As originally published in the IPC APEX EXPO Conference Proceedings.

The Effect of Radiation Losses on High Frequency PCB Performance

John Coonrod, Rogers Corporation, Advanced Circuit Materials Division

This paper is an extension of an IPC paper [1] presented last year which addressed microwave insertion loss of common PCB transmission line circuits. Insertion loss of these circuits is made up of 4 components; conductor loss, dielectric loss, radiation loss and leakage loss. The previous paper focused on conductor loss and dielectric loss, whereas this paper will address radiation loss.

Radiation losses can be a disruptive force for many different reasons. Designs which are sensitive to EMI (ElectroMagnetic Interference) can be affected by radiation loss of a circuit and specifically how the radiated energy may corrupt neighboring circuits. Also the performance of loss-sensitive systems can be impacted with the addition of radiation loss when it is not fully considered. Finally, broadband high frequency RF and millimeter-wave applications certainly have issues with radiation loss and designers expend many efforts to account for these losses.

Background

The difficulty of predicting and accounting for radiation loss stems from the fact there are many variables associated with this loss mechanism. Radiation loss is dependent on frequency, effective dielectric constant, circuit thickness, impedance transitions, discontinuities, signal launch, by-product of spurious modes and circuit design. Complicating the issue is that radiation loss is often a combination of several of these dependencies. In an effort to demonstrate the different aspects of radiation loss in the following work, many of these variables will be nullified in order to illustrate the impact of one particular factor of radiation loss. This will be done several times with different factors to show several key dependencies of radiation loss when considered individually.

There are three common PCB constructions used in high frequency RF applications. These constructions are shown in figure 1 and are microstrip, coplanar and stripline.

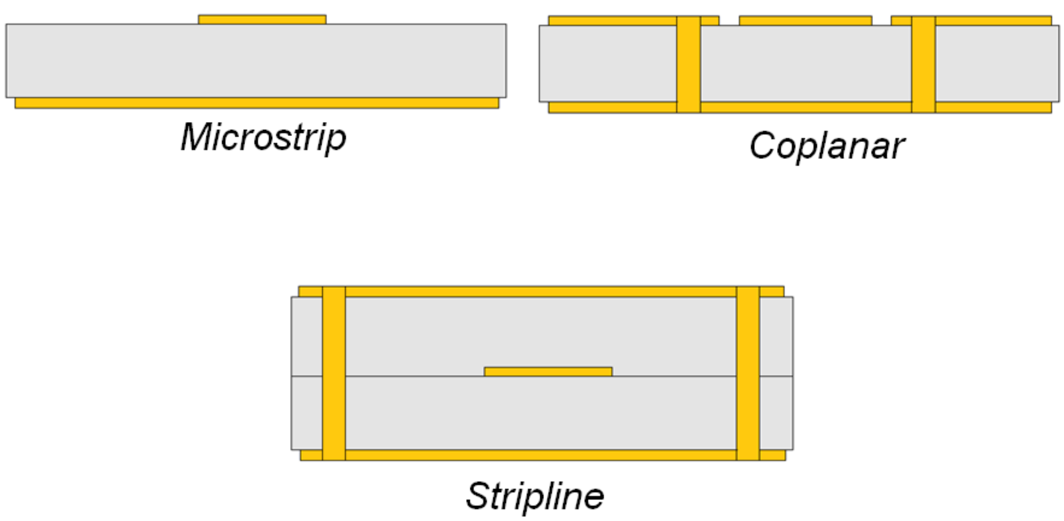

Figure 1. Three common PCB constructions used in high frequency PCB applications.

The microstrip configuration is probably the most common and is often a portion of a multilayer PCB as the outer layers of the circuit. The stripline configuration is also very common and has some advantages over microstrip. One advantage when designed correctly is no radiation loss of the stripline circuit, whereas the microstrip circuit is prone to these losses.

There are many different variants of the coplanar configuration and what is used most often in high frequency PCB applications is a grounded coplanar waveguide (GCPW) also known as a conductor backed coplanar waveguide (CBCPW). This configuration is shown in figure 1 and referred to simply as coplanar.

The foregoing work will focus on microstrip configurations and some information regarding GCPW circuitry. Stripline will not be addressed. A general expression for insertion loss of transmission line circuits built as these configurations is given:

$$\alpha_T = \alpha_C + \alpha_D + \alpha_R + \alpha_L \tag{1}$$

The total loss (insertion loss, α_T) is made up of conductor (α_C), dielectric (α_D), radiation (α_R) and leakage losses (α_L). Leakage losses are typically associated with semiconductor grade materials and are normally not an issue for high frequency PCB circuits. There are exceptions with some high power applications, however for the scope of this paper leakage losses will be dismissed.

