of ground connection are much more severe. Because of their larger size, it is typically not possible to directly access ground connections from the edge of the RF circuit block as in the example shown above.

3D module structures

Three common 3-D module structures are shown in Figure 4. All of these package types use a silicon substrate to fabricate the Integrated Passive Devices (IPDs) required for the front-end. These circuits are generally balun transformers (used to convert signals from single-ended to differential or vice-versa) band-pass filters (mainly used in receivers to pass only the desired band and block undesired out-of-band interference) and low-pass filters (mainly used in transmitters to eliminate harmonics from the output spectrum.) Active RFICs are electrically connected to the silicon substrate by either flip-chip solder bumps or by wire bonds.

A major issue in the design of such modules is the ground reference. In receiver circuits, RF signals originate at an external antenna, and in transmitter circuits the RF power is delivered to an external antenna. In both cases, the ground reference for the signal is usually a ground plane on the system board. This ground is typically a continuous, or nearly continuous, plane of metal with a low RF impedance. Below 10GHz, where most applications of silicon IPDs are found, this effectively approximates an ideal ground plane.

In 3D module structures like the examples shown above, it is usually desirable to add a secondary ground plane on the IPD substrate. Since the silicon substrate does not have a thru-via, local connections to the main system ground are not possible. (Emerging TSV technologies may change this, but at present this is not in widespread use.) Instead, the local ground plane on the IPD is connected to the system ground around its periphery, either by wire-bond or flip-chip connections. Because the dielectrics on the silicon IPD are thin, microstrip designs are not generally feasible. So, this local ground plane is usually a coplanar structure, with openings to accommodate the components. As such, it is less ideal than the system ground plane, or even a microstrip arrangement with a solid metal ground. Openings in the coplanar ground constitute inductive discontinuities that give rise to RF voltages on the “ground” plane, which is not necessarily well grounded.

In addition to the non-ideal characteristics of coplanar ground structures, the peripheral connection of the IPD ground to the system ground is imperfect. The inductance of wire-bond connections in particular, and flip-chip connections to a lesser extent, causes the connection to be progressively worse at higher frequencies. This is not a major problem for small IPDs, like the example shown in Figure 2. These circuits are so small that they do not have their own secondary ground plane, but instead make local connections to the main
system ground. If the IPD is not too large, any part of the IPD circuit is sufficiently close to an edge to conveniently connect to a wire-bond pad.

**Performance effects in Flip-Chip Modules**

In the case of a large CSMP substrate, individual circuit blocks are often located much farther from the edge. So they use a coplanar ground. Figure 5 shows a micrograph of a low-pass filter circuit used in a CSMP module with a coplanar ground structure. The RF signal is connected into and out from the circuit using coplanar waveguides. In this arrangement, the circuit ground connections are not individually made to the main system ground reference, but instead are made locally to the secondary coplanar ground, which is itself connected at its periphery to the main system ground. This indirect method of ground connection can have a significant impact on the circuit’s performance.

Figure 6 shows a simplified schematic of the low-pass filter example in Figure 5. The basic circuit uses a 5th-order elliptic architecture. (This same circuit architecture was used in the discrete IPDs shown in Figure 1.) The two series inductors, $L_1$ and $L_2$, and the three shunt ground capacitors, $C_{G1}$, $C_{G2}$ and $C_{G3}$, constitute the basic ladder structure of a conventional low-pass filter. The elliptic characteristics are obtained from capacitors $C_{Z1}$ and $C_{Z2}$, which resonate with the inductors to create two transmission zeros (attenuation poles) above the pass band. These zeros are used to enhance the harmonic attenuation of the filter.

Unlike the ideal prototype circuit upon which this design is based, however, the shunt capacitors do not connect directly to ground. Instead, the capacitors are connected locally to the coplanar ground of the IPD. This local ground is subsequently connected to the main system ground. Even ignoring the non-ideal properties of coplanar ground, the connection between the coplanar ground and the main system ground invariably adds some inductance, through either the connecting bond wires or solder bumps. These connections can have a significant effect of the performance, particularly the harmonic attenuation of low-pass filters.

Figure 7 shows simulated response of this circuit for different values of the ground interconnection inductance $L_{GND}$. The circuit in this example was designed for operation in the WiFi/Wimax bands around 2.4-2.5GHz. The pass-band performance of the filter is not appreciably affected by small amounts of ground-return inductance, but the higher frequency attenuation shows substantial degradation.

