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AND8200/D

Design Considerations for ESD/EMI Filters: I

(Almost Everything You Wanted to Know About EMI Filters and Were Afraid to Ask)

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APPLICATION NOTE

Background

With data transfer rates and clock frequencies ever increasing and encroaching on radio frequencies, the need for EMI filters has increased dramatically. At the same time, the portability of electronic devices such as PDAs, laptop computers, MP3 players and so on, the need to protect those devices from ESD has also increased because use of peripheral interfaces. Some time ago these two functions were independent of each other with discrete components making the EMI filter and a separate device to provide the ESD protection. As consumer electronics continue to get smaller, the ESD and EMI filter functions were gradually integrated together.

For ESD protection, a TVS diode would typically be used. TVS diodes are nothing more than Zener diodes that have been modified to handle large currents for very brief periods of time while “clamping” at some voltage slightly over the peak operating voltage. A TVS diode with a clamping voltage of 6.0 V would be used in applications where the maximum operating voltage was 3.3 V. There is a limit however, to the TVS’s ability to clamp. That is a given TVS can only dissipate so much power in those short periods of time. A way to improve this is to increase the size of the TVS. A larger device can handle more current than a similar but smaller device, but at a cost, more capacitance.

EMI filters have in the past been RC or LC pi networks using individual surface mount components. One of the main problems with using discrete components is that there are parasitics that can negatively affect the performance of the filter. Overall, most commonly used EMI filters for portable applications take a form similar to that of Figure 1.

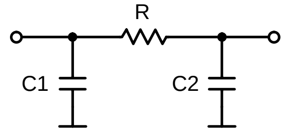

Figure 1. Common RC “Pi” Filter

These filters are typically low pass filters that reject signals over 800 MHz. The component values for the filter are typically determined by how much attenuation is needed in the rejection band (800 MHz to 2.5 GHz for example), the source and load impedances, and how much the desired signal can be attenuated without losing data.

In applications where both ESD and EMI filtering are required, it is advantageous to choose a single integrated ESD/EMI filter that replaces a number of discrete surface mount components. Where board space is a premium, a single integrated component is hard to beat. If cost is an issue, a solution composed of a number of discrete surface mount components cannot compete with a single integrated ESD/EMI filter. If it is decided that a single integrated ESD/EMI filter is to be used, a number of questions need to be answered.

How much ESD protection is required?
What are the EMI filtering requirements?
What is the acceptable Group Delay?
How much insertion loss can be tolerated?
What type of package should be used?
What else?

ESD Protection

How much ESD protection is needed? As much as possible would be the correct answer. Having too little ESD has the obvious consequences of inviting damage to sensitive components that were meant to be protected. Is there such a thing as too much ESD protection? For high speed data lines and high frequency lines the answer is “yes”. The component that does the work for ESD protection is the TVS or Transient Voltage Suppressor. One of the ways to measure a TVS’s ability to suppress ESD is in the form of the IEC 61000−4−2. Level 4 is required for accessible contacts on cell phones. For further details refer to AND8074/D.

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Table 1. IEC 61000−4−2 Ratings for ESD Protection by Contact Discharge and Air Discharge for Various Voltage Levels

1a − Contact Discharge		1b − Air Discharge	
Level	**Test Voltage (kV)**	**Level**	**Test Voltage (kV)**
1	2	1	2
2	4	2	4
3	6	3	8
4	8	4	15
X*	Special	X*	Special

*"X" is an open level. The level has to be specified in the dedicated equipment specification. If higher voltages than those shown are specified, special test equipment may be needed.

On an integrated circuit a TVS would be in the form of a special Zener diode. The breakdown voltage of the Zener diode (BV_Z) would be determined by the breakdown voltage of the IC (BV_{IC}) being protected. In applications where a single TVS diode is used for ESD and EMI protection, BV_Z needs to be less than BV_{IC}. The limiting factor comes to be the current handling/power dissipation capability of the diode. Smaller diodes can not handle as much instantaneous current as a larger diode. A TVS's ability to protect against ESD is, for the most part, a function of how much instantaneous power can be dropped across the diode. To get more ESD protection from these Zener diodes, the physical size of the diode has to increase. The result of increasing the size of the diode is that the junction capacitance also increases. The dual nature of the Zener diode is that it not only takes care of the TVS role, but it also behaves as a capacitor (Figure 2).

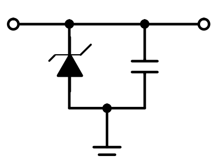

Figure 2. The Zener diodes also function as capacitors to ground.

The capacitance of the diode is a key component of the EMI filter used in the tuning of the filter which will be discussed later. Another consideration when determining the size of the TVS/Zener diode and its capacitance, is its effect on the Group Delay or Envelope Delay. For ESD protection it comes down to:

1. What level of ESD protection is needed?
2. What is the breakdown voltage of the circuit to be protected?
3. What effect will the capacitance of the diode have on the EMI filter?

The level of ESD protection needed will be determined greatly by the environment of the end function. Data lines that interface with external components such as USB or FireWire peripherals will need greater ESD protection than data lines that do not have an external interface. Data lines that have connectors to interface with external components quite often can face ESD contact discharges of a few thousand volts, enough to destroy sensitive electronics. The internal data lines also need a degree of ESD protection. Electronics that have moving parts; scanners, flip phones, etc., can generate enough internal charge to cause an ESD discharge.

Many high−speed data lines operate at 3.3 V. These systems however, typically can handle a little over two times this for very brief periods of time. So an acceptable breakdown voltage, (V_{BZ}), would be around 6.2 V to 7.2 V, as in the SM05C. For systems operating at 5.0 V, the acceptable breakdown would be 13.3 V to 15 V, as in the SM12C and so on. The Zener breakdown voltage should only be used as a reference as to what voltage the diode will clamp to when subject to ESD. For normal operating conditions, data signals should not have amplitude greater than what is called the Reverse Working Voltage Maximum (V_{RWM}). Figure 3 shows a typical IV curve response for a Zener diode. V_{RWM} is a voltage point where the diode is not considered to be conducting, but beyond it and before the breakdown voltage, enough current is conducting to cause a nonlinear response. By keeping the data signals below this point will prevent any substantial distortion to the data.

The capacitance of a TVS is determined by two main factors. The first, as discussed before, is the ESD rating. The more power a TVS has to dissipate, the larger the surface area of the diode, and therefore a larger capacitance. Also, if a larger breakdown voltage is desired, then a reduction in capacitance is to be expected. A TVS with a breakdown voltage of 14 V, and the same surface area as a diode with a 6.5 V breakdown voltage, will have a lower capacitance. The effect the diode capacitance has on the EMI filter is direct in terms of signal group delay which will be discussed later.

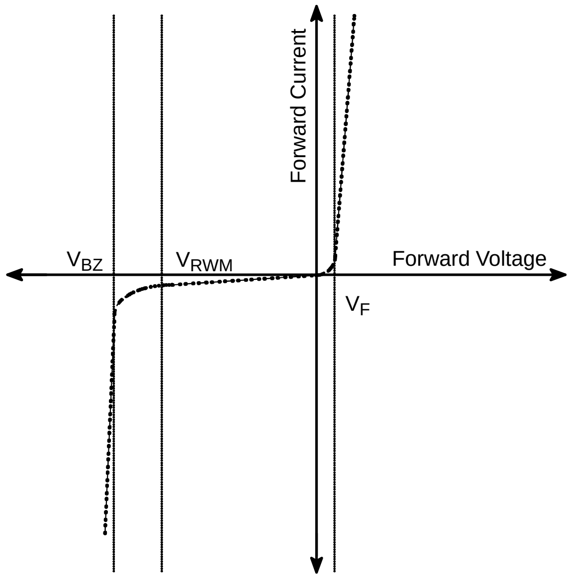

Figure 3. A typical IV characteristic of a Zener diode. Shown is the Zener breakdown voltage (V_{BZ}), the reverse working voltage maximum (V_{RWM}), and the Zener forward voltage.

EMI Filtering

The function of EMI (Electromagnetic Interference) filters is to attenuate or reject signals and harmonics that are generated by clock signals or data lines that may interfere with radio frequencies. The frequency range that typically needs safeguarded is usually 800 MHz to 2.4 GHz (Figure 4), sometimes higher or lower in frequency. The reason that this band of frequencies is important, is that this band of frequencies is the domain of most cell phones and many wireless networks. In many cases, the radio signal strength that cell phones receive is so weak that external "noise" can disrupt a call causing fading or even dropped calls. This noise is not only generated outside of the cell phone, but in many cases is generated within the cell phone itself, from the digital components as described above.

