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Historical Rationale for Military EMI Limits		

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Historical Rationale for Military EMI Limits

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**ELECTRONIC INDUSTRIES ASSOCIATION
ENGINEERING DEPARTMENT
2001 PENNSYLVANIA AVE, N.W.
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The G-46 Committee will continue to devote its efforts to the advancement of the knowledge and understanding of the subjects of electromagnetic environmental effects. Therefore, we invite your comments on this publication and welcome your suggestions concerning publication of additional information that would aid in the effective conduct of any activity relative to training and education within the electromagnetic environmental effects disciplines.

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PREFACE

Almost fifty years have passed since the advent of formalized military specifications requiring EMI qualification of electrical/electronic equipment. From a few specialized emissions measurements an entire industry has grown up. In the nature of such an evolution the limits and procedures have become codified and passed on from program to program. The form has been kept, but in many cases the functionality has been lost. Some of the most basic limits and methods are largely proof of compliance requirements, with a sometimes questionable relationship to system level EMC concerns. This was recognized as long ago as 1967. The top level EMC specification, MIL-E-6051D, *EMC Requirements, Systems* (still current at this writing) stated,

“3.2.4.1 Subsystems/equipments. Unless otherwise specified in the contract, subsystems/equipments shall be designed to meet the requirements of MIL-STD-461 and MIL-STD-462. *Since some of the limits in these standards are very severe, the impact of these limits on system effectiveness, cost, and weight shall be considered.* Proposed modifications to the limits shall be included in the EMC plans for the system and subsystems/equipments...” (Author’s italics) [1]

The purpose of this document is to foster an *engineering issues oriented* understanding of military EMI requirements. There are several reasons that this is important. The MIL-STD-461D and MIL-STD-462A revision process is currently nearing completion. There is little understanding of how the current (MIL-STD-461C) requirements came about. MIL-STD-461D promises to be in some cases a radical departure from the past, but in others cases quite traditional. The MIL-PRIME philosophy being incorporated into a new system level electromagnetics effects standard will place a premium on the understanding of engineering issues behind EMI/EMC requirements. Imposition of MIL-STD-461 requirements in the past has placed the responsibility for EMC design on the Procuring Activity. Suppose a weapon system were to have a system level incompatibility due to MIL-STD-461 compliant equipment interaction. Given a reasonable configuration equipment installation, the contractor could make a case requiring the Procuring Activity to pay for the fix. Under MIL-Prime philosophy, the contractor will achieve system level EMC in part by *system application specific* tailoring of MIL-STD-461 requirements. Such tailoring will become a more visible responsibility of the contractor. [2] There is thus a need for a broader understanding of the engineering issues behind EMI requirements.

Rationale for the limits and test procedures are unknown to the majority of industry’s EMC engineers. Review of specifications and papers written in the formative years of the discipline reveals strong interest in the design of limits and procedures. There has been a shift away from this aspect of EMI control towards a *design to specification compliance* emphasis. Optimum techniques for designing to compliance are very important. However, if the EMI limits/procedures are themselves not optimized, then the best the cleverest designer can hope to achieve is to minimize the cost, space, and weight impact of the improper requirement. The EMC discipline is extremely empirical in nature. In general, the control of

EMI has been preceded by an EMI incident. The problem is recognized, diagnosed, and fixed. Then, with hindsight, controls are instituted to prevent reoccurrence. A statement of general validity concerning military EMI performance requirements is that they are artifacts of a particular solution to a specific historical problem. In some cases, the historical problem is gone but the controls remain.

Several valuable efforts along the lines of this document have been previously rendered by C.B. Pearlston (1967), Al Parker (1990) and Dave Cofield (1991). The Pearlston report compares and contrasts EMI specifications of the day. Mr. Parker, a key supplier of EMI test equipment and a pillar of the industry since the '50's, delivered his historical perspective at an IEEE EMC chapter meeting. These reports are both included as appendices to this document. Mr. Cofield's informative history "Fifty Years of EMC in the Army" is available for all to read in the 1990 Interference Technology Engineering Master - ITEM. Mr. Cofield has been active in Army EMC since the early days of MIL-STD-461A, and is a member of the Tri-Service committee revising MIL-STD-461/462. When MIL-STD-461D and MIL-STD-462A are released, they will contain appendices with the rationale behind each requirement and test method, as well as suggestions for appropriate tailoring.

Present EMI limits and test methods have evolved from earlier specifications. These are listed and discussed. Many of the limits and methods derived in the 1950s are no longer applicable to today's technology and should be replaced. Suggested changes are listed for emission and susceptibility considerations, and alternative methodologies are discussed.

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INTRODUCTION

The EMI control program exists like an insurance policy, to replace unknown (and potentially large) program performance, schedule and budget impacts with a predictable level of effort which may be factored into the cost and schedule of the program. It is a continuing responsibility to ensure that the premiums reflect the value of the insured item and the risks involved. This document looks at what it takes to insure EMC considering electromagnetic spectrum usage in the 1990s, vs. current EMI control techniques. It will be shown that the conducted emissions/susceptibility limits and test methods overlook some very important factors and that the low frequency radiated (emission) limit and test method (emission/ susceptibility) is quite dated.

A condensed and abridged genealogy of military EMI/EMC specifications and standards is presented in Figure 1. The phrase "E pluribus unum" is as good a description of the evolution of EMI standards as it is of United States history. What is important to understand is, as usual, the reasons behind the standards and their evolution, rather than the names and dates. Nevertheless, a knowledge of the standards, their evolution and the time frame is useful. An excellent source is in Volume 1 of the older (1971) Don White EMC handbook series, pages 4.13ff. [3]

Some Milestone Specifications

This section traces the technical evolution of Air Force specifications. The author, whose background is aerospace, would like this section to be inclusive of all Service specifications. At the present time, this has not been possible. Readers of this document with pertinent information to share are asked to communicate with the author via the EIA G-46 committee.

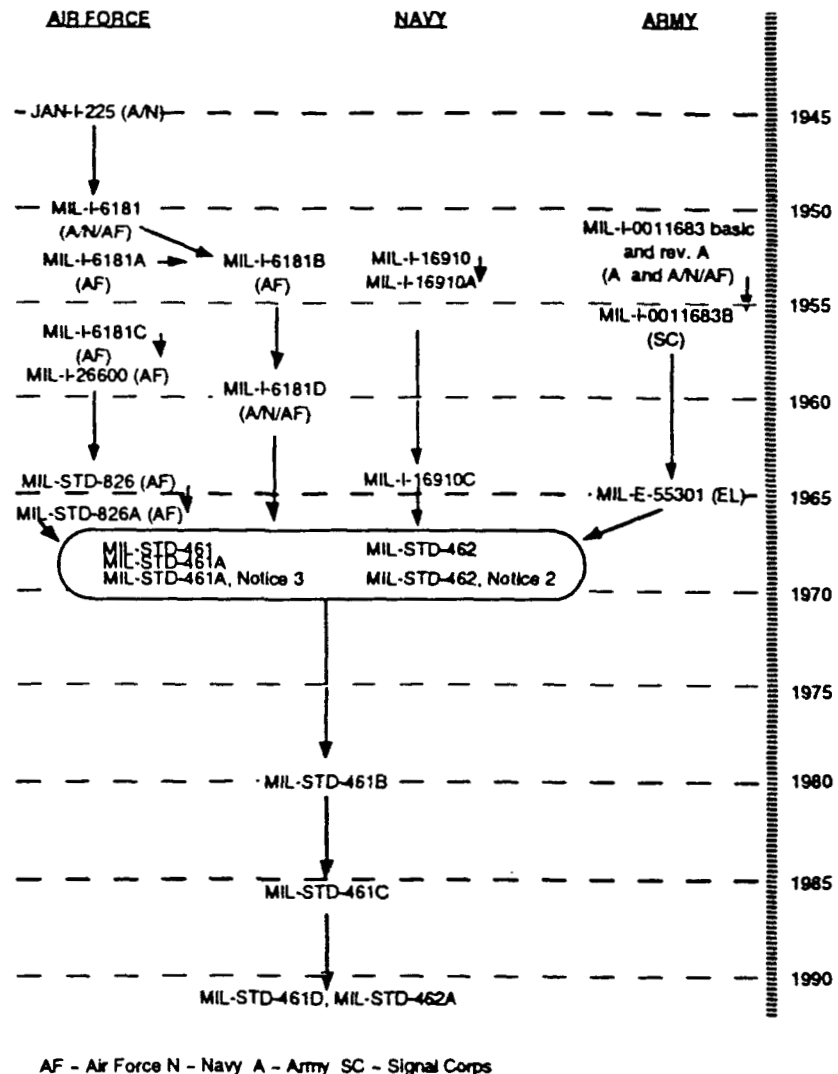


Figure 1: Abbreviated genealogy of military EMI specifications and standards

JAN-I-225 and MIL-I-6181**The Beginning [4]**

JAN-I-225 was a test method for measurement of RF conducted and radiated noise from 150 kHz to 20 MHz. The source impedance for conducted emission measurement purposes used 8 μ F capacitance from the power feeder to ground to bypass the power supply, and 10' leads from the capacitors to the measurement point. The Ferris noise meter type 32B was specified. The RF input was twinaxial, as opposed to today's single ended coax inputs. Such an input would have had a higher resistance than 50 Ω . A 10 pF blocking capacitor was used to isolate the receiver input from the power source, and convert the required 24" twisted pair receiver input to coax, for the purpose of minimizing stray pickup. This technique was adopted for MIL-I-6181. Radiated measurements require use of the 41" rod. Shield rooms were homemade at that time, and instructions are included in the case of broadcast reception to tune to another nearby frequency. No limits are included in JAN-I-225. The original concept was to put limits in equipment procurement specifications. Note: This specification was for general aircraft, but it antedates the creation of the Air Force from the WWII era Army Air Force. Hence the specification designation Joint Army-Navy (JAN).

CE and RE limits in MIL-I-6181 were derived from measurements made on a B-29 in much the same configuration as specified in JAN-I-225, which is based on these B-29 measurements. During WWII, the Boeing plant in Wichita, Kansas produced four B-29s per day. With this mass production set up, there were enough peculiarities between individual aircraft, that, like automobiles, some were considered good, and some were lemons. In particular, some had noisy communications systems, and some were quiet. One aircraft that was particularly quiet was the subject of

the test that preceded and influenced both JAN-I-225 and MIL-I-6181.

MIL-I-6181B (and NADC-EL-5515) [5]

MIL-I-6181B is noteworthy not only in itself, but also for the associated explanation of the test methods and limits. This explanation is included in report NADC-EL-5515, released in 1955. A copy of this model study is included as an appendix to this document. It is highly recommended that the serious EMC professional read this work. Although the results are somewhat dated, the problem solving approach is invaluable. The author, Mr. William Jarva, laid the foundation of today's EMC practice with this specification. His report, including such items as the use of the 41" rod antenna, the antenna-EUT separation, NB/BB discrimination, absence of magnetic requirements and many other test set up details, explains the how and why of military EMI testing yet today. It is astonishing to the author (and part of the motivation for writing this document) how many misinformed explanations EMC engineers will give for the hows and whys of EMI limits and test methods. Most of these pseudo-explanations are less than satisfactory as a motivation for significant design effort to achieve compliance. Mr. Jarva's report is elegant and simple and provides complete engineering justification for each method and limit. That these limits are not necessarily applicable today, 40 years later, is not surprising and in fact accurately predicted by Mr. Jarva, who gave the precise conditions under which MIL-I-6181B limits and methods would need revision. Mr. Jarva's report is the subject of detailed discussion in this document.

MIL-STD-826: An Attempt at a Tri-Service Standard

MIL-STD-826 was meant to be a Tri-Service standard, but it never gained acceptance outside the Air Force. It is noteworthy for departing from

voltage based CE measurements to current measurements. There was an emphasis on automated EMI testing, even to the point of showing radiated test set ups with all frequency band antennas in place in the shield room connected to a coaxial switching network and automated EMI receiver. MIL-STD-826 extended emission control below 150 kHz. Previous Air Force specifications started at 150 kHz.

MIL-STD-461, MIL-STD-462:

The Tri-Service Standard

Heavily influenced by MIL-STD-826, MIL-STD-461/462 finally became the Tri-Service standard. One year after its release, it was modified as

revision A, and heavily modified by rev. A Notice 3 in 1970. Revision A Notice 4 (the Army version) is noteworthy in that, alone of the Services, it required LISNs instead of feedthrough capacitors as standard source impedances for power line conducted emission data. The Services eventually all went their own way, releasing Service unique Notices. Finally, in 1980, MIL-STD-461B combined all the -461A Notices. MIL-STD-461B, and the current MIL-STD-461C (1986 effectivity) are organized into Parts, some of which are Service unique, and some of which apply to particular types of equipment.

ISSUES

The balance of this document discusses various engineering issues of EMC, and the relationship the EMI limits and test methods bear to these issues. The author does not claim that this is an exhaustive analysis, only that these are the issues most often confronted by the author and his colleagues. Readers are welcomed and requested to contact the author with constructive criticism. Especially valued are inputs of historical interest by those who made the history. The author can be contacted through the EIA G-46 committee.

Narrowband/Broadband: An Important Distinction Obscured by Automated Test Procedures

Narrowband (NB) and broadband (BB) signals must be so qualified relative to a measurement bandwidth. With the exception of a pure sinusoidal continuous wave (narrowband) and an ideal delta function (broadband), no signal may be intrinsically defined as NB or BB. In terms of the receiver bandwidth, a narrowband signal is one which has only one spectral component in the receiver bandwidth at the tuned frequency. Likewise, a repetitive broadband signal may be ideally defined as having more than one spectral component in the receiver bandwidth. There are three practical checks for NB/BB determination:

- Increase the measurement bandwidth. If the signal level does not change the signal is NB. If the signal *power* changes in direct proportion to the bandwidth, i.e., 10 db per decade bandwidth change, the signal is incoherently BB (thermal in nature). If the signal *voltage* increases in proportion to the band-

width change, i.e., 20 dB per decade bandwidth change, then the signal is impulsively or coherently BB.

- Tuning off the frequency of maximum amplitude by two impulse bandwidths. A change in amplitude of 3 dB or more implies a NB signal.
- The pulse repetition rate may be compared to the measurement bandwidth. If the repetition rate is less than the measurement bandwidth, the signal is BB.

A narrowband signal is one whose spectral component at the tuned frequency lies within the bandwidth of the measuring device. Thus, increasing the measurement bandwidth does not increase the measured signal amplitude, and in fact decreases the signal to noise ratio due to a higher noise floor. The spectral components of a broadband signal, in contrast, do not all fall within the measurement bandwidth, and therefore *the amplitude can only be expressed as the measured amplitude per the measurement bandwidth*. Usually the bandwidth is normalized to an agreed upon reference such as 1 MHz. This standard came about with the advent of the Stoddart NM-10/20/30 receivers and the Empire Devices NF-105, both of which measured broadband signals by a signal substitution method employing impulse generators calibrated in terms of dB μ V/MHz. The measurement technique was as follows. The technician, using headphones, listens to the detected audio output of the EMI receiver as he tunes through the controlled spec-

trum. When a signal is detected, it is identified to be either continuous wave - like (CW) or impulsive. Depending on the type, either the carrier or field intensity detector is utilized (CW signal) or the peak detector (impulsive). The measured signal is replaced by a calibrated impulse generator (IG). This signal source is adjusted until the EMI receiver meter deflection is the same as for the measured signal. Since a broadband source is used for the substitution, a PEAK detector must be used, even if the signal to be measured is cw. The IG can be used only since each receiver tuning head has a fixed IF bandpass, and thus the IG response can be compared to a cw limit through knowledge of that IF bandwidth. However, since various stages of the RF and IF can and will mistrack, accounting for such mistracking is necessary. This necessitates calibration factors which vary with frequency. The signal source setting is recorded vs. frequency. Corrections for the pickup device losses and/or transducer factor are added and the result compared to the limit. Again depending upon the type of signal, the corrected signal level is compared to either the narrowband or broadband limit. Notice that there is no broadband correction factor to normalize the receiver bandwidth to 1 MHz. This has already been done by recording the impulse generator setting. To the extent that the EMI receiver measurement bandwidths adequately model the victim receivers requiring protection, this is an elegantly simple, and relatively foolproof test procedure. This test procedure began to be misinterpreted with the advent of automated EMI receiver systems, also known as surveillance systems. White Electromagnetics, Inc. built the first such system from receivers built by the Communications Electronics company, the forerunner to Watkins-Johnson. It was followed by the very successful Fairchild FSS-250 Spectrum Surveillance System. These systems used a flatbed plotter so a pen could follow the excursions of the meter display as a ramp generator drove the pen across the paper tracking with the frequency scan. The plotter

paper was precalibrated by inputting known signals (impulse generators/signal generators) at levels corresponding to the limit of interest adjusted for the pickup device characteristics. Narrowband and broadband limits had to be hand drawn on the paper using the machines' response to the known inputs for guides. When the data was taken, excursions above the limit(s) had to be examined and evaluated to determine which limit was applicable. This added complexity was a source of confusion. This confusion was only aggravated by the increasing use of spectrum analyzers. The first spectrum analyzer to gain acceptance for EMI testing purposes was the Hewlett Packard 8550 series, using the HP141T mainframe. This was followed by the hugely successful, highly automated HP856X series, which at the present time seem to be replacing true EMI receivers at many test facilities. Now all connection to the original procedure for broadband sources has been lost. In fact, the procedure has been reversed. The signal is not listened to, and the normalization of a broadband signal to a 1 MHz bandwidth amplitude is done by analysis. Thus, a VLF broadband signal measured in dB μ V with a 1 kHz bandwidth is routinely extrapolated to increase by the 1st power (not the square root) of the ratio of 1 MHz to 1 kHz. This requires an assumption of linear extrapolation of signal spectrum occupancy across three orders of magnitude. In the original procedure, the receiver response to a signal identified to be broadband was compared to the response to a calibrated broadband source. Now a signal demonstrated to be broadband with respect to a 1 kHz measurement bandwidth is *assumed* to be purely impulsive with much greater than 1 MHz bandwidth. This is clearly a questionable practice at best. The very units of broadband measurements, dB μ V/MHz, imply an impulsive signal. Only impulsive signals can be described with such units. An incoherent broadband source would be described by dB μ V/Hz^{1/2}. Electronic equipment, either analog or digital, hardly has the impulsive characteristics of ignition systems.

