EMI Filter Design of a Three-Phase Buck-Type PWM Rectifier for Aircraft Applications.

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***Abstract*—An EMI filter for a three-phase buck-type medium power pulse-width modulation rectifier is designed. This filter considers differential mode noise and complies with MIL-STD-461E for the frequency range of 10kHz to 10MHz. In industrial applications, the frequency range of the standard starts at 150kHz and the designer typically uses a switching frequency of 28kHz because the fifth harmonic is out of the range. This approach is not valid for aircraft applications. In order to design the switching frequency in aircraft applications, the power losses in the semiconductors and the weight of the reactive components should be considered. The proposed design is based on a harmonic analysis of the rectifier input current and an analytical study of the input filter. The classical industrial design does not consider the inductive effect in the filter design because the grid frequency is 50/60Hz. However, in the aircraft applications, the grid frequency is 400Hz and the inductance cannot be neglected. The proposed design considers the inductance and the capacitance effect of the filter in order to obtain unitary power factor at full power. In the optimization process, several filters are designed for different switching frequencies of the converter. In addition, designs from single to five stages are considered. The power losses of the converter plus the EMI filter are estimated at these switching frequencies. Considering overall losses and minimal filter volume, the optimal switching frequency is selected.**

Keywords: Three Phase Rectifier, EMI Filter, High power factor.

I. Introduction

The input filter in a PWM rectifier system has three purposes: 1) to ensure sinusoidally shaped input currents by filtering the switching-frequency harmonics; 2) to attenuate the electromagnetic interference with other electronic systems; 3) to avoid susceptibility to electromagnetic emissions from surrounding systems and itself [1], [2], [3]. While designing an EMI filter for a power electronic system, the applicable EMI standards need to be considered.

Typically in industrial applications, the standard to comply with is CISPR 22 class B [4]. The frequency range considered by this standard reaches from 150kHz to 30MHz. In [1], [5], systems with a switching frequencies (f_s) of 28kHz and 18kHz respectively have been designed. These f_s have been chosen because they are sufficiently higher in comparison with the grid (50 or 60Hz). In addition, the first, second, third, fourth, and fifth harmonic of the f_s are out of the range of CISPR 22 class B; thus, the first harmonic to consider in the input filter design is the sixth harmonic at 168kHz (when f_s = 28kHz). This can be seen in the fig 1.

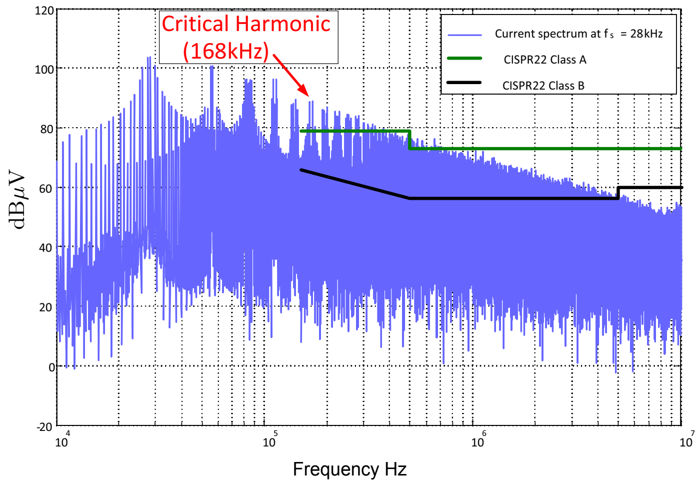


Fig. 1. Spectrum current using 28kHz switching frequency

This work introduces new considerations in the input filter design for a three-phase buck-type pulse-width modulation rectifier (fig. 2(a)) for aircraft applications. In this application the standard to comply with is MIL-STD-461E [6]. This standard is more restrictive than the CISPR 22, regulating a wider range of frequencies from 10kHz to 10Mhz. Fig. 2(b) shows the limits for MIL-STD-461E, CISPR 22 class A, and CISPR 22 class B. Due to the frequency range of the MIL-STD-461E and the fact that switching frequencies below 10kHz would not be an optimal design, the rectifier switching frequency must be inside of the range. Therefore, the input filter must be designed in order to attenuate the switching frequency.

