A Look at Europe’s Thirst for SPECTRUM
Harmonic Suppression of Edge Coupled Filters Using Composite Substrates

Bandpass filters are employed in numerous microwave applications such as communication systems, radar and test equipment. Microstrip edge coupled structures are often used for bandpass filters because they are relatively easy to implement in printed circuit board (PCB) technology. These filters have many advantages, however, they suffer from an inherent issue with harmonic spurious responses. There are many techniques to address the suppression of these responses, however, each have their own concerns. The harmonic suppression procedure outlined in this article has minimal impact on design complexity and has some additional benefits for the microstrip PCB features, which are often in addition to the filter pattern on some designs. The design process will be outlined, models defined and circuit performance will verify the procedure.

The center bandpass frequency \(f_0\) of a microstrip edge coupled filter will have spurious responses at even harmonic frequencies, that is unwanted filter responses at \(2f_0\), \(4f_0\), \(6f_0\), etc. Specifically, the band at \(2f_0\) has a detrimental effect on symmetry of the intended passband and can impact the bandwidth. The spurious responses occur because the phase velocities of the even and odd mode for each filter element are different.

There are many methods to suppress the \(2f_0\) mode and most of these methods attempt to equalize the phase velocity of the even and odd modes. The use of over-coupling\(^1,2\) is done to extend the phase length of the odd mode to equal the even mode phase velocity. Another method\(^3,4\) uses capacitors to extend the travel path of the odd mode. A sawtooth pattern or wiggly coupled\(^5\) segment is used to lengthen the odd mode path, while maintaining a relatively unchanged even mode path and equalizing the phase velocities. Additionally, suspended microstrip configurations have been employed\(^6\) to equalize the modal phase velocities. All of these methods introduce their own set of capabilities and limits, however, in general, increase the design cycle difficulty. In the case of the suspended microstrip, shown in Figure 1, the design cycle is less impacted, but achieving proper dimensional control in the circuitry with effective multilayer via hole interconne-
tions can be extremely challenging.

The current work offers a variation on suspended microstrip and utilizes a composite substrate copper clad laminate. The laminate uses two dielectric layers, which have significantly different relative permittivity ($\varepsilon_r$). As opposed to the suspended microstrip, this composite laminate concept can be processed through normal PCB fabrication processes, have plated through hole vias and are assembled like most other PCBs.

**COMPOSITE LAMINATE DEFINITION**

The copper clad laminate used to fabricate the microstrip circuitry should ideally be the same type of substrate, however, with very different $\varepsilon_r$. Several variations have been evaluated, but only a composite laminate substrate, using layers of RO3010™ and RO3003™ materials, will be discussed here. The choice of these materials is due to their proven ability to be used in a variety of high volume and multifaced PCB configurations. Additionally, these materials are from the same substrate family, so there are no concerns for material compatibility issues. The RO3010 material has a suggested Design Dk ($\varepsilon_r$) of 11.2 and the RO3003 substrate has a suggested Design Dk of 3.0. This difference in $\varepsilon_r$ is enough to alter the odd mode phase velocity to approach an equalization of the even mode.

The composite laminate uses the dielectric layer with the high $\varepsilon_r$ adjacent to the signal plane and the low $\varepsilon_r$ material next to the ground plane as shown in the figure. There is a ratio consideration for the thickness of the different $\varepsilon_r$ materials to be used in the composite laminate. A comprehensive study for even-odd mode equalization regarding edge coupled microstrip structures suggests some general design rules for frequencies below 20 GHz. When using layered dielectrics with $\varepsilon_r$ ratio of 3.5:1 to 4.5:1, the thickness ratio should be 2:1, with the thicker substrate being the high $\varepsilon_r$ material and placed against the signal plane. Recent experiments have shown that the high $\varepsilon_r$ material can be thinner than the low $\varepsilon_r$ material and still have benefits regarding altering the odd mode phase velocity. Even though this configuration is not optimum for a near perfect equalization of the odd and even mode phase velocities, it does have enough effect on the odd mode velocity to help minimize the $2f_0$ response.