More details regarding conductor loss and dielectric loss were given in last year's IPC paper[1] from this author and radiation losses can be defined by the following simple expressions:

$$\alpha_r = 60\left(\frac{2\pi h}{\lambda_0}\right)^2 F(eff) \tag{2}$$

$$F(eff) = 1.0 - \frac{\varepsilon_{eff} - 1}{2.0\sqrt{\varepsilon_{eff}}}\log\left(\frac{\sqrt{\varepsilon_{eff}} + 1.0}{\sqrt{\varepsilon_{eff}} - 1.0}\right) \tag{3}$$

$$F(eff) = \frac{\varepsilon_{eff} + 1.0}{\varepsilon_{eff}} - \frac{(\varepsilon_{eff} - 1.0)^2}{2.0\varepsilon_{eff}^{\frac{3}{2}}}\log\left(\frac{\sqrt{\varepsilon_{eff}} + 1.0}{\sqrt{\varepsilon_{eff}} - 1.0}\right) \tag{4}$$

The symbols α_r, h, λ_0 and ε_{eff} are the radiation loss, height (thickness) of the substrate, wavelength in free space and the effective dielectric constant respectively. When using equation 2 to solve radiation loss, equation 3 must be used when a matched transmission line is considered and equation 4 must be used for an open-circuit line. Equations 2, 3 and 4 were taken from a paper[2] which evaluated the radiation loss of different microstrip circuits and discontinuities. Of the different circuits evaluated in this study, equations 3 and 4 will be used for different scenarios to demonstrate the radiation loss behavior of different circuit designs.

Some of the dependencies of radiation loss can be easily seen in equation 2 and one issue is the thickness of the circuit (h) can have a dramatic impact on this loss. A thicker circuit will have significantly more radiation loss than a thinner circuit. Additionally applications at higher frequencies (smaller λ_0) will also have increased radiation loss as seen in equation 2 where λ_0 has an inverse relationship to α_r. Circuits used at very high frequency such as millimeter-wave (mmWave), which is greater than 30 GHz, typically use thin substrates to help offset the radiation loss issue.

It is less intuitive when looking at equations 2, 3 and 4, but a circuit with a low effective dielectric constant will yield more radiation loss than a circuit with a higher effective dielectric constant. The term effective dielectric constant refers to the combination of the dielectric constant (Dk) for the substrate and air (Dk ≅ 1) when considering circuits such as microstrip or coplanar. Basically with all other issues set equal, a circuit using a substrate with a high Dk will have less radiation loss than a circuit using a low Dk substrate.

As a quick side note and an exception to the concern of radiation loss in this paper, there are some circuit designs that desire high radiation loss. These are antenna applications which have radiating elements sometimes built as microstrip patch antenna PCB's. These circuits often use thick substrates with low Dk and this material combination causes more radiation of the RF energy.

Continuing the discussion of dependencies for radiation loss is the subject of impedance transitions. There are several aspects of microwave circuit design where impedance transitions are necessary. An example would be a microwave power amplifier PCB where the power amp chip has a low input impedance (typically less than 2 ohms) and the circuit board is usually at 50 ohms. The designer will often utilize an impedance transforming network/s to minimize the reflected energy at these impedance transitions. However there are many tolerances associated with the network performance and if the impedance transitions are not smooth there will be reflected energy. Some amount of this energy gets reflected back to the source on by the circuit transmission line, however, some amount of energy will be radiated at that impedance transition. The radiated energy will be loss and that energy has to go somewhere, so the radiated energy could corrupt other circuit conductors causing EMI issues.

Another item which is sometimes related to impedance transitions is signal launch. The transition of the RF energy from the connector to the PCB often has reflections due to impedance transitions or wave propagation mode transitions. RF connectors are typically coaxial in design which means their wave propagation mode is TE (Transverse Electric) and the dominate mode for planar configurations such as PCB's is a TEM (Transverse Electrical Magnetic); more specifically microstrip and coplanar configurations have a quasi-TEM wave propagation mode because the fields use both the dielectric and air causing the wave not to be a pure TEM mode. When the energy of the RF signal has to transition from a TE wave propagation mode to a quais-TEM mode, there are stray reactances at that transition. These reactances are sometimes accompanied with impedance transitions and this can cause reflected energy which generates radiated energy. Accordingly, designers with EMI sensitive circuitry very often put a lot of emphasis on the connector areas of a circuit in the design phase.