II. Design of the Input Filter

A. Converter Topology

The EMC input filter is designed for a three-phase, three-switch, current source (buck-type) PWM rectifier system, fig 2 (a), with sinusoidal input current, direct start-up, and overcurrent protection in case of an output short circuit. In [7] and [8] this topology has been considered for the realization of the input stage of high-power telecommunications rectifier

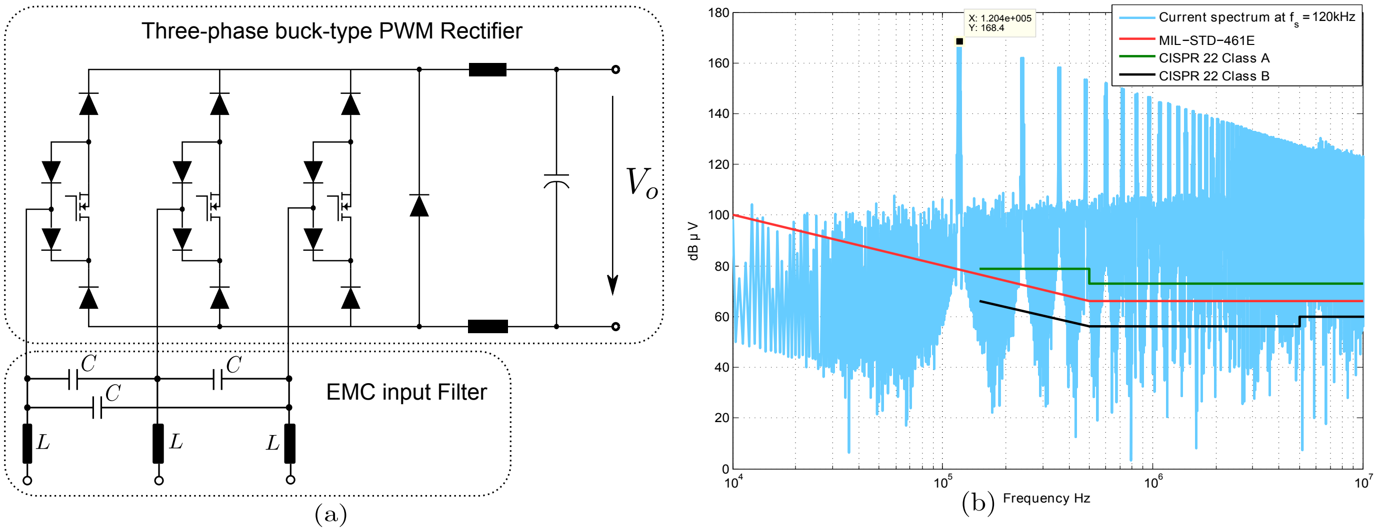


Fig. 2. (a) Three-phase buck-type PWM rectifier Topology and EMI input Filter, (b) Current spectrum of the rectifier at 10kW without input filter

modules. All of these benefits have prompted the authors to introduce this rectifier in aircraft applications.

In this work, a 10kW system for an aircraft application will be designed.

B. Cutoff frequency of the input Filter

In order to know the desired attenuation, the topology needs to be simulated without the input filter. Fig 2 (b) shows the measured input current spectrum. This current is measured utilizing a line impedance stabilizing network (LISN).

With a switching frequency (f_s) of the converter at 120kHz, the first harmonic has an amplitude of 168.4 dBμV and the MIL-STD-461E limit is 84 dBμV. Considering a margin of 6 dB, the required attenuation is $168.4 - 84 + 6(\text{Margin}) = 96.4\text{dB}\mu\text{V}$. The cut-off frequency as a function of attenuation and the switching frequency is given by:

$$\omega_{cutoff} = \frac{1}{\sqrt{L \cdot C}} = \frac{2\pi \cdot f_s}{\sqrt{10^{\text{Att[dB]}/(20n)}}} \quad (1)$$

$$L \cdot C = \frac{10^{96.4\text{[dB]}/(20n)}}{(2\pi \cdot 120\text{kHz})^2}, \quad (2)$$

where n is the number of the filter stages. Eq. 2 indicates the value of the product $L \cdot C$ as a function of the required attenuation at a certain frequency. To design the inductive and capacitive values, it is necessary to take into consideration the power factor of the rectifier.

