CONTROLLING RESONANCES IN PCB-CHASSIS STRUCTURES

Tim Williams

Elmac Services, PO Box 111, Chichester, UK PO19 5ZS

ABSTRACT

Many electronics products are built using printed circuit boards (PCBs) bolted to a metal chassis. The PCB is typically fabricated with a ground plane which is electrically connected through pillars to the chassis. This structure creates a shorted transmission line which exhibits a half wave resonant frequency according to its major dimensions. This paper discusses the theoretical aspects of RF coupling to this structure for emissions and immunity, and then presents experimental results which show actual resonant amplification of emissions followed by an evaluation of proposals to minimise the resonance. This leads to specific design recommendations that can be applied directly by product designers.

INTRODUCTION

Many electronics products are built using printed circuit boards (PCBs) bolted to a metal chassis. High performance PCBs are fabricated with a ground plane which is connected either directly or via capacitors through the mounting pillars to the chassis. At frequencies below structural resonance this creates an excellent low transfer impedance assembly which contributes substantially to good EMC performance.

However, at the same time this forms a shorted transmission line which exhibits a half wave resonant frequency according to its major dimensions. At this frequency, and at higher order multiples, any circuit currents couple to the structure efficiently and radiated emissions (and susceptibility) are at a maximum.

A similar effect occurs on the PCB itself, in the transmission line formed by the power and ground planes on a multilayer board. Considerable attention has been paid to modelling and describing this phenomenon [3] – [7] together with investigating ways of dealing with it. There is less attention given to the equivalent effect at chassis level, perhaps because this is seen as the province of mechanical designers rather than relating to the electronics of the product. Nevertheless there can be a severe coupling problem around the principal resonant frequency, manifesting as worsened emissions or immunity as this frequency is approached, but this problem is by no means intractable and the practical solution is surprisingly simple.

RESONANT FREQUENCY OF A SHORTED TRANSMISSION LINE

The theory governing transmission line structures is well established. A pair of plates with air dielectric and the cross-sectional geometry shown in Figure 1 will exhibit a characteristic impedance given by [1]

$$Z_0 = 377 \cdot \frac{h}{W} \cdot \frac{1}{(1+(h/\pi \cdot W)) \cdot (1+\ln(2\pi\, W/h))} \qquad (1)$$

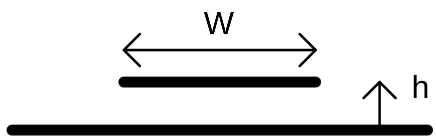

Figure 1 Parallel plate cross section

If the bottom plate is taken to be the chassis and the other the PCB, dominated by its ground plane(s), the impedance of this structure is determined by the separation height between the two and the width of the PCB.

As it stands the resonant frequencies of this structure could be derived from the simple expression for the resonance of a cavity of length L and width W metres, where m and n are integers representing the modes [7]:

$$F = 150 \cdot \sqrt{\left(\frac{m}{L}\right)^2 + \left(\frac{n}{W}\right)^2} \qquad (2)$$

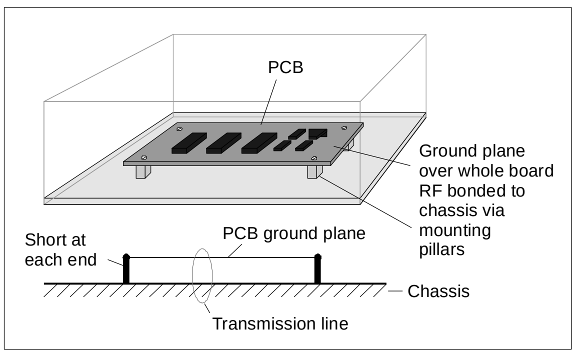

Figure 2 Transmission line equivalent circuit

But when the ground plane is bonded to the chassis plate at the corners of the board, this has the effect of applying a near short circuit to each corner of the line, as shown in Figure 2. As a first approximation, neglecting the inductances and the relative position of the bonding pillars, it might be reasonable to suppose that the first major resonance of the assembly would be related to the corner-to-corner distance across the plate:

$$F = 150 \cdot \sqrt{\frac{1}{L^2 + W^2}} \qquad (3)$$

Applying this equation to the actual measurements described later gives a good approximation, within 10% for the geometries listed, to the measured resonant frequency, and because of its simplicity is recommended for preliminary assessment. But it does not give an accurate answer and the deviations from the measured values are not consistent for different geometries. A true prediction of the resonance must factor in the inductance of all the pillars and their position with respect to the PCB boundaries. This is a job for numerical modelling rather than an analytical expression.