Spurious wave propagation modes can occur when a resonance is set-up within a circuit and an additional wave is generated. The created wave is typically unwanted and can be disruptive to the main quais-TEM wave that is desired to propagate on the PCB. As these spurious waves interact with the intended wave or some circuit features, they can also generate radiated energy. More detailed information on signal launch, spurious modes and wave propagation can be found from a paper recently presented on mmWave PCB concerns [3].

Circuit design can play a significant role on radiation loss issues and specifically the type of circuit configuration is important. There is a typical trend of circuit configuration based on frequency for high frequency RF applications. Microstrip circuit configurations are common at microwave frequencies, however, at mmWave frequencies it is more common to employ GCPW circuit configurations. This circuit design migration is mostly due to the minimizing spurious modes as well as radiation losses. There is a hybrid of these configurations where in the signal launch area the circuit will be configured as a GPCW and away from that it will be a microstrip circuit. This design is called a coplanar-launched microstrip and it is an attempt to utilize the best attributes of both circuit types. Microstrip circuits generally have lower insertion loss at microwave frequencies than GCPW circuits, however at mmWave frequencies, micro strip circuits suffer increased radiation loss whereas the GCPW performs much better. When designed correctly, a coplanar launched microstrip circuit will minimize the reflections and stray reactances in the signal launch area yielding a microstrip circuit that performs better at higher frequencies by having less radiation loss.

An excellent paper[4] which gives practical design guidance for microstrip, coplanar-launched microstrip and GCPW shows radiation loss for a microstrip circuit can be improved by the use of the coplanar-launched microstrip configuration. Additionally this paper shows negligible radiation loss for an optimum designed GCPW circuit out to 50 GHz.

Results of experimental data

Multiple experiments were conducted to demonstrate the concepts discussed regarding radiation loss. The first set of experiments employed the use of microstrip gap coupled resonators as shown in figure 2.

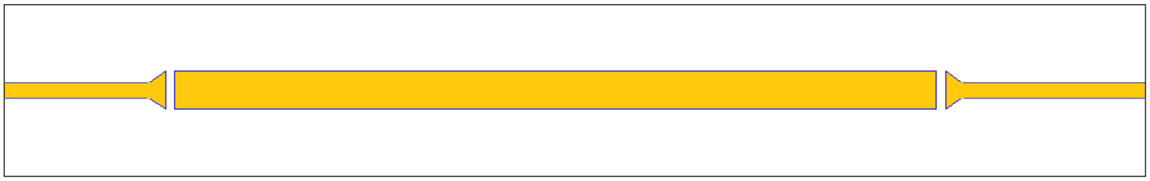

Figure 2. Top view of a microstrip gap coupled resonator.

The resonator was designed to be a ½ wavelength resonator with the lowest node being at 1 GHz. The coupling was such as to obtain good resonant peaks at multiples of 1 GHz while maintaining a return loss better than 15 dB. A return loss greater than 15 dB ensures the loaded Q will have little impact on the resonator response, which essentially means the connectors, cables and calibration has less influence on the resonator performance. The loss data generated by measuring this circuit was from evaluating the loaded Q of the resonator. The unloaded Q is related to the loaded Q and is made up of different components which are associated to the different loss components as can be seen in the following equations:

$$Q_L = \frac{f_0}{BW_{-3dB}} \tag{5}$$

$$Q_0 = \frac{Q_L}{\left(1 - 10^{\frac{-IL}{20}}\right)} \tag{6}$$

$$\alpha_T = \frac{\beta}{2Q_0} = \frac{2\pi}{2\lambda_g Q_0} \tag{7}$$

$$\alpha_R = \alpha_{T\_open} - \alpha_{T\_enclosed} \tag{8}$$

The Q_L value is the loaded quality factor (Q) of the measured resonator and Q_0 is the unloaded Q. The guided wavelength related to the effective dielectric constant of the circuit is λ_g and β is the propagation constant. The α_T value is the total loss of the resonator, when using equation 7. The radiation loss (α_R) is then determined by measuring the total loss of the resonator when it is open (α_{T_open}) and subtracted from the loss of the same resonator when it is enclosed ($\alpha_{T\_enclosed}$). This is assuming the difference in total loss between the open and enclosed resonator is due to radiation loss only.