C. Input Capacitor Consideration in industrial application

In [7] and [9] the input capacitor is designed in order to limit the reactive power of the rectifier. Eq. (3) gives the maximum value for the input capacitor C as a function of the reactive power (in percentage of the nominal power P_N). Usually this power is limited to (5..10%) of the rated power in order to ensure high power factor.

$$C \leq \frac{(0.05...0.1) \cdot P_N}{\omega \cdot U^2_{N,l-l,rms}} = 1.67...3.04\mu\text{F}, \quad (3)$$

where ω is the grid frequency and $U_{N,l-l,rms}$ is the line to line input voltage (RMS). In aircraft applications, $\omega = 2\pi \cdot 400$ rad/s and $U_{N,l-l,rms} = 115\sqrt(3)$V. Thus, a good value for the capacitor is 1μF because the capacitance is lower than 1.67μF, eq. (3). Once the capacitance is fixed, the filter inductance can be calculated with eq.(2); therefore, the inductor value is 452μH using a two stage filter.

Fig. 3 (a) shows the equivalent circuit for the rectifier including the two stage filter impedance seen from the grid. The analytical expression of the impedances are as presented in eqs. (4) and (5)

$$Z_{eq1stg} = \jmath\omega L + \frac{1}{\jmath\omega C + 1/R} \quad (4)$$

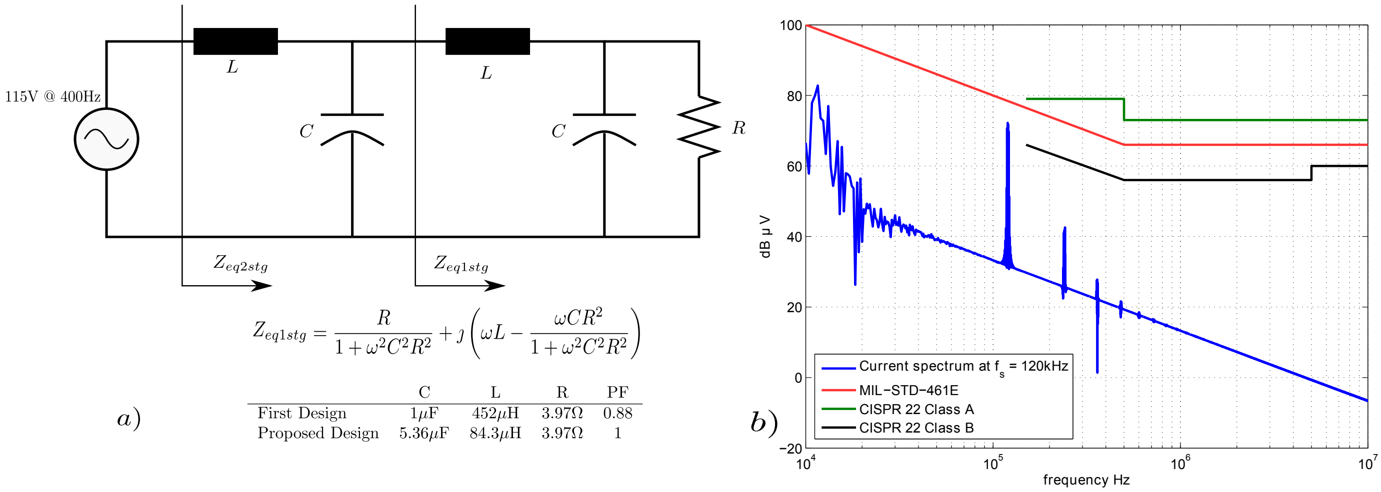
$$Z_{eq2stg} = \jmath\omega L + \frac{1}{\jmath\omega C + 1/Z_{eq1stg}}, \quad (5)$$

where Z_{eq1stg} and Z_{eq2stg} are the equivalent impedances for single stage and two stage filters respectively. The resistance corresponding the output power for 10kW at 115V is $R = 4\Omega$.