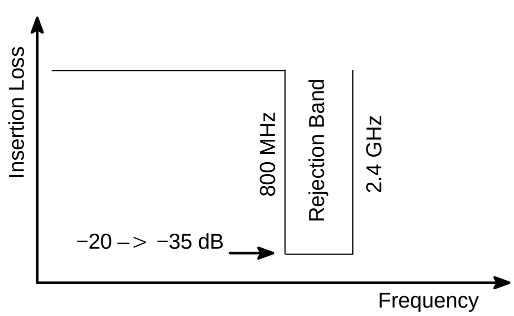

Figure 4. The amount of attenuation in the rejection band varies from application to application. The band of frequencies though, stays the same regardless of the application.

Before we can continue with discussing the filters, we need to define a few terms:

Insertion Loss (IL) is used here to describe the transmission coefficient between two points in a circuit often described in terms of dB. When examining S parameters, S_{21} is often described as insertion loss. Insertion Loss and S21 will be used interchangeably from here on out.

dB is typically used to describe a ratio of power levels. Where the insertion loss of a circuit with P_{OUT} and P_{IN} would be expressed as $IL = S_{21}(dB) = 10\log\left(\frac{P_{OUT}}{P_{IN}}\right)$ in dB or if V_{OUT} and V_{IN} are used with identical loads, insertion loss would be expressed as $IL = S_{21}(dB) = 20\log\left(\frac{V_{OUT}}{V_{IN}}\right)$ also in dB.

The **Pass Band** is the range of frequencies that are allowed to "pass" through a filter with minimal attenuation. For our purposes it starts from DC and ends at the cut off frequency.

Cut Off Frequency is the frequency at which the signal strength is 3.0 dB less than it is Pass Band. 3.0 dB of attenuation equates to half the original signal power.

The **Rejection Band** is the range of frequencies that are to be filtered out or attenuated to a desirable level. For low pass filters this range is all frequencies over the cut off frequency.

The term **Decade** is used from time to time with reference to frequency. A decade is a change in frequency by a factor of ten. For example, a change in frequency from 100 Hz to 1000 Hz is one decade. Any change in frequency, 10^n Hz to 10^{n+1} Hz, is a decade. This is used quite often on logarithmic scaling.

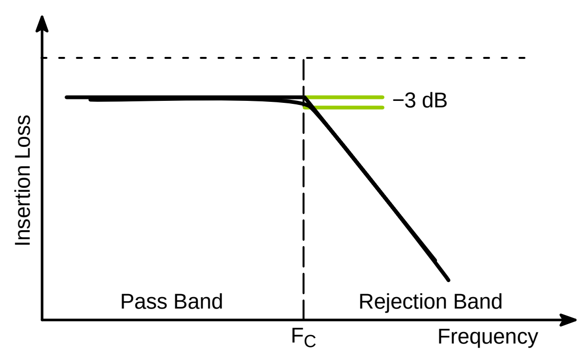

Figure 5. The insertion loss curve for an ideal low pass filter showing the pass band, rejection band, and where the cut off frequency (F_C) would be located.

For a given circuit, such as a filter, the response or "transfer function" can be described as an ABCD matrix function G(ω). For our purposes, this circuit, as in Figure 6, will be analyzed in a 50 Ω environment, so the source impedance (RS) and load impedance (RL) is 50 Ω. The transfer functions then can be analyzed in terms of insertion loss (S_{21}).

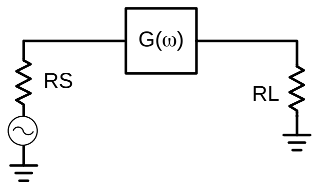

Figure 6. Any filter can be described as transfer function (G(ω)) in a circuit with a load RL and source with internal impedance of RS.

Low cost EMI filters typically come in one of three types. The first is a simple capacitor to ground (Figure 7). This topology relies upon the size of the capacitor and the source and load impedances.

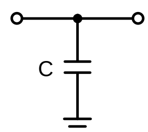

Figure 7. A capacitor to ground shown as a two−port network that can be described by a transfer function.

This is the most basic configuration. A single capacitor to ground is usually used in one of two cases. A large capacitor is used to filter out all but the lowest frequencies, or a very small capacitor to ground is used to filter only the highest frequencies while providing some level of ESD protection. The cut off frequency is determined by the capacitance and the network impedance.

The transfer function for a capacitor to ground can be expressed as $G_1(\omega)$ in Equation 1. The ABCD matrix takes into account the voltages and currents at each port. This transfer function only takes into account the capacitor to ground and nothing outside.

$$G_1(\omega) = \begin{vmatrix} 1 & 0 \\ j\omega C & 1 \end{vmatrix} \qquad \text{(eq. 1)}$$

The term "jω" is used to describe two things. Rather than using 2π, ω is used. For many the concept of $\sqrt{-1}$ or "j" is difficult to grasp. This denotes a difference in phase between the voltage applied to a reactive component such as a capacitor and the current in the capacitor. In time varying systems such as filters, this phase difference can play a critical role.

From this equation we are able to derive the insertion loss in the form of Equation 2. In this step the transfer function is inserted into a network (Figure 6) with source and load impedances. From this point the behavior of a capacitor to ground can be determined. From here on out Z_O (network impedance) will be 50 Ω. There are cases where Z_O may be smaller or larger.

$$S21_1(\omega) = \frac{1}{1 + j\omega\left(\frac{Z_O C}{2}\right)} \qquad \text{(eq. 2)}$$

Then there is the basic RC filter where there is a single series resistor with a single capacitor to ground as shown in Figure 8. This configuration is also a first order filter but has the added advantage allowing more control over determining the cut off frequency. The drawback to this is with the series resistor there will be loss in the pass band (the range of frequencies that are not to be filtered out).

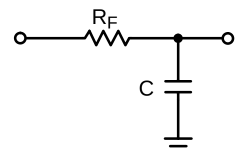

Figure 8. A basic first order RC filter. The series resistance allows for more control of the cut off frequency than the simple capacitor to ground.

The addition of the series resistor R_F adds a little more complexity to the transfer function $G_2(\omega)$. Even if you are not too familiar with these types of matrixes, it is apparent from Equation 3 that the addition of a single component can have substantial impact on the response. In this transfer function we can see the interaction of just R_F and the capacitor C.

$$G_2(\omega) = \begin{vmatrix} 1 + j\omega R_F C & R_F \\ j\omega C & 1 \end{vmatrix} \qquad \text{(eq. 3)}$$

As before the insertion loss is derived from the transfer function as shown in Equation 4. From here we can see the interaction of RF and C, but how the loads affect the response curve.

$$S21_2(\omega) = \frac{2}{\left(2 + \frac{R_F}{Z_O}\right)\left(1 + j\omega C\left(\frac{R_F + Z_O}{2 + \left(\frac{R_F}{Z_O}\right)}\right)\right)} \qquad \text{(eq. 4)}$$

Finally there is the most commonly integrated EMI filter, the RC "Pi" filter. Figure 9 shows the arrangement of the filter with the capacitor to ground, series resistor, then a second capacitor to ground. Again this filter will have some pass band−loss, but the advantage is this topology approaches that of a second order filter with a maximum rate of attenuation of 40 dB per decade starting at its cut off frequency. This allows for a higher cut off frequency, giving a wider pass band and more attenuation in the rejection band using the same amount of total capacitance and resistance as the first order RC filter.

