Mitigating Radiated Emissions of Power Feeders
On-board Electric Aircraft

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Abstract—The implementation of all-electric aircraft (AEA) will face several engineering challenges mainly due to its high power requirements. At its core lies the electric powertrain, consisting of the battery, inverter, and motor. Non-ideal behavior of components, cables, and other structures will act as a propagation path for Electro Magnetic Interference (EMI). The analysis of EMI in the design phase is considerably complicated due to structural and geometric design, thus, hard to predict. This paper is a first step towards proposing the attenuation of radiated emissions from power feeders by optimizing the switching behavior of converters. Thus, contributing to the overall attenuation level and reducing the performance requirements of power line filters.

Index Terms—all-electric aircraft, radiated emission, electromagnetic interference (EMI), electromagnetic compatibility (EMC)

I. INTRODUCTION

The implementation of all-electric aircraft (AEA) will face several engineering challenges mainly due to the high power requirements. At its core lies the electric powertrain, consisting of the battery, inverter, and motor (Fig. 1). To enable flight, high voltage and current levels are required onboard, while for secondary power distribution a greater number of onboard power converters are introduced. Due to the increased power ratings, together with higher switching frequencies, power density can be increased and inherently create a harsher electromagnetic environment [1], [2].

The usage of wide bandgap semiconductors is required to increase efficiency and reduce thermal loss by utilizing faster switching transitions. An increased electric and magnetic field strength can be expected due to the higher di/dt and dv/dt. Non-ideal behavior of components, cables and structures will act as a propagation path for the Electromagnetic Interference (EMI) [3], [4]. The analysis of EMI in the design phase is considerably complicated due to Common Mode (CM) propagation path being dominated by structural and geometric design thus, hard to predict. In [4] two simulation methods for EMI evaluation of a three-phase inverter is presented, covering the mechanisms responsible for most of the EMI of such a device. It can be seen how CM EMI is being generated by the switching mechanism in the inverter, with one of its propagation paths along the cable towards the motor. An optimized modelling method for evaluating the conversion of CM currents into radiated electric field for a single-wire above the ground-plane was proposed in [5]. Mitigation approaches are often aimed at a single component by either reducing the total CM EMI being generated by the inverter, reducing the radiative nature of the cable, or even reducing the parasitic capacitance of the motor completing the current loop. The interaction between the components of a power train is important for a trade-off analysis and eventual optimization.

In this paper, the intent is, as a first step, to analyze the expected radiated emission by considering the interaction of the power converter and power feeders to the load (motor), leading to optimization. The CM currents originating from the switching behavior of the power converters are typically mitigated by implementing power line filters to attenuate the excessive noise. In aerospace, this has a heavy penalty as the filter contributes significantly to the bulkiness and weight of systems. Considering cables with pre-specified length, their radiating behavior could be characterized by a certain transfer function between the input currents and radiated electric fields. Thus, optimal switching frequencies could be selected such that the spectral content of the input would be complementary to the spectrum of that transfer function. When implementing such an approach, cables will behave as complementary filtering elements, contributing to the overall attenuation level. Thus, lower performance requirements would be required for the implemented filters. This paper is structured as follows. In Sec. II the relation between radiated emissions and power feeders in an AEA is presented. Sec. III presents the theoretical background on Linear Time-Invariant (LTI) systems.

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Sec. IV the measurement setup representing the interaction between converter and power feeders is described. As for Sec. V the measurement results are shown demonstrating the proposed effect on the attenuation of the electric field. Finally, Sec. VI presents the conclusions gathered from the trends seen in radiated emissions.

II. Radiated EMI and Power Feeders

Electric powertrains are designed respecting a number of predefined constraints. The physical aspects of their topology are determined by application, as in an AEA, where the position of its elements is determined in the design phase. Thus, power feeders will have fixed dimensions interconnecting switching devices and electric motors. When in operation, the fast switching of converters will introduce voltage and current transients to cables, which are likely to result in compatibility issues, when not mitigated properly. As the length of power feeders increases, radiated emissions can become a concern, especially with the increase in switching frequencies and switching speed. Radiated emissions are defined by RTCA DO-160 as occurring in the range of 100 MHz – 6 GHz for airborne systems. The correlation between $f_{sw}$ and the length of power feeders was investigated using four different conductor lengths and switching frequencies. Higher $f_{sw}$ are expected to increase the BW of the signal that will be exciting the conductors [8]. While longer rise times are expected to reduce the BW. Additionally, longer elements will reduce the frequency that will start radiating. This assessment proposes the attenuation of radiated electric fields, reducing the demand of power line filters, which contributes to a reduction in bulkiness of such devices on a component level.

