ELIMINATION OF THE DEGRADING EFFECTS OF LOAD AND LINE IMPEDANCES FOR POWER LINE EMI FILTERS

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Received September 14, 1976

Presented by Prof. Dr. F. CSÁKI

Introduction

It is a well-known fact that the electro-magnetic interference (EMI) filtering differs considerably from ordinary filtering. Essential differences are as follows:

1. EMI filters usually operate under mismatched conditions. The reason is that the line impedances are rarely known and vary widely for different lines to be filtered. In addition, the impedances usually vary in the frequency range of the stopband. These facts explain why conventional filter theory is of little value in the field of EMI filtering.

2. Considerable ac or dc power has to be fed through the EMI filters. This means the energy level of signals to be passed with negligible attenuation may be much higher than that of the signal to be considerably attenuated. Thus saturation effects may be of importance.

3. EMI filters have to operate over many — in some cases up to eleven — decades of frequency. Over such wide frequency ranges filter elements cannot be expected to behave as predicted and cannot be represented by their lumped-circuit equivalents.

4. The delay phase shift caused by the filter is insignificant for most of the EMI filters.

Facts listed under 1, 2 and 3 make EMI filtering more difficult while the fact in 4 facilitates EMI filter design.

The mismatched conditions can degrade the effectiveness of EMI filters and can result in dangerous effects. So the inserted filters may provide high gain in the stopband rather than the attenuation desired.

This paper describes a method of eliminating the effects of the unknown line and load impedances. The basic feature of the method is that any type of filter has to be attached at the input and output terminals by a specific *RL* or *RC* two-port. Proper design of the components of these "isolating" two-ports provides the conditions $Z_G \approx 0$ and $Z_L \approx \infty$ (Z_G — generator impedance; Z_L — load impedance) for the filter independently of the real line and load impedances.

Method of eliminating the effects of the termination impedances

The method will be illustrated on hand of a simple L-type filter; it will be seen, however, that it is applicable also in the case of more complex filters because only the two reactances (at the input and output terminals) are of importance for the method to be described.

Let us consider the circuit shown in Fig. 1. If a high attenuation is desired in the frequency range $(\omega_1; \omega_2)$ where ω_1 and ω_2 are the lower and the upper

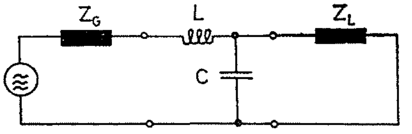


Fig. 1

limit, resp. of the frequency range for the stopband then L and C must be chosen to satisfy the inequality

$$\frac{1}{LC} \ll \omega_1^2 \, . \tag{1}$$

The dangerous effect of a mismatched situation can be illustrated with the following considerations. Be $Z_G \approx 0$ and $Z_L \approx \infty$, then the attenuation is:

$$A \approx \omega^2 LC \quad \text{if} \quad \omega > \omega_1 \tag{2}$$

Now, be $Z_G \approx 0$ and $Z_L = j\omega L_L$. Thus the attenuation becomes:

$$A = \left| 1 + \frac{L}{L_L}(1 - \omega^2 L_L C) \right| \tag{3}$$

If L_L satisfies the condition

$$\frac{L}{\omega_2^2 LC - 1} \leq L_L \leq \frac{L}{\omega_1^2 LC - 1} \tag{4}$$

then the attenuation is zero at a certain frequency in the stopband and it means infinite gain!

Similar problems arise if the impedance of the generator is capacitive. Obviously, if the reactances at the filter terminals meet equal reactances of the opposite sign at any frequency within the alleged stopband, then the effect of the filter will be degraded. If the load and line impedances are reactances with the same sign as those of the filter at the terminals, then the effectiveness of the filter increases.

Now, let us attempt to eliminate the interfacial effects (resonances) of the load and line impedances by inserting a two-port between the line and the filter and between the load and the filter as shown in Fig. 2. The task is to determine the values of L_1, R_1, C_2 and R_2 with the assumption that the attenuation of the system at any values of Z_G and Z_L will not be less than that provided by the two-port in the center (the original L-type filter) under the ideal conditions $Z_G = 0$ and $Z_L \approx \infty$.