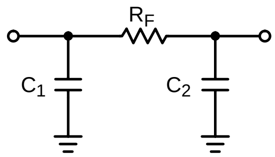

Figure 9. Standard Configuration for an RC "Pi" Filter

Because of the addition of a second capacitor, the realization of the transfer function becomes a bit more complicated. Equation 5 shows the interaction of all three components with each other.

$$G_3(\omega) = \begin{vmatrix} 1 + j\omega C_2 R_F & R_F \\ j\omega C_1 + j\omega C_2 + (j\omega C_1)(j\omega C_2)R_F & 1 + j\omega C_1 R_F \end{vmatrix} \quad \text{(eq. 5)}$$

Now to find the insertion loss for the for the Pi filter we look to Equation 6. Again the interaction of the filter changes when the filter is placed into a system. In effect the environment that the filter is placed changes the frequency response.

$$S21_3(\omega) = \frac{2}{\left(2 + \frac{R_F}{Z_o}\right)\left(1 + j\omega C_1\left(\frac{R_F + Z_o}{2 + \left(\frac{R_F}{Z_o}\right)}\right) + j\omega C_2\left(\frac{R_F + Z_o}{2 + \left(\frac{R_F}{Z_o}\right)}\right) - \omega^2 C_1 C_2\left(\frac{Z_o R_F}{R_F + Z_o}\right)\right)} \quad \text{(eq. 6)}$$

One could add more components to the filter in an attempt to improve the filtering response while maintaining a certain total resistance and capacitance in an attempt to improve filtering. For the most part there is no significant gain to be made by doing this. In this respect, this is one of the limitations of RC filters. Only under ideal circumstances can an RC "Pi" filter achieve a frequency response of a second order filter, regardless of the number of segments.

The significance of the "order" of a filter is the rate at which it attenuates over frequency. The rule is the greater the order, the higher the rate of attenuation. As a rule, there is 20 dB of attenuation per order. A first order filter will have an attenuation of 20 dB per decade after the cut off frequency. A second order filter will have an attenuation rate of 40 dB per decade after the cut off frequency. This continues with 3^{RD} order at 60 dB per decade and so forth.

Now let us make a comparison of the three filter types just described. For the three instances, the cut off frequency will be set at 100 MHz. For the first filter F_C occurs when $\omega_C\left(\frac{Z_o C}{2}\right) = 1$, where $\omega = 2\pi F_C$. That means, solving for C, the capacitance is 63.7 pF. For the second filter we will make $R_F = 42\ \Omega$, the reason for this choice will become clear later on. To find the capacitance we use the equation:

$$\omega_C C\left(\frac{R_F + Z_o}{2 + \left(\frac{R_F}{Z_o}\right)}\right) = 1$$

Then solving for C and find C = 49.1 pF. For the third filter R_F will also equal 42 Ω. Setting C_1 and C_2 to the same value and doing some calculations (Appendix D), we find that C1 = C2 = 29.4 pF.

Now let's compare the three filters with the component values that were calculated. Start off by looking at Equations 2, 4, and 6 when $\omega \approx 0$. From Equation 2 we find $|S21_1(\omega)| \approx 1$, from Equation 4 we get $|S21_2(\omega)| \approx \frac{1}{\sqrt{2}}$, and from Equation 6 we get $|S21_3(\omega)| \approx \frac{1}{\sqrt{2}}$. The significance of $\frac{1}{\sqrt{2}}$ is that when put into Log form, $\left(20 \log_{10}\left(\frac{1}{\sqrt{2}}\right) = -3\ \text{dB}\right)$. From this we see that filter 1 does not attenuate incoming signals in the pass band while both filters 2 and 3 attenuate half of the power of signals in their pass bands. Pass band attenuation is one of the drawbacks to RC type filters. Now let us look beyond the cut off frequency and compare the three filters at some frequency in the rejection band. For this case this comparison will be done at 1.0 GHz. Calculating the insertion loss for the first filter we find $S21_1$=−20.8 dB. The second filter with the series resistor has better attenuation with $S21_2$ = −23.1 dB. The third filter that uses the RC Pi network has even better attenuation in the rejection band with $S21_3$ = −31.7 dB. So in this case by sacrificing some signal strength in the pass band, the RC Pi filter has better attenuation in the rejection band. For the first filter to have the same attenuation as the Pi filter at 1.0 GHz, it would need a cut off frequency of less than 31 MHz with a capacitor of 200 pF. The second filter would also need some changes to meet the 1.0 GHz attenuation of the Pi filter. The cut off frequency could be lowered by either increasing the series resistance or increasing the capacitance to ground. Increasing the series resistance has the obvious disadvantage of attenuating the pass band more. With this knowledge in hand, Figure 10 shows how the three different filters generally behave when they are required to have the same cut off frequency.

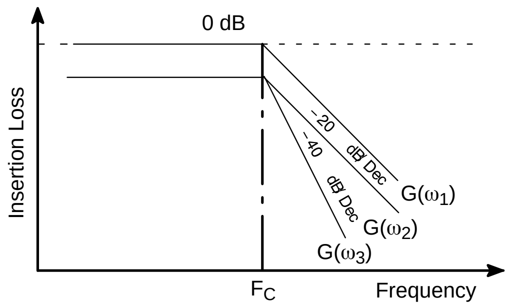

Figure 10. A comparison of the frequency responses of the three types of filters. All with the same cut off frequency and both RC filters with the same series resistance value.

When determining a filter topology and what component values are needed, there are few considerations that need to be kept in mind. Because this a filter, how much attenuation is needed in the rejection band? This is dependant on the application. Typical applications call for −25 dB to −35 dB of attenuation between the frequencies of 800 MHz to 2.4 GHz. What is the pass band bandwidth? This determines the cut off frequency. The bandwidth is determined by the type of signal to be passed. An audio signal would have a smaller bandwidth than say a high speed digital line. Finally, if an RC filter is to be used, how much pass band attenuation is acceptable? This is ultimately determined by the sensitivity by the receiving end of the network.

Once a filter topology has been determined, other aspects of the filter need to be considered, such as the filter's "Group Delay" and how the filter's package changes the insertion loss response of the filter.

Group Delay/Time Response

What is Group Delay? Group delay τ simply put, is the negative rate of change in phase of a linear time varying system with respect to ω ($2\pi \times$ frequency). Not so easy to grasp. To explain it would be easiest to use an example. Let's start off with a case when group delay is constant across all frequencies. There are a few idealized cases when this happens. The best example is the response from a lossless transmission line with matched impedance $Z_O = \left(\sqrt{\frac{L}{C}}\right)$. Matched impedance is a condition where a transmission line's series inductance and capacitance to ground interact such that there is no loss due to either inductance or capacitance. The basics are, a signal is applied to the input of a transmission line and some *time* later the signal is received at the other end. This time difference between when the signal is applied to the input and received at the output, can be called the time delay or "Group Delay". Under these ideal conditions, if the velocity of the signal and length of the transmission line are known, the time it takes the signal to travel can be calculated. But what does this have to do with the change in phase of the signal? Let's take a look at a transmission line with three signals being applied each at different frequencies. At the input side we can say that the three signals have the same phase $\theta = 0$. As the three signals travel down the transmission line to the output side, as in Figure 11, it can be seen how their phases change along the transmission line. Remember, the signal amplitude along the transmission line $= A\cos\theta$ and $\theta = \frac{2\pi}{\lambda}$ where λ = the wave length of the signal.

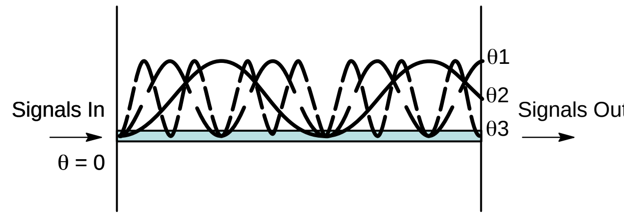

Figure 11. Three signals on a transmission line. For most applications in the frequency range discussed, the frequency is low enough where the signal wave length is many times longer than the circuit itself.

The wave length of a sinusoidal signal is inversely proportional to frequency ($\lambda \propto 1/\text{frequency}$). So as the frequency changes for a signal on our ideal transmission line, so changes the wave length and so changes the phase angle. In an ideal transmission line the phase angle changes at a constant rate with respect to the frequency as shown in Figure 12.

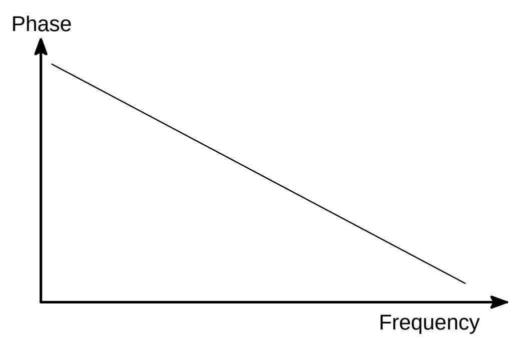

Figure 12. Idealized transmission lines have a phase angle that changes at a constant rate over frequency.