What probably should have happened when spectrum analyzers began to be used is that the broadband limits at low frequencies should have been stated as dB μ V/kHz, as in the earliest specifications. Then the large extrapolations would have been eliminated.

Differentiation between impulsive and random noise is not necessary if the only analysis to be performed is translating from EMI test data to predicted noise levels coupled into a potential victim receiver's IF bandwidth. Regardless of the normalization factor, the right answer will be obtained, as long as the noise characteristic assumed by the analyst is the same as that used in the EMI data reduction, and the disparity between the test impulse bandwidth (IBW) and victim IBW is not too great. For example, assume that an E-field intensity of 30 dB μ V/m is measured at 25 MHz, using a receiver with a 10 kHz IBW. If the signal is determined to be broadband with respect to the IBW, then normalization can be calculated as 30 dB μ V/m + 20•log (1 MHz/10 kHz) = 70 dB μ V/MHz/m. The noise field intensity available to a different receiver with a 25 kHz IBW may be calculated as 70 dB μ V/MHz/m + 20•log (25 kHz/1 MHz) yielding 38 dB μ V/m. If in this example the test IBW had been 1 kHz, the measured level would have been 20 dB less, but the correction factor would have been 20 dB greater, and it wouldn't theoretically matter. But as a practical matter, the large difference between the test IBW and the victim IBW adds uncertainty that the signal is truly BB. It is easier to extrapolate from 10 kHz to 25 kHz than from 1 kHz to 25 kHz. For this reason proposed MIL-STD-461D will specify minimum BW for emission measurements. MIL-STD-461D is the first military EMI standard to eliminate BB/NB discrimination and specify instead receiver BW. Some of the pre-MIL-STD-461 standards and specifications required the use of particular EMI receivers, which is tantamount to specifying receiver bandwidths; but this has not been done for at least 25 years.

It must be pointed out that receiver IF bandwidth is not the only issue of concern when addressing receiver susceptibility. Receivers are not only sensitive at the IF stage. Depending on design, the RF mixer stage may be susceptible to a much larger spectrum than the IF bandwidth. An extreme example is the AN/ARC-182, a VHF/UHF communications transceiver. The AN/ARC-182 is essentially an AN/ARC-164 (UHF transceiver) with a mixer which up converts the entire military VHF band into one UHF channel. Now it is very important to know just how *broadband* the EMI is, because the RF front end bandwidth of our victim is over 100 MHz! Stretching reality to make a point, if noise field intensity just meeting the RE02 BB limit from 60 MHz to 160 MHz were available at the ARC-182 antenna, the total field intensity would be over 100 mV/m. With a typical antenna factor of 10 dB over this frequency range, the mixer power would be -10 dBm. Either the mixer would be near saturation, or front end attenuation protecting the mixer would desensitize it to weaker signals.

It may come to pass at some future time that these terms will be of historical interest only, since the current revision process leading towards MIL-STD-461D and MIL-STD-462A is deleting narrowband/broadband criteria for categorizing emissions. For the record, it should be understood that the purpose of this significant change in EMI test and characterization is due to frustration on the part of the Services with the lack of understanding of these concepts demonstrated by practitioners in the field [2]. The author is no stranger to this feeling, being employed at the time of this document in a similar position, and having much the same responsibilities as the EMC engineers employed by the DoD. A philosophical question might be posed as to whether EMI limits and tests should be simplified to the point where an acceptable percentage of decent test reports are received; or whether the data necessary to support system level EMC analysis should be demanded

from contractors who have the responsibility to hire or train qualified personnel. The author believes the latter option promotes excellence in the field, while the former option, by reducing procedures to the lowest common denominator, produces "proof of compliance" data only. It is interesting to note that the recent drive by NAVAIR to certify EMC engineers and technicians was precisely for the purpose of promoting excellence in the field. It would seem that various forces are pulling in different directions within the Services (although the author must point out that the Tri-Service committee writing MIL-STD-461D and MIL-STD-462A is in agreement that the NB/BB distinction must go.)

Technical Rationale for Conducted EMI Requirements

(portions of this section are adapted from a paper by the author and Mark J. Nave, EMC Services) [6]

Control of power line conducted emissions (CE) is one of the fundamental requirements in any EMI standard. There are three reasons for specifying CE limits on the power bus. These reasons are to control the level of voltage ripple, electric field radiation, and noise currents on the structure (control of platform skin potential).

Discussion of Voltage Ripple

Most commonly used electrical equipment requires power from a voltage source; that is, a constant - voltage source. Fluctuations, or distortions of the supplied voltage level, by whatever mechanism, may disrupt the normal operation of the equipment. One goal of CE specifications is to limit distortions of the supplied voltage level by limiting the voltage drop across the power bus caused by each electrical load. This EMI coupling method is often referred to as common impedance coupling. The bus impedance is characterized by three distinct regions: resistive, inductive, and transmission line. At lower frequencies the bus

behaves in a purely resistive manner. At some frequency, the LR corner frequency, the bus inductive reactance equals the bus resistance. Above this frequency the bus impedance is predominantly inductive. At even higher frequencies, where the bus dimensions approach $\lambda/8$ in length, the distributed capacitance becomes significant, and the bus behaves as a transmission line. Figure 2 shows the impedance of a typical bus vs. frequency.

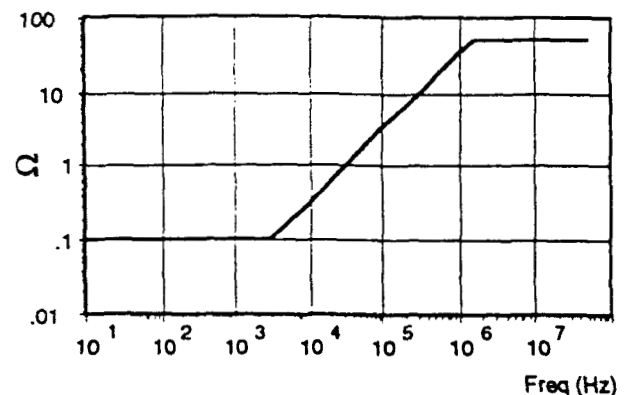


Figure 2: Generic power bus impedance vs. frequency curve

Bus Resistive Frequency Range

The bus impedance is resistive at frequencies below the LR corner. Over this frequency range traditional power quality considerations are applicable. That is, wire gauge is specified to support the expected load current with a specified voltage drop. Higher current users simply drive larger gauge wiring.

Bus Inductive Frequency Range

As frequency increases, a breakpoint occurs where the bus inductive reactance just equals its resistance. This breakpoint is dependent on the bus configuration, which determines the inductance. Above this breakpoint the bus impedance rises smoothly at 20 dB per decade. In this region, changing wire gauge and physical configuration

has a slight effect on bus impedance; but the dependence of bus inductance on wire diameter and wire separation is logarithmic. Above the LR corner frequency, therefore, the bus impedance cannot be realistically reduced to any arbitrarily desired level. Once the bus physical configuration is defined, the bus inductance is fixed. Control of voltage drop across the bus must be achieved by limiting the user currents.

The EMI control methodology above the LR corner frequency range stands in direct contrast to that used below the LR corner frequency. At low frequencies the coupling problem is unilaterally solved by increasing wire gauge. At higher frequencies the coupling problem is unilaterally addressed by limiting EMI currents in absolute magnitude.

The Power Bus as Transmission Line

At a frequency where the bus length approaches $\lambda/8$, the bus impedance levels out to be the transmission line characteristic impedance of $\sqrt{L/C}$, where these values are now the *distributed* inductance and capacitance of the transmission line. Figure 2 shows a generic bus impedance vs. frequency plot.

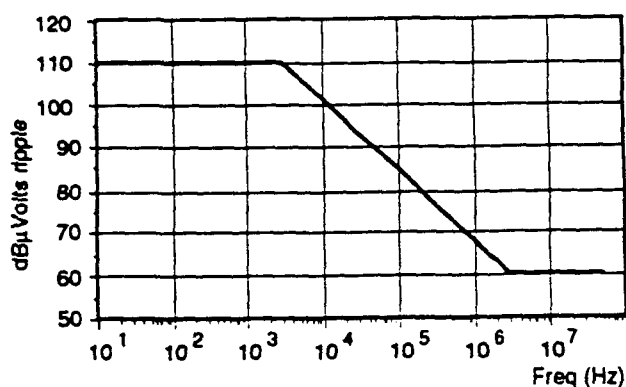


Figure 3: *Generic ripple voltage limit*

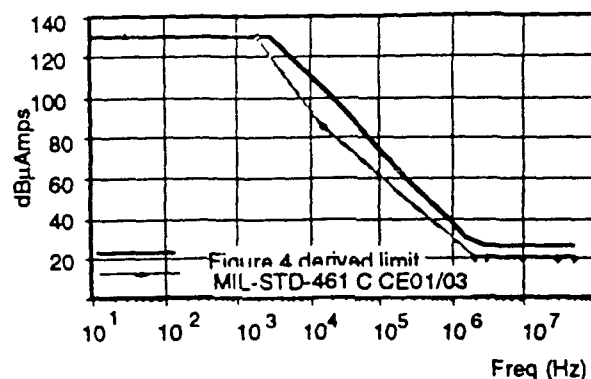


Figure 4: *Comparison of MIL-STD-461C CE limit derived from Figures 2 and 3*

To illustrate, a noise current limit similar to MIL-STD-461 basic through revision C may be derived by using the bus impedance curve of Figure 2 with a ripple voltage specification, as shown in Figure 3. [5-9] To derive the current limit, one may simply divide the ripple voltage of Figure 3 by the bus impedance of Figure 2. The results are shown and compared to the MIL-STD-461 limits in Figure 4. The voltage ripple limits of Figure 3 were picked to provide a suitable safety margin between emission and susceptibility levels of sensitive equipment at the time MIL-STD-461 and forerunner standards were derived. Inspection of Figures 2 through 4 reveal that, below about 1.6 kHz, 3 Amps of load induced noise current may be accommodated. Above 1.6 kHz, the noise current decreases at 40 dB/decade of frequency from 130 dBμA to 20 dBμA at 3 MHz.

Since the wire gauge feeding the load provides the same source impedance for low frequency noise currents as for power current, the power quality ripple approach allocates the equipment noise current on the basis of a percentage of the bus maximum power frequency current. Military CE limits in this frequency range (CE01) did not apply this power quality approach. The CE limit was on a per equipment, not per Amp basis, as shown above. [7-9] Parts of MIL-STD-461B/C

do apply a per Amp CE01 limit. [10,11] MIL-STD-461C CE limits are shown in Figure 4 with the derived limit included for the sake of comparison.

Concluding this discussion of voltage ripple control, it is noted that CE limits may be derived analogously with power frequency voltage drop analysis, taking into account the complex impedance of the bus, in addition to its resistance. It is important to understand that in the foregoing discussion, the noise currents were assumed to flow in the bus in the same path as the power frequency current.

Discussion of Electric Field Radiation from the Power Bus

Currents flowing on a conductor generate electric, magnetic, and electromagnetic fields. Within the dimensions of a typical platform (airplane, tank, etc.), the E & H fields arising from power frequency currents follow the basic laws for the static case of electric field lines of force between objects at different potentials and the right hand rule for calculating the direction of the magnetic field surrounding a current carrying conductor. At higher frequencies where the bus length and the separation from the bus begin to approach a wavelength, radiation of a traveling wave becomes efficient. Figure 5 shows E & H field vectors emanating from a single feeder above groundplane return configuration over the frequency range where CE limits are imposed.

Control of these fields is important in frequency bands where sensitive receivers operate, in order to avoid desensitization or outright interference. Because the E & H fields are proportional to the voltages and currents present on the bus, these must be controlled to a degree compatible with the platform receiver systems' sensitivity. That is, the platform CE-generated *unintentional* radiation must not interfere with the platform's desired spectrum usage.

Besides a voltage and current dependence, the radiated fields are highly dependent on the physical layout of the power bus. The single feeder above a groundplane found in most metal structure platforms is the most efficient radiator per unit noise current. Some platforms utilize dedicated wires for return current. Radiation from such buses is much smaller per unit noise current since the return current is constrained to flow in a path much nearer the feeder current.

The relationship between cm currents and electric field radiation is developed in [12]. It would be fortuitous if CE limits sufficed to optimally control electric field RE. Incidentally, if cm from *both* power and signal lines were controlled per the algorithms in [12], then the RE02 test below 50 MHz could be eliminated. Control of cm, separately from dm currents, would be a superior method of RE control from the point of view of test repeatability. This test is similar to CE04 of MIL-STD-462. [13, 14] The author feels that this method, extended to 100 MHz, plus perhaps surface current measurements on enclosures, would

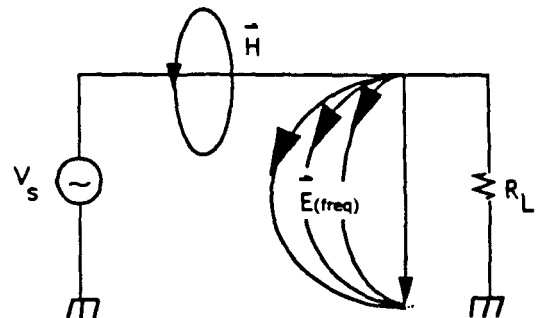


Figure 5: Static E & H fields from single wire above groundplane return

be a superior requirement below 100 MHz, for accuracy and repeatability in bare wall shielded enclosures. See also SAE ARP 1972, "Recommended Measurement Practices and Procedures for EMC Testing", Sections C and J for a similar approach. [15]

Discussion of Structure Currents (Vehicle Skin Protection)

Power frequency currents returning on metallic structure typically see a milliohm or sub-milliohm resistance on that path. From the point of view of wire weight, cost, and voltage available at the load, returning current on structure is the minimum cost option. If it is desired to provide for above structure current return, then, other things equal, the length of the bus wiring is doubled, and the wiring cross-sectional area must be doubled as well. Thus, the mass of the wiring is quadrupled. The cost of the wiring is more than doubled.

At higher frequencies, however, inductance, skin effect and the tendency of higher frequencies to radiate around structural discontinuities combine to significantly raise the impedance of the structure return path. If significant voltages are dropped across a portion of structure used as an antenna groundplane, this will couple into the antenna-connected receiver (assuming the interference is in-band). Over the frequency range controlled by CE limits (below 50 MHz for MIL-STD-461C and earlier releases of MIL-STD-461, and below 30 MHz for proposed MIL-STD-461D and pre-MIL-STD-461 CE limits), antenna groundplanes can consume the entire platform. Thus, although power frequency currents may be allowed to flow on structure, care must be taken to force noise current return paths to remain in the equipment.

CE Measurement Procedures

The measurement of CE in a precise and accurate fashion requires that the source impedance of supplied power be standardized over the frequency range of interest. If the source impedance is too high, an artificially low current will be measured.

Source Impedance Standardization for CE Current Tests

If noise currents are to be measured, then the source impedance should be as low as possible, in order that the unit under test, acting as a current sink, sees a minimum impedance voltage source. The 10 μF feedthrough capacitor described in SAE ARP 936 is the military standard for this purpose under MIL-STD-461 basic through revision C. [16] The 10 μF feedthrough capacitor places 40 Ω of reactive impedance line to ground on the power bus at 400 Hz. Thus it draws about 3 Amps from a nominal 115 Vac bus. At 15 kHz and above (test method CE03) it is a virtual short circuit. However, the use of a one meter power line 5 cm above a groundplane between EUT and capacitor changes the source impedance as shown in Figure 6.

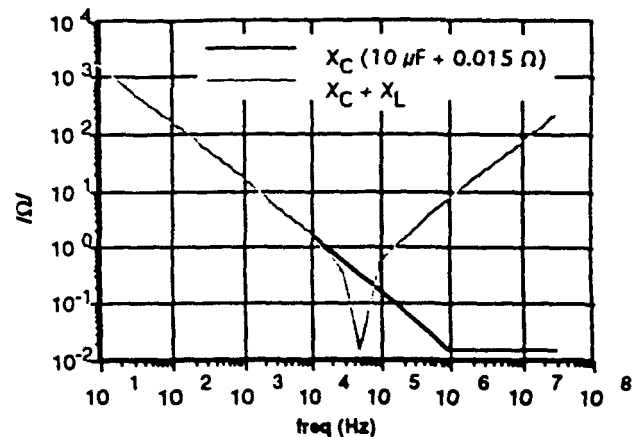


Figure 6: Source impedance of 10 μF capacitor and capacitor with 1 meter wire

The use of the 10 μF feedthrough capacitor for test method CE01, 30 Hz to 15 kHz, is even more problematical, because at lower frequencies the feedthrough capacitor does not effectively short across the power source impedance. This is presently uncontrolled under MIL-STD-462. For example, a nominal 4 Amp 28 Vdc input switched mode power supply (SMPS) failed CE01 by draw-

ing over 3 Amps of ripple current. Placing 10,000 μF across the 10 μF feedthrough capacitors reduced the ripple current to negligible levels. The SMPS had gone into oscillation due to the high source impedance (EMI filter line inductance) between the power source and EUT. Further specification of the power source is necessary. A reasonable control would be to specify a maximum voltage ripple at the equipment under test (EUT) during the current measurement. If the ripple were too high, then the source impedance would have to be lowered. This is an operational description of the power source impedance which can apply to any situation.