With this filter, the power factor of the system is only 0.88 ($\cos(\angle Z_{eq2stg})$), but according to eq. (3) should be higher than 0.99. Therefore, the power factor does not correspond with the design considerations because eq. (3) does not include the effect of inductor, which can be neglected for a grid frequency of 50 or 60 Hz. However, in aircraft applications the grid is 400Hz [10]; at this frequency the effect of the inductor can not be neglected anymore.

D. Proposed Consideration for the input capacitor

In order to know the influence of the single stage L-C filter on the power factor, the real and the imaginary part of the impedance needs to be considered separately according to eqs.



$$Z_{eq1stg} = \frac{R}{1+\omega^2C^2R^2} + \jmath\left(\omega L - \frac{\omega C R^2}{1+\omega^2C^2R^2}\right)$$

	C	L	R	PF
First Design	1μF	452μH	3.97Ω	0.88
Proposed Design	5.36μF	84.3μH	3.97Ω	1

Fig. 3. (a) Equivalent circuit for the rectifier including the input filter seen from the grid. (b) Current spectrum using the proposed filter.

(6) and (7),

$$\Re\{Z_{eq1stg}\} = \frac{R}{1+(\omega\cdot C\cdot R)^2} \quad (6)$$

$$\Im\{Z_{eq1stg}\} = \omega L - \frac{\omega C}{R^{-2}+\omega^2C^2}. \quad (7)$$

The unity power factor is obtained when $\Im\{Z_{eq1stg}\} = 0$. With this condition, and the cut-off frequency being as it is in eq. (1), the filter component is as presented in eqs. (8) and (9)

$$C = \frac{1}{R\sqrt{\omega_{cutoff}^2 - \omega^2}} \quad (8)$$

$$L = \frac{1}{\omega_{cutoff}^2 C}. \quad (9)$$

Thus, the power factor of the rectifier is unity using a single stage input filter. In addition, if $(\omega\cdot C\cdot R)^2 << 1$ (this applies when the capacitance is in the order of μF), the equivalent impedance for the single stage filter is approximately R ($Z_{eq1stg} \approx R$). If $Z_{eq1stg} \approx R$, then $Z_{eq2stg} \approx Z_{eq1stg}$ according to eqs. 4 and 5, and $\Im\{Z_{eq2stg}\} = \Im\{Z_{eq1stg}\} = 0$. Therefore, independent of the amount of the filter stages, when designing the filter according to eqs. (8) and (9), the power factor at full power is unity.

Using this proposed design method, the filter capacitance and inductance are $C = 5\mu$F and $L = 84\mu$H. In comparison with the classical design, the proposed design is smaller due to the inductor value. Fig. 3 (b) shows the current spectrum of the rectifier including the designed input filter. The current spectrum complies with the MIL-STD-461E in all the frequencies of the range.

E. Power factor depending on the power demand of the load

The Three-Phase Buck-Type PWM Rectifier is a two quadrant converter; when the input voltage is positive, the input current is positive and when the input voltage is negative the input current is negative. In addition, the rectifier is controlled in order to obtain sinusoidal wave form currents proportional to the input voltage. For this reason, the rectifier has a resistive behavior at low frequencies (grid frequency), and cannot deliver or absorb reactive power. Because of this, the power factor depends on the input filter and the power demand of the load.

In principle, the filter was designed in order to obtain unitary power factor at full power (10kW). The black line in fig 5 (a) shows the behavior of the power factor in full range of the power demand; however, from 5kW to 10kW the power factor is relatively high (higher than 95%). On the other hand, if the filter is designed at 5kW, the range of the high power factor increases from 2.5kW to 10kW. This can be seen in the red line in fig 5 (a). If the filter is designed at 1kW, the range is reduced from 0.5kW to 2kW, showed with the blue line in fig 5(a). The figs 5 (b) and (c) show the same curves for three and four stage filters respectively. The high power factor range increases with the number of stages in the filter. In low power demand, the power factor is inevitably low because the equivalent resistance of the rectifier is neglected in comparison with the impedance of the input filter; thus, the system is practically reactive.

In order to have high power factor in the wide range of the power demand, the input filter has to be designed in half of the nominal power (in this case 5kW). However, generally the power factor is measured at full power; therefore, in this paper, the filter is designed in order to obtain unitary power factor at full power.