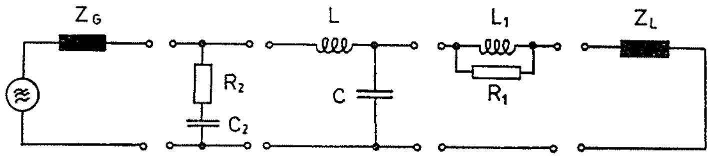


Fig. 2

Assume that Z_G and Z_L can vary in the range

$$\begin{aligned} 0 &\leq \operatorname{Re} Z_G < \infty \\ -\infty &< \operatorname{Im} Z_G < \infty \\ 0 &\leq \operatorname{Re} Z_L < \infty \\ -\infty &< \operatorname{Im} Z_L < \infty \end{aligned}$$

Output port

Assume $Z_L = a + jb$ where $0 \leqslant a < \infty$ and $-\infty < b < \infty$. Thus the impedance shown in Fig. 3 is:

$$Z(j\omega) = \left(\frac{\omega^2 L_1^2 R_1}{\omega^2 L_1^2 + R_1^2} + a\right) + j\left(\frac{\omega L_1 R_1^2}{\omega^2 L_1^2 + R_1^2} + b\right) . \tag{5}$$

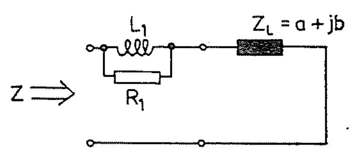


Fig. 3

The attenuation of the filter in the center practically does not decrease in the stopband if

$$|\, Z(j\omega) \,| \gg \frac{1}{\omega C} \tag{6}$$

or

$$|Z(j\omega)| > \varepsilon \frac{1}{\omega C} \tag{7}$$

where $\varepsilon \gg 1$.

(7) can be written in the form

$$\left(\frac{\omega^2 L_1^2 R_1}{\omega^2 L_1^2 + R_1^2} + a\right)^2 + \left(\frac{\omega L_1 R_1^2}{\omega^2 L_1^2 + R_1^2} + b\right)^2 > \varepsilon^2 \frac{1}{\omega^2 C^2} \tag{8}$$

It can be shown that the left side of (8) is at its maximum in the possible range of a and b for

$$a = 0$$
$$b = -\omega L_1 R_1^2/(\omega^2 L_1^2 + R_1^2)$$

leading to the inequality

$$\frac{\omega^2 L_1^2 R_1}{\omega^2 L_1^2 + R_1^2} > \varepsilon^2 \frac{1}{\omega^2 C^2} . \tag{9}$$

Solving (9) we obtain

$$f_1(\omega) < R_1 < f_2(\omega) \tag{10}$$

where

$$f_1(\omega) = \frac{2\omega L_1 \varepsilon}{\omega^2 L_1 C + \sqrt{\omega^4 L_1^2 C^2 - 4\varepsilon^2}} \tag{11}$$

$$f_2(\omega) = \frac{\omega L_1}{2\varepsilon}\left[\omega^2 L_1 C + \sqrt{\omega^4 L_1^2 C^2 - 4\varepsilon^2}\right] . \tag{12}$$

Thus

$$\{f_1(\omega)\}_{\max,\omega} < R_1 < \{f_2(\omega)\}_{\min,\omega} . \tag{13}$$

From the form of $f_1(\omega)$ and $f_2(\omega)$ it is evident that f_1 is at its maximum and f_2 at its minimum in the range $(\omega_1; \omega_2)$ if $\omega = \omega_1$ therefore

$$f_1(\omega_1) < R_1 < f_2(\omega_1) . \tag{14}$$

Another inequality has also to be satisfied, that is

$$\omega^4 L_1^2 C^2 - 4\varepsilon^2 > 0 \tag{15}$$

or

$$L_1 > \frac{2\varepsilon}{\omega_1^2 C} \tag{16}$$

A third condition is also to be fulfilled, namely the output port, shown in Fig. 4, must have no considerable gain. If the maximum gain is required to be less than μ, the inequality

$$\frac{|Z_L|}{|Z_1 + Z_L|} < \mu \tag{17}$$

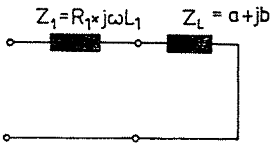


Fig. 4

has to be satisfied.