Using a large EMI filter bypass capacitance to modify the bus source impedance also changes the MIL-STD-462 prescribed standardized source impedance in the lower frequency range of requirement CE03. This is not desirable, since it can cause a specification exceedance that would not exist if the source impedance were the 10 μF capacitors alone. A solution here is to place just enough inductance in the lines between the large capacitor and the feedthroughs to decouple the parallel shunting effect. This should only occur above a nominal impedance and frequency. It (fortuitously/intentionally?) works out that the breakpoint frequency between CE01 and CE03, 15 kHz today, 20 kHz pre-MIL-STD-461B, is where the 10 μF capacitor has a reactance of 1 Ω . If enough inductance is placed in the line to also give a series reactance of 1 Ω at this frequency, then we have effectively decoupled the two different purpose capacitors at and above this frequency. The inductance that performs this function is 8 μH .

Starting with a feedthrough capacitor, we have progressed to two line to ground capacitors connected by a series inductor. In essence, we have developed a *line impedance stabilization network*. The next section develops this concept further.

Source Impedance Standardization for CE Voltage Tests

If noise voltages are of interest, then a non-vanishing power source impedance is desired. A Line Impedance Stabilization Network (LISN) provides this function. A typical LISN is schematically shown in Figure 7. The LISN provides a low source impedance at the power frequency (at current levels below its rating) and increasingly decouples the source impedance from the bus at higher frequencies. See Figure 2 for a generic LISN impedance curve. The high frequency plateau is usually 50 Ω .

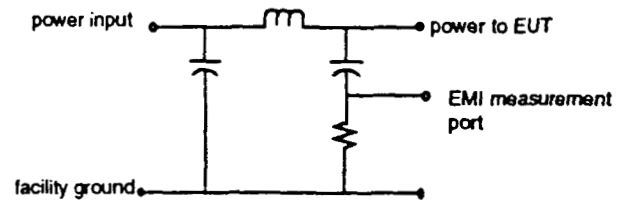


Figure 7: LISN schematic

Conduction Modes: Common Mode and Differential Mode

There are basically two types of noise present on a power bus: differential mode (dm) and common mode (cm). The design of a power line filter is also done with cm and dm parts. The design of each of the two different sections is slightly different, a reflection of the different requirements of each.

Single Phase Differential Mode

Differential mode noise is the simplest kind of noise. DM noise current flows in the same path and direction as the power frequency current, as depicted in Figure 8a.

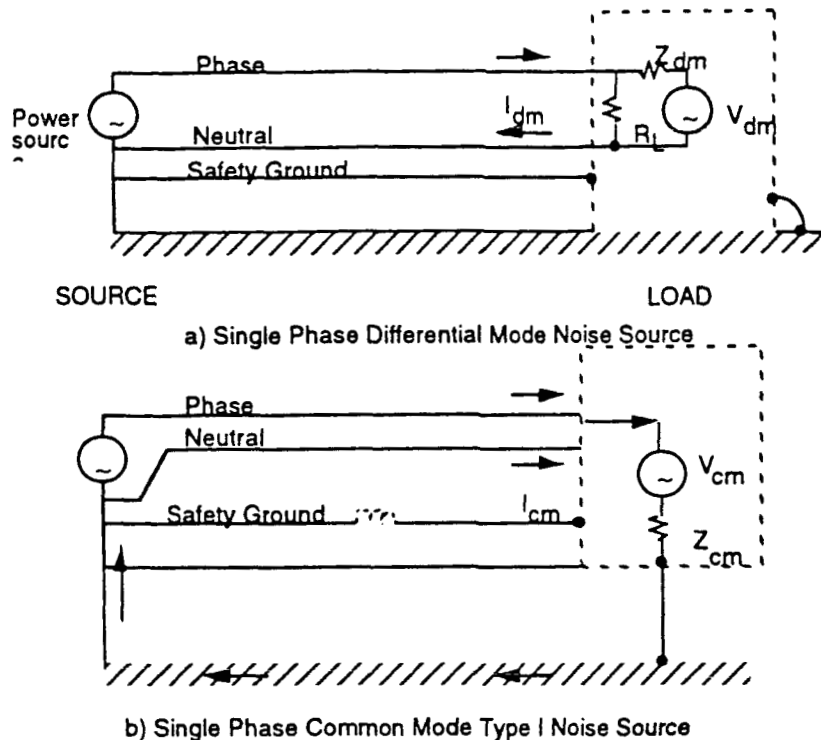


Figure 8: Single phase dm and cm noise
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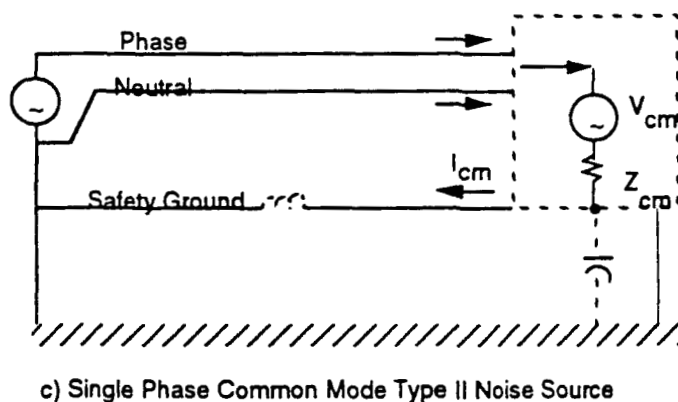


Figure 8: Single phase dm and cm noise (continued)
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Differential mode noise is also referred to as the normal, or longitudinal mode. It is also characterized by the currents flowing in the phase and neutral lines being 180° out of phase.

Single Phase Common Mode

Common mode noise is characterized by the noise currents flowing in the feeder and return lines being in phase. This propagation mode is shown in Figure 8b and 8c. There are two sub-categories of common mode noise referred to as type I and type II. Figure 8b illustrates common mode type I noise. In this mode the common mode noise source is referenced to the equipment chassis, which is in turn grounded to the reference plane. The impedance of the reference ground plane is much less than the inductance of the safety ground. Consequently, the return current path is through the ground plane. In the vast majority of cases, common mode type I is the cause of radiation problems, given good PCB design. This is because the large loop area enclosed by the current path makes a fairly efficient radiator. The differential mode currents, in contrast, enclose a very small loop area, the product of the length of the cable and the separation between conductors.

Common mode type II typically occurs when the chassis is isolated from the reference plane. Common mode type II is depicted in Figure 8c. When the equipment chassis is isolated from the reference plane, a

parasitic capacitance (on the order of 10's of pF) is in series with the return path through the reference plane. This high impedance capacitance results in the safety ground now appearing as the lowest impedance return path, hence carrying most of the current. The radiating loop area enclosed by the current path of common mode type II is much smaller than that of common mode type I, and therefore much less of a radiation problem. In real systems, both modes are simultaneously present to some degree.

Conduction Mode Isolation

The rationale in proposed MIL-STD-461D for limiting higher frequency conducted emissions (measuring both modes simultaneously) is to limit the radiation induced by the cm currents. [2] A problem arises when the dm emissions dominate the cm emissions. Under these circumstances one must "unnecessarily" use excessive dm filtering to bring the total emissions within the required limits. Isolation of conduction modes may provide the information and the means of optimally specifying and controlling CE.

Test Methods

The need and purpose of independently measuring cm emissions has been well established. Those familiar with testing will be familiar with the current probe method presently used for isolating modes shown in Figure 9. Figure 10 shows how the same mode rejection can be performed with LISNs, using differen-

tial mode rejection networks (DMRN) connected to LISN ports (with the unique benefit of working equally well in the time domain as the frequency domain.)

Specifying and controlling conducted emissions is like solving one equation with two unknowns; the unknowns being the conduction modes, cm and dm. One mode must be attacked at a time. For a LISN based measurement such as under the proposed MIL-STD-461D CE03, the DMRN makes this possible, essentially reducing the problem to solving one equation with one unknown. [19]

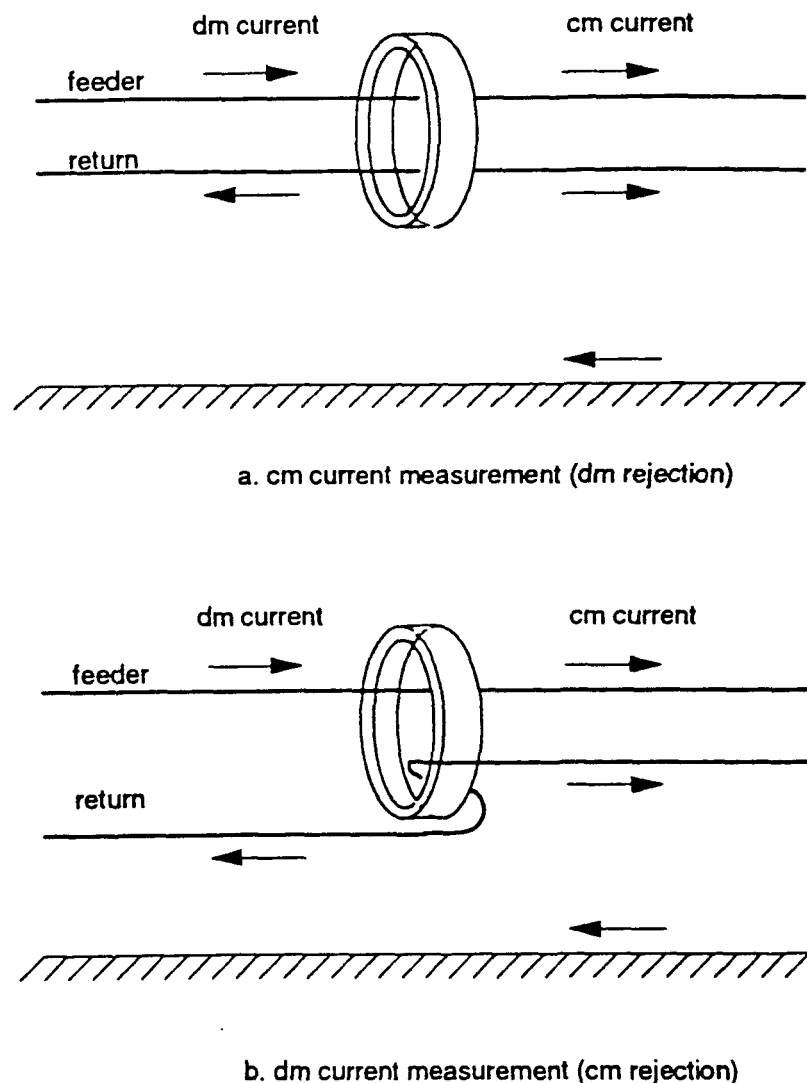


Figure 9: Mode selection/rejection with current probes [17]

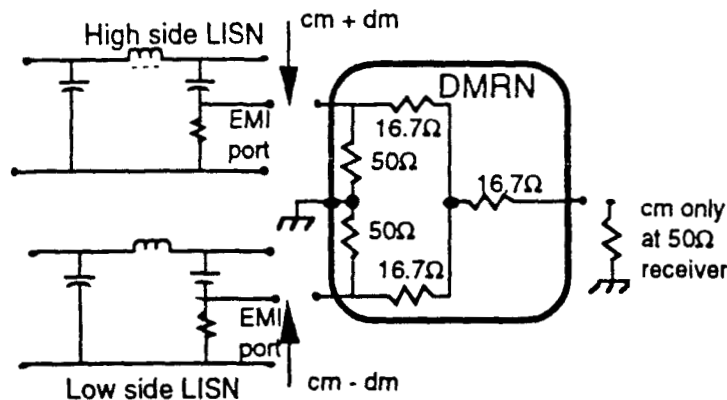


Figure 10: *DM rejection/CM selection with a DMRN* [18]

Historical Military Conducted EMI Control

Motivation for control of power line CE has been established. Methods for measurement of CE have been discussed. A review of actual EMI standard test set ups for measuring CE is now instructive. References 20-22 (obsolete), allowed for line to line measurements if the return path was not grounded at the test sample (no LISN). Every other CE voltage measurement in the frequency domain has been line to ground. [23-35]

Current measurements are made separately for each line. [13,14,36,37] The first instance of a LISN being specified for CE measurement is in MIL-I-6181B, released in 1953. [25] MIL-I-6181B/C/D and another Air Force unique standard, MIL-I-26600, all required the use of one LISN per power supply *feeder*. [25-27, 32] Return was routed through the metal chassis of the LISN, using the ground studs. Figure 7 shows a typical LISN topology. Figure 11 shows a CE test set up diagram from MIL-I-6181B.

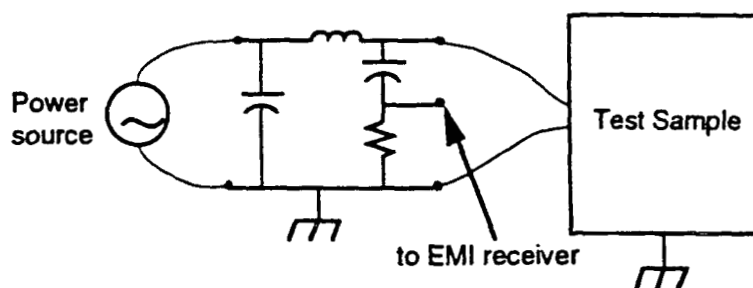


Figure 11: *Single LISN per feeder test set up per MIL-I-6181B*

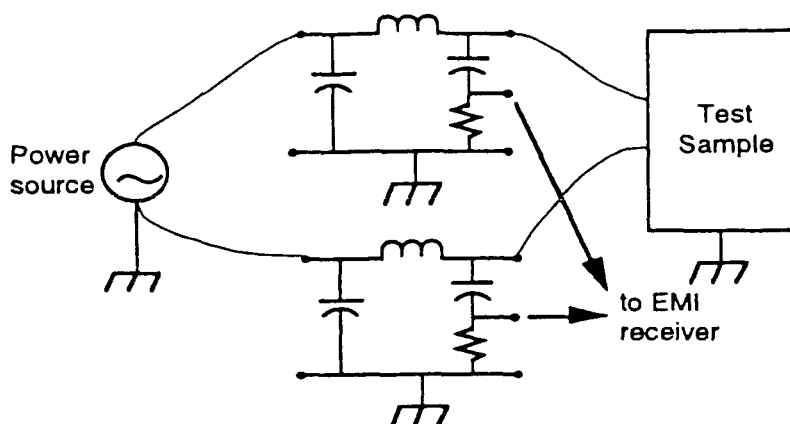


Figure 12: Functional equivalent of test set up under MIL-STD-826 and MIL-STD-461

Navy and Army EMI standards of this pre-MIL-STD-461 era called out LISNs for every ungrounded power conductor, as shown in Figure 12. [30, 34] When MIL-STD-826 and MIL-STD-461 replaced the voltage limits with current limits and replaced the LISN with the $10\mu\text{F}$ feedthrough capacitor, it was the Army/Navy philosophy of one impedance standardizing device in each ungrounded power conductor which was adopted. The limits, however, did not change significantly until MIL-STD-461A, Notice 3 significantly relaxed the low frequency portion of CE03.