The losses associated with the conductor (α_C) and dielectric (α_D) loss were determined from an excellent paper on micro strip characterization from Hammerstad and Jensen [5]. The conductor loss calculation had a multiplier applied to α_D to account for the effects of copper surface roughness on conductor loss per Morgan [6].

The resonator was measured initially without an enclosure (open to the environment) which allowed radiated energy to be lost from this circuit. The radiated energy for this design will mostly be from the two gap coupled areas between the 50 ohm feed lines and the resonator element (wide conductor in the middle). Later the same resonator is measured with a grounded metal enclosure which captures the radiated energy so the radiation loss is nullified. For this first experiment the material used to make the circuits was a 30mil thick TMM®4 laminate. The reason a laminate from the product family of TMM materials was used for this experiment and most of the following, is due to this material being available with many different Dk values ranging from about 3.5 to 12.2. Some experiments will look at performance differences when considering different Dk values and using laminates that share the same base substrate, with the main difference being Dk, reduces possible variables due to circuit materials.

Figure 3 shows two screen shots of a microstrip gap coupled resonator at node 2 or approximately 2 GHz. Figure 3a is the initial measurement and figure 3b is the measurement with the resonator circuit inside the grounded enclosure.

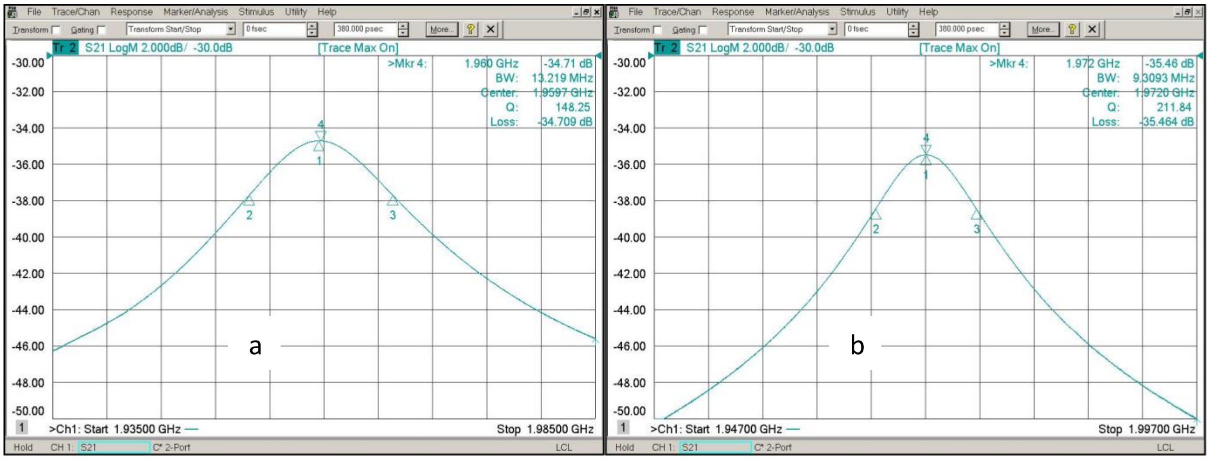

Figure 3. Two screen-shots of the same microstrip gap coupled resonator using a substrate with a Dk of 4.5 (a) initial measurement and (b) is the measurement after the circuit is tested within a grounded enclosure.

There are a couple items to consider in figure 3. The Q values are different and the center frequency is different. Again this is the same physical resonator circuit, however being tested with and without a grounded enclosure. The impact of radiation loss is apparent by the difference in Q.

Using a lower microwave frequency for the measurement, allows more accurate determination of the conductor loss. At 2 GHz the copper roughness effects are well defined by the use of Hammerstad and Jensen formulas combined with the Morgan multiplier. Additionally the Tanδ term (dissipation factor or Df) necessary to solve for the dielectric loss was measured on the same piece of material used to make the resonator circuit, at ≅ 2.5 GHz and per IPC-TM-650 2.5.5.5c[7] test method. The total loss of the resonator circuit was determined to be 0.270 dB and the component attributed to radiation loss was 0.081 dB.

The same experiment was repeated with another material in the same product family but with a much higher Dk value. That material was TMM13i with a Dk value of 12.2 and screen-shots of the measured resonator are given in figure 4.