It is easy to show that the left side of (17) is at its maximum for

$$a = 0$$
$$b = -\omega L_1$$

and thus

$$R_1 < \{\omega L_1 \sqrt{\mu^2 - 1}\}_{\min, \omega} = \omega_1 L_1 \sqrt{\mu^2 - 1}. \tag{18}$$

The parameters of the "isolating" four-pole terminal have to satisfy the inequalities (14), (16) and (18). All conditions are featured by not containing the upper limit of the frequency range of the stopband.

Input port

Let us assume again that $Z_G = \alpha + j\beta$ where $0 \leqslant \alpha < \infty$ and $-\infty < \beta < \infty$.

The impedance shown in Fig. 5:

$$Z(j\omega) = \frac{(\alpha - \omega\beta R_2 C_2) + j(\omega\alpha R_2 C_2 + \beta)}{(1 - \omega\beta C_2) + (j\omega\alpha C_2 + \omega R_2 C_2)}. \tag{19}$$

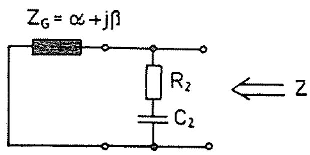


Fig. 5

The attenuation of the filter in the center practically does not decrease in the stopband if

$$|Z(j\omega)| \ll \omega L \tag{20}$$

or

$$|Z(j\omega)| < \frac{1}{\varepsilon}\omega L \tag{21}$$

where ε has the same meaning as in (7).

Inserting (19) into (21) it is seen that the left side of (21) is at its maximum for

$$\alpha = 0$$
$$\beta = (1 + \omega^2 R_2^2 C_2^2)/\omega C_2$$

hence

$$\frac{1 + \omega^2 R_2^2 C_2^2}{\omega^2 R_2 C_2^2} < \frac{1}{\varepsilon}\,\omega L\ . \tag{22}$$

Solving (22) we obtain

$$\frac{2\varepsilon}{\omega C_2}\,\frac{1}{\omega^2 L C_2 + \sqrt{\omega^4 L^2 C_2^2 - 4\varepsilon^2}} < R_2 < \frac{\omega^2 L C_2 + \sqrt{\omega^4 L^2 C_2^2 - 4\varepsilon^2}}{2\omega C_2 \varepsilon} \tag{23}$$

Again, it is evident that $\omega = \omega_1$ gives the critical values. On the other hand, the following condition must also be satisfied

$$\omega^2 L C_2 > 2\varepsilon$$

or

$$C_2 > \frac{2\varepsilon}{\omega_1^2 L}\ . \tag{24}$$

Investigating the gain of the input port, R_2 must satisfy

$$\frac{R_2^2 + 1/\omega^2 C_2^2}{(R_2 + \alpha)^2 + (\beta + 1/\omega C_2)^2} < \mu^2\ . \tag{25}$$

The left side of (25) is at its maximum for $\alpha = 0$ and $\beta = -1/(\omega C_2)$, thus

$$R_2 > \frac{1}{\omega_1 C_2 \sqrt{\mu^2 - 1}}\ . \tag{26}$$

It is to be noted that (9) and (22) can also be derived in another way. We show it for (22) but the procedure is the same for (9).