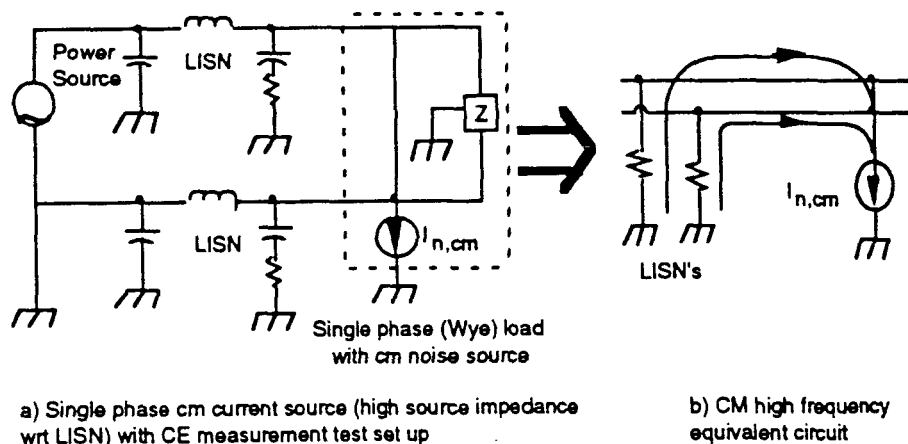
Structure Current Return vs. Above Ground Current Return

The above noted test set up differences are very important. In the case of current returning on

platform structure, the Air Force set up of one LISN per feeder was the best simulation. Filter design is straightforward and codified into MIL-F-15733. [38] Filter attenuation measurements are easily performed in an unbalanced 50 Ω test set up per MIL-STD-220A.[39]

If, however, a current return path is provided above ground, then the test method of providing a standard source impedance in each conductor is a better simulation. This is the method of the early Army/Navy and present CE measurement. But in this type of test set up, it is important to note that (if present) both dm and cm voltages/currents are measured simultaneously. See Figure 13 which also illustrates an interesting side effect to using one LISN per conductor, rather than routing return through the LISN ground terminal binding

Figure 13:
Two LISN test set up showing cm and dm currents.
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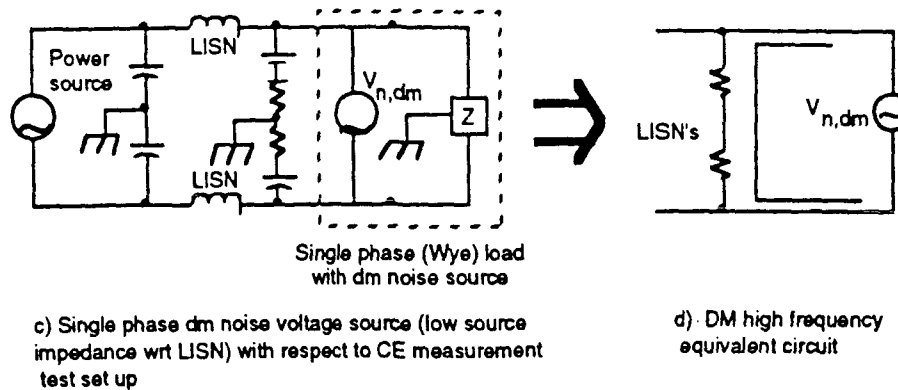


Figure 13b: Two LISN test set up showing cm and dm currents. (continued)
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posts. Recall that the LISN was originally designed for the test set up of Figure 11. Regardless of mode, in such a configuration the LISN presents the source impedance characteristic of Figure 2. But with the test set up of Figure 12, the dm source impedance is doubled (inductors in series), and the cm source impedance is halved (inductors in parallel), shown in Figure 13b. Thus at high frequencies, the use of two 50 Ω LISNs results in a dm source impedance of 100 Ω , while the cm source impedance is 25 Ω . The effect on the dm impedance is opposite that expected of the *in situ* case, because the dm inductance is much smaller than the cm inductance (due to relative loop dimensions.)

From the point of view of platform level compatibility, the difference between cm & dm noise is purely a function of the geometry of the bus. If the bus is laid out in the single feeder above structure return as shown in Figure 14, then what are normally considered cm & dm currents flow in identical paths (same as the power frequency current) and have identical effects on voltage ripple, electric field radiation from the bus, and structure currents. In the case of a dedicated wire return above ground as shown in Figure 15, however, dm current flows in the feeder and return the same as the power frequency, while cm current is drawn from both of the feeder and return wires and returned on structure. In this case the dm

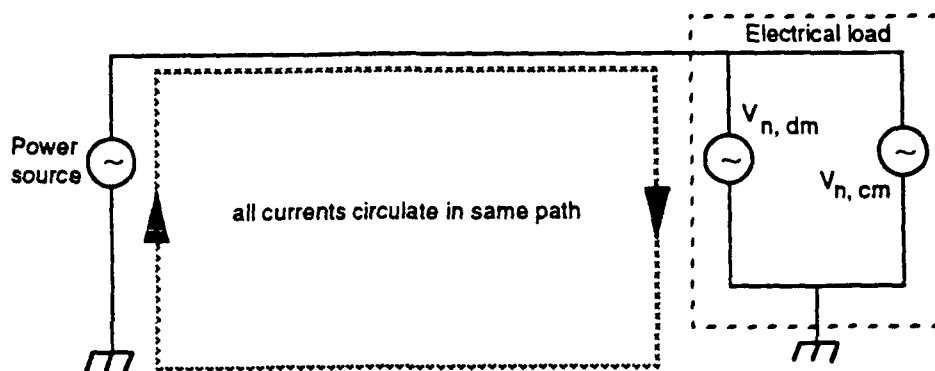


Figure 14: Noise current circulation in structure return bus

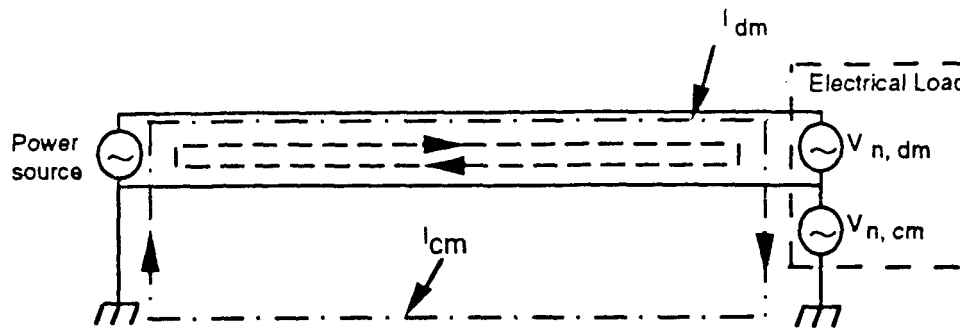


Figure 15: Circulation path of noise currents in above ground current return bus

currents contribute much less to the electric field radiation because of the vastly reduced radiating loop area. Therefore dm currents can be controlled with ripple voltage as the main constraint. The cm current must be controlled as before.

It was previously remarked that an above ground current return bus imposes large cost and weight impact to the platform. By taking advantage of the lower dm radiation efficiency, dm currents may be controlled based on bus voltage ripple considerations only. The dm section of the equipment EMI filter may then be reduced or even eliminated for light loads. If all bus loads are thus treated, some of the previously mentioned impacts may be ameliorated. To give a numerical example, consider the following. The CE voltage limit as measured in a LISN is specified to be 1 millivolt at the power supply switching frequency of 500 kHz (proposed level under MIL-STD-461D to control radiation from common mode emissions). Use of an above ground current return bus allows differential mode emissions to be controlled by load induced voltage ripple levels only. The ubiquitous military HF receiver, AN/ARC-190, picked as likely the most sensitive HF load on the bus, can stand a CS02 level of 100 mV at its tuned and image frequencies. Allowing for a 6 dB safety margin, the load induced dm noise from our power supply can be as large as 50 mV. This is a 34 dB relaxation.

Another example is the control of CE in space

systems. Most or all space systems have above ground power distribution. Furthermore, intentional use of the spectrum (receivers/transmitters) occurs at frequencies well above that controlled by CE limits, typically in the military UHF for manned systems, and microwave bands for unmanned systems. Here the dm CE control may be relaxed as above, but furthermore there is no need to control cm CE (except to the level that crosstalk between adjacently routed cables is not a problem.) Some of these space systems have stringent weight control programs. For instance, the Atlas rocket skin is so thin that it must be supported with compressed air until fuel is installed, since it cannot support its own weight. And yet, most (if not all) of these systems have above ground power distribution and complete CE control per MIL-STD-461 or derivative specifications, such as MIL-STD-1541. [40] Significant weight savings could be achieved in these systems if just the CE control *necessary* was exerted.

There is a problem with MIL-STD-461 basic through revision C and probably including revision D CE limits not existing under the older limits controlling ripple from 150 kHz to 25/30 MHz. The *original intent* for CE was to protect a victim from any single culprit sharing a common bus. Thus the limit was imposed on each EUT regardless of load current draw. Since only linear power supplies existed at this time, the high current source of differential mode noise was filter capacitor charging. This noise is related

harmonically to the line frequency and is mostly gone by 100 kHz, below the controlled spectrum. The remaining source of higher frequency noise is diode reverse recovery, which can be treated with small, lightweight and less costly components, relative to lower frequency, high current noise. MIL-STD-461 conducted requirements, and MIL-STD-462 conducted test methods (above 150 kHz) are written to protect against two types of interference. The conducted *emission* limits and test method protect power line connected receivers from in band voltage ripple induced onto the power line by the steady state operation of the equipment under test (EUT). The conducted *susceptibility* limit covering the frequencies above the audio band ($f > 15$ kHz) forced all new equipment, including receivers, to have sufficient power line filtering that there would be no need for equipment to meet these stringent CE limits. In essence, MIL-I-6181 basic forced new equipment to meet a 1 mV CS level, while holding CE levels to 50 μ V (HF plateau of the CE limit.) Later revisions made both the CS and CE limits higher. The important point to realize is that the CE limits were a *grandfather clause* protecting existing receivers with marginal power line filtering and unshielded antenna lead-ins. Once the older receivers were removed from service, replacement models meeting the intent of the CS02 limits would allow for much less stringent CE03 limits. CE limits have become successively looser, but there still exists large safety margins between CE03 induced ripple and CS02 levels. For instance, the AN/ARC-190 HF receiver meets a 100 mV CS limit at the tuned frequency. This is 46 dB above the ripple induced in a 50 μ A bus by the 10 μ A CE03 HF limit. On a similar note, the broadband CE limit as a *voltage ripple control* is equally outdated. If there is a problem with broadband *dm* CE in an above ground power bus, then it will manifest itself not as a frequency domain issue, but as time domain ripple. Such a requirement is best implemented by deciding on a reasonable or worst case power source impedance

(i.e. LISN) and limiting the time domain ripple induced in the LISN by the steady state operation of the EUT. It is not at all obvious that such a limit is necessary. Victim equipment will have been qualified to CS01/02, which should reflect normal ripple levels (see pertinent paragraph below). The CS06 requirement should be more than sufficient to allay concerns of susceptibility to short duration noise of amplitude greater than the normal ripple level.

Transient Conducted Emissions

A time domain test currently in vogue, but slated for obsolescence under the proposed MIL-STD-461D is requirement CE07, equipment power line turn-on/off and mode switching transients. This requirement is somewhat unique in that no MIL-STD-462 test method exists for compliance verification. CE07 first appears in MIL-STD-461B, released in 1980. MIL-STD-462 has not been revised since the basic release which was concurrent with MIL-STD-461 basic (1967, although many Notices have been issued). No CE07 procedure has ever been included in these Notices. Prior to MIL-STD-461B, and subsequent to MIL-STD-461C (planned post-1992) the replacement for CE07 was/will be CE03T (for transient). This is a frequency domain requirement that limits the signal amplitude at a set of defined frequencies to some larger value tied to the normal CE03 limit. Pre- MIL-STD-461B the limit was 20 dB above the CE03 broadband limit (CE03BB) for mode-switching transients, and 50 dB above the CE03BB limit for on-off transients. Such a frequency domain requirement exists for the purpose of protecting receivers. The CE07 requirement exists to protect power quality. The rationale under MIL-STD-461D for changing back to the frequency domain is that there have been no problems associated with the CE07 performance of equipment. [2] This is definitely not the case for space systems. For various reasons, turn-on transients are of greater interest, and test methods have been

worked out. The author wrote the CE07 test method for the Space Station Freedom Program. With perhaps some bias, he feels it addresses all the pertinent issues. The procedure is included as an appendix to this document, since there is no military specification to cite as a reference. The interested reader will notice that the specified LISN is qualitatively different than that discussed for military CE tests. Figure 16 below shows the topology of such a LISN. This type of LISN was developed to model an above ground power bus. All bus modeling elements are in series with the bus. Notice that this two wire LISN does present a 50 Ω source impedance at high frequencies.

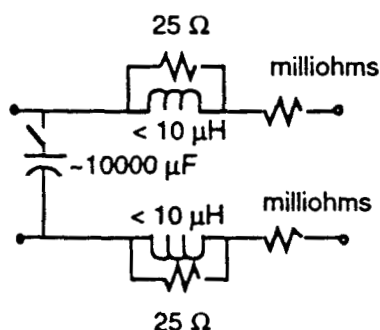


Figure 16: LISN for above ground power distribution

The capacitor is local energy storage. Its ac decouples the LISN from the rest of the power bus, including shield room mounted EMI filters. The capacitor guarantees that voltage drop measured across the LISN is due to the inductive and resistive elements in the LISN, and not in uncontrolled impedances within the shield room power bus. On dc power buses, the capacitor can be several thousand microfarads. On an AC bus, it can be on the order of 10 μF or less. What is necessary for an AC simulation is a stiff supply. It is also very important on an AC bus that the EUT be cycled on/off at the peak of the power waveform, in order that maximum current be switched. This is difficult and time consuming without proper test equipment. What is necessary is a

“peak sensing switch” which is analogous to a zero crossing switch except with a 90° phase shift. The purpose of the peak sensing switch is exactly opposite of the purpose of the zero crossing switch.

One source of confusion which has existed, and which may be part of the reason for the demise of CE07 as a requirement, is how to make the measurement with military/commercial style LISNs. The transient of interest occurs differentially between the feeder and return. Military/commercial LISN measurements are made line to ground. The author solved this issue by terminating the EMI ports with 50 Ω dummy loads and measuring differentially from the feeder LISN power output port to the return LISN power output port. This technique captures the entire waveform, but inherently changes the high frequency impedance plateau to 100 Ω , not 50 Ω .

If the measurement is made from the EMI port into an oscilloscope, the oscilloscope input must be bypassed by 50 Ω , or the LISN will appear a very high source impedance indeed. Assuming this is done, we have built a high pass filter which severely attenuates transient waveforms lasting longer than 50 μs . 50 μs is the boundary between short transients controlled under MIL-STD-461 and longer surges/sags controlled under power quality specifications. Figures 17 and 18 show the different LISN connections and the resulting roll off of long duration transient amplitude.

Most CE07 test set ups reviewed by the author measured off the LISN EMI port, not the power port. It would seem that there exists, built into that procedure, a bias against long duration transients. Perhaps this is the source of the observation that no CE07 problems have been observed. Of course, it may also be that no problems have been reported from the field. But if transient load induced effects on power quality are not controlled in the time domain, then on what basis do the power quality specifications present and guarantee normal operational transient limit curves?

Low Frequency Conducted Emissions

Below 150 kHz, the impedance of the bus becomes small enough that it is not necessary to impose differential mode CE control with the intent of protecting a CS01/02 qualified victim from a single culprit; whereas above this frequency the bus

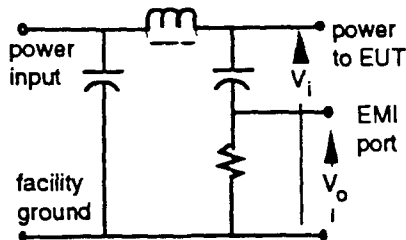


Figure 17: Possible LISN connections to measure transients

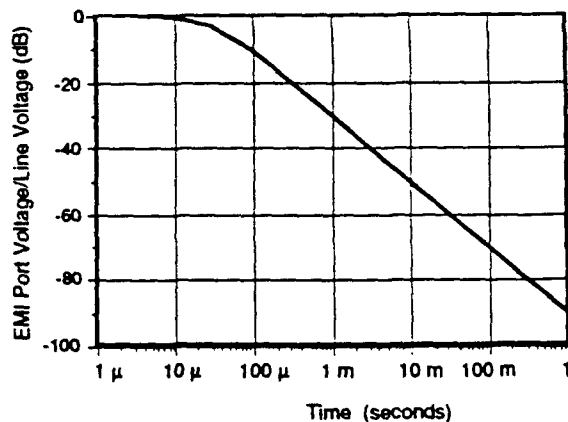


Figure 18: Ratio of LISN EMI port voltage to line voltage as a function of transient duration

wiring can be modeled entirely as a transmission line (viz. the 5 μ H LISN). With low impedance coming into play, control of differential noise current must be related to load current. In other words, the scope of EMI CE control now overlaps the frequency range where power quality concerns control load induced ripple. MIL-STD-461 basic and revision A did not recognize this. MIL-STD-461B/C in some parts tailor CE01/03 limits by load current, but not for all parts. Thus the part 2 CE01 limit is difficult to understand. It is really no limit at all for a 1 Amp load, since it allows 3 Amps of noise current below 2 kHz. It is too strict for a large load. It is definitely not designed for an ac load: a suite of ac loads meeting the CE01 limit would cause serious voltage waveform distortion. Figures 19 and 20 show compliant CE01 test data taken from a 400 Hz load and the resultant time domain voltage distortion with a source impedance being the reactance of 8 μ H bypassed by 50 Ω . The 50 dB per decade slope of the limit above 2 kHz would generate a complex filter if the load just met the limit at 2 kHz.