In the context of modelling decoupling capacitors on power/ground plane structures, it has been proposed [5],[6] that the structure is modelled as a mesh of interlinked x-y transmission lines with the capacitors placed as L-C-R elements at each appropriate node in the mesh, depending on their physical position. Exactly the same approach could be taken here, with the bonding pillars shown as inductances placed within the mesh of elemental transmission lines. This method would give the impedance at each nodal position and therefore the most significant resonant frequencies. Future work along these lines may be reported in due course.

Any RF source within the PCB ground plane – that is, any operating circuit on the PCB – has the potential to couple with the common mode circuit offered by the transmission line. Maximum coupling with current sources will occur at low impedance nodes, for instance near the bonding pillars, while maximum voltage coupling will occur at high impedance nodes such as in the centre, away from the pillars.

This description of the electrical consequences of the structure suggests a number of possible ways to reduce its impact. These can be summarised as

- increase the number of bonds and reduce their separation distance, to raise the resonant frequency but without removing the resonance
- load the bonds with lossy components, to damp and if possible eliminate the resonance

The rest of this paper discusses some results of experimental investigations of these approaches.

EXPERIMENTAL SETUP

In order to investigate the emissions profile of a PCB/chassis structure the assembly shown in Figure 3 was constructed. The larger plate formed the chassis reference and the smaller was excited at the emissions frequency. The excitation was provided by a tracking generator at a constant level so that a direct and continuous plot could be obtained of emissions versus frequency. Since the impedance of the feed to the excited plate could vary widely over the frequency range, the feed was taken via a 10dB attenuator pad to minimise the standing waves on the feed cable and consequent ripple on the emissions pattern.

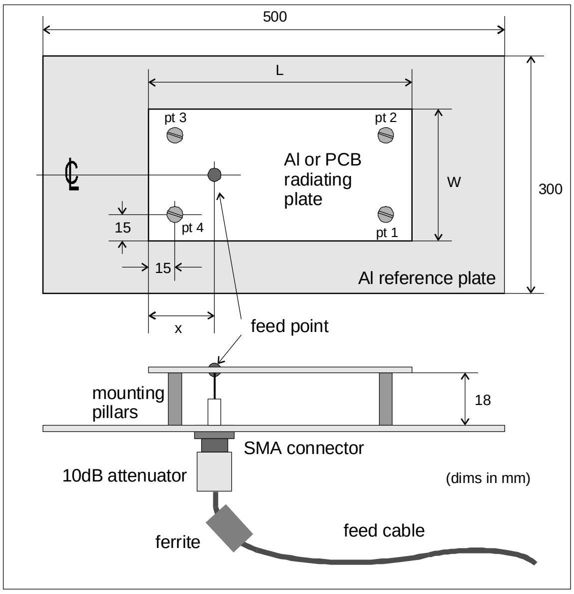

Figure 3 Test structure

Excitation was provided in general by connecting the outer of the coax feed to the reference plate and the inner, after the 10dB attenuator, directly to the radiating plate. The feed amplitude into the attenuator was set at 100dBμV. The feed point could be positioned towards one end of the plate, between two mounting pillars, or in the exact centre of the plate. In practice when the feed was applied to the centre of the plate the line resonance was substantially damped by the 50Ω attenuator impedance and the resonant characteristic was not seen. Results are not reported for this condition, although it is inherently a good example of the damping approach described later.

When the feed was taken close to the injection point but not actually connected, i.e. forming a coupling stub of about 15mm height, emissions were below the noise floor except very near to the resonant frequency. This therefore allowed the resonant frequency of the structure to be determined accurately with no loading effects from the feed. It also demonstrates the effect of localised coupling: if instead of the signal injected via the SMA connector, we visualise a disturbance source on the PCB with a high dV/dt (such as a microprocessor package), it is easy to see that the structural resonance shown in Figure 6 will cause substantial coupling at the resonant frequency.