Having the composite laminate with a thin layer of the high $\varepsilon_r$ material and a larger thickness of the low $\varepsilon_r$ material is beneficial for achieving a composite z-axis (thickness) $\varepsilon_r$ value that is closer to the low $\varepsilon_r$ material value. For microwave features that predominately use the z-axis $\varepsilon_r$ properties of this laminate, such as transmission lines, this enables wider conductor widths for less conductor losses. Essentially a composite laminate can be constructed, which will give a nearly equal even-odd mode phase velocity for suppressing the $2f_0$ mode, have a high $\varepsilon_r$ value near the signal layer to give improved bandpass performance of coupled features and have a lower z-axis $\varepsilon_r$ in order to maintain wider conductor widths which will yield less conductor loss.

**EVEN AND ODD MODE OPERATION FOR EDGE COUPLED MICROWAVE FEATURE**

For a microstrip edge coupled feature, the even and odd mode phase velocity can be approximated by:

$$V_{P,even} = \frac{c}{\sqrt{\varepsilon_{eff,even}}}$$  (1)

$$V_{P,odd} = \frac{c}{\sqrt{\varepsilon_{eff,odd}}}$$  (2)

$$\varepsilon_{eff,even} = \frac{C_{even}}{C_{even,air}}$$

$$\varepsilon_{eff,odd} = \frac{C_{odd}}{C_{odd,air}}$$

$$C_{even} = C_p + C_f + C_{gF}$$

$$C_{odd} = C_p + C_f + C_{gA} + C_{gd}$$

where $V_P$ is the phase velocity for either the even or odd mode, $\varepsilon_{eff}$ is the effective dielectric constant for each even or odd mode and $C$ is the various capacitances. $C_{even,air}$ is the capacitance of the microstrip structure when air is used as the substrate for the even mode and the same nomenclature applies to the odd mode. $C_p$ is the parallel plate capacitance, $C_{fr}$ is the fringing capacitance, $C_{gF}$ is the fringing capacitance in the even mode only at the magnetic wall, $C_{gA}$ is the gap capacitance due to coupling in air, $C_{gd}$ is the gap capacitance in the dielectric and $c$ is the speed of light in free space. **Figure 2** gives a visual depiction of the even-odd mode operation of coupled microstrip features.

When considering the even mode operation, it can be seen that the phase velocity will be affected by the dielectric properties of the laminate only. The odd mode phase velocity associated with the quasi TEM wave propagation is affected by the laminate as well as the capacitive gap coupling in the air ($C_{gd}$). The odd mode phase velocity will be faster than the even mode just by the nature of using air in the gap coupling. The composite laminate can slow the odd mode phase velocity by the use of a thin layer of high $\varepsilon_r$ substrate. The capacitive gap coupling ($C_{gd}$) using the high $\varepsilon_r$ material, in combination with the capacitance gap coupling ($C_{ga}$) in the air can be made to approximate the phase velocity of the even mode.
OUTLINE OF FILTER DESIGN PROCEDURE

Several different edge coupled filters were designed in order to evaluate different composite laminate configurations. Furthermore, filters were designed at different frequencies and bandwidths to evaluate the effects of coupling as well as E field depth penetration into the higher εr material.

The filter design, which will be discussed in more detail, is a 5 element (n = 5), Chebychev filter with a passband ripple of less than 0.5 dB, a fractional bandwidth of 0.07 and a center frequency of 1.88 GHz. The procedure for defining the circuit features is a common ladder circuit, lowpass prototype method defined in many microwave texts.9,10

The n = 5 prototype parameters for this filter are:

\[ g_0 = g_6 = 1.0 \]
\[ g_1 = g_5 = 1.7058 \]
\[ g_2 = g_4 = 1.2296 \]
\[ g_3 = 2.5408 \]

The design equations are given by:

\[ \frac{J_{01}}{Y_0} = \frac{\pi}{2} \frac{\text{FBW}}{g_0 g_1} \]
\[ \frac{J_{01}}{Y_0} = \frac{\pi \text{FBW}}{2} \frac{1}{\sqrt{g_0 g_{n+1}}} \]
\[ \frac{J_{n,n+1}}{Y_0} = \frac{\pi \text{FBW}}{2} \frac{1}{\sqrt{g_n g_{n+1}}} \]

are the elements of the ladder-type low pass prototype with a normalized center frequency equal to 1. FBW is the fractional bandwidth of the filter. \( J_{n,n+1} \) are admittances for the J-inverters and \( Y_0 \) is the characteristic admittance of the terminating lines.