A successful ac power quality specification for low frequencies would specify harmonic currents as a function of fundamental current and harmonic number.

$$I[(2n+1)f_0] = \frac{\beta \cdot I(f_0)}{2n+1}$$

$$I[(2n)f_0] = \frac{I(2n+1)}{100}$$

where $n = 1, 2, 3, \dots$

$\beta < 1$ ($\beta = 1$ at very low frequencies, approaches 0.3 after LR corner frequency)

f_0 = power frequency

$I[mf_0]$ = current at harmonic frequency
($m = 1, 2, 3, \dots$)

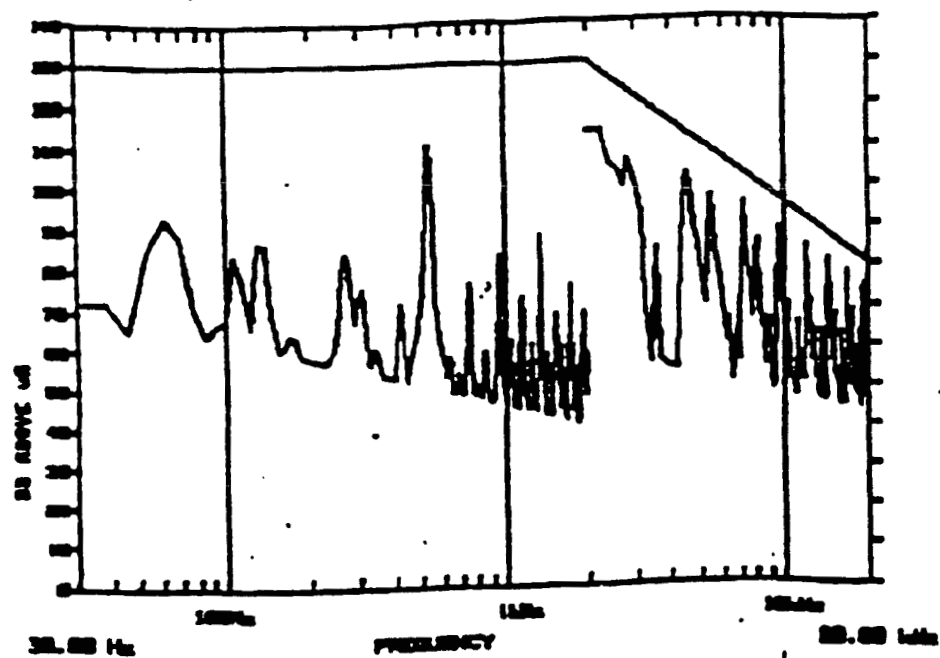


Figure 19: CE01 test data from a 400 Hz load

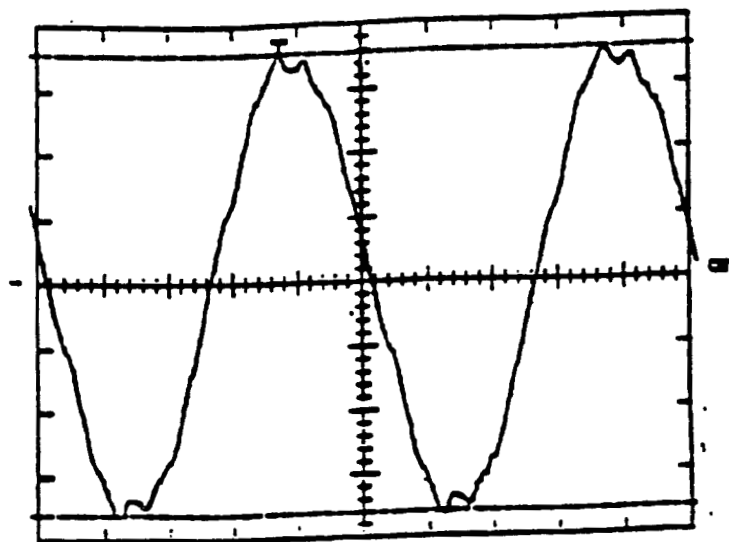


Figure 20: Time domain distortion on 400 Hz voltage waveform due to current harmonic distortion shown in Figure 19

Notice that even harmonics are constrained much more tightly than odd harmonics. Even harmonics distort only one half cycle of the waveform. This causes problems for equipment controlling the power quality.

With the advent of switched mode power supplies (SMPS) in the late 1960s, a whole new form of high frequency (relative to linear supply) differential noise was created. Now *intentional current waveforms are noise currents* which must be limited. Further, there are significant sources of cm currents. As explained in the section on separation of conduction modes, the test method gives a total current only. In essence, there is one equation for two unknowns. Common mode current control, if it is necessary, is for the control of electric field emissions and limitation of platform structural potential differences as explained below. Control of dm noise currents above a few Megahertz is problematical. It is accepted in the industry that noise phenomena above ~30 MHz do not efficiently propagate through a bus. So there is little effect compared to the lower frequencies. Potential CE EMI victims can easily meet 100 mV immunity levels. Should dm CE in an above ground power bus be limited to 6 dB below 100 mV at these high frequencies? This would be a 40 dB relaxation from the current 10 μ A limit. Do we need such a limit? The author does not recall ever seeing such a dm exceedance. A reasonable tailored stop frequency for a dm limit might be the greater of ten times the SMPS frequency or 2 MHz. This would suffice to control dm effects adequately. The limit could be 1 mA at 2 MHz and above to meet 50 mV ripple level.

Signal Line Conducted Emissions

All issues and Notices of MIL-STD-461 prior to MIL-STD-461B required signal lead common mode currents to meet the same requirements as power leads (except that with power leads the requirement was imposed on both cm and dm

current.) The purpose was obviously to control radiation. It makes sense to control signal lead emissions to the same level as power line emissions, if conduction modes are not taken into account, and if radiation is the problem to be controlled. MIL-STD-461B and revision C essentially deleted these CE02/CE04 requirements, by making them a subset of CE01/03, not requiring the test for cables between equipments in the same subsystem, and calling out the need for Procuring Activity to invoke imposition of these requirements. The rationale the author has heard for this change is "If there is self compatibility between the components of a subsystem, then why impose a test further limiting noise between already compatible equipments?" The answer is that the test had nothing to do with self compatibility of the cable under test (CUT) connected equipments; the test controlled radiated emissions from the CUT. This same conduction mode mix-up is inherent in MIL-STD-461 itself, where limit exemptions are granted for exceedances at frequencies which are the spectrum of intentional cable signals. The test requires a current probe around the entire cable bundle - it is a common mode requirement. Intentional signals with any frequency spectrum will be differential mode. Thus the exemption is unnecessary and unwarranted. If power line CE control at higher frequencies is for the protection of antenna connected receivers, it does not make sense to leave signal lead CE uncontrolled. A lack of appreciation of the causes and effects of different conduction modes is the culprit in both the erroneous exemption, and (in the opinion of the author) in the unfortunate obsolescence of the signal line CE requirements.

Many engineers feel that controlling power and signal common mode currents is a superior method of low frequency radiated emission control. If cm currents were controlled to 100 MHz, and radiated emissions were measured above 100 MHz, then, with the aid of absorber material, RE mea-

surements would be much more repeatable. (For structure return bus configurations, powerline dm noise currents would require the same limits as cm emissions.)

Audio Frequency Conducted Susceptibility

In the audio frequency bands where the CS limit overlaps the power quality control of ripple, there is presently little agreement between the two. For example, compare the MIL-STD-461C CS01 limit to the aircraft, ground vehicle and ship power quality ripple limits. See Figures 21 and 22 for comparisons.

Audio frequency CS testing was introduced under MIL-I-6181B for 28 Vdc loads only. The limit was 2 Vrms from 60 Hz to 10 kHz. MIL-I-6181C/D opened up the test applicability to any equipment requiring susceptibility testing and the limit increased to 3 Vrms over a frequency range of 50 Hz to 15 kHz. MIL-STD-826 was the first standard to state the limit as a percentage of line voltage.

Audio frequency CS levels should be directly relatable to power quality ripple limits. It is not optimal to have equipment designers having to weigh the limits of two different specifications to determine which is the worst case driver for his design. MIL-STD-461 CS01 could defer to the appropriate power quality ripple level by stating the limit as 6 dB above the ripple level. The test method needs no change in this frequency range.

RF Conducted Susceptibility

Test method and limit CS02 capacitively couples 1 Volt of ripple to the power conductors over a frequency range of 50 kHz to 400 MHz. As mentioned earlier, it is not easy to reconcile this requirement with CE03. Nor is it necessary to view these requirements as mirror images im-

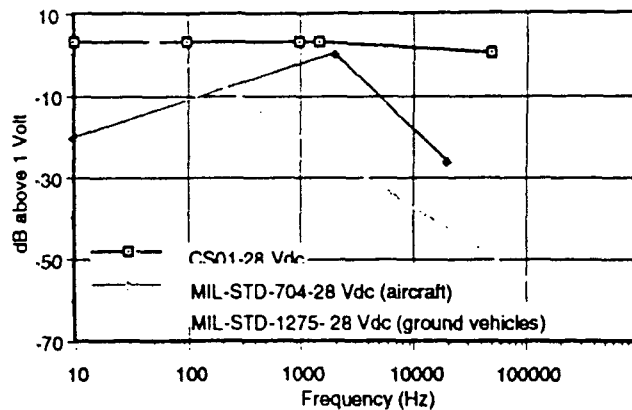


Figure 21: CS01 vs. power quality ripple requirements for 28 Vdc power

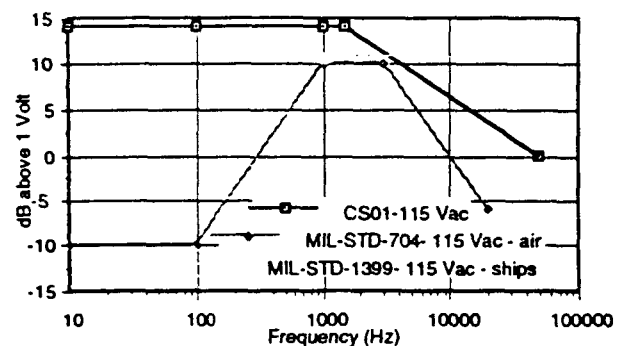


Figure 22: MIL-STD-461C CS01 vs. power quality ripple requirements for 115 Vac

posing a huge safety margin. CS02 is considered by some to protect the victim against ripple due to the effects of high intensity radiated field to power bus coupling. If so, then it is problematical that signal cables are not also so protected. This seeming contradiction should be eliminated under the proposed MIL-STD-461D, which replaces and augments CS02 with swept CW and damped sinewave bulk current injection (BCI). [2] BCI is imposed on all cables, power and signal, associated with the EUT. The reader, knowledgeable in EMI requirements, might at this point ask why, regardless of CS02 or BCI, we must overlap CS requirements with RS requirements. After all, the CS02/swept CW BCI frequency range is but a subset of the RS03 spectrum. The answer lies in the limitations of RS qualification testing. Figure 23 illustrates two problems inherent

in shield room RS testing. Figure 24 illustrates a third. Inspection of Figure 23, a typical low frequency RS03 test set up, demonstrates limitations of shield room radiated testing. The shield room/groundplane EUT set up severely limits the capability to calibrate the field. To maximally stress the EUT, the test set up must be plane wave illuminated at the right angle of incidence. Within one wavelength of a generator as large as the generator-EUT separation, in the proximity of the shield room wall and groundplane, this might be considered difficult. Evidence of this is obvious to anyone who has ever run one of these tests. Just about any desired field strength is achievable simply by moving around the E-field sensor. Next notice the "antenna." It is not an antenna but an E-field generator, i.e., it generates a field similar to that from the E-field fringing around the end of a parallel plate capacitor. The attached 50 dummy load is necessary in many cases because the high impedance of the E-field generator (ca-

pacitor) alone would cause excessive VSWR to the signal source. Because the power is dissipated in the load rather than the "antenna", there is no power in the field. There exists the specification Volts/meter, but the Amps/meter associated with a plane wave are missing. Coupling from field to wire from such a high impedance source is loaded by wire connected loads of lower impedance than the field. Since the EUT, at one meter separation from the "antenna", is very much in the near field, the field impedance approaches that of the source. Typical cable connected circuits will have impedances at or below 100 Ω , which greatly loads and thus reduces the coupled voltage from the field. Only such circuits as high source impedance sensors driving instrumentation/operational amplifiers might see the actual unloaded field coupled voltage. It should be noted that RS04 (requirement obsolete under MIL-STD-461B, method still valid under MIL-STD-462) required a parallel plate for E-fields below 35 MHz. A uniform

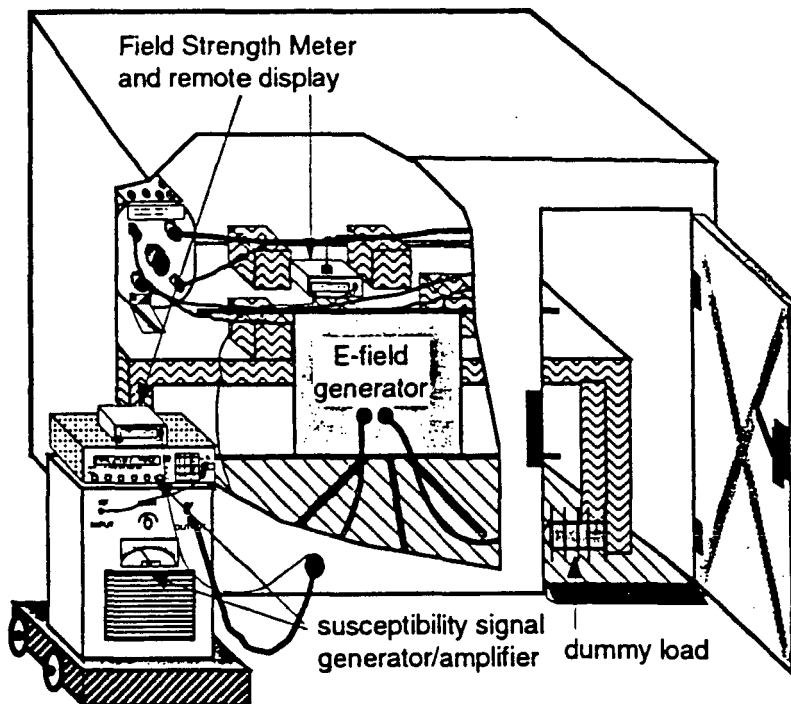


Figure 23: Shield room low frequency RS03 test set up

377 f field, perpendicular to the plates, is generated inside the device. Unfortunately, this method has become optional under MIL-STD-461B/C.

Figure 24 demonstrates that, other things equal, an electrically short cable's coupling efficiency is proportional to length. Under MIL-STD-461 basic through revision C the minimum interconnecting cable length was 2 meters suspended 5 centimeters above the groundplane for radiated tests. Figure 24 compares the coupled voltage from a plane wave with E-field polarization perpendicular to the long dimension of the wire for 2 and 10 meter wires. The conclusion is that if the platform installed wiring is longer than the tested wiring, the qualification test did not adequately stress the wire connected circuits. By applying a theoretically derived BCI susceptibility signal level, this coupling efficiency problem is side stepped.

Care must be taken, however, to not over specify the BCI current. The electromagnetic pulse (EMP) requirements first included in MIL-STD-461B and then revision C included cable and pin drive current spectra implying very long cables. Figure 25 shows a representative curve from MIL-STD-461C, part 2.

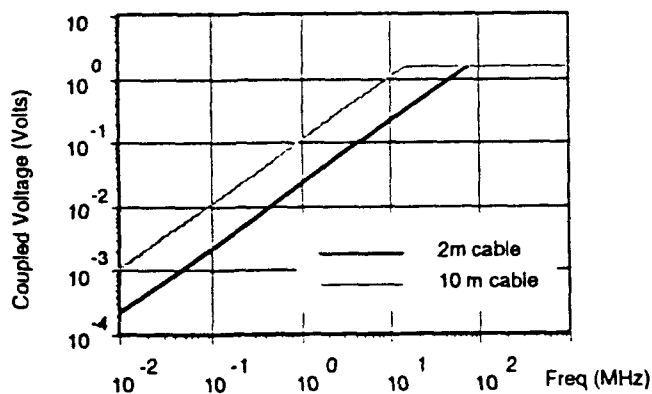


Figure 24: Field to cable coupling physics

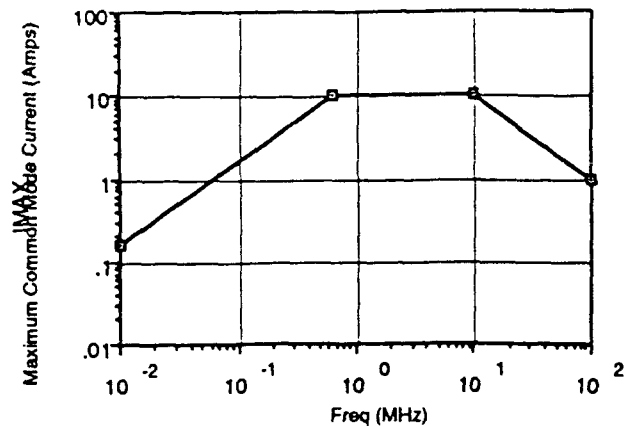


Figure 25: MIL-STD-461C CS10 pin injection current limit

The low frequency 20 dB per decade rise to a peak of 10 Amps at 630 kHz implies that the current carrying cable is 240 meters long. At over twice the length of a Boeing 747, that is much longer than any likely weapon system, other than a ship. If we tailor this specification for a 24 meter long platform, which is large enough to encompass all fighter aircraft, helicopters, and most ground vehicles, then the limit would be as shown in Figure 26.

The maximum protection is now not necessary below a frequency of 6.3 MHz and the amount of protection necessary is 20 dB less below 630 kHz. At these frequencies, the protection will in general be shielding because many circuits intentionally occupy this spectrum and filtering and even transient suppression may interfere with the desired signal. If transient suppression is required, the amount of power to be dissipated at 6.3 MHz is 20 dB less than at 630 kHz. A 20 dB reduction in protection may mean the difference between an ordinary twisted shielded pair and a requirement for an overbraid, or between a small board level component and a significant real estate impact.

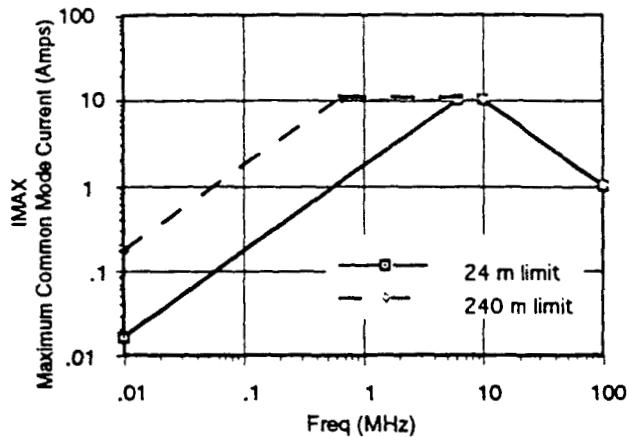


Figure 26: 24 m platform tailoring vs. basic CS10 limit

BCI drive currents are to be calibrated into prescribed loads, noting the forward power necessary from the signal source. Under MIL-STD-461D, the prescribed load is expected to be a short wire terminated at either end in 50 Ω or 100 Ω . The actual cable under test (CUT) is more typically a shielded bundle above a groundplane, terminated in milliohms. This gives rise to a qualitatively quite different coupled current vs. frequency than the specification calibration curve.