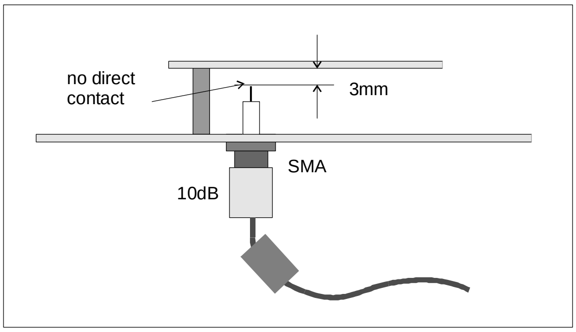

Figure 4 Indirect coupling

This assembly was measured with a BiLog antenna on an open site at 3m distance with no ground plane and no height scan. Maximum E-field emissions as expected were found with the assembly flat and with the major dimension edge-on to the antenna in vertical polarization.

Repeatability of the measurements was established by taking five consecutive sweeps with minor layout differences introduced between sweeps. The standard deviations of the five sweeps demonstrated the reliability of each result. In all cases reported here, the standard deviations around the principal resonance were low, although at frequencies above this – say greater than 600MHz – more substantial variability was seen, attributable partly to cable effects and partly to site imperfections.

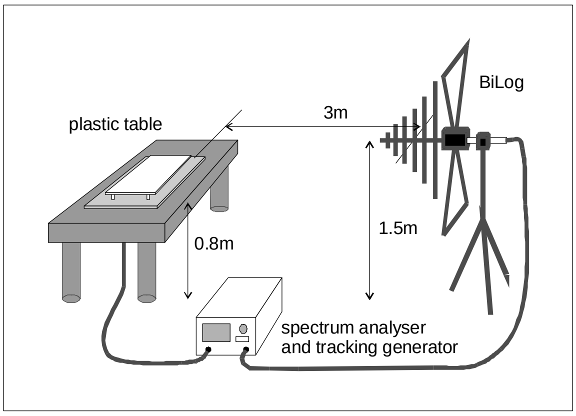

Figure 5 Measurement setup

EXPERIMENTAL RESULTS

The following graphs show selected results from the above setup and for the different sized radiating plates as given in Table 1. Lines A-E are aluminium plates and Line F is a double-sided copper clad PCB. The calculated resonant frequency is taken from equation (3), and the distance x refers to Figure 3.

Reference	**Length mm**	**Width mm**	**x mm**	**Calc F_{res} MHz**
Line A	300	50	15	493.20
Line B	450	100	90	325.40
Line C feed A	300	100	15	474.34
Line C feed B			85	
Line D	300	200	15	416.03
Line E	450	200	90	304.60
Line F	220	100	10	620.70

Table 1 Plate dimensions

Shorted lines: corner pillars only

This configuration used four metal mounting pillars in each corner with no other connections. The first plot shows the results of indirect injection according to Figure 4. Calculated resonant frequencies are in general higher than actually observed, suggesting that mounting pillar inductance and loading is causing a downward shift. The resonant amplification is clearly evident.