In the response shown in Figure 7, for $L_{GND}=0$ the two poles in the attenuation above the pass-band are clearly visible, one located at about 4.5GHz and the other at about 6GHz. However, even for ground-return inductance as small as 50pH the degradation in the attenuation is quite pronounced. To put this value into perspective, the inductance of a typical flip-chip solder bump is about 40pH, and a typical bond-wire is about 500pH. With a modest number of solder bump connections in parallel, low ground inductance values can be obtained, but it requires a large number of bond wires to achieve net inductance substantially smaller than 50pH.

The reason for the extreme sensitivity of the stop-band attenuation to ground-return inductance is illustrated in Figure 8. At low frequencies, the inductors in the circuit have a low impedance and the capacitors have a high impedance. Consequently, the RF signal passes directly through the path formed by two inductors. In this situation the ground-return path through $L_{GND}$ is isolated and has no appreciable effect.

At high frequency, the signal flow is very different. In this case, the impedance of the shunt capacitors is low, and that of

![Figure 5: Micrograph of a low-pass filter in a CSMP module, showing the ground connections in the IPD circuits with secondary coplanar ground.](image5)

![Figure 6: 5th-order elliptic low-pass filter circuit.](image6)

![Figure 7: Simulated response for the example circuit shown in Figures 3 and 4 for various values of ground-return inductance.](image7)
the series inductors is high. Especially near the resonant frequencies of the transmission zeros, the series path through the filter is blocked. In this situation, the shunt capacitors direct the signal current to ground. The pathway for this shunting effect is through the ground-return inductor, \( L_{GND} \). This creates a small voltage drop, \( v_{\text{leak}} \), across this inductance. Because the impedance in the ground return path is shared by both the input and the output, this voltage that develops in the ground return path interconnections is the leakage path for small signals to pass directly from input to output. Even if the series elements in the filter are perfectly blocking, this leakage mechanism will allow some of the signal at the input to show up at the output. So, the key problem in high frequency response arises from the interconnection inductance. However, such inductances are only problematic when they are part of a shared impedance path between the input and output.

The reason for the extreme sensitivity is simply that in the stop bands of the low pass filter the relative signal amplitude is low – typically -20 to -40dB. Even very small levels of signal leakage coupled through the shared impedance of the ground interconnection can be a significant contribution at these low levels.

The results discussed above suggest that the attenuation performance is limited by the voltage leakage through the shared ground return inductance. In practice, however, this limitation can be offset to some extent by retuning the component values in the rest of the circuit. So, for example, the small signal that is transmitted through the main circuitry of the filter can in some cases be adjusted to offset the leakage component through the ground return inductance, resulting in a high level of attenuation. This tuning is generally too complicated to be done analytically, but instead is achieved through simulation and numerical optimization [5]. The sensitivity of the design to ground return impedance, as illustrated in Figure 7, indicates the importance of modeling this impedance and including it in the design and optimization procedure.

The example module circuit described above was used in a 3D flip-chip CSMP module. In this module the ground connection between the IPD and main system board had 7 solder bumps located within less than 1mm of the RF input and output ports, giving a ground return inductance \( L_{GND} \) of about 7pH. (Ground connections that are located far from these ports are much less effective, since the added distance of the current path is equivalent to added inductance.) The circuit was then optimized using the procedure described in [5]. Measurement and simulation are compared in Figure 9.

The measured characteristics of the flip-chip CSMP low-pass filter are summarized in Table 1.

<table>
<thead>
<tr>
<th>Parameter (dB)</th>
<th>Freq (GHz)</th>
<th>min</th>
<th>typ</th>
<th>max</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pass band ins. loss</td>
<td>2.4-2.5</td>
<td>0.4</td>
<td>0.4</td>
<td>0.4</td>
</tr>
<tr>
<td>Pass band ret. loss</td>
<td>2.4-2.5</td>
<td>21</td>
<td>23</td>
<td>26</td>
</tr>
<tr>
<td>2f0 attenuation</td>
<td>4.6-5.0</td>
<td>25</td>
<td>30</td>
<td>35</td>
</tr>
<tr>
<td>3f0 attenuation</td>
<td>7.2-7.5</td>
<td>27</td>
<td>28</td>
<td>29</td>
</tr>
<tr>
<td>4f0 attenuation</td>
<td>9.6-10.0</td>
<td>22</td>
<td>22</td>
<td>23</td>
</tr>
</tbody>
</table>