III. APPLICATION TO THE OPTIMIZATION OF THE INPUT FILTER

In [1] the switching frequency for this topology is selected at 28kHz because the fifth harmonic (140kHz) is out of the standard range (150kHz - 30MHz). However, in aircraft applications this method of hiding the switching frequency harmonies below the standards frequency range can not applied because the MIL-STD-461E starting at 10kHz.

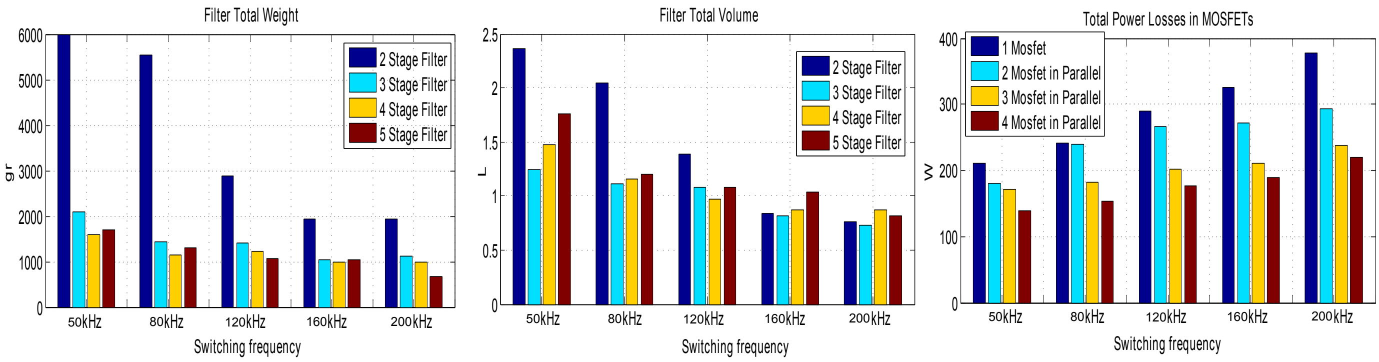


Fig. 4. (a) and (b) Volume and Weigh estimation for single, two, three, four and five stage. (c) Power losses in semiconductor using single, two, three, four MOSFETs in parallel

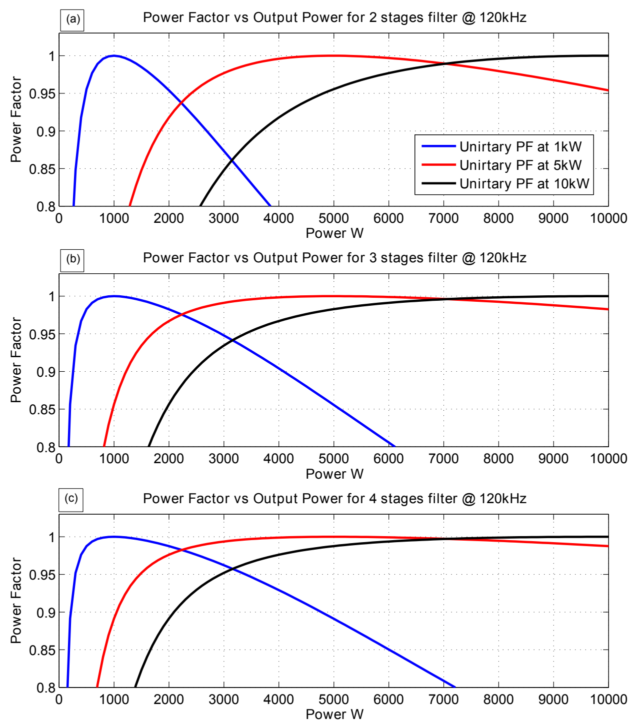


Fig. 5. Power factor depending to the power demand. (a) for two phase filter, (b) for three phase filter, (a) for four phase filter.

The switching frequency will be determined by the trade-off between volume/weight and power losses, for that the size/weight of the filter and the losses are going to estimated for different switching frequencies; thus, to obtain a design with a good balance between size and losses.

A. Weight and Volume estimation of the filter

For multiple-stage LC filters the minimum volume is achieved by using the same cut-off frequency for all stages [11]. Table I shows the cut-off frequencies from single to five stage filters for different switching frequencies in order to comply with MIL-STD-461E. The cut-off frequencies for single stage filters are close to or lower than the grid frequency, which makes it impossible to employ a single stage filter solution.