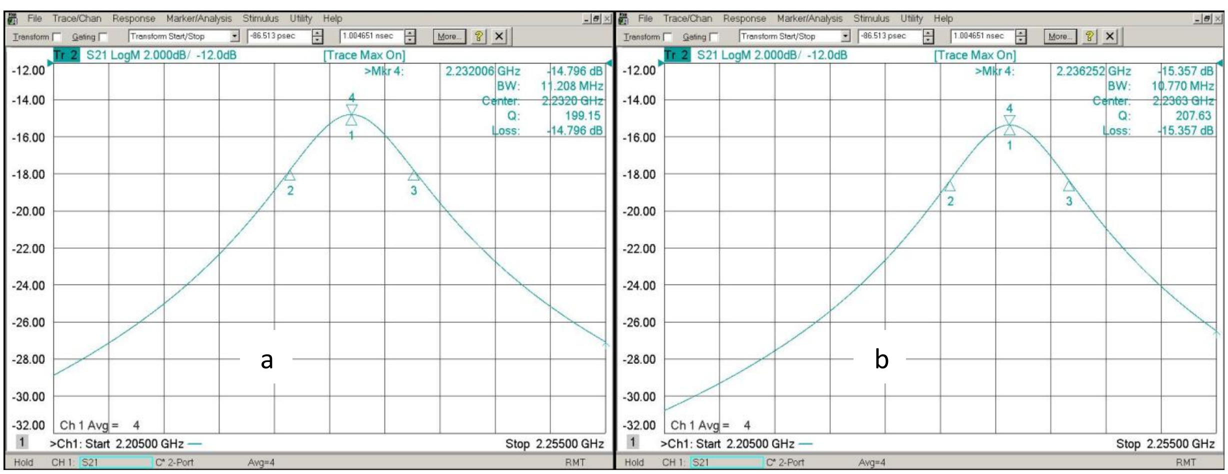

Figure 4. Two screen-shots of the same microstrip gap coupled resonator using a substrate with a Dk of 12.2 (a) initial measurement and (b) is the measurement after the circuit is tested within a grounded enclosure.

A remarkable difference is seen in figure 4 when testing a resonator circuit using higher Dk material as compared to figure 3 which used lower Dk material. Some points to consider in figure 4 with the resonator using the high Dk material, is a minimal shift in the power level shown on the y-axis, a slight change in the center frequency and much less difference for the Q value of the resonator circuit when tested with and without the grounded enclosure. The loss calculations were done in the same manner as explained for the data shown in figure 3 and the total loss was 0.168 dB and the loss associated with the radiation effects was 0.005 dB.

The difference in radiation loss between the resonator using the low and high Dk material is significant, however the numbers may not be intuitive to those working in the PCB industry. As a reference only and one which may give better intuitive insight for the PCB technologist, is to consider the differences of Q as if it was calculated as dissipation factor or Tanδ; this would equate to a dissipation factor difference of 0.0018 for the low Dk circuit when tested with and without the enclosure and a Df difference of 0.0002 for the high Dk circuit. Radiation loss is one reason that microwave material characterization needs to be carefully done when using a microstrip circuit that is not enclosed, because faulty values for Df can be reported.

Great care was taken to ensure the resonator circuits were as similar as possible. The circuits used the same signal launch approach, which was a coplanar-launched microstrip and used the same connectors. Additionally, much effort had to be put into the grounded enclosure in order to ensure valid loss measurements.

Coplanar-launched microstrip transmission line circuits were tested with and without the enclosure initially to evaluate how the enclosure behaved over a wider band of frequencies. The initial enclosure design was found to have spurious wave propagation modes due to "waveguide modes". The original enclosure allowed a waveguide mode to propagate and that interfered with the desired TEM mode of the transmission line. The interference increased the loss and also caused very noisy insertion loss curves. After modifying the enclosure, the radiation losses had an acceptable agreement with the models. Figure 5 is the insertion loss curves for transmission line circuits using the lower Dk material and showing the difference of the initial fixture design before and after being modified.