The desired even and odd mode impedances of each element of the filter are determined by:

\[ (Z_{oc})_{n,n+1} = \frac{1}{Y_0} \left[ 1 + \frac{J_{n,n+1}}{Y_0} + \left( \frac{J_{n,n+1}}{Y_0} \right)^2 \right] \]
\[ (Z_{oc})_{n,n+1} = \frac{1}{Y_0} \left[ 1 - \frac{J_{n,n+1}}{Y_0} + \left( \frac{J_{n,n+1}}{Y_0} \right)^2 \right] \]

Once the even and odd mode impedances for each element are defined, the circuit pattern dimensions are determined by the use of a free software from Rogers Corp., MWI-2010. This software uses closed form equations to determine the even and odd mode impedance with a given circuit geometry and material properties. The software also gives the wavelength and for each element, the length of the element will be \( \frac{\lambda}{4} \) the wavelength, minus the extension due to fringing. Another method for determining the length of each element is given by:

\[ l_n = \frac{\lambda_n}{4} \left( \sqrt{\varepsilon_{re}} \right)_{n} \left( \varepsilon_{ro} \right)_{n} \]

where \( \varepsilon_{re} \) is the even mode effective dielectric constant, \( \varepsilon_{ro} \) is odd mode and \( \Delta l_n \) is the length extension due to field fringing.

The information in Table 1 was found by using the MWI-2010 software for a microstrip edge coupled model, with an enhancement to include composite substrates. The phase velocity of the different segments of the filter was found and av-

<table>
<thead>
<tr>
<th>Laminate</th>
<th>Odd Mode</th>
<th>Even Mode</th>
<th>Difference Odd/Even</th>
</tr>
</thead>
<tbody>
<tr>
<td>Homogenous</td>
<td>1.75</td>
<td>1.59</td>
<td>1.10</td>
</tr>
<tr>
<td>Composite</td>
<td>1.68</td>
<td>1.65</td>
<td>1.02</td>
</tr>
</tbody>
</table>

The results from these models are shown in Figure 3. The filter performance for the composite laminate is using a 0.127 mm (5 mil) layer of RO3010 substrate and 0.635 mm (25 mil) layer of RO3003 substrate, with the signal plane next to the RO3010 material. The homogeneous laminate is using a single substrate with a \( \varepsilon_r = 4.13 \).

It can be seen that the range of frequencies in the \( 2f_0 \) response region for the microstrip filter using the homogeneous substrate has an insertion loss of approximately –5 dB and the filter using a composite substrate has an insertion loss of ~32 dB. Therefore, the spurious \( 2f_0 \) response is significantly reduced with the use of the composite laminate. Furthermore, it can be seen that the bandwidth of the filter using the composite laminate is wider and the \( 2f_0 \) response has a much narrower bandwidth, compared to the filter using the homogeneous laminate.

MEASURED RESULTS

The composite laminate was made with the two substrates previously described, RO3010 and RO3003 materials. The homogeneous laminate used a TMM®4 copper clad laminate of the same thickness as the composite laminate. The choice of that material is based on the εr being near the desired value and the material being readily available.

The information in Table 1 was found by using the MWI-2010 software for a microstrip edge coupled model, with an enhancement to include composite substrates. The phase velocity of the different segments of the filter was found and av-

![Fig. 3 Comparison of simulated S21, using the same geometry with different substrates.](image-url)
Fig. 4 Physical PCB conductor layout of the microstrip filter.