III. Theoretical Background

When considering a radiated power transfer from a conductor toward a receiver, the transfer function will be the ratio between the input current and measured electric field. In order to represent such interaction, a Linear Time-Invariant System (LTI) (Figure 1) was selected.

![Fig. 1: Block diagram of an LTI System](image)

This represents the relation of an input, $X(\omega)$, with the output of the system, $Y(\omega)$, based on linearity and time-invariance constraints. The transfer function $H(\omega)$ is defined as in (1).

$$H(\omega) = \frac{Y(\omega)}{X(\omega)} \quad (1)$$

In the time domain, an input digital signal can be represented by a trapezoidal pulse $x(t)$ which can be seen in (Figure 2). This waveform can be described as having a period ($T$), pulse width ($\tau$), and rise and fall times ($\tau_R, \tau_F$).

![Fig. 2: Contributors of the LTI system](image)

The frequency spectrum envelope $X(\omega)$ of such a pulse is defined by its parameters (Fig. 2b). It contains two slope rates, 20 and 40 dB, bounded by $\tau$ and $\tau_R/\tau_F$ values. From now on called first and second break frequencies. The resulting output signal will be dependent on this input signal and the transfer function $H(\omega)$ (Fig. 2c), which has a slope of 20 dB for lower frequencies as long as the cable is electrically short [9]. In Accordance to [8], for a near-field coupling the output signal envelope would present a trapezoidal format, with both slopes at a 20 dB rate (Fig. 2d). Which is the result of combining Fig. 2b and Fig. 2c.

In this paper it was accomplished by reducing the $f_{sw}$ of power converters and increasing the rise/fall time, taking into consideration the length of power feeders and their behavior.
IV. MEASUREMENT SETUP

A measurement setup was created, utilizing a Teledyne T3AFG120 function generator to emulate in a controlled manner the Pulse Width Modulation (PWM) waveforms created at the output of typical inverter systems with a simple trapezoidal clock signal, albeit with a lower power level. An eight-channel oscilloscope was used to simultaneously measure the output of the generator, the output of a loop antenna, and the 2x three axis of the two E-field probes. The probes and cables were placed inside a Faraday cage (anechoic chamber), while the measurement equipment was placed outside.

A. Source

The signal generator is connected to cables that are suspended above a ground-plane and mounted as is depicted in Fig. 3. Four different cable lengths are used:

- CL01 = 20 cm
- CL02 = 30 cm
- CL03 = 45 cm
- CL04 = 60 cm

The cables are scaled down by a factor of ten, thus representing lengths of 2 m to 6 m. This was done, to accommodate the size of the anechoic chamber that was used. The scaling of the cables also implies that the frequency of the PWM has to be scaled with a factor of ten, i.e., switching frequencies of typical power electronics are in the range of tens of kilohertz but will be represented by signals in the range of hundreds of kilohertz. Note that the anechoic chamber is used as a Faraday cage and is not properly absorbing the field at these low frequencies. The setup is built to represent a possible cross-talk scenario without a victim cable or cable bundle, by placing the measurement probes in the near-field of the "radiating" cable. All cables were terminated with 50 Ω. Note that the current therefore has an identical shape as is the case for the voltage applied. With respect to an AEA, this representation of the inverter with a motor is an oversimplification, as the current should be more triangular in nature due to the inductive loading by a motor. The measurement performed here is more of a conceptual nature, and therefore inductive and high power loading is considered to be beyond its scope.

B. E-Field

Two in-house developed E-field probes [10] were placed at approximately 10 cm and 1 m. The setup inside the chamber can be seen in Fig. 4. The probes utilized are described in detail in [10], and a schematic representation with the pinout diagram can be seen in Fig. 5. All three axis are separately sampled via a 1:10 voltage probe with a 4824A Picoscope. To obtain the total E-field strength, they are added according to (2).

\[ E_{\text{tot}} = \sqrt{E_X^2 + E_Y^2 + E_Z^2} \]  

(2)

The voltage measured needs to be converted to received power (or eventually E-field strength), for simplicity the conversion from the RSSI detector to received power is used. The conversion between power and e-field is a multiplication factor but requires a calibration step, which is outside of the scope of this paper.