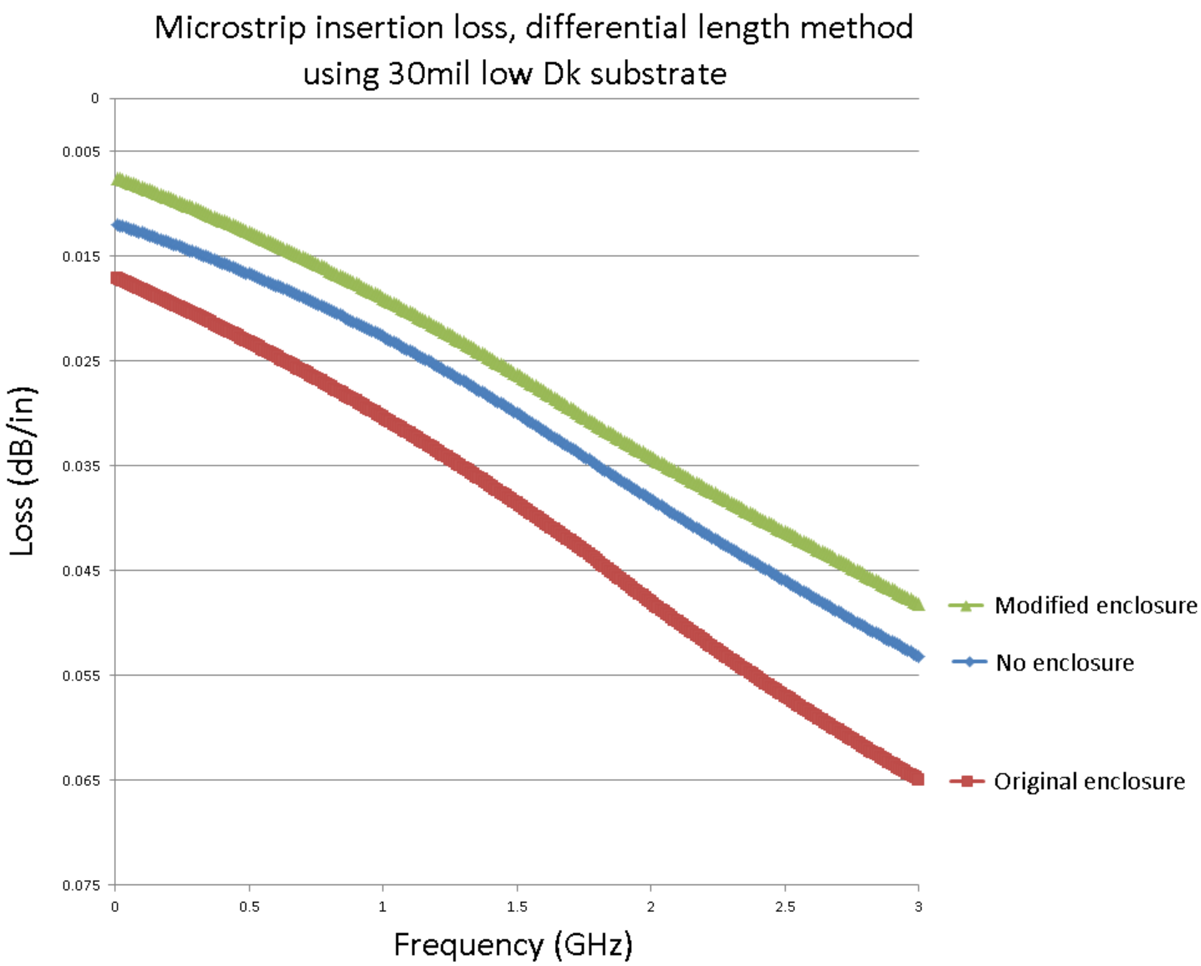

Figure 5. Microstrip insertion loss of transmission line circuits using low Dk materials, with no enclosure, with enclosure of the original design and the modified enclosure.

Due to poor signal launch of the design and in an effort to not confuse that issue with an interaction of radiation loss, higher frequency data beyond 3 GHz was not obtained in this experiment. The difference in the insertion loss curves which is related to radiation loss can be seen in the comparison of the curves labeled "No enclosure" and "Modified enclosure". The curve with

"No enclosure" is open and prone to radiation loss whereas the "Modified enclosure" does not allow radiation loss to escape the system. The circuit still radiates energy inside the enclosure, however, the radiated energy is shunt to ground and is part of the ground return path. Since this ground return path of the enclosure may not be as pure as the ground plane of the circuit, there can be more conductor losses associated with the enclosure. This study did not attempt to account for these potential losses. Figure 6 shows a picture of the resonator circuit using the high Dk substrate along with the lid of the enclosure.