Author's Calculation of BCI Current for CS10

Given the CS10 pin drive specification of Figure 25; it is the current induced on a 240 m 300 transmission line by a 50,000 V/m plane wave. If that same transmission line is now shielded with milliohm shield terminations, the impedance is no longer constant but a function of frequency. If the shield dc resistance is 10 m Ω per meter, and the inductance is 1 μ H/m, with the same transmission line characteristic as before, then the coupled current will appear as in Figure 27, which includes the CS10 pin drive current for comparison. Note that the only difference is at low frequencies, where there is a "short circuit" phenomenon due to the small termination impedance and the mono-

tonically decreasing inductive reactance. The large low frequency current requires an extra control on the BCI test beyond the pre-recorded forward power. There must also be a maximum CUT induced current limit. For swept CW type BCI tests, this limit has in the past been set at about 1 Amp. This phenomenon should be of test interest only, since at these low frequencies the transfer impedance of the shield is close to its dc resistance. In the test chamber, with a 2 meter test cable, a transfer impedance of 20 m Ω , the 3 Amp "short circuit" current will induce only 60 mV of common mode noise.

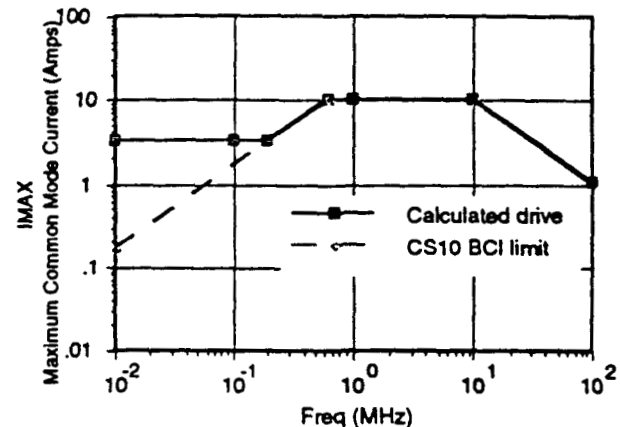


Figure 27: Actual shield currents from BCI

Technical Rationale for Radiated EMI Measurements

This section discusses issues related to fundamental physics of electromagnetic radiation. These matters were discussed in NADC-EL-5515 at the inception of the EMC discipline. Because they are physics, and not technology related, they are as important and valid today as they were in 1955 and as they will be tomorrow. The matters considered in this section straddle the gauntlet between the test engineer and the EMC analyst. Some people are comfortable with the issues discussed

in this section; for them it is intuitively obvious and not deserving of the detailed discussion presented. For others, the differences between far field antenna measurements and near field probing are less clear. For these readers, this section is very important. Following the usage of NADC-EL-5515 (including a very lucid discussion of these same matters, *sans* examples; see appendix C) the author will hereinafter refer to EMI “antennas” by the more technically and semantically correct term pickup device, and replace the antenna factor terminology with “transducer factor”. It is absolutely imperative that the practicing EMC engineer understand these concepts. Without this knowledge, it is impossible to understand military EMI radiated measurements.

In order to specify limits and test methods for radiated EMI control, the problem of interest must be reduced to a model which can be evaluated in a standardized, repeatable manner. For instance, the FCC’s concern in controlling unintentionally radiated EMI is that one consumer’s culprit not interfere with his neighbor’s victim. Interference between the same consumer’s culprit and victim is not considered to be serious, since the consumer has control over that situation. Since consumers do not live in shield rooms, the only attenuation available between culprit and victim antenna is path loss between the consumer and his neighbor. Interference to broadcast reception is controlled over the frequency range 450 kHz to 1 GHz, although radiated measurements are made only from 30 MHz to 1000 MHz. Below 30 MHz, conducted emission control is exerted. Thus the test site is the well known open area test site (OATS) with 3, 10, and 30 meter antenna - EUT separations. This allows for far field measurements (except in the extreme case of a 30 MHz tuned dipole on a 3 meter site.) The only extrapolation made is in the limit when the antenna - EUT separation is different than that for which the limit is given. The extrapolation is the square law dependency of the plane wave. De-

pending on the polarization, this is not always accurate. For the most accuracy, one must test at the same separation used by the FCC, *and using the same antennas*. Indeed, if emanations are measured within 6 dB of the limit, tuned dipoles are mandated. Broadband antennas are allowed in general for the sake of efficiency. Although antenna factors/gains/ effective apertures can be used to accurately predict the different signals out of, say, a tuned dipole and logperiodic exposed to the same plane wave, in practice deviations from a plane wave cause unpredictable variations.

The very concept of antenna factor as defined in MIL-STD-461 A and SAE ARP 958 is based on a plane wave illumination of the receive antenna, even though the transmit antenna is placed one meter distant and the wavelengths are as long as 15 meters (biconical - 20 MHz - 200 MHz.) [41] The derivation is quite simple, given the plane wave assumption. Effective aperture is the ratio of power at the antenna terminals into a matched load to the incident field power intensity:

$$A_{eff} (m^2) = \frac{P_r}{P_d} = \frac{G\lambda^2}{4\pi}$$

where

- P_r = received power load (Watts)
- P_d = power density (Watts per square meter)
- G = antenna gain (numeric)
- λ = wavelength (meters)

For a 50 antenna load, and a plane wave, the effective aperture equation may be rewritten as

$$A_{eff} = \frac{\frac{V_r^2}{50}}{\frac{E^2}{377}} = \frac{G\lambda^2}{4\pi}$$

where

- V_r = received voltage at load (Volts)
- E = electric field intensity (Volts/meter)

Solving this equation for the antenna factor yields

$$AF(m^{-1}) = \frac{E}{V} = \frac{9.73}{\lambda \sqrt{G}}$$

In log form,

$$AF(dB^{-1}) = 20 \cdot \log \left(\frac{E}{V} \right) = 19.77 - 20 \cdot \log (\lambda \sqrt{G})$$

Evidence that the plane wave assumption is less than accurate may be divined from antenna fac-

tors given for the same antenna at one and three meter spacing. The data in Tables I and II is gleaned from the 1987 EMCO (The Electro-Mechanics Company) catalog for their Model 3104 biconical antenna and Model 3146 logperiodic antennas, respectively. The author has inserted the antenna separation to wavelength ratio columns so that the antenna factor convergence as a function of this ratio may be more easily discerned.

Freq (MHz)	1m AF (dB)	3m AF (dB)	1m/λ	3m/λ	Freq (MHz)	1m AF (dB)	3m AF (dB)	1m/λ	3m/λ	Freq (MHz)	1m AF (dB)	3m AF (dB)	1m/λ	3m/λ
20	9.3	15.3	0.0667	0.1	85	8.6	8.0	0.283	.85	150	18.4	15.1	0.5	1.5
25	12.1	18.1	0.0833	0.25	90	9.5	8.4	0.3	.9	155	20.9	18.1	0.5167	1.55
30	12.8	14.4	0.1	.3	95	10.9	11.2	0.3167	.95	160	20.7	15.8	0.5333	1.6
35	13.3	12.8	0.1167	.35	100	11.4	12.1	0.3333	1	165	19.7	16.2	0.55	1.65
40	13.8	12.5	0.1333	.4	105	12.8	13.0	0.35	1.05	170	18.0	18.3	0.5887	1.7
45	14.0	12.9	0.15	.45	110	15.3	15.4	0.3687	1.1	175	18.5	16.4	0.5833	1.75
50	15.1	12.1	0.1667	.5	115	14.3	14.0	0.3833	1.15	180	18.0	17.0	0.6	1.8
55	13.0	11.3	0.1833	.55	120	14.0	13.4	0.4	1.2	185	17.7	17.2	0.6167	1.85
60	12.8	10.7	0.2	.6	125	15.1	12.8	0.4167	1.25	190	17.2	17.5	0.6333	1.9
65	10.6	9	0.2167	.65	130	14.8	12.7	0.4333	1.3	195	17.2	17.0	0.65	1.95
70	8.8	7.8	0.2333	.7	135	15.0	12.7	0.45	1.35	200	16.8	17.2	0.667	2
75	7.8	7.6	0.25	.75	140	15.6	12.8	0.4667	1.4					
80	7.8	7.5	0.2667	.8	145	17.1	13.0	0.4833	1.45					

Table I: One and three meter antenna factors for EMCO 3104 biconical antenna

Freq (MHz)	1m AF (dB)	3m AF (dB)	1m/λ	3m/λ	Freq (MHz)	1m AF (dB)	3m AF (dB)	1m/λ	3m/λ	Freq (MHz)	1m AF (dB)	3m AF (dB)	1m/λ	3m/λ
200	17.0	12.7	0.667	2	525	19.2	19.3	1.75	5.25	850	23.7	23.5	2.833	8.5
225	14.8	12.2	0.75	2.25	550	19.5	19.8	1.833	5.5	875	24.0	23.9	2.917	8.75
250	14.3	13.3	0.833	2.5	575	19.7	19.9	1.917	5.75	900	24.4	23.8	3	9
275	14.9	14.3	0.917	2.75	600	20.7	20.0	2	6	925	24.3	24.0	3.083	9.25
300	16.8	15.9	1	3	625	21.5	20.3	2.083	6.25	950	23.8	24.0	3.167	9.5
325	17.5	15.0	1.083	3.25	650	22.0	20.8	2.167	6.5	975	24.4	24.9	3.25	9.75
350	18.7	15.4	1.167	3.5	675	21.6	21.9	2.25	6.75	1000	25.1	25.2	3.333	10
375	17.5	16.1	1.25	3.75	700	21.8	22.1	2.333	7					
400	17.1	17.1	1.333	4	725	22.1	22.1	2.417	7.25					
425	17.4	17.6	1.417	4.25	750	22.7	21.9	2.5	7.5					
450	18.4	18.0	1.5	4.5	775	22.8	22.0	2.583	7.75					
475	19.8	18.7	1.583	4.75	800	22.8	22.0	2.667	8					
500	20.5	19.3	1.633	5	825	22.8	23.2	2.75	8.25					

Table II: One and three meter antenna factors for EMCO 3146 logconical antenna

Although there is a trend towards convergence at higher frequencies, there is not a good relationship over the entire frequency range, since even a 200 MHz wavelength is 1.5 m. The antennas are coupling capacitively/inductively as well as by radiation. By contrast, a smaller, higher frequency antenna does achieve the far field at one meter as evidenced by the good high end agreement in the Table II data shown on the preceding page.

The above tables demonstrate that when probing the near field it is crucial that the test antenna used be similar to the device of interest (that which needs protection by the standard) and that the test pickup device be located the same separation as exists between the EUT and antenna in the anticipated *in situ* condition. Extrapolations from measured field intensity data from one meter to a field intensity at some other distance are simply not valid below about 1 GHz.

Another example directly from MIL-STD-461A reinforces the above and also shows that orientation is important. The RE01 loop construction is given in paragraph 5.2.1 of that standard. The transducer factor (relating the measured $\underline{B}$ -field to the voltage out of the loop) is shown in Figure 1B of the standard (reproduced here as Figure 28.) That curve is derivable from Faraday's law of electromagnetic induction, *and the assumption of constant B-field intensity penetrating the loop at a right angle to its plane.* Absent such a well behaved field, two different loops will yield different measured field strengths.

Faraday's law states that the voltage induced in a loop is the time rate of change of the magnetic flux penetrating the plane of the loop. Lenz's law states the induced voltage opposes the flux which created it (the principle of the inductor.)

$$V = -n \frac{d\phi}{dt}$$

$$\text{where } \phi = \underline{B} \cdot d\mathbf{a}$$

n = number of wire turns around the loop

The transducer factor relates the induced voltage V to the originating field, $\underline{B}$. The transducer factor is derivable from the relationship:

$$V = -n \frac{d \left(\int \underline{B} \cdot d\mathbf{a} \right)}{dt}$$

It is the expression for ϕ that necessitates the use of the proper pickup device at the proper separation when probing the near field. Consider that in the general case neither the magnitude gradient of $\underline{B}$ over the loop plane, nor its angular relationship to the plane are known. And yet, in order to get a transducer factor, some relationship must be assumed so that the integral may be evaluated. The relationship assumed is that of constant $\underline{B}$ field magnitude and perpendicular direction over the plane of the loop:

$$\underline{B} \cdot d\mathbf{a} = B A_{\text{loop}}$$

Since the only time dependent parameter is the frequency dependence of $\underline{B}$, the transducer factor expression reduces to

$$f(V,B) = 2 \pi n A_{\text{loop}}$$

Plugging in the loop area and number of wire turns from paragraph 5.2.1 of MIL-STD-461A into this equation yields the curve of the cited Figure 1B, shown in Figure 28 below. The purpose of this exercise is now at hand. If we have two loops, differing only in loop area, immersed in a field of constant intensity and angular direction, then the calculated transducer factors for both loops will give the correct flux value, as shown in the center of the solenoid of Figure 29. *The number of flux lines penetrating the loop at right angles is exactly proportional to the loop area.* But if we probe the field outside the solenoid, as shown in the upper portion of Figure 29, then the larger loop will yield an apparently smaller field strength, since the field now has a

nonzero gradient and the number of extra flux lines penetrating the larger loop is no longer a linear function of its larger loop area as assumed in the calculated transducer factor. Finally, if one of two identical loops is tilted differently than the other, as shown in the lower portion of Figure 29, vastly different measurements will result.

Since EMI measurements are imposed on incidental, not intentionally generated electromagnetic fields, the general case is analogous to the leakage lines around the solenoid, not the well behaved field inside. Probing the near field of an electric dipole imposes the same constraints as the magnetic case, as shown in Figure 30. The probe used must model the victim, or the measurement cannot be correlated with the actual problem.

Figure 30 graphically demonstrates that different electric field pickup devices (short, bold right-most dipole vs. co-located longer, dashed dipole) can give uncorrelatable results when probing at identical locations, because the pickup devices see a different average intensity over their *physical* apertures, while the only analytical correlation possible is due to differences in *effective* aperture. It also demonstrates that the same pickup device (small dipoles just to the right of the inner curved field lines) will give different and uncorrelatable results when placed in different portions of the field, because the field intensity varies with position in an unknown manner (shown here by the varying degree to which the pickup devices are tangential to the electric field lines.) Again, the average intensity across the *physical*

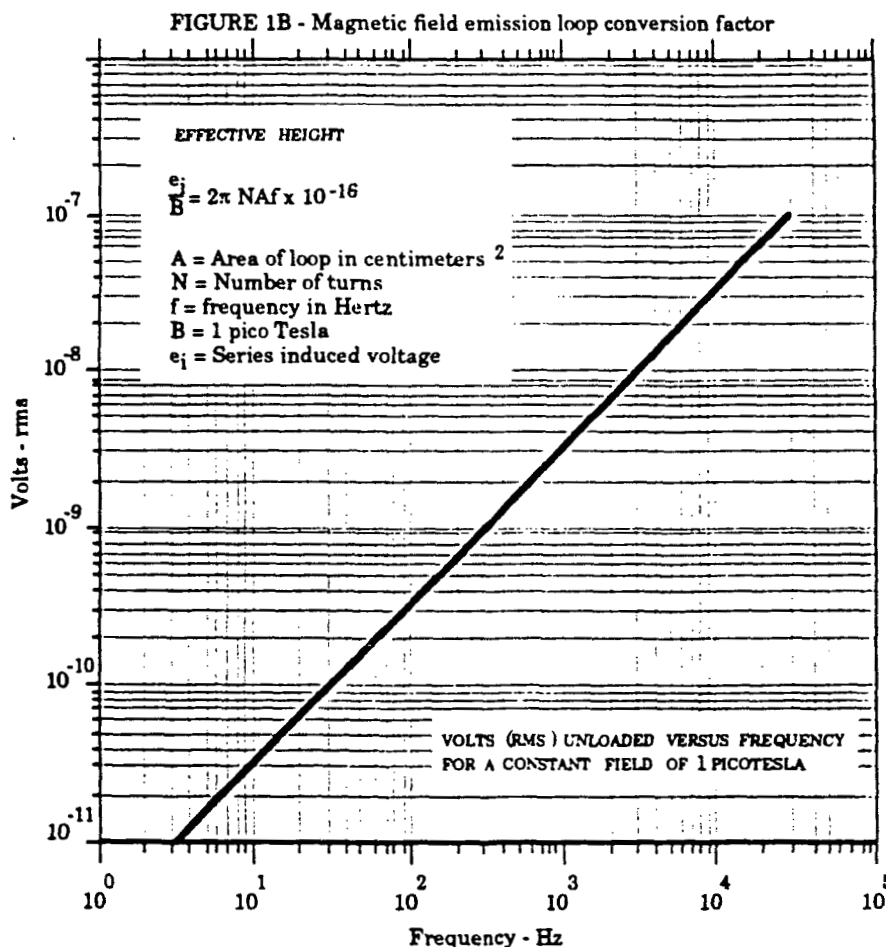


Figure 28:
 Reproduction of
 MIL-STD-461A
 Figure 1B
 (AT205 loop
 antenna factor)

aperture has changed, while the *effective* aperture is constant. Finally, for the sake of the contrast, the only case in which pickup device type is not important is shown. This case is that of the two different pickup devices to the left of the straight perpendicular field line. Since the pickup devices are calibrated in such a field, their use *in this case only*, will yield identical field strength measurements. Naturally, operating inside a bare wall shield room, at resonance frequencies an electromagnetic hall of mirrors, only exacerbates such problems.