Now we can make the relationship between the change in phase with respect to frequency and group delay. As stated before, group delay is the negative of the slope of the phase angle over frequency as expresses by Equation 7.

$$\tau(\omega) = -\frac{d\theta(\omega)}{d\omega} \quad \text{(eq. 7)}$$

The Mathematical Definition of Group Delay

If we use Equation 7 on the frequency response from Figure 13, the resulting group delay is constant (Figure 14) across all frequencies. What this means is, when τ (ω) is constant for all frequencies, a signal that is put in comes out unchanged time at τ later (Figure 14).

Figure 13. An idealized flat group delay across frequency. This is a result of a constant rate change of the system's phase angle across frequency.

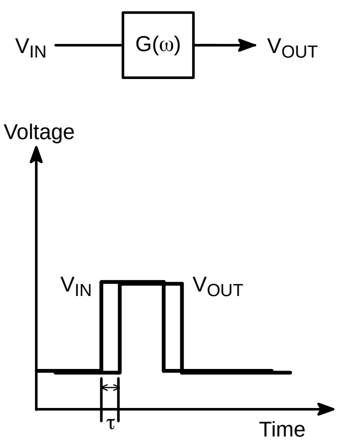

Figure 14. If all signal components that make up a pulse waveform at the input of a system are delayed by the same amount of time, the resulting output signal will be unchanged.

Now that we know what a constant group delay does to a signal, the next question should be what happens to data when the group delay is not constant?

If we look at a typical data stream as in Figure 15, we see a nice series of square waves that are comprised of a fundamental and a series of harmonics. When all of the components maintain the proper phase angle with respect to each other, there is no change in the signal.

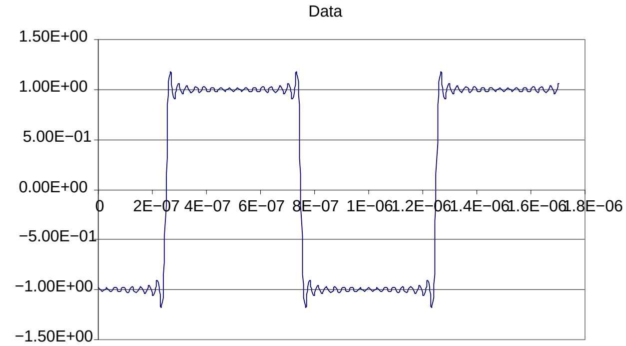

Figure 15. A Realistic Pulse Train with No DC Offset

Mathematically this data stream can be described as a series of Cosines as in Figure 16.

$$S(t) = \frac{4k}{\pi} \sum_{n=1}^{\infty} \frac{\mathrm{Cos}((2n-1)\omega_0 t)}{(2n-1)} (-1)^n$$

Figure 16. Mathematical Expression for a Pulse Train with Fundamental Frequency ω

But what if the signal is put through a system where only the group delay is affected. A realistic group delay response from a low pass filter is shown in Figure 17. In this case though, there will not be any of the associated attenuation to the signal. Low pass filters, especially RC filters, tend to have a constant group delay at the lower frequencies and then the group delay rolls off and approaches zero.

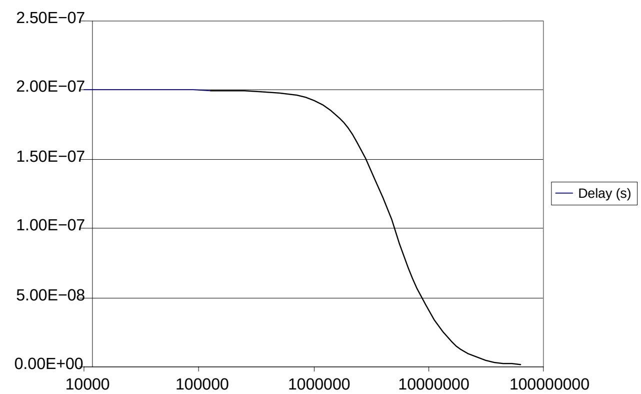

Figure 17. Simple Group Delay Characteristics for an RC Filter

The affect of this change in group delay can be dramatic depending on where the greatest amount of change occurs, with respect to the frequency components of the data signal. In this case the data signal is at 1.0 MHz, which is about the same frequency that our system's group delay changes. Because there is no attenuation, the affects can be seen in Figure 18. In this case, the pulse form is still recognizable, but the distortion is very noticeable.

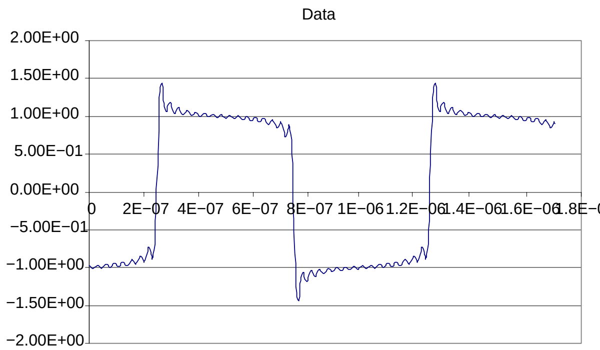

Figure 18. The resultant pulse train after going through the system with Figure 17's group delay response without any signal attenuation.

The effects on the data stream amounts to a shift in phase for each component of the signal dependant on the harmonic of the signal and where it falls in the phase shifting system.

$$S(t) = \frac{4k}{\pi} \sum_{n=1}^{\infty} \frac{\mathrm{Cos}((2n-1)\omega_0 t + \emptyset_n)}{(2n-1)} (-1)^n$$

Figure 19. Mathematical expression for a pulse train with fundamental frequency ω and a phase shift for each component of the data signal.

Group delay for the most part is a consequence of the charge and discharge times of reactive circuit elements (capacitors and inductors). For our purposes it comes directly from the series resistance, shunt capacitances, and the source and load impedances. In short, group delay is similar to an RC time constant. Looking at it this way you see that the "time constant" changes proportionately to R and C and so does group delay. In effect, the velocity of some frequencies change because of the interaction with the R_F and C. The result is some components of a signal arrive at the end of a filter at different times. For low pass RC filters, the group delay decreases with frequency.

The only effective way to improve group delay is to reduce the resistance, capacitance, or both. Unfortunately, this directly impacts the filtering characteristics of the filter. There are other options but they tend to be more costly or technically impractical. One of these methods is to use an active filter topology. These can be very effective in filtering, and offer high−speed responses, but tend to be costly. The other alternative is to use inductances in the filter. In place of the resistor an inductor can be used. Unfortunately, again, for the inductor to be of any practical use, it needs to be large enough to offer attenuation at higher frequencies, and have half way decent Qs.

Unidirectional or Bidirectional Zener Diodes

There are a few that need to be kept in mind when using an RC Pi filter. One of those is whether to use unidirectional or bidirectional diodes for the filter. When making this decision has it been determined if the filter is to be used in a unidirectional or bidirectional configuration? And, do the components that need ESD protection have symmetrical break down characteristics, that is, does the circuit break down at the same magnitude in the negative polarity as it does in the positive polarity? The reason this is important is because ESD can be either positive or negative and not taking this into account could result in damaged components. To understand this in greater detail, we need to evaluate three Pi filter configurations under two different ESD events.

The first of these examples is a Pi filter using unidirectional Zener diodes as in Figure 20. For these examples we say that the capacitance for each diode is the same, and that R_F is an arbitrary value.

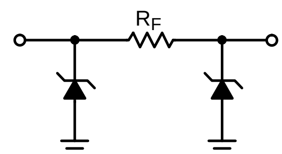

Figure 20. A Bidirectional Pi Filter Using Unidirectional Zener Diodes

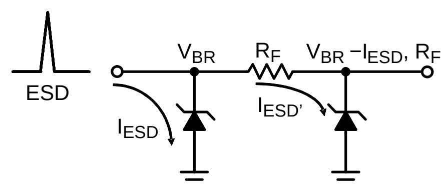

Figure 21. A bidirectional ESD/EMI filter with unidirectional Zener diodes being subjected to ESD with positive polarity.

In this configuration, when the filter is subjected to ESD with a positive polarity, there is an associated surge of current (Figure 22). This surge hits the leading Zener diode. The voltage across the diode increases until it reaches its breakdown voltage. At this point, most of the ESD current (I_{ESD}) flows through the first diode and is dissipated. There is, however, a small amount of current, ($I_{ESD'}$), that is conducted across R_F and the second Zener diode, resulting in a slightly lower voltage across the second diode (Figure 23).