In Fig. 6 a short period of a typical measurement is shown. As can be seen, the received power receives a pulse at the moment of the PWM transition. This is to be expected as the field is being generated by the change in voltage and current, i.e., the dv/dt and di/dt.
implementing trapezoidal pulses of controlled amplitudes, duty time, and maximum magnetic field versus switching frequency. Sec.V to obtain the trend of maximum emission versus rise time. These maxima are then combined in an individual measurement. These maxima are then combined in Sec.V to obtain the trend of maximum emission versus rise time, and maximum magnetic field versus switching frequency.

V. MEASUREMENT RESULTS

The evaluation of switching frequencies was performed by implementing trapezoidal pulses of controlled amplitudes, duty cycle (D), and rise/fall times (τ_{R/F}). For this assessment, sixteen different switching frequencies were evaluated: 10, 15, 20, 25, 30, 40, 50, 65, 75, 85, 100, 200, 350, 500, 650, and 750 kHz. For each frequency five rise times (assuming rise=fall time) were implemented, 10, 20, 30, 40, and 50 ns, resulting in 80 measurements per cable. The pulse amplitude was of 2.5 V, with a 50% duty cycle.

A. Reduction in switching frequency

As expected, the reduction in f_{sw} resulted in lower E-field levels, as seen in Fig. 8.

This is explained by the shift-left of the first break frequency (f_{B1}) in the input spectra (Fig. 2). The 20 dB slope initiates earlier and is extended towards the second break frequency (f_{B2}), reducing the overall spectral amplitude. The attenuation is perceived only at higher-frequencies, having no effect for f ≤ f_{B1}. For frequencies below 200 kHz the response is nearly flat with a small increase rate for the E-field amplitude. This is most likely due to lower power at that specific frequency range.

In addition, longer cables presented overall higher emission levels. This is expected as the length of the radiating structure is proportional to the increase in E-field [8].

When evaluating the radiated magnetic field (H-field), the reduction in f_{sw} presented a nearly flat response up to f ≤ 100 kHz which is the maximum operating range for the antenna (Fig. 9). Above that, the H-field level drops as the fundamental is outside the region of interest as seen in Fig. 7.
Finally, for every assessment, the amplitudes within the region of interest remained the same, with differences only occurring above 100 kHz.

Fig. 9: Max. H-field with increasing switching frequency - $f_{sw}=10$ ns

B. Increase in rise time

When increasing $\tau_R$ the E-field levels reduced considerably (Fig. 10). As expected, longer rise times will cause $f_{B2}$ to shift left, while $f_{B1}$ stays unchanged.

This will reduce the bandwidth and anticipate the 40 dB slope, which reduces the input spectral amplitude (Fig. 2). The longer the $\tau_R$, the more the shape of a pulse will tend to a sine wave, hence, lower harmonic content. In addition, voltage overshoots at the input will reduce as $\tau_R$ increases, contributing to the attenuation of radiated emission levels. Resulting in attenuation of up to 8 dBm.

While for the H-field, the increase in $\tau_R$ presented no reduction (Fig. 11), and where the flat response was also mentioned for the change in $f_{sw}$. This is likely related to the low input power at lower excitation frequencies, as also seen for the E-field measurements (Fig. 8), resulting in an insufficient net received voltage at the probe.

For the E-field, the overall trend was also noticed for the remaining assessments with every combination of $f_{sw}$ and $\tau_R$. When combined, both approaches resulted in a significant attenuation of E-field levels. Nevertheless, as seen in Fig. 10 the increase in $\tau_R$ is the most effective. While the reduction in $f_{sw}$ presents more prominent results only at higher frequencies.

VI. CONCLUSION

This paper demonstrated the relation between cable length and radiated field arising from its unintentional antenna behavior. The assessments demonstrated that the radiated field interfering with neighbouring systems can be attenuated by changes in the switching behavior. With a characterized transfer function for a cable, the radiated field can be attenuated by only knowing the input pulse parameters. It is achieved with the reduction in switching frequencies and an increase in rise time. Finally, the implementation of E-field probes provides the estimation of emissions without the need of a specific victim (cable), which facilitates not only controlled measurements but in situ as well. This is a first step towards the development of a mitigation technique that will tie cable length and switching frequency, defining design limits contributing to the reduction of radiated emissions.

REFERENCES