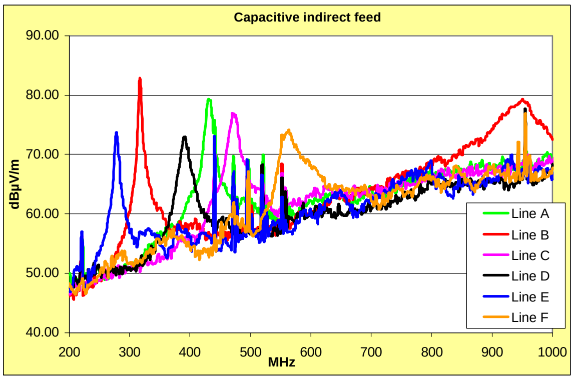

Figure 6 Indirect connection as per Figure 4

Measurements on the same plates with a direct connection from the 10dB attenuator to the plate (Figure 7) showed resonant peaks at similar frequencies but much broader in bandwidth due to loading, and, of course, at a higher overall level. These peaks in the 250–600MHz range are directly related to the PCB size and aspect ratio and fall exactly in the range at which emissions from high-speed digital circuits are significant. The loading effect and consequent widening in bandwidth is greater the further in from the end of the plate is the connection. It is noticeable that the amplitude of the major peak is almost independent of the size of the plate. In fact, coupling to transmission lines is determined principally by the separation distance of the plates rather than by their length [2]; the distance used in the experiments was 18mm, a wider separation would give a higher amplitude and vice versa.

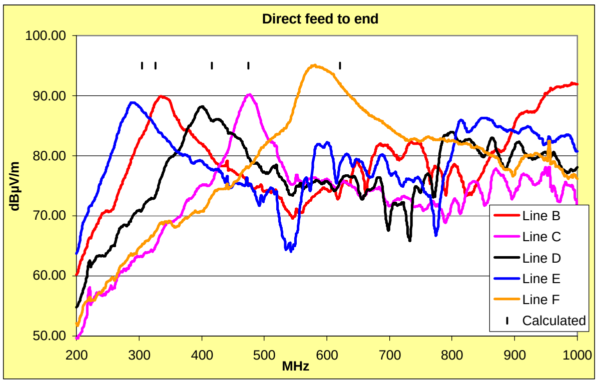

Figure 7 Direct connection

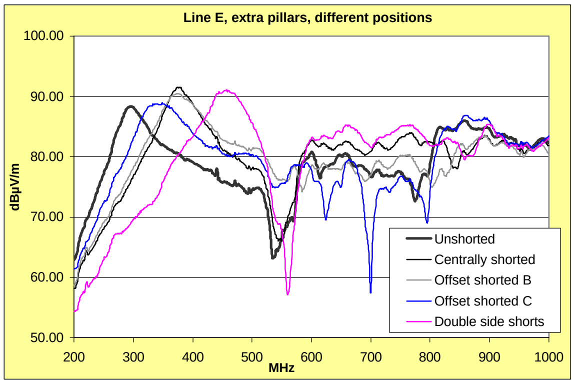

Figure 9 Extra pillars – Line E

Shorted lines: adding pillars

One method of changing the resonance would be to include extra mounting pillars at places within the structure other than the corners. This would modify the transmission line structure and would be bound to increase the resonant frequency. This is not necessarily because the half-wave distance between shorting points is now less; since the pillars have appreciable inductance their impact on the resonance is not straightforward. However, experimental results of adding one or more pillars to two of the lines characterised earlier shows that although the frequency shift does indeed happen, the peak amplitude of the emission level is not affected – it is merely shifted in frequency. This is sometimes known as the "waterbed" effect – a change improves coupling at one frequency while worsening it at another.

The following figures show the effect of adding extra pillars on lines C and E. The positions referred to in the graphs are illustrated in Figure 3. Line C was measured with two feed positions, feed A being at one end, feed B being inboard by 8.5cm. The loading effect (increasing the bandwidth of the resonance) of the inboard location is clearly visible, both with and without the shorting pillars.

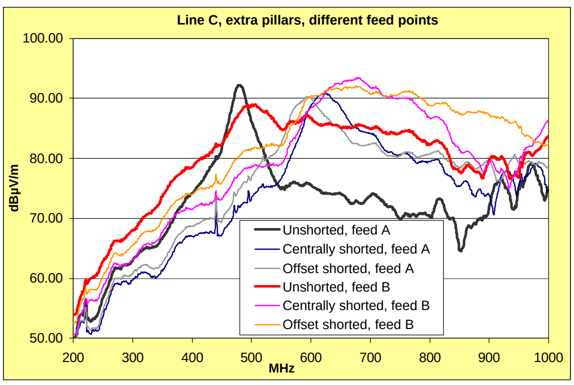

Figure 8 Extra pillars – Line C

Loaded lines

Instead of attempting to short the transmission line formed by the PCB and the chassis, by merely adding mounting pillars – which as we have seen is not particularly successful – the connections could be replaced by resistive loading. This should damp the line rather than allow it to resonate. If the damping value equals the line Z_0, the line should be perfectly matched and no resonance at all should exist.