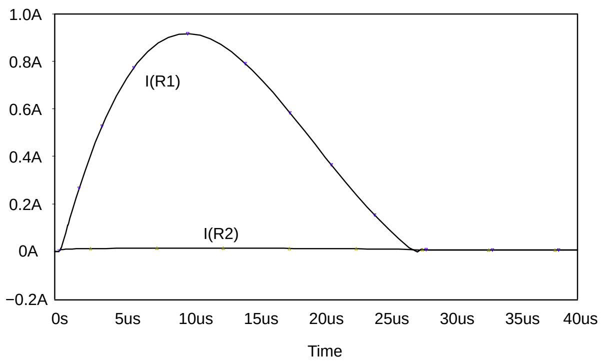

Figure 22. ESD surge current as seen by the leading port of the ESD/EMI filter.

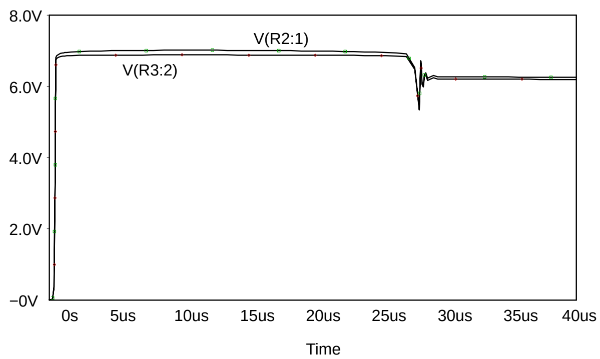

Figure 23. The associated voltages across each of the Zener diodes. There is only a small difference between the voltages across the two Zener diodes.

Because this is the most common configuration for Pi filters, this will be used as the baseline for the rest of the configurations.

Now, if the same filter is subjected to ESD with a negative polarity (Figure 24), the response is a bit different. When the surge current (I_{ESD}, Figure 25) hits the Zener diode and it begins to conduct in the forward direction (V_F). Again, some small amount of current ($I_{ESD'}$) conduct though R_F and the second diode (Figure 26).

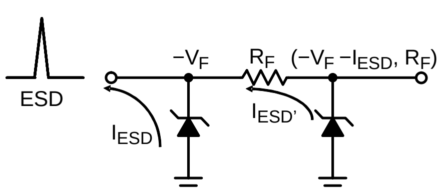

Figure 24. The same bidirectional ESD/EMI filter with unidirectional Zener diodes subjected to ESD with negative polarity.

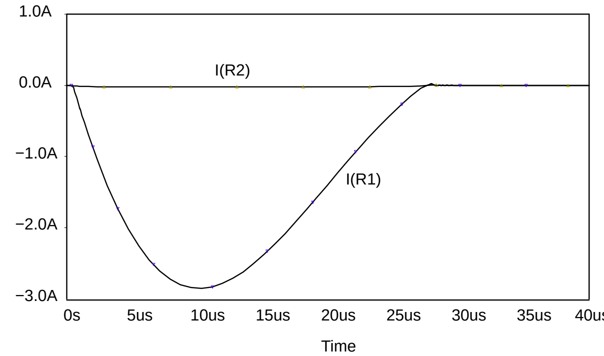

Figure 25. The ESD surge current with negative polarity. The current is substantially higher because the forward conduction of the Zener diode offers less resistance.

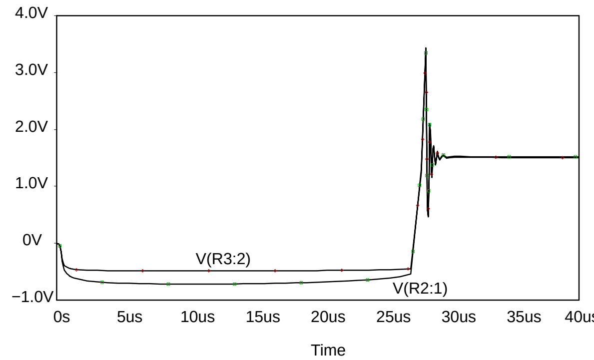

Figure 26. The voltage across the two diodes is significantly lower than before. The second diode's voltage drop is less than the first's because it is not fully in conduction.

The Pi filter with the unidirectional Zener diodes work very well with data signals that have a DC offset with the peak voltage less than the breakdown voltage of the Zener diodes. Because of the forward conduction of the Zener diodes, when a negative voltage is applied, this configuration is not well suited for data signals that are reference to zero volts.

With process technology improvements, it is now relatively simple to create back−to−back Zener diodes. But rather than immediately looking at an ESD/EMI filter with both diodes in a back−to−back configuration, it may be best to first look at having a back−to−back Zener diode on one side of the filter, and a unidirectional Zener diode on the other, as shown in Figure 27.

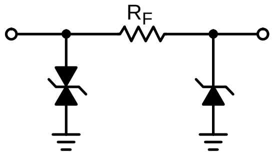

Figure 27. In this configuration, the filter would typically be used for a data moving in a singe direction. The ESD/EMI filter with back−to−back Zener diodes on one side, and a unidirectional Zener diode on the other, can be used bidirectionally.

Applying a similar ESD surge with positive polarity as in Figure 28, the back−to−back Zener diodes and the amount of current that enters the leading port, is less than that of the filter with just the unidirectional Zener diodes. This is because there the combination of the reverse breakdown voltage (V_{BR}) and the forward voltage (V_F) to account for. This is assuming that the breakdown voltage of the Zener has not been modified to adjust for the addition of the forward bias voltage drop of the mirrored diode.

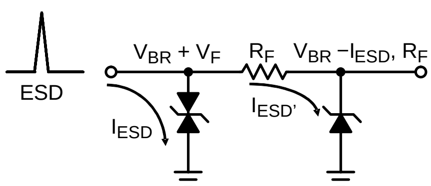

Figure 28. An ESD/EMI filter with back−to−back Zener diodes on the leading port, and a unidirectional Zener diode on the second port, subjected to ESD with positive polarity.

Regardless, if there has been an adjustment made to the breakdown voltage of the Zener diodes, one of the results is that there will be an asymmetric breakdown characteristic between the two sides of the filter. The breakdown voltage of the single unidirectional Zener will be less than the back−to−back Zeners. This will need to be kept in mind if data is being moved in both directions through the filter. Also, because of the lower breakdown voltage of the unidirectional Zener, the likelihood of that Zener diode conducting is greater than in the previous configuration. This difference in breakdown voltages also means that more current can conduct through R_F and the unidirectional Zener diode shown in Figure 29.

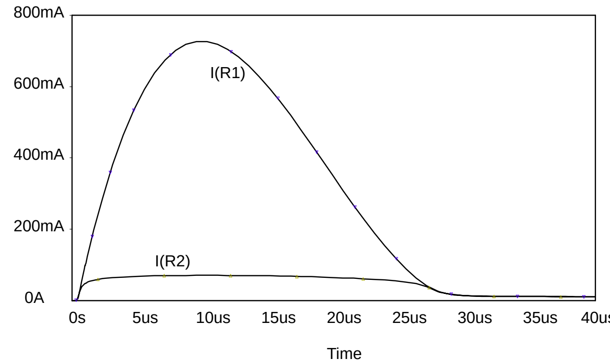

Figure 29. The ESD surge current seen by the leading port of the ESD/EMI filter with an associated current conduction through the second Zener diode.

When used to filter data moving in a single direction, this configuration has the advantage of dissipating most of the energy at the leading back−to−back Zener diodes, and then further reducing the voltage at the unidirectional diode as seen in Figure 30.

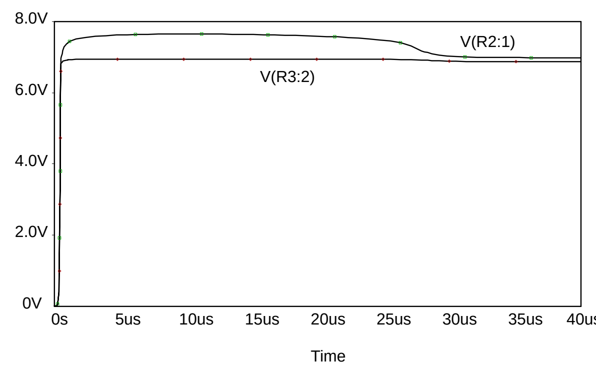

Figure 30. The different voltage levels at the two ports of the ESD/EMI filter. The leading back−to−back Zener diodes have a higher breakdown voltage than the unidirectional Zener.