Figure 6. Resonator circuit (a) using high Dk material shown, with the original lid design (b) of the enclosure and (c) the modified lid.

The lid of the enclosure in figure 6 is prior to the modification. During testing the resonator circuit was flipped from its view in figure 6 so the circuit side was down and inside the recessed area of the lid. Additionally, conductive electrical tape was used to seal the edges of the circuit around the lid and finally the base plate (not shown) was attached as well. This set-up looked like a metal box with the threads of the RF connectors protruding out of each end of the box.

The enclosure was modified by adding randomly oriented raised ridges on the inside of the lid. This alters the waveguide effect so the waveguide modes are not allowed to propagate. The addition of the ridges made a significant improvement on reducing the noise of the wideband insertion loss curves for the transmission line circuits.

The following experiment shows the difference that radiation loss contributes to insertion loss when considering good and poor signal launch. Using a single copper clad laminate, four circuits were fabricated. There were 2 sets of circuits and each set had a long and short length microstrip transmission line. Within each set, the long and short transmission lines were exactly the same in every detail, except for the length. The known difference in length allows for a test method where insertion loss of the actual circuits can be determined without the effects of the connectors and signal launch. This method is defined in a paper regarding different microwave test methods[8]. Most of the effects of the connectors and signal launch are subtracted out of the reported insertion loss, however, it is unlikely that all of the effects are completely removed.

The point of the following experiment is to show that even though the majority of the effects of the signal launch are removed with the differential length method, the radiation losses associated with the signal launch remain. Figure 7 shows two sets of circuits which were made on the same low Dk material, however one set had good signal launch and the other was relatively poor.

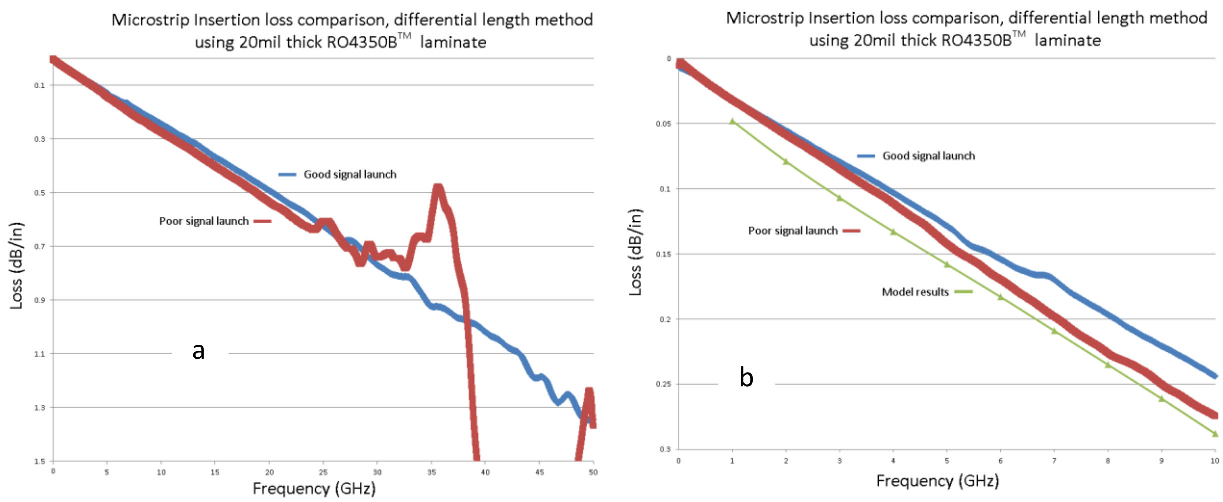

Figure 7. Two sets of circuits were made from the same sheet of copper clad laminate, with the only difference being the signal launch design. Insertion loss curves shown with both sets across a (a) wide range of frequencies and (b) a lower range of frequencies.