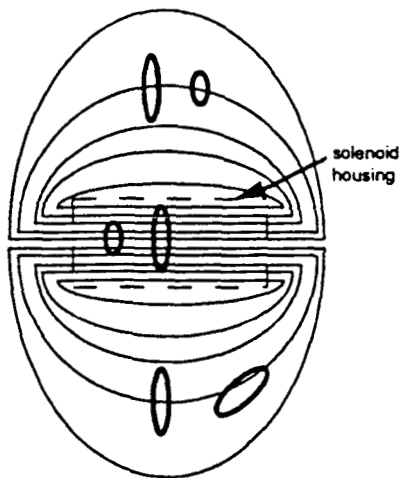


Figure 29:
*Probing the
magnetic
field in and
around
a solenoid*

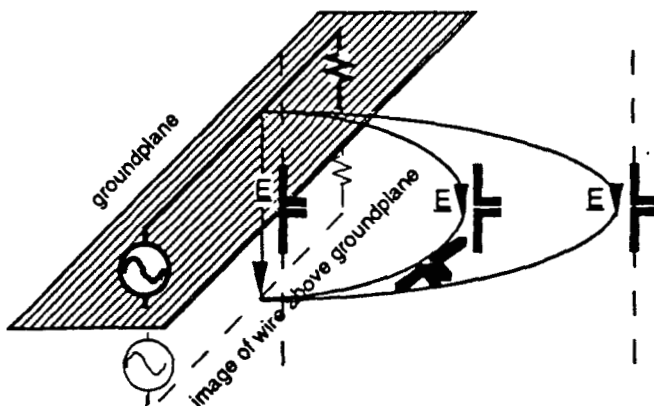


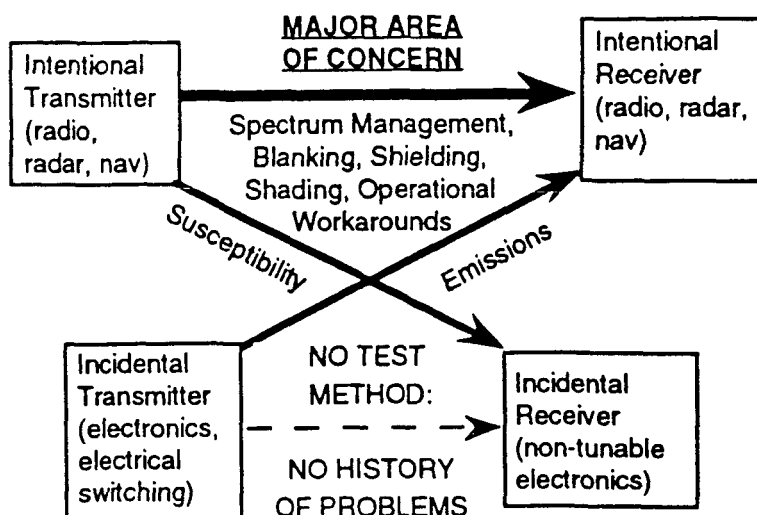
Figure 30: *Near field probing*

Historical Military Radiated EMI Requirements

Traditional MIL-STD-461 radiated requirements, and MIL-STD-462 radiated test methods are written to protect against two types of interference. The first is disruption of receiver systems (communications, navigation, radar) by incidental (unintentional) sources of electromagnetic energy. The second is protection of non-antenna connected electronics from high power intentional transmitters of electromagnetic energy (communications, navigation, radar). Figure 31 is a graphical display of traditional military radiated EMI concerns.

The most prevalent problem is the intentional transmitter frequency or harmonics thereof interfering with same platform or nearby platform receivers. Reducing this problem to an acceptable level is the major portion of system level EMC design. The two paths labeled susceptibility and emissions are controlled by RS03 and RE01/RE02. The non-interaction shown at the bottom of the figure is the subject of some confusion. As stated elsewhere herein, there is no history of such problems and thus no standardized control to protect against it. However, there are many engineers who labor under the misconception that RE02 and RS03 are complementary requirements, and that both must be enforced in order to ensure general non-antenna connected subsystem compatibility. That is, if RE02 and RS03 are not imposed, some non-transmitting culprit might radiate enough to upset some non-receiving victim. In particular, the civilian space EMC community seems to hold this relationship as self evident. This includes NASA, ESA, and even (hearsay) MIR Space Station (Soviet) EMC engineers. That this type of non-RF culprit/victim interaction does not and cannot occur is a matter of experience, and can be analytically demonstrated. The author's experience, including that of his professional colleagues, does not include a

Figure 31:
*Focus of military
radiated EMI concerns*



single such incident. Non-RF equipments operating at power levels within a few orders of magnitude of each other will not interact, since the amount of radiated power from an *incidental or unintentional* culprit radiator is about one millionth of the consumed power, and thus well below the threshold of susceptibility of the nearby victim. Inspection of the test methods will also demonstrate that there was never an attempt to make RS03 data comparable to RE02 data. RE02 data is taken at one meter from the EUT. RS03 field intensities are established at the EUT. Thus there is no way to compare the RE02 emissions from a culprit to the RS03 performance of a victim, unless the victim just happens to be one meter from the culprit. The following analysis should suffice to demonstrate the impossibility of accidentally radiating at RS03 levels (1 Volt/meter). A tuned dipole is the most efficient radiator possible accidentally. How much power at the antenna terminals is necessary to establish a 1 V/m field intensity at one meter? Solution of the radar equation yields a power of 20 mW, or 13 dBm. By comparison, one of the most energetic signals routed is that of the MIL-STD-1553 databus, at 15 Volts into 75 Ω . This is a power of 3 W. Nearly 1% of the total time domain signal power would have to radiate in order to approach 20 mW. This does not happen.

Misunderstanding of the relationship of RE02 and RS03 has led many to believe that one or both of the limits are too severe. Comparison of the sub millivolt/meter RE02 levels with the Volt/meter RS03 levels is at fault. Figure 32 shows the erroneous and actual safety margins involved. There is no frequency dimension to the x-axis. This is a snap shot at any individual frequency (zero span mode). The left hand side shows the erroneous comparison. The right hand side shows that the safety margins are really between

- the RS03 qualification level and the actual threat field intensity
- the RE02 level and the field intensity which will interfere with the victim receiver.

Radiated Electric Field Emissions

Imposition of radiated EMI performance requirements on subsystems dates back to the 1940s. The first EMI performance requirements (hereinafter referred to as limits) were emissions limits designed to preserve the performance of the antenna connected communications and navigation electronics in use at the time. No EMI susceptibility

limits had been imposed on these (WWII era) receivers. The design of these receivers (using unshielded transmission lines between antenna and electronics) left them extremely susceptible to both conducted and radiated EMI. Figure 33 demonstrates the radiated emission problem. Electric field emissions from unintentional radiating elements in the immediate vicinity of the unshielded antenna lead-in coupled interference into the connected receiver.

The coupling path was not through the externally mounted antenna, which was well shielded from noise generated inside the all aluminum fuselage. As non-antenna connected electronic systems became more prevalent in military platforms, the new electronics had to be designed to be compatible with the older susceptible antenna connected equipment. The new (antenna and non-antenna connected) equipment, however, was designed with immunity to interference levels orders of magnitude greater than the old equipment. Newer antenna connected electronics in particular were designed to use shielded transmission line (coax) between antenna port and antenna. It was recognized by the early 1950s that the emissions limits performed a grandfathering function to allow continued use of the older receivers. It was expected that when the then current inventory of government furnished equipment (GFE) receivers was exhausted that the emission limits would be revised to reflect the immunity levels of the newer generation of receivers. As will be seen, the minor changes to the limits which have occurred do not reflect this *original intent*.

The MIL-STD-462 RE02 test method is still based on the susceptibility of an antenna lead-in wire to near field interferences. The key to understanding the test method is that at one time a separation of one meter (originally, one foot) between antenna lead-in wire and EUT was precisely the configuration of interest [5]. The configuration of the test set up that defined the RE02 limit and test method is shown in Figure 34 per the description in [5], based on the installation depicted in Figure 33. In Figure 34, the BC-348Q WWII era voice receiver was connected to a dummy load or antenna mounted outside the shielded chamber by an unshielded antenna lead-in wire. This test set up models the actual situation shown in Figure 33.

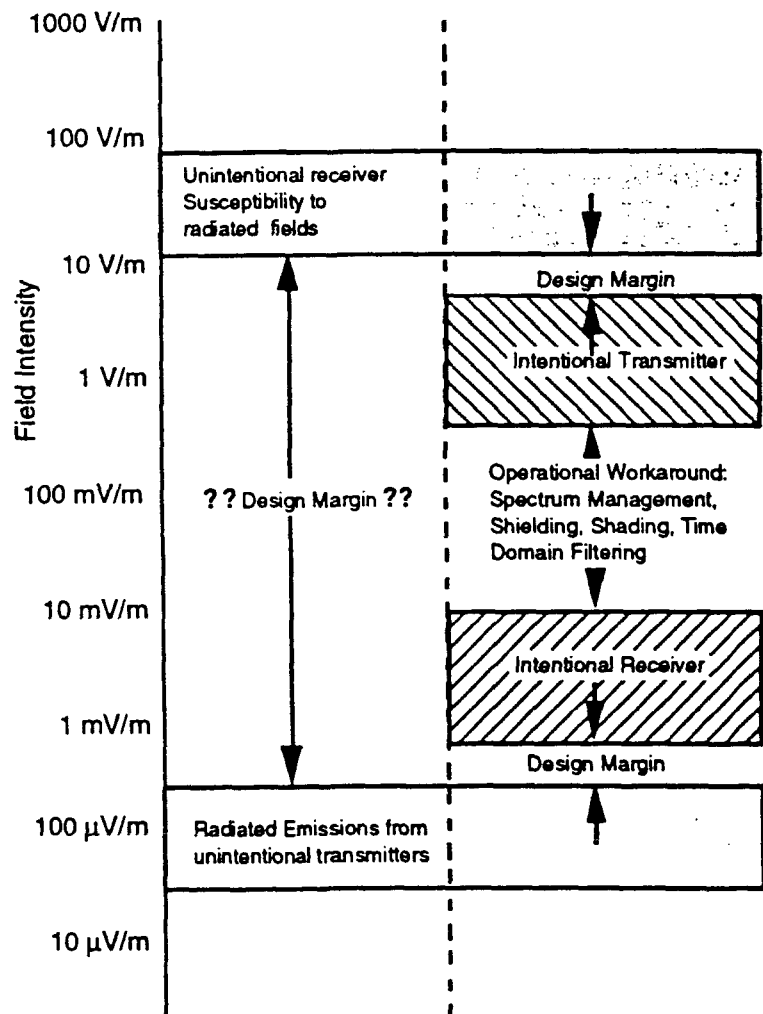


Figure 32: Radiated safety margin analysis

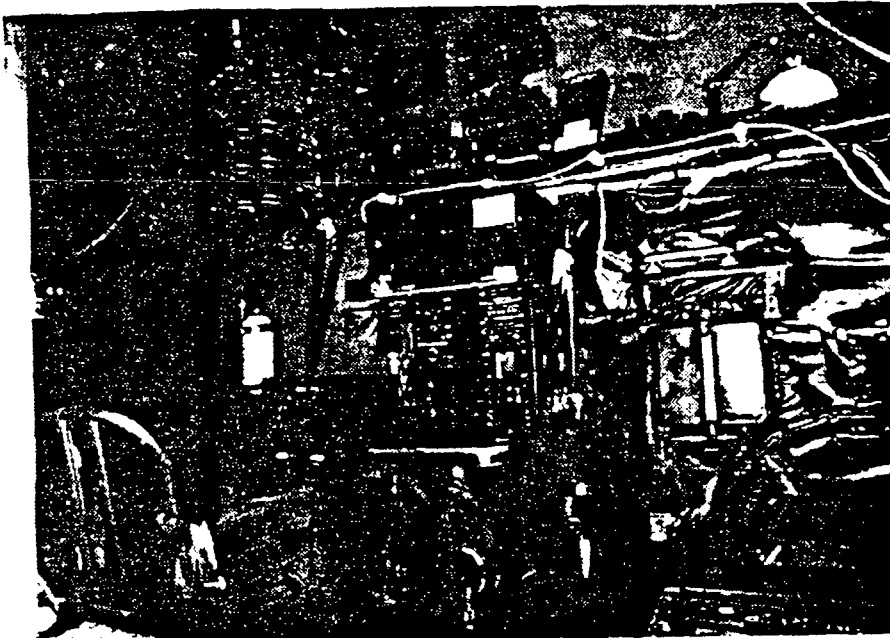


Figure 33:
*Origin of the Air Force
radiated emission
problem/solution*

B-36 Photo
Courtesy of the
Air Force Museum
Wright Patterson AFB
Dayton, Ohio

B-36 Peacekeeper (WWII design bomber, operational first post war decade) radio room (radio operator faces aft). Radio and antenna tuners are mounted on tray in center of picture, aft of chair. White wire on white standoffs is unshielded antenna lead; standoffs are high voltage insulating ceramic. Antenna lead is white because it is bare copper wire with 3/8" dia. white ceramic bead insulators, strung on antenna lead like beads on a necklace. The antenna lead penetrates the fuselage through a ceramic bulkhead several inches in diameter. The other antenna lead is grounded directly to aircraft structure.

This short lead-in wire acted as an electrically short (high source impedance) electric field pick up device at low frequencies. Thus, the 41" rod was selected as an accurate model of such a victim at low frequencies, and tuned dipoles at frequencies where the lead-in wire could resonate. Various common noise generators of the day were exercised in the room, equidistant from the unshielded lead-in and the pickup device connected to the EMI receiver. Measurements were made with the EMI receiver tuned to frequencies where the BC-348Q evidenced susceptibility to the operation of the noise generators. These noise generators (ignition systems, brush motors/ generators, and others) were uniformly broadband sources and were recognized as such. Continuous wave (CW) or narrowband limits were derived from broadband limits and a knowledge of the EMI receiver bandwidths. Note that the "field strength" - a nebulous concept at best inside an all metal room with a pickup device as large as the

separation between it and the field source - was of no interest in this problem. The entire issue was neatly sidestepped by specifying the type of pick up device used and levying an antenna induced specification. There was an acute sensitivity on the part of the investigators to the need for standard pickup devices in order to assure accuracy. It was specifically prohibited to substitute different pickup devices and account for the difference by use of transducer factors. (The very use of the word antenna is a misnomer in this context; note that [5] uses the word "pickup device" in place of antenna. The idea of antenna factor in those days was related only to the losses of networks necessary to match the pick up device to the 50 Ω receiver input, not the present inverse relationship to antenna effective height.) Room size relative to the longest extension of the dipole antenna was specified to further standardize the test [32]. (Although the requirement is plainly stated in [32], the author did not realize the import until he

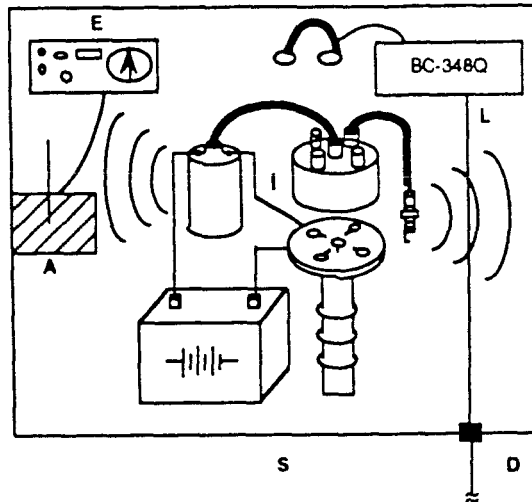


Figure 34: Original configuration for RE02

A-antenna (pick up): 41" rod, ≤ 20 MHz; dipole, $20 \leq f \leq 1$ GHz
D-12' whip antenna
E-EMI receiver
I-ignition system and other common noise sources
L-unshielded antenna lead-in
S-shield room

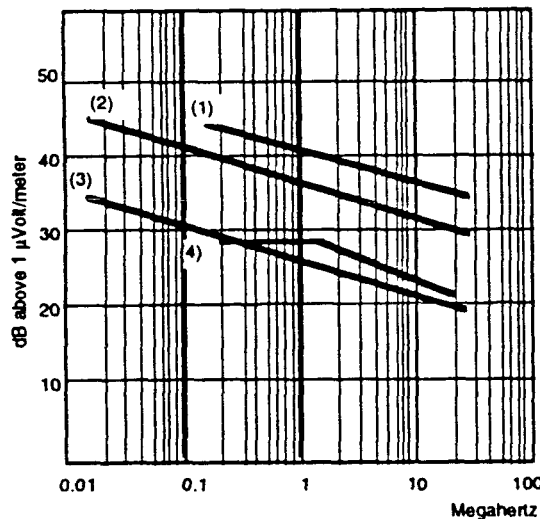


Figure 35:

Comparison of narrowband radiated emissions limits below 25 MHz, 1953-1989

- (1) RTCA/DO-160C, Section 21, 1989
- (2) MIL-STD-461C, RE02 (Air Force), 1986
- (3) MIL-STD-461C, RE02 (Army), 1986
- (4) MIL-I-6181B, Radiated Emissions, 1953

visited the superb vintage EMC Science Center in Glendale, California, the test facility of Mr. John Peters. A no longer used Ace Engineering shield room was 17.5' long by 10.5' wide. The groundplane ran along the long dimension of the room. The author asked Mr. Peters about the "odd" dimensions of the room. Mr. Peters' reply was to the effect of "How else could it be?" In the days of tuned 35 MHz dipoles horizontally polarized and placed 1' from the EUT, this room size was one of Ace's most popular, since it was compact but still slightly larger than the requirement of MIL-I-26600.) The lowest frequency of interest for Air Force purposes was 150 kHz, the lowest frequency for Automatic Direction Finding (ADF). The extension of RE02 to 14 kHz was

a Navy requirement when Tri-Service standards replaced Service unique standards. To this day commercial aerospace EMI requirements are specified at and above 150 kHz only. [42]

As indicated, excellent historical reasons exist for the RE02 test set up of MIL-STD-462 and limit of MIL-STD-461. The MIL-STD-461 limit and MIL-STD-462 procedures for RE02 are bound together and form a "molecular" test requirement. That is, the limit and the test method *together* form the smallest possible component of a valid test, according to the original criteria. If performance against the RE02 limit is measured in some other fashion, or if the procedure is adopted for testing against a limit based on a different set of

criteria, then you no longer have a valid test. A few notes on this subject.