Fig. 5 PNA test results for the same circuit pattern using composite and homogeneous laminates.

eraged. The physical PCB conductor layout for the filter is shown in Figure 4. The circuits were fabricated, assembled, tested and the measured results are shown in Figure 5. Tr2 is the $S_{21}$ curve for the filter using a homogenous laminate and Tr4 is the $S_{21}$ curve for the filter using the composite laminate. The markers were placed at two frequencies for each $S_{21}$ curve. Marker 1 is the center frequency and marker 2 is the peak of the $2f_0$ response for the filter using the homogenous laminate. Marker 2 has a $|S_{21}|$ peak of –7.3 dB. Marker 3 is the center frequency and marker 4 is the peak of the $2f_0$ response for the filter using the composite laminate, the peak of this $2f_0$ response is –31.3 dB. Although there are some differences from the measured results as compared to the model shown in Figure 3, the significant reduction of the spurious $2f_0$ response is verified.

The main difference of the measured results, as compared to the model, is a frequency shift. The composite laminate uses two substrates and the high $\varepsilon_r$ substrate has a wider $\varepsilon_r$ tolerance, which is typical of dielectric material with high $\varepsilon_r$ values. The high $\varepsilon_r$ material was within the manufacturers specification, however it had a slightly higher $\varepsilon_r$ value than the nominal 11.2 given as the Design Dk. This accounted for the center frequency shift. The other frequency shift associated with the filter using the homogenous material is a natural response of using the available TMM4 material, which had a slightly higher $\varepsilon_r$ than the desired composite $\varepsilon_r$ of 4.13. The TMM4 material has a Design Dk ($\varepsilon_r$) of 4.5 and this too accounts for the frequency shift.

Other differences between the measured results and the model are attributed to the PCB manufacturing effects and tolerances regarding the circuit conductor patterns. These differences are likely the cause for the peaks of the $2f_0$ responses to be slightly different on the measured circuits as compared to the EM model.

BRIEF OVERVIEW OF EXTENDED EVALUATIONS

Other circuits were modeled and evaluated with regards to the same material considerations. Some of the other issues evaluated were 3 element and 7 element filters, as compared to the 5 element filter shown previously. Intuitive assumptions were verified where less filter elements (3 element) had less benefit of the composite material for the $2f_0$ suppression. More filter elements (7 element) had a slight improvement over the 5 element results shown.

Additionally, there were alterations of the 5 element filter with changes in coupling, bandwidth and center frequency. Experiments were also performed using thinner composite materials, using a larger range of $\varepsilon_r$. These materials used the same high $\varepsilon_r$ material ($\varepsilon_r = 11.2$) and the low $\varepsilon_r$ material had an $\varepsilon_r = 2.2$. Due to the wide range of circuit patterns involved with the different filter elements and the material considerations, only general observations can be given here for the composite laminate:

1. Tightly coupled elements will have more benefit than loosely coupled.
2. Higher frequency filters can use a composite laminate with a thinner layer of the high $\varepsilon_r$ material.
3. Wider bandwidth can be achieved when using a larger ratio of $\varepsilon_r$ values.
4. As the laminate gets thinner, a firm definition on the ratio of the high/low $\varepsilon_r$ material becomes more critical.

The composite laminate procedure shown here can be very successful to minimize the spurious $2f_0$ responses, however thorough EM modeling is highly recommended. The filter models shown here demonstrated the concepts, but they were not optimized.

CONCLUSION

It has been shown that a composite laminate using the proper ratio of high/low $\varepsilon_r$ material can significantly reduce the spurious $2f_0$ harmonic re-
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response of a microstrip edge coupled bandpass filter. It was also illustrated that the composite laminate can achieve wider bandwidth. In comparison to the many other techniques for reducing the $2f_0$ harmonic response, the composite laminate does not increase the design complexity and is capable of most common PCB configurations.

Additionally, there are other benefits to this technology, since many microwave PCB designs with filter features also have many other microwave structures. The added benefit of the composite laminate shown here is when the thinnest effective high $\varepsilon_r$ material is used to reduce the $2f_0$ response, the overall z-axis $\varepsilon_r$ value remains relatively low and this allows the other PCB features to have wider conductors. The wider conductors will allow other microwave features to have less conductor loss as well as improved PCB fabrication yields.

The composite laminate allows the designers to easily and significantly reduce the $2f_0$ harmonic response, increase the bandwidth and still maintain a lower z-axis $\varepsilon_r$ value to allow other features on the microwave PCB to have lower losses.

References