Now there is the second case with this configuration being subjected to ESD with negative polarity, shown in Figure 31. With the same type of ESD applied, the surge current behaves differently than before.

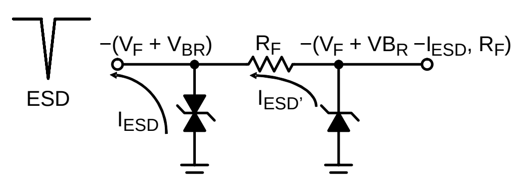

Figure 31. An ESD/EMI filter with back−to−back Zener diodes on the leading port and a unidirectional Zener diode on the second port subjected to ESD with negative polarity.

Much of this behavior has to do with the value of R_F. This is because the breakdown voltage of the back−to−back diodes is much larger than the forward voltage of the unidirectional Zener. If R_F is small enough, most of the energy from ESD with a negative polarity will be dissipated in the unidirectional Zener. The current through R_F in this case, will be dependent on the difference in voltage across the two ports.

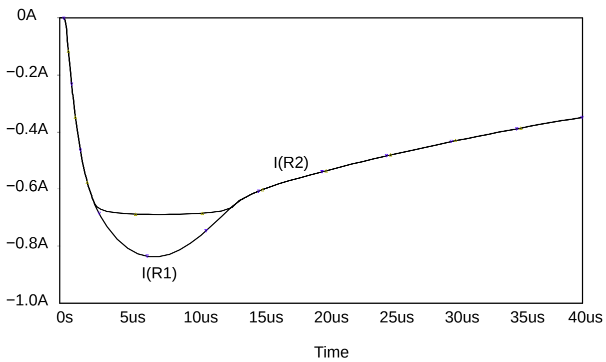

Figure 32. The ESD surge current behavior for ESD with a negative polarity in an ESD/EMI filter with a back−to−back Zener diode on the leading port and a unidirectional Zener on the second port. The current through the unidirectional Zener can be reduced by increasing the value of R_F.

So long as one of the two, the back−to−back or unidirectional Zener diodes dissipates the ESD energy, the second port should not go below VF of the unidirectional Zener seen in Figure 33.

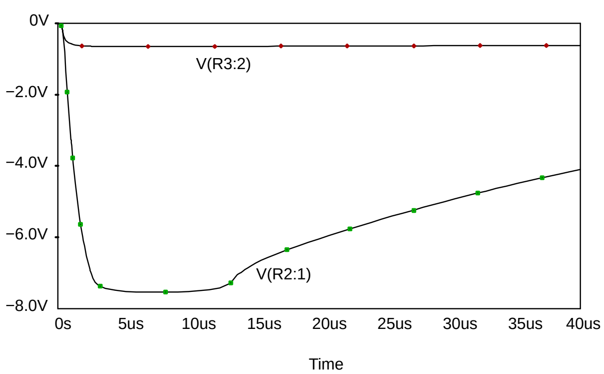

Figure 33. The voltage levels at both the back−to−back Zener diode and at the unidirectional Zener diode.

Even though this configuration can be used for data transfer in both directions, it is best used for cases where one side needs more protection than the other.

The final configuration for the Pi ESD/EMI filter is using back−to−back Zener diodes for both sides (Figure 34). This configuration is best suited for audio applications where the signal can have either positive or negative voltages with respect to ground.

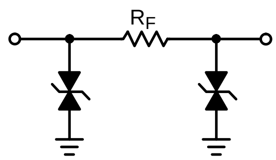

Figure 34. In this configuration the filter would typically be used for audio signals. The ESD/EMI filter with back−to−back Zener diodes on both sides of the filter can be used bidirectionally.

Again, ESD with positive polarity (Figure 35) causes a current surge (Figure 36) to hit the ESD/EMI filter. As with the previous configuration, the voltage is clamped to V_{BR} + V_F. Also, as before, most of the ESD energy is dissipated at the leading back−to−back Zener diodes. The difference is that there is not as much difference between the leading diode's and the second diodes voltages. This is because the higher overall breakdown voltage of the second back−to−back diode is the same as the leading back−to−back Zener diode, therefore not allowing nearly as much current to flow through R_F.

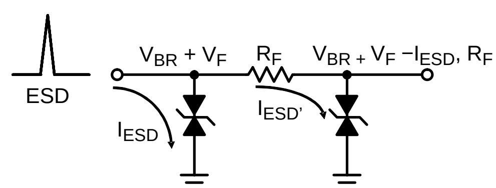

Figure 35. An ESD/EMI filter with back−to−back Zener diodes and on the second port subjected to ESD with positive polarity.

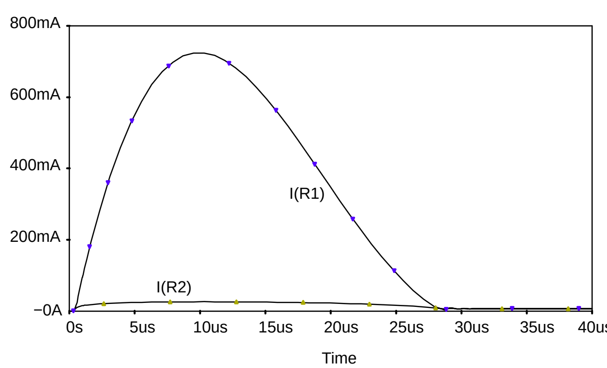

Figure 36. The ESD surge current seen by the leading port of the ESD/EMI filter with an associated current conduction through the second Zener diode.

If R_F is increased, the amount of current will then subsequently decrease, and the voltage difference between the two ports will also decrease.

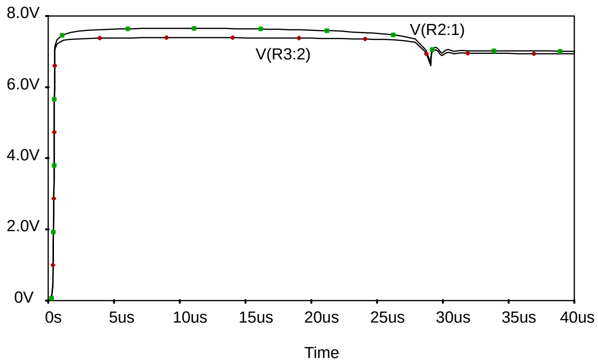

Figure 37. The different voltage levels at the two ports of the ESD/EMI filter.

Finally, when ESD with a negative polarity is applied as in Figure 38, it is seen that the surge current (I_{ESD}) and the current through RF ($I_{EDS'}$) is identical to the case where the ESD has positive polarity.

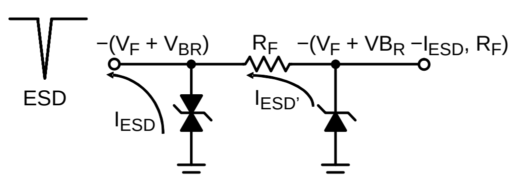

Figure 38. An ESD/EMI filter with back−to−back Zener diodes and on the second port subjected to ESD with negative polarity.

This of course assumes that the characteristics of the Zener diodes are identical, regardless of their orientation. Realistically, this is not often the case (for vertically integrated Zener diodes).

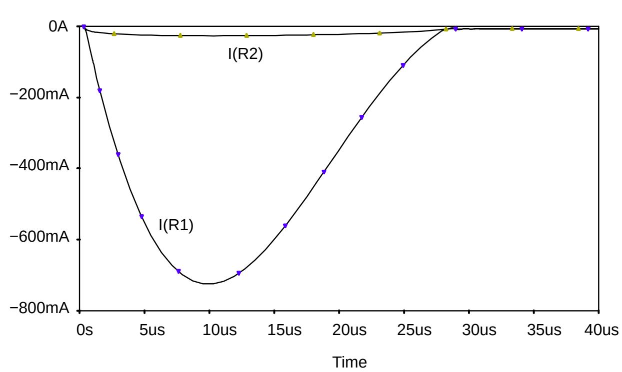

Figure 39. The ESD surge current seen by the leading port of the ESD/EMI filter with an associated current conduction through the second Zener diode.