In figure 7a it can be seen that at about 25 GHz the set of circuits with the poor signal launch start to have noise in the insertion loss curve and it gets worse as the frequency increases. This behavior is typical for a 20mil thick microstrip transmission line with a relatively low Dk and the noise is due to spurious parasitic modes interfering with the desired quasi-TEM mode of the circuit. The same circuit and material, but with a good signal launch, has much smoother insertion loss curve out to 50 GHz. The better signal launch minimizes the severity of the spurious modes, however there is something else to consider. If an engineer were to evaluate figure 7a and determine to cut-off the data at 10 GHz as shown in figure 7b, the data may be misrepresented since the insertion loss curves are relatively well behaved at this range of frequencies. Although the noise due to the poor signal launch is not seen in figure 7b, the impact due to the poor signal launch does affect the insertion loss curve at these lower frequencies and causes the circuit with the poor signal launch to have more loss. Again, these circuits used the very same sheet of material and were made within inches of each other so the material properties are assumed to be nearly identical. A difference in the insertion loss curves may be wrongly assumed to be a material related issue. As an additional side note, poor return loss can also cause the insertion loss to be worse. However, for the circuits tested in figure 7, all of them had good return loss (better than 15 dB) out to 20 GHz or more, so the range of concern up to 10 GHz shown in figure 7b is not related to poor return loss.

If an engineer were to use the data shown in figure 7b to back calculate the dissipation factor without considering radiation loss, the results would have an error. The difference of loss at 10 GHz of 0.242 dB/in. for the good signal launch circuits compared to 0.274 dB/in. of the poor signal launch circuits will have a Df difference of 0.0009. Considering the material used in figure 7 had a Df value tested at 10 GHz of 0.0034 using the IPC-TM-650 2.5.5.5c method, the assumed and erroneous Df could be as high as 0.0043.

There is another important issue to consider when trying to back calculate the Df from transmission line insertion loss besides radiation loss and signal launch and that is the impact of copper surface roughness. There are very few models which have the capability to accurately report the effects of copper surface roughness on insertion loss over a wide range of frequencies. Additionally, all copper types have a natural variation of copper surface roughness and this can vary from one circuit to another, when using the same material. Without knowing the actual copper surface roughness of a circuit, this can create a significant variable especially on thinner laminates such as 10mils thick or less. In general it is not considered a good practice to attempt extraction of the materials' Df by microstrip insertion loss measurements due to the many variables associated with radiation loss, signal launch and copper surface roughness.

Lastly, the green curve shown in figure 7b is a model generated by a software called MWI-2010 which is available for free download from Rogers Corporation. This software uses the equations from the Hammerstad and Jensen paper as well as the Morgan paper previously mentioned. An additional routine was added to this software to include equations 2, 3 and 4 from this paper for evaluation of radiation losses. It can be seen in figure 7b that the output of this model is somewhat conservative, however relatively accurate.

In summary, there are many variables which can impact radiation loss and only a few of the more easily determined aspects were evaluated in this paper. Equations 2, 3 and 4 have been found to be relatively accurate for predicting radiation loss for microstrip circuitry and at microwave frequencies. However, they cannot account for several real life scenarios. The equations are simple enough to enter into a spreadsheet and use as a quick reference in concert with a field solver or MWI-2010 which gives effective dielectric constant. In cases where equations 2, 3 and 4 were found inaccurate, they have been shown to overestimate the amount of radiation loss which is a conservative error. As an example, throughout this study most results were within 5% or better between the measured value and the radiation loss predictions, with more error found at mmWave frequencies.

[1] John Coonrod, "Insertion Loss Comparisons of Common High Frequency PCB Constructions", IPC APEX EXPO 2013, Feb. 2013

[2] Abouzahra, Mohammad Deb, and Leonard Lewin, "Radiation from Microstrip Discontinuities", *IEEE Transactions on Microwave Theory and Techniques*, Vol. MTT-27, No. 8, August 1979, pp. 722-723.

[3] John Coonrod, "Electrical Challenges for PCB Millimeter-Wave Applications", PCB West 2013 Conference and Exhibition, Sept. 2013

[4] B. Rosas, "Optimizing Test Boards for 50 GHz End Launch Connectors: Grounded Coplanar Launches and Through Lines on 30-mil Rogers RO4350B with Comparison to Microstrip," Southwest Microwave, Inc., Tempe, AZ, 2007, www.southwestmicrowave.com.

[5] E. Hammerstad and O. Jensen, "Accurate models of microstrip computer aided design", 1980 MTT-S Int. Microwave Symp. Dig., May 1980, pp. 407-409.

[6] S. P. Morgan, "Effect of surface roughness on eddy current losses at microwave frequencies," *J. Applied Physics*, vol. 20, pp. 352–362, Apr. 1949.

[7] IPC-TM-650 Standard Test Methods, http://www.ipc.org

[8] John Coonrod, "Understanding the Variables of Dielectric Constant for PCB Materials used at Microwave Frequencies", European Microwave Week 2011, Oct. 2011.