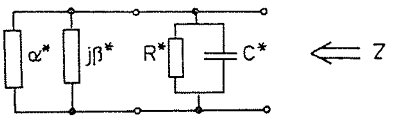
Let us transform the series circuit R_2C_2 into a parallel circuit R^*C^*. Values of R^* and C^* are:

$$R^* = (1 + \omega^2 R_2^2 C_2^2)/\omega^2 R_2 C_2^2 \tag{27}$$

$$C^* = (1 + \omega^2 R_2^2 C_2^2)/\omega C_2\ . \tag{28}$$

The impedance of the generator can also be replaced as shown in Fig. 6. Thus,

$$Z = \frac{1}{\dfrac{1}{\alpha^*} + \dfrac{1}{R^*} + j\left(-\dfrac{1}{\beta^*} + \omega C^*\right)}\ . \tag{29}$$



$$Z_G = \frac{1}{\frac{1}{\alpha^*} + \frac{1}{j\beta^*}} \qquad Z^* = \frac{1}{\frac{1}{R^*} + j\omega C^*}$$

Fig. 6.

Z is at it maximal value if $\alpha^* = \infty$ and $\beta^* = -1/\omega C^*$ and $|Z|_{\max} = R^*$, hence

$$R^* = \frac{1 + \omega^2 R_2^2 C_2^2}{\omega^2 R_2 C_2^2} < \frac{1}{\varepsilon}\, \omega L \ , \tag{30}$$

the same as (22).

Practical design consideration

The above method of eliminating the line and load impedances is effective only in the case if the required values of the components in the "isolating" ports are not too high in comparison with the values of the components in the filter. Comparing the inequality (1) with the inequalities (16) and (24) shows that L_1 and C_2 have not to be chosen generally greater than L and C. It means they meet the economical and practical requirements. The investigation of the conditions for R_1 and R_2 shows that they can be satisfied unless ε is extreme high and μ extreme low.

To illustrate the conditions above let us consider the following example. Fig. 7 shows a simple filter which provides min. 40 dB attenuation above 160 kHz ($\omega \approx 10^6\ s^{-1}$) if $Z_G = 0$ and $Z_L = \infty$. Now, let us determine the elements of the isolating ports so that the attenuation does not decrease at any random value of the load and line impedances.

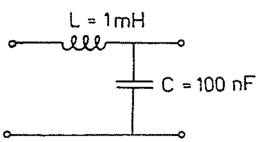


Fig. 7

Let us choose $\varepsilon = 10$, $\mu = 2$ and $\omega_1 = 10^6\ s^{-1}$. Thus from (16) $L_1 > 2 \cdot 10^{-4}$ H. Be $L_1 = 3 \cdot 10^{-4}$ H and (14) and (18) yield:

$$115\,\Omega < R_1 < 768\,\Omega$$
$$R_1 < 520\,\Omega.$$

From (24) $C_2 > 2 \cdot 10^{-8}$ F. Be $C_2 = 3 \cdot 10^{-8}$ F, thus from (23) and (26):

$$12.7\,\Omega < R_2 < 87.5\,\Omega$$
$$19.3\,\Omega < R_2$$

Finally, if $R_1 = 500\,\Omega$ and $R_2 = 50\,\Omega$ and the filter is assumed to operate in a power line with a voltage of 220 V and to pass a current of 10 A then R_1 passes a current of about 1.25 mA and R_2 passes a current of about 2.08 mA. This means that the dissipated power on them at the line frequency (50 Hz) is negligible.

Conclusions

This paper presented a method for eliminating the effects of the load and line impedances on the effectiveness of power line EMI filters. Using the proposed "isolating" ports the filter can be treated as if it would terminate at $Z_G = 0$ and $Z_L = \infty$. This fact is of great importance since it allows the application of the conventional filter theories.

The calculation required for determining the elements of the isolating ports is very simple. Filters designed in the suggested way can be effectively used in practice.

On the other hand, the theory presented here may be of use in designing lossy filters where R_1 and R_2 are not lumped components but represent the magnetic and dielectric losses in the coil (inductor) and in the capacitor.

Summary

Power line EMI filters often operate under mismatched conditions. This leads to difficulties in the operation and design of EMI filters. A practical method of eliminating the degrading effects of the mismatched conditions is outlined. *RL* and *RC* four-poles provide the elimination of the effects of the load and line impedances if values of their components satisfy the conditions derived.

Acknowledgement

The author wishes to express his thank to I. Ipsits, Associate Professor at the Dept. of Automation, Technical University Budapest, for his encouragement, advice and constructive criticism on this work.

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