**Performance effects in Wire-Bond Modules**

Low-pass filter design in wire-bond modules is much more difficult because the inductance of a typical wire bond is more than an order of magnitude greater than that of a typical solder bump. It is not generally effective. Moreover, in large modules the ground return current tends to flow preferentially in the bond wires adjacent to the corresponding signal. This is the lowest inductance path and, consequently, the preferred low impedance path. This becomes especially true at higher frequencies.
Distributed effects on the coplanar ground and non-uniform current flow in multiple wire bonds make it essential to use detailed electromagnetic simulation and modeling of the IPD and its wire bonds. In general, the coplanar ground plane of a silicon module substrate is not an equipotential, so the circuit diagram of Figure 6 is an oversimplification. Figure 10 shows a sketch of a coplanar ground plane with its wire bond connections. If the RF input port, signal current $i_{\text{IN}}$, flows into the module, and ground return currents $i_{G1}$, $i_{G2}$, etc. flow out along the numerous ground wires. In general, the return current tends to flow most strongly in the wires immediately adjacent to the signal wire. Mutual inductance makes this the lowest impedance return path, and the circuitous path required for current in the more distant wires makes their effective inductance even higher.

If the input and output ports of the module are located far apart, as shown in the figure, the shared inductance between them becomes small. Very little of the input port’s return current will be present near the output port and vice versa. So, the shared inductance effects in wire bond modules are not as bad as the higher wire bond inductance might suggest, but it is still important to model them and take them into account. Two good design practices that are indicated by these considerations are to place ground return wire bonds close to the signals, and keep the input and output ports as far apart as possible.

Even with such measures, however, the ground integrity in wire-bond modules is not as good as in flip-chip, and it is more difficult to obtain good high frequency. The same low-pass filters circuit architecture in the two different cases will generally perform better in a flip-chip module.

Another useful technique in wire-bond modules is to add transmission zeros in the stop band. This is illustrated in Figure 11. This circuit uses the same basic 5th-order elliptic filter circuit as in the previous examples, but it also includes one or more shunt resonators. Like the added capacitors in the elliptic filter circuit, these also add transmission zeros in the stop band, but they operate in a fundamentally different way. The added capacitors in the elliptic filter circuit resonate with the main filter inductors to create a high impedance that blocks signal transmission. In the shunt resonators, the capacitor and inductor are in series, and at resonance they create a low impedance that short circuits the signal path to ground.

In circuits with imperfect ground, shunt resonators behave differently because at their resonant frequency they inject current directly into the coplanar ground structure, rather than reflecting it back to the source. Figure 12 shows measured and simulated characteristics of an example of such a design, used in a wire-bond module. The filter is similar to the example in Figure 5, but it also includes one added shunt resonator. The resonator gives rise to the added transmission zero seen just below 8GHz.

The performance characteristics of this low-pass filter are summarized in Table 2. It can be seen by comparison with Table 1 that the attenuation is significantly higher than for the flip-chip example. (This is accompanied by somewhat higher pass-band insertion loss, which is a typical tradeoff in low-pass filter designs.) These results demonstrate that despite the
problems of signal leakage from imperfect ground integrity, it is possible to achieve good low-pass filter performance in wire-bond modules.

Table 2: Measured characteristics of a low-pass filter used in a wire-bond module.

<table>
<thead>
<tr>
<th>Parameter (dB)</th>
<th>Freq (GHz)</th>
<th>min</th>
<th>typ</th>
<th>max</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pass band ins. loss</td>
<td>2.3-2.7</td>
<td>0.9</td>
<td>0.9</td>
<td>0.9</td>
</tr>
<tr>
<td>Pass band ret. loss</td>
<td>2.3-2.7</td>
<td>16</td>
<td>20</td>
<td>30</td>
</tr>
<tr>
<td>$2f_0$ attenuation</td>
<td>4.6-5.4</td>
<td>34</td>
<td>36</td>
<td>40</td>
</tr>
<tr>
<td>$3f_0$ attenuation</td>
<td>6.9-8.1</td>
<td>43</td>
<td>48</td>
<td>53</td>
</tr>
<tr>
<td>$4f_0$ attenuation</td>
<td>9.2-10.8</td>
<td>28</td>
<td>29</td>
<td>31</td>
</tr>
</tbody>
</table>

Conclusions

We have presented examples of silicon IPD low-pass filters implemented in a variety of configurations. In CSMP modules, the added ground return impedance between system ground and the secondary ground plane inside the module has been shown to significantly degrade the high frequency attenuation of the filters.

Despite this limitation, good filter performance can usually be obtained by adjustment in the component values, taking the ground return characteristics into account. In flip-chip modules where ground return impedances are usually small, this adjustment is small. In wire-bond modules the added ground return impedance is significantly larger. In such cases it may be necessary to modify the filter architecture to obtain equivalent performance. An example was presented using an added shunt resonator, resulting in very good harmonic attenuation.

References