The concept of damping is hardly new. Relatively recently, it has been considered for general transmission line and cavity situations [8] and also for decoupling on PCB ground/power plane structures [6],[7]. There is nothing particularly radical in extending it to a PCB/chassis structure.

The effect of resistance loading at all four corners, so that the line is effectively ungrounded, on line F is shown in Figure 10. The lengthwise Z_0 of this particular structure is given by (1) as 54Ω. Thus two 100Ω resistors at each end should match the line reasonably well. In fact it can be seen that while 100Ω is marginally better, there is little difference in effect between 47Ω and 100Ω, suggesting that the actual value is not critical.

This damping configuration clearly reduces the peak emission by about 15dB, which is evidently a worthwhile gain, but at the expense of raising emission levels either side of resonance. This is to be expected since the radiating plate now carries RF voltage along its entire length. This may be acceptable in some situations where only the emissions at the resonant peak are troublesome. On the other hand, there will undoubtedly be cases where such an increase cannot be tolerated. There may also be functional or safety reasons to avoid a totally ungrounded configuration and ensure that at least one point is directly, not resistively bonded to the chassis.

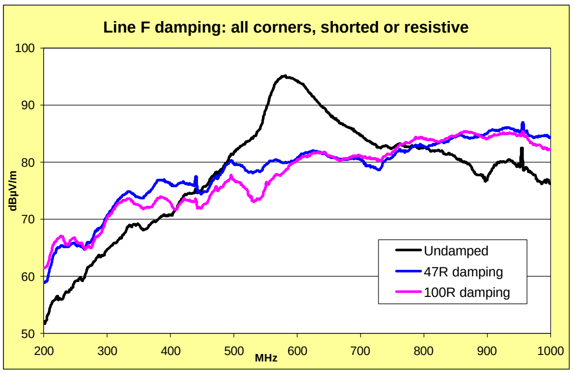

Figure 10 Resistive loading on all four corners

This therefore raises the question of what is the impact of having one or more of the corners directly grounded, and the others resistively damped. This impact was investigated for Line F with resistors of 47Ω as shown in Figure 11.

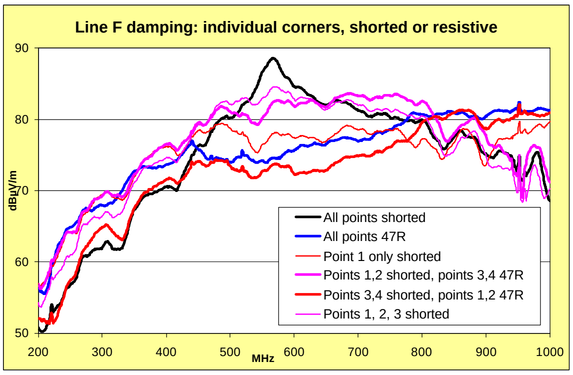

Figure 11 Resistive loading on individual corners

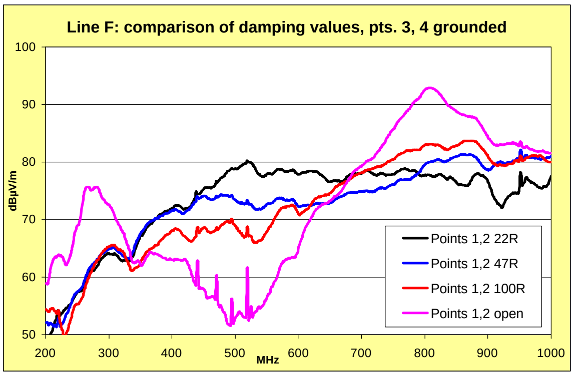

Figure 12 Effect of far end resistance value

These graphs should be read in conjunction with Figure 3 which shows the corner points in relation to the feed point. The best situation, slightly better in fact than having resistors on all corners, is that with points 3 and 4 shorted and points 1 and 2 resistively damped. Points 3 and 4 are nearest to the feed and points 1 and 2 are furthest. If the situation is reversed, so that the points furthest from the feed are shorted, then the overall coupling is around 10dB higher over most of the frequency range, although the highest resonant peak is still reduced.