TABLE I

CUT-OFF FREQUENCY OF THE FILTER FOR SINGLE TO FIVE STAGES AND DIFFERENT SWITCHING FREQUENCIES IN ORDER TO COMPLY MIL-STD-461E.

cutoff freq	50kHz	80kHz	120kHz	160kHz	200kHz
Single stage	308Hz	392Hz	467Hz	555Hz	618Hz
Two stages	3.9kHz	5.6kHz	7.5kHz	9.4kHz	11.1kHz
Three stages	9.2kHz	13.6kHz	18.9kHz	24.2kHz	29.1kHz
Four stages	14kHz	21.2kHz	30kHz	38.8kHz	47.1kHz
Five stages	18kHz	27.6kHz	39.6kHz	51.5kHz	63kHz

Filters with two to five stages are designed using equations (1), (8), and (9). Then, with the inductance and capacitance values from resulting from these cut-off frequencies, the size and weight of the different filter solutions can be estimated.

For the weight estimation, the weight of the magnetic cores and of the wound wire are considered. The weight of the capacitors is neglected, since it is much lower than that of the magnetic components.

For the volume estimation, first, the total surface area is calculated by taking the sum of the all the elements; this number is multiplied by 1.5. Then, the volume is obtained by multiplying the height of the highest component by the total surface area to get the boxed volume.

The figs. 4 (a) and (b) show the estimation results. A two stage filter is not practical because the volume and weight are considerably bigger than for three, four, and five stage filters. Four and five stage filters provide only minimal improvements (if any) compared to three stage filters, and the number of components (and parasitic couplings between components) is much higher. Consequently, the three stage filter appears to be the best solution.

B. Power losses estimation

The conducting losses are estimated using the current stresses in the semiconductor [9]. The switching losses are estimated for considered switching frequencies [12] and [13]. These calculations have been carried out using one, two, three, and four MOSFETs in parallel in order to decrease the conducing losses. For every combination of f_s and the number of MOSFET in parallel, the optimal device, with respect to the power losses, has been selected from of a database available components. The power losses of the best MOSFET for every combination switching frequency vs MOSFETs in parallel are shown in the fig. 4 (c).

For the same number of MOSFETs in parallel, the total losses increase with the switching frequency, due to switching losses. The total losses decrease with the number of MOSFET in parallel; from one MOSFET to two MOSFETs in parallel (and from two to three MOSFETs in parallel), the power losses decrease considerably. However, to change from three to four MOSFETs in parallel, the power losses are only marginally decreased. In addition, the reliability decreases with the number of MOSFETs; therefore, it is apparent that using three MOSFETs in parallel is good option. The total losses increase consistently with the increase of the frequency; however, the weight does not decrease consistently. In the fig. 4 [a], the filter at 80kHz and 120kHz, as well as the filter at 160kHz and 200kHz, have practically the same weight using a three stage filter. Therefore, 80kHz and 160kHz are better switching frequencies because they have better ratios of weight losses. In order to be conservative in efficiency, the switching frequency selected is 80kHz.

IV. Conclusion

This work introduces new considerations in the input filter design for a three-phase buck-type pulse-width modulation rectifier for aircraft applications. For this type of applications, the EMI standard to comply is MIL-STD-461E. This standard is more restrictive than the CISPR 22 because of its frequency range, and it is not recommendable to consider a switching frequency below the standard range. The switching frequency has an impact in the trade-off between size of the input filter and the power losses; in order to obtain an optimum switching frequency the volume/weight and power losses have been estimated for a 10kW system. According to these estimations, the best trade-off between volume/weight and power losses is at 80kHz using three filter stages and three MOSFETs in parallel because the reduction of the filter size from 80kHz to 120kHz is marginal in comparison with the increment of the losses.

In addition, the classical design method for the input filter considers only grid frequencies of 50Hz or 60Hz; however, when the grid frequency is 400Hz, the power factor for the system is not close to unity. This work proposes a new consideration in the filter design in order to obtain a unitary power factor at full power independently of the number of the filter stages.

Currently a 10kW three-phase buck-type pulse-width modulation rectifier prototype is being built.

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