1. It was recognized by Mr. Jarva at this time (circa 1953) that the unshielded antenna lead-in wire was already obsolete. In fact, the same specification, MIL-I-6181B, based on the findings contained in NADC-EL-5515, mandated coaxial antenna inputs for all new receivers. But the military inventory at the time had to be protected, and so the specification was levied. There is a comment in NADC-EL-5515 to the effect that once the inventory of the older configuration receivers is exhausted, that a new radiated emission limit will be necessary.

The Figure 35 comparison of the 1953 MIL-I-6181B radiated limit with the 1986 MIL-STD-461C revision reveals very little difference, especially against the Army (unrelaxed) RE02 limit. Apparently Mr. Jarva's recommendation was not heeded. (For the sake of simplicity, only the rod antenna range below 25 MHz has been shown in Figure 33. Comparison of older tuned or near tuned dipole based tests vs. today's biconical show a large *paper* difference that is not physically real. This is because the large size of these dipoles yields a very efficient pickup relative to the biconical. This is misleading since at one foot separation from the EUT it is hardly plane wave illuminated as is assumed by the antenna factor/effective aperture equations.) Therefore, there are serious questions as to the applicability of the 41" rod as a pickup device except possibly at frequencies where it adequately models the high impedance of an HF probe type antenna (2-30 MHz). This is related to the high impedance characteristic mentioned above. The rod pickup device is modeled as a Thevenin voltage source with the source impedance being the reactance of ten picofarads. The voltage source is equal to half the electric field intensity. Table III gives the source impedance vs. frequency, and Figure 36 shows the model.

From these numbers the transducer factor is derived [43]. Passive matching of the rod to 50 Ω by tuning out the 10 pF with a series of inductors yields the actual voltage delivered into a 50 ohm receiver port, as shown in Figure 37. Alternatively, the lossy passive matching of the rod

frequency (MHz)	0.01	0.1	1	10	30
Reactance (Ω)	1.6M	160k	16k	1.6k	531

Table III: 41" rod antenna source impedance

pickup device source impedance to 50 ohms can be replaced with high impedance active circuitry which allows for an open circuit measurement. These impedances are necessarily much greater than the rod source impedance as shown in Table III. Comparison of the antenna induced voltages of Figure 37 and/or the impedances of Table III with typical victim thresholds of susceptibility and circuit impedances do not show a good correlation. The conclusion is inescapable; the rod antenna, as predicted in [5], has outlived its usefulness (with the possible exception of modeling the HF probe, as mentioned above, from 2-30 MHz; and this scenario postulates a fairly open platform, with little or no shielding/shading between installed equipment and antenna.)

2. Because the most susceptible victim at the time of release of MIL-I-6181B and NADC-EL-5515

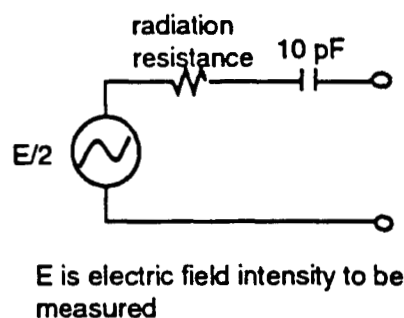


Figure 36:
41" rod antenna vs. frequency model

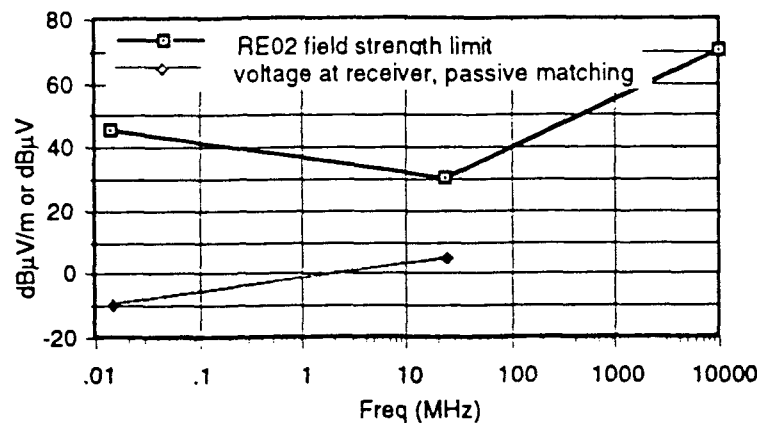


Figure 37: 41" rod antenna induced voltages into 50

was sensitive to purely electric fields with essentially no magnetic component necessary, the issue of magnetic EMI was not considered to be of great importance. However, the issue was specifically addressed. Mr. Jarva writes that although it would be nice to measure both the electric and magnetic components that it was felt that this would needlessly complicate the test. Mr. Jarva makes the very important point that when probing the near field of the EUT it is critical that the probe model the victim of interest, and that the probe - EUT separation model the actual separation. *Otherwise, the measurement is meaningless.* There was no history of magnetic interactions between closely spaced rack mounted equipment where the magnetic fields might be expected to be strongest. NADC-EL-5515 specifically states that strong magnetic sources will have to be handled on a case by case basis. There was no correlation found between measurements made near the EUT with a small loop and measurements made a foot distant with larger loops. Thus it was deemed to be uneconomical to standardize control of magnetics i.e., box to box compatibility issues, in the Air Force.

3. Mr. Jarva had a relatively simple radiated phenomenon to model and control. One may quibble about the accuracy of shield room tests, but at least the pickup device, and pickup device-

EUT separation was determined. The data was usable as is: no extrapolations necessary. This situation was exactly analogous to present commercial FCC EMI practice. The limits and procedures are an end in themselves and bear directly upon the probability of an EMI incident. With the advent of the coax antenna lead-in, the test method was no longer valid and the problem became coupling directly to victim antennas. This problem cannot be standardized as neatly since the coupling is highly platform dependent. The coupling from the interior of a windowless B-52 to the centerline mounted UHF blades on the bottom of the fuselage is quite a different path loss than that between the electronics in a wide open attack helicopter and its antennas. There is no panacea for this problem. *The best that can be had is an accurate, repeatable test with reasonable limits that by experience contribute to EMC in the general case.*

Radiated Magnetic Field Emissions

MIL-STD-461, -462 also include magnetic radiation requirements and test methods RE01 and RS01. There is also an obsolete limit designated as RE04. The test method is not obsolete because MIL-STD-462 has not been revised as of this writing. It should be noted that RE01 and RS01

were written to protect Navy VLF *receivers*. RE04 was written to protect a variable- μ magnetometer, a sensitive *receiver* covering 20 Hz - 50 kHz. The point here is that the existing magnetic requirements protect receivers, just as does the existing electric field requirement. The military has not found it necessary or cost effective to standardize magnetic emission control in the general case of system level compatibility. Nor has the commercial aerospace industry, which controls only electric field emissions from 150 kHz to 1215 MHz (RTCA/DO-160C section 21 [42]), with the exact same test set up as MIL-STD-462

The origins of RE01/RS01 and RE04 are recounted here. There are several issues of procedural interest, even though the RE01 limit has changed and the rationale is no longer the same, and, as mentioned earlier, the RE04 limit is obsolete.

RE01/RS01 [44]

1. Background

Source data for the ancestor of RE01/RS01 was gathered by Mr. Moser, Mr. P. J. Johnson, Mr. Irv Grey, Mr. Bill Prysner, Dr. Ralph Showers, and Mr. Gerry Barker for MIL-I-16910C Amendment 3, circa 1965. The work was done by the Navy Underwater Sound Laboratory (now known as U. S. Navy Underwater Systems Center), Electromagnetic Environment and Optical Systems Division. The concern was specific to submarine VLF receivers and sonar interference from low impedance magnetic field ambients interior to submarines.

1.1 Origin of Requirement

In particular, the AN/BRR-3 VLF receiver was the victim of primary concern. This receiver was installed on Polaris and attack class submarines. Mr. Moser explained that in terms of absolute

sensitivity, underwater receivers are limited solely by internally generated noise, since sea water has an attenuating effect and therefore there is no atmospheric limitation. Thus, the design of an underwater receiver may proceed in the direction of ultimate sensitivity without concern for atmospheric noise limits. The BRR-3 was classified confidential and was a state of the art receiver. It was, without doubt, the most sensitive VLF receiver of its time.

According to Mr. Moser, concern was expressed about in band interference to the receiver from adjacently installed electrical/electronic equipment functioning as nonintentional in band emitters. The susceptibility of the BRR-3 was determined over its operating VLF range of 14 to 30 kHz by use of a Helmholtz coil. That threshold of susceptibility (TOS) became the RE01/RS01 limit over the VLF band. For measuring emissions, the existing Stoddart AT205 RE01 loop as later described in MIL-STD-461A, paragraph 5.2.1 was chosen, and a loop/EUT separation of 7 cm was specified. Dr. Ralph Showers played a role in this. The receiver used at the time was the Hewlett Packard 302 wave analyzer (a wave analyzer being a spectrum analyzer with a meter deflection instead of a CRT display.)

The low frequency (60 Hz to 400 Hz) portion of the RE01/RS01 limit was determined by making emissions measurements using the AT205 loop at 7 cm spacing from various locations interior to Fleet Ballistic Missile and Attack submarines such as generators and fluorescent lamp ballasts located in the general vicinity of the BRR-3, but at much greater separation from the BRR-3 than 7 cm. The $1/r^3$ variation of the magnetic field close to a small source was not really accounted for in setting the limits for RE01/RS01. A straight line (semi-log paper, dB above one pico Tesla vs. log frequency) connects the power frequency magnetic emission limit to the VLF limit. This accounts for the slope of the original limit line.

Note that the modern (MIL-STD-461C) RE01 limit has been significantly raised, with a corresponding rise in RS01 levels, plus the addition of a 20 dB safety margin, whereas before the safety margin could be 0 dB. However, the 7 cm test spacing requirement for both RE01 and RS01 reduces the need for any expressed safety margin.

2. Some Ramifications with Respect to Test Setup

As an aside, conversation with Mr. Al Parker of Solar revealed that the VLF front end of the NM-10 receiver utilized an RF transformer and thus provided a balanced input to the AT205 loop which, in turn, was in the configuration of a shielded, balanced two port device, with the shield being terminated at one end by the NM-10, and at the other end making connection to the loop electrostatic shield. None of this information is contained in MIL-STD-461A. Nowadays, the various manufacturers of RE01 loops (A. H. Systems, Eaton, Electrometrics, EMCO, and Solar - checked by the author as of 1990) all use an unbalanced coax output to complement the unbalanced coaxial inputs of the various receivers commonly in use. One side of the loop wire is referenced to the electrostatic shield in the loop fixture itself without use of a balun. According to Mr. Parker, at VLF there is no significant change in noise immunity between these two cabling methods. Mr. Parker also stated that the 13 cm loop was originally designed for 10 kHz to 30 kHz and not for use down to 30 Hz. Mr. Parker's present setup for RE01 uses an EMC-10 (battery operated) in the room with the loop. If the receiver is not battery powered, or for other reasons must be removed from the measurement site, then care must be taken to not ground the coax at the shield room wall penetration, lest a ground loop pollute the desired signal at the (grounded) receiver. This was not a concern with the original loop/receiver design. Instructions to this effect should be called out in the RE01 test procedure.

Another important point with respect to the original twinaxial input is that it was a much higher impedance than the present 50 Ω . In fact, it did not load the output of the AT205 loop at all. Inspection of Figure 1B reproduced from MIL-STD-461A in this document will reveal that the loop conversion factor was specified in the unloaded condition. With today's 50 Ω input spectrum analyzers and receivers this unloaded case is no longer realized. The source impedance of any loop is

$$Z_{\text{loop}} = R_{\text{winding}} + j 2 \pi f \cdot L_{\text{loop}}$$

where, R_{winding} is the resistance of the loop wire, and

L_{loop} is the inductance of the windings.

As developed previously, the voltage output of a loop in a constant magnitude and perpendicular magnetic field density B is,

$$V = 2 \pi f n A_{\text{loop}} B_{\text{mag}}$$

The complete expression for a loaded loop conversion factor is then,

$$V = \frac{2 \pi f n A_{\text{loop}} B}{\sqrt{\left[1 + \frac{R_{\text{winding}}}{R_{\text{load}}}\right]^2 + \left[\frac{2 \pi f L_{\text{loop}}}{R_{\text{load}}}\right]^2}}$$

For the AT205 loop configuration, the parameters are

$$R_{\text{winding}} = 10 \Omega, \text{ and} \\ L_{\text{loop}} = 340 \mu\text{H}$$

The load impedance is 50 Ω , so the correction factor is:

$$CF = \sqrt{1.44 + (4.21 \text{E-}5 \cdot f(\text{Hz}))^2}$$

Graphically, the correction factor by which to multiply the curve of MIL-STD-461A Figure 1B is shown in Figure 38. Today, it is much more

common to specify the loop conversion factor in terms of indicated dBpT per dB μ V output. For such a curve, the correction from unloaded to 50 load is as shown in Figure 39.

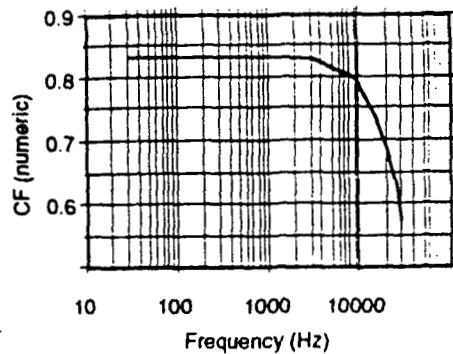


Figure 38: Multiply curve of Figure 28 by this curve to arrive at 50 loaded AT205 factor

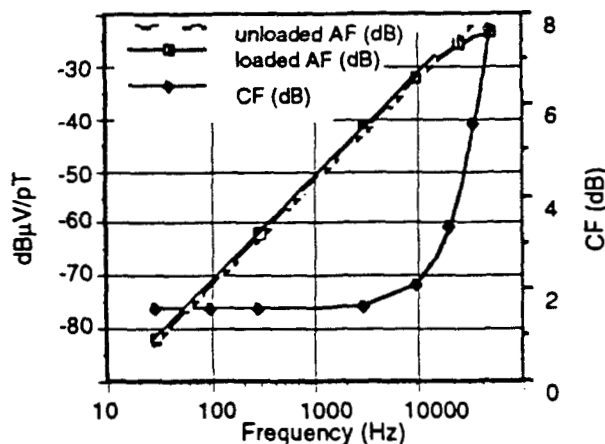


Figure 39: Loaded and Unloaded AT205 antenna factors, with expanded scale for correction factor

3. Conclusions and Recommendations

Similar to RE02, RE01/RS01 has its roots in a well defined problem and an appropriate solution. Unlike RE02, which has been applied universally on a Tri-Service basis and NASA, RE01 has been relegated to the relatively restricted world of submarines, sonar, and submarine communications, whence it originated.

The concept of an RE01/RS01 requirement pair for general platforms is valid and useful, but the limits should be determined on a case by case basis. It would be damaging to argue that all procured electronics should meet a strict RE01 requirement to give maximum flexibility to platform integrators; the efficient specification would guarantee that an unintentional magnetic field emitter would be side by side compatible with an unintentional magnetic field receptor, and that the emitter would be compatible with the sensitive receiver, given reasonable zoning controls that also form the basis of the general RE01 limit. At the Line Replaceable Unit (LRU) level there is no more need for across the board standardized magnetic performance limits than in Mr. Jarva's day. However, with increasing levels of electronics integration, magnetic control may need to be exerted. An example is the Line Replaceable Module (LRM) concept, in which printed circuit boards from a variety of sources are qualified as end items in themselves and will be integrated into a complete weapon system only by a prime contractor.

RE04 [44]

1. Background

Mr. Moser recalled that RE04 was written around a sensor developed at EMCO in the 1960's. He suggested the author get in touch with Mr. Michael Hart, president of EMCO. Mr. Hart referred to page 73 of the current (copyright 1987) EMCO catalog and the Model 6640B Magnetic Field Intensity Meter. This device is a microprocessor controlled direct descendant of the original. The principle of operation has not changed.

1.1 EMCO Model 6640 Magnetic Field Intensity Meter, Principle of Operation