Figure 12 shows the effect of varying the resistor value at the far end to the short, from 22Ω up to an open circuit. The curve for an open circuit clearly shows the transformation of the line from a half wave to a quarter wave resonant structure, with peaks now at around 270 and 800Mhz. Although the 580MHz peak has been comprehensively demolished, this has been achieved at the cost of substantial problems elsewhere in the spectrum. A load between 47 and 100Ω is still optimum for the majority of the spectrum region.

DISCUSSION

This experimental work has concentrated on the radiated coupling from a metal plate artificially excited by an RF connection at a particular location on the plate. A real PCB is of course more complicated than this. It will probably not be perfectly rectangular, and the bonding pillars will be located according to many criteria both mechanical and electrical, and not necessarily at the corners. Areas of particular noise generating potential or sensitivity will be located around the board, not just at one point, their position mainly determined by functional considerations.

These issues make it hard to apply generic rules for positioning and damping of bonding pillars to give minimum transmission line resonant coupling. Yet it is clear that correct application of such rules would give a highly worthwhile improvement in the EMC of many products. From the investigation reported here, the most that can be advised would be

- directly bond the PCB ground plane to the chassis at one or more locations nearest to the worst interference source or most susceptible circuit
- apply resistively loaded bonds at other locations, sizing the resistor values so that the total (parallel) resistance roughly matches the calculated transmission line Z_0

But this advice is quite likely to conflict with other EMC or functional requirements, for instance the absolute requirement for direct bonding at the interfaces.

In fact modelling the ground plane/chassis structure as a mesh of transmission lines, and iterating the placement of bonding pillars, resistive damping and critical noise sources for minimum resonant impedance, seems likely to offer a more robust way of designing these layout aspects. Developing an automated algorithm for this process would be a useful contribution to the EMC design toolbox.

CONCLUSIONS AND RECOMMENDATIONS

This paper has shown that resonant coupling from a PCB-chassis structure exists, can be predicted, and more importantly can be controlled. A product designer faced with either an emissions or immunity problem at or around a particular frequency can easily identify whether this frequency corresponds to the major transmission line resonance of a given PCB-chassis assembly.

If it does, the resonance can either:

- be shifted in frequency, by manipulating the positioning of extra mounting pillars, to a frequency at which the circuit is expected to be benign; or,
- be damped, by loading the pillar connections according to the geometry of the structure.

Either or both of these practices can be implemented at the product design stage as a matter of standard procedure, when laying out the PCB.

REFERENCES

[1] EMC: Electromagnetic Theory to Practical Design, P A Chatterton & M A Houlden, Wiley, 1992, Annex E

[2] Coupling of external electromagnetic fields to transmission lines, A A Smith, Wiley 1977

[3] Study of the Ground Bounce caused by Power Plane Resonances, S Van den Berghe et al, IEEE Trans EMC, vol 40 no 2, May 1998

[4] Power Bus Decoupling on Multilayer Circuit Boards, T H Hubing et al, IEEE Trans EMC, vol 37 no 2, May 1995

[5] Developing a Decoupling Methodology with SPICE for Multilayer Printed Circuit Boards, C B O'Sullivan et al, IEEE EMC 1998, Denver, August 24-28 1998, pp 652-655

[6] Reducing Simultaneous Switching Noise and EMI on Ground/Power Planes by Dissipative Edge Termination, I Novak, IEEE Trans. AP Vol 22 No 3, August 1999 pp 274-283

[7] Analysis on the Power/Ground Plane Resonance Damping and Radiated Field Reduction with the RC Termination, S Fujio et al, IEEE EMC 2001, Montreal, August 13-17 2001

[8] Damping Transmission-Line and Cavity Resonances, C E Baum & D P McLemore, 12th Intl Symp EMC, Zurich, February 18-20 1997, pp 239-244 paper 45H4

This paper was first published in the proceedings of EMC EUROPE 2002, International Symposium on EMC, September 9-13, 2002, Sorrento, Italy pp 305-310