For practical purposes it is easiest to have these characteristics be the same.

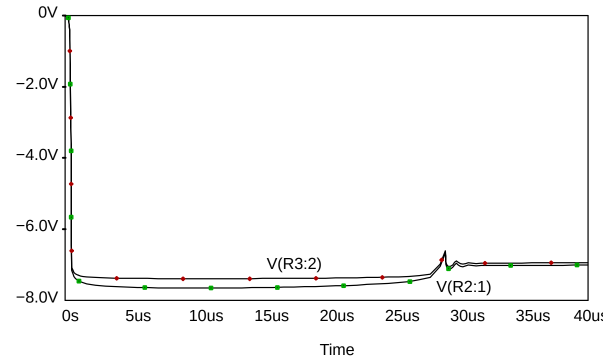

Figure 40. The different voltage levels at the two ports of the ESD/EMI filter.

As shown, an ESD/EMI filter with back−to−back Zener diodes on both ports have nearly identical characteristics for both positive and negative ESD. This also means the filter can be used for bidirectional data flow. With the magnitude of the clamping voltages nearly the same for positive and negative ESD, this configuration is good for data transmission that does not have a DC offset and where electrical components have nearly symmetrical ESD sensitivities. For the same reasons this configuration would not work well in conjunction with devices that breakdown at a much lower magnitude negative ESD than they would to positive ESD.

The common configuration uses the unidirectional Zener diodes. This is because most data has some DC offset, and the data source and the receiving end typically is more sensitive to ESD with negative polarity. Another reason is that unidirectional Zener diodes are much easier to integrate vertically than are back−to−back Zener diodes. Vertical back−to−back Zener diodes require a few more process steps and are not as cheep to produce. ESD/EMI filters that use back−to−back Zener diodes are best suited to data that has both positive and negative voltage components. Audio signals are a good example of this.

Signal Response/Insertion Loss

In band insertion loss is the amount of attenuation done to the data signal through the filter. The amount of insertion loss is determined by the series resistance of the filter. At lower frequencies, before the cut off frequency, the circuit should behave like a voltage divider (Figure 41).

The amount of in band insertion loss that can be tolerated is ultimately determined by the system designer. Because the filter acts as a voltage divider at lower frequencies, the sensitivity of the receiving circuit needs to be considered. The filter is doing no good if it attenuates the data signal below a level the receiver can detect.

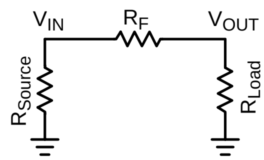

Figure 41. Low Frequency Equivalent Circuit

Packaging

There are a number of considerations to be made regarding the choice of package for the ESD/EMI filter. These decisions are typically based on the package size and footprint, cost, and electrical performance. The electrical performance of the package is rarely considered, though it becomes critical at higher data rates.

Early on, the only package type available was some kind SOT type package. Functionally this worked fine, especially if it used an exposed die pad. The electrical characteristics were far superior to a discrete component solution. The drawback though, is it has a large footprint. It is still smaller than an array of surface mount components, but the demand for smaller would relegate SOT type packages to a small assortment of filters.

The next innovation for ESD/EMI filers came about with the introduction of the Flip−Chip. The flip−chip package is not really a package at all. It is more a lack of package. What has been done is in place of wire bond pads and wire bonds to a lead−frame; small solder bumps are placed directly onto the die. The die is then flipped so that the solder bumps can make contact with the circuit board and the solder is then reflowed to make contact. In applications where component size (smaller the better) is important, this was a breakthrough. Not only are flip−chip devices smaller than a discrete solution or SOT solution, in many cases the electrical performance tended to be better. In other cases the performance was not as good. The issue is grounding. If the flip−chip has plenty of ground bumps, this is not much of an issue. When ground bumps are sparse though, is when filtering is degraded. This has to do with the amount of inductance contributed by the solder bumps. The ground on a flip−chip is less than ideal. The flip−chip ground makes contact to the board ground by way of the solder bumps. This adds inductance in series with the capacitance to ground as shown in Figure 42.

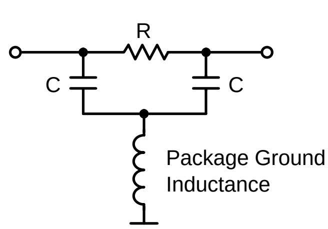

Figure 42. Regardless of the package being used, there is a certain amount of inductance in the ground path, some types of packaging introduce more than others.

What this does is create an LC circuit that has a self resonance. This self resonance behaves with band rejection qualities. At low frequencies the bump inductance behaves as an electrical short. But there comes a frequency where $\omega L_{Bump} = \frac{1}{\omega C_{effective}}$, where $C_{effective}$ is the effective capacitance of the filter, and this is a self resonant point. At this point the filter will not attenuate any further. Above this frequency though, shown in Figure 43, the capacitors behave like electrical shorts and the bump inductance offer more impedance and the result is the amount of attenuation decreases.

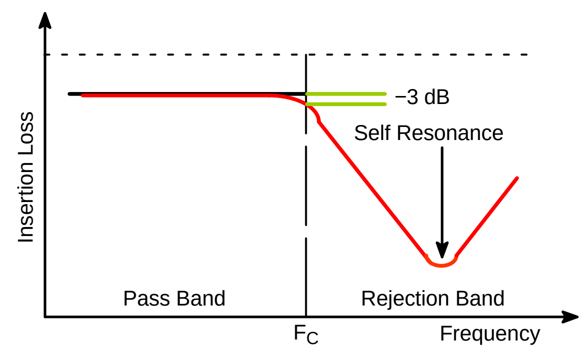

Figure 43. The effect of the parasitic inductance to ground causes a self resonance that causes the filtering response to behave more as a band rejection filter allowing higher frequencies to pass.

This type of response is seen in almost all ESD/EMI filters. What makes the flip−chip less attractive, is that this resonance typically takes place near the center of the rejection band. If there is enough inductance caused by the solder bumps, this response can reduce the amount of attenuation across the rejection band. The severity of this response is determined by the amount of capacitance in the filter, and the number of solder bumps. Simply put, the more solder bumps there are, the less inductance to ground, and therefore that inductance only becomes a factor until higher frequencies.

The next step in package development comes in the form of the DFN (Dual Flat No lead). What makes the use of the DFNs so special, is that the filters are assembled similarly to standard lead frames, but are no larger than the flip−chips. Additionally, the DFNs don't have the issue of inductance to ground. This means the frequency response does not have the self resonance in a region of frequencies that need to be attenuated.

When determining what type of package to use, the inductance of the ground path needs to be taken into account. More inductance in the ground path equates to a self resonance that occurs at a lower frequency. A similar filter with less inductance in the ground path has a self resonance at a higher frequency with the added benefit of better attenuation in the rejection band. This means that two filters with identical capacitances and series resistances, but with different packaging, could have very different frequency responses. Figure 44 shows how a filter's response changes with variations in the ground path inductance.

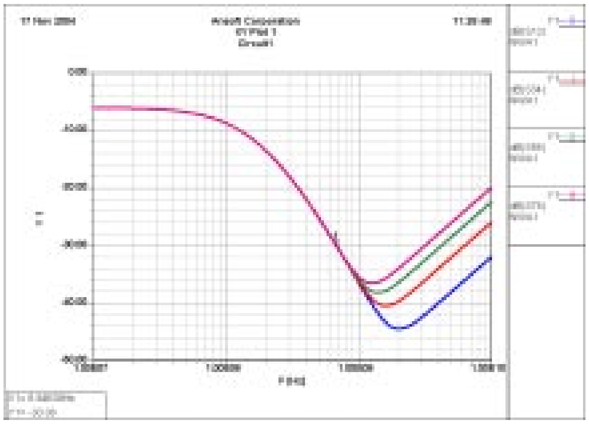

Figure 44. Self resonance changes with ground path inductance. The response with the least amount of attenuation has a ground path inductance of 40 nH where the response with the greatest attenuation has a ground path inductance of only 10 nH.

Other Issues to Consider

The number of lines is ultimately up to the end user. But there are a few things that may be helpful in the decision making process. First off, what is the total number of lines that need to be protected? This is more of a practical matter because most ESD/EMI filters come in multiples of two. There are a few exceptions but this is rare. Ultimately, when choosing a part or parts, to choose an EMI filter that meets the filtering requirements.

Cross talk is rarely an issue with most EMI filters because analog cross talk is least at the low frequencies in the pass band. Also, as cross talk increases with frequency, it is usually attenuated by the filter receiving the cross talk. It does become an issue when the pass band insertion loss is large. If the insertion loss is large enough, and the analog cross is substantial, line−to−line interference can occur. Cross talk is typically so insignificant that it is just overlooked.

Bringing It Together

Now having considered all of the above, you should have a good idea what type of ESD/EMI filter is needed. Having considered ESD, EMI, Group Delay, and Insertion Loss, the general characteristics and component values should be known. Finally, when it is determined that unidirectional or bidirectional Zener diodes are needed for a particular data type, what type of package to use and how many data lines are needed, a part can simply be ordered. Unfortunately most ESD/EMI filters come in relatively standard values of capacitance and series resistance. Most of these values however, do fulfill most needs. There are cases though where non−standard values can be built. Currently RC filters are able to fulfill the ESD/EMI filtering needs for today's data rates, but as data rates increase, this type of filter will need to make way for filters using inductors.

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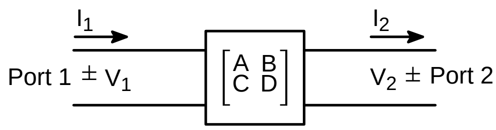
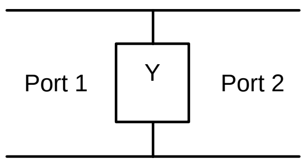
Appendices

Appendix A: Explanation of the ABCD Matrix



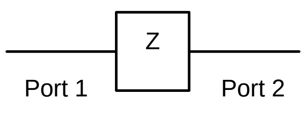
$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix}$$

$$A = \left.\frac{V_1}{V_2}\right|_{I_2 = 0} \qquad B = \left.\frac{V_1}{I_2}\right|_{V_2 = 0} \qquad C = \left.\frac{I_1}{V_2}\right|_{I_2 = 0} \qquad D = \left.\frac{I_1}{I_2}\right|_{V_2 = 0}$$



A = 1 B = 0
C = YD = 1

Shunt elements are handled in their admittance form (a component's admittance is the reciprocal of its impedance). If shunt element is a capacitor, then Y = jωC. This then yields the matrix $G_1(\omega) = \begin{vmatrix} 1 & 0 \\ j\omega C & 1 \end{vmatrix}$ as shown in Equation 1.



A = 1 B = Z
C = 0D = 1

In cases where there is a series, element Z is the impedance of that element. In the cases where a series resistance is used, $Z = R_F$. This then yields $G'_1(\omega) = \begin{vmatrix} 1 & R_F \\ 0 & 1 \end{vmatrix}$.

$$G_1 * G'_1 = G_2(\omega) = \begin{vmatrix} 1 + j\omega R_F C & R_F \\ j\omega C & 1 \end{vmatrix} \qquad \text{Equation 3}$$

If $R_F = 0$, then Equation 3 is equivalent to Equation 1.

$$G_1 * G'_1 * G_1 = G_3(\omega) = \begin{vmatrix} 1 + j\omega C_2 R_F & R_F \\ j\omega C_1 + j\omega C_2 + (j\omega C_1)(j\omega C_2)R_F & 1 + j\omega C_1 R_F \end{vmatrix} \qquad \text{Equation 6}$$

Appendix B: Calculating S21 (Insertion Loss)

Calculating insertion loss from an ABCD Matrix.

$$S_{21} = \frac{2}{A + \left(\frac{B}{Z_0}\right) + CZ_0 + D} \qquad \text{Equation 2}$$

$$S21_1(\omega) = \frac{1}{1 + j\omega\left(\frac{Z_0 C}{2}\right)} \qquad \text{Equation 4}$$

$$S21_2(\omega) = \frac{2}{\left(2 + \frac{R_F}{Z_0}\right)\left(1 + j\omega C\left(\frac{R_F + Z_0}{2 + \left(\frac{R_F}{Z_0}\right)}\right)\right)} \qquad \text{Equation 6}$$

$$S21_3(\omega) = \frac{2}{\left(2 + \frac{R_F}{Z_0}\right)\left(1 + j\omega C_1\left(\frac{R_F + Z_0}{2 + \left(\frac{R_F}{Z_0}\right)}\right) + j\omega C_2\left(\frac{R_F + Z_0}{2 + \left(\frac{R_F}{Z_0}\right)}\right) - \omega^2 C_1 C_2\left(\frac{Z_0 R_F}{R_F + Z_0}\right)\right)}$$

Appendix C: Phase Angle and Group Delay

$$\theta(\omega) = \mathrm{Tan}^{-1}\left[\frac{(\mathrm{IM}(S_{21}))}{(\mathrm{RE}(S_{21}))}\right]$$

$$\theta_{21}(\omega) = \mathrm{Tan}^{-1}\left[\omega C \frac{R_F + Z_0}{2 + \frac{R_F}{Z_0}}\right]$$

Equation is valid for both single capacitor to ground and series resistor with capacitor to ground configuration. For the single capacitor configuration RF = 0.

$$\theta_{21}(\omega) = \mathrm{Tan}^{-1}\left[\frac{\omega A}{1 - \omega^2 B}\right] \qquad A = (C_1 + C_2)\left(\frac{R_F + Z_0}{2 + \frac{R_F}{Z_0}}\right)$$

$$B = (C_1 C_2)\left(\frac{Z_0 R_F}{2 + \frac{R_F}{Z_0}}\right)$$

Appendix D: Example

Single Capacitor to Ground

From Equation 1

$G_1(\omega) = \begin{vmatrix} 1 & 0 \\ j\omega C & 1 \end{vmatrix}$ Equation 2 is derived $S21_1(\omega) = \dfrac{1}{1 + j\omega\left(\frac{Z_o C}{2}\right)}$

The cutoff frequency occurs at $\omega_C = 2\pi F_C$ and $F_C = 100$ MHz. The cutoff frequency is defined as the frequency where the signal strength is reduced by 3.0 dB. For this case it occurs when $\omega_C\left(\frac{50C}{2}\right) = 1$, the 50 comes from the system impedance. Solving for C we get a capacitance of 63.7 pF.

Series Resistor with Capacitor to Ground

As before, the transfer function is found to be (Equation 3) $G_2(\omega) = \begin{vmatrix} 1 + j\omega R_F C & R_F \\ j\omega C & 1 \end{vmatrix}$. From that, Equation 4 is found $S21_2(\omega) = \dfrac{2}{\left(2 + \frac{R_F}{Z_o}\right)\left(1 + j\omega C\left(\frac{R_F + Z_o}{2 + \left(\frac{R_F}{Z_o}\right)}\right)\right)}$. Solving for C using the formula $\omega_C C\left(\dfrac{R_F + Z_o}{2 + \left(\frac{R_F}{Z_o}\right)}\right) = 1$ it is found that C = 49.1 pF when $R_F = 42\ \Omega$ and $F_C = 100$ MHz.

RC "Pi" with two capacitors to ground with a series resistor. The process begins in the same manner with Equation 5 $G_3(\omega) = \begin{vmatrix} 1 + j\omega C_2 R_F & R_F \\ j\omega C_1 + j\omega C_2 + (j\omega C_1)(j\omega C_2)R_F & 1 + j\omega C_1 R_F \end{vmatrix}$ and from that Equation 6 is found $S21_3(\omega) = \dfrac{2}{\left(2 + \frac{R_F}{Z_o}\right)\left(1 + j\omega C_1\left(\frac{R_F + Z_o}{2 + \left(\frac{R_F}{Z_o}\right)}\right) + j\omega C_2\left(\frac{R_F + Z_o}{2 + \left(\frac{R_F}{Z_o}\right)}\right) - \omega^2 C_1 C_2\left(\frac{Z_o R_F}{R_F + Z_o}\right)\right)}$.

In this case, the two capacitance values will be equal, and the series resistance R_F will be 42 Ω. Solving for the capacitances the "real" component of the denominator is set to equal the "imaginary" component. Doing this we get $1 - \left(\omega^2 C^2 \dfrac{(Z_o R_F)}{\left(2 + \frac{R_F}{Z_o}\right)}\right) = 2j\omega C \dfrac{(R_F + Z_o)}{\left(2 + \frac{R_F}{Z_o}\right)}$. Solving for C we get C = 29.